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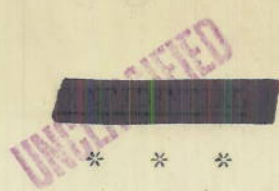
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ELECTRONICALLY SCANNED PANORAMIC RECEIVER  
(5-10C Kilocycles)

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## 1. ABSTRACT

This report concerns the development and preliminary tests of an electronically scanned sonar panoramic search receiver having a frequency range from 5 to 100 kilocycles. Within the range covered, the probability is 100 per cent that the receiver will indicate the frequency of any supersonic carrier with a pulse width as short as 3.3 milliseconds and a recurrence rate as low as one cycle per second.

Preliminary laboratory and field tests indicate that the completed model generally satisfies the original specifications.

This receiver was developed and ready for preliminary shipboard tests within three weeks after the construction phase of the problem was begun. Because of the need for this type of equipment, preliminary circuit research and development had to be terminated earlier than was consistent with best judgment and practice. The X-1 Sonaromic Receiver, therefore, is to be considered a crash model, not a refined complete operational instrument. In view of this, recommendations for improvements in future models are included in this report.

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### 3. INTRODUCTION

Reference (a) assigned the problem of developing a panoramic type receiver for use in the reception and identification of audio and supersonic frequencies transmitted through water. The proposed receiver was to be similar in principle to the OBH Modulation Analyzer, but eliminating certain unessential features of the OBH. Its contemplated use was in connection with a suitable sound head on submarines or surface craft, to establish quickly the existence and characteristics of received signals especially those of higher frequency than presently available gear might receive. With this in mind, the problem details of Reference (a) requested frequency coverage from 0 to 120 kc, sector scan of any 10 kc band within this range, and audio output to a headset corresponding to this band. The equipment was to be capable of indicating signals having pulse widths as narrow as 3 milliseconds and repetition rates as low as one per second. Other requirements of somewhat less importance were also indicated. The receiver as constructed satisfies in general the broader requirements of the problem but because of the "crash" nature of the development certain compromises were made between requirements and construction. Chiefly these are a reduction of the frequency range to 5-100 kc and an increase in size to a 15 inch cube from the request for approximately one-half this size.

Independent of the high frequency performance of the equipment, the virtue of panoramic reception of underwater sound signals may be realized from the following considerations. A submarine often receives its earliest indication of the presence of an enemy vessel by reception of echo-ranging from the latter. Reception at maximum distance implies that the receiving head is trained in the proper direction and that the receiver is tuned to the proper frequency. Essentially a coincidence of three events is required, namely, correct training, correct frequency, and these two at the time that the enemy ping reaches the submarine. Elimination of the necessity for tuning by panoramic means requires only that two events be simultaneous and may reduce greatly the time before detection.

### 4. DETAILED DESIGN AND OPERATION

Numbers on components mentioned in this section refer to the schematic diagram, Plate 14.

#### 4.1 General Electrical and Mechanical Features

The X-1 Sonaromic Receiver is an electronically scanned panoramic receiver covering the audio and supersonic ranges from 5 to 100 kc. Its operation may be understood by reference to the block diagram of Plate 13. The receiver is automatically tuned by means of a reactance tube across the local oscillator tank circuit. The receiver can be made to scan the entire range from 5 to 100 kc; or it can scan any

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10 kc portion of the 5 to 100 kc band; or it can be set to any spot frequency within the broad range for listening purposes. These three receiver settings correspond to i-f bandwidths of 8 kc, 12.5 kc, and 8 kc, respectively. The receiver has two scanning speeds. On the Broad range from 5 to 100 kc, a 300 cycle sweep is used, and on the narrow 10 kc range, a 100 cycle sweep is used.

The reactance tube is electrically scanned by means of a saw-tooth voltage generated in a thyatron circuit. Inasmuch as the saw-tooth voltage is used to sweep the reactance tube, and is also used to sweep the electron beam across the screen of the cathode ray tube, there is automatic synchronization between the position of the electron beam and the frequency to which the receiver is tuned.

Signal input to the equipment is made by means of either of two Type AN (Amphenol) connectors on the front panel. A third connector is an all-purpose jack used for feeding large test signals in to the receiver. The other two input jacks are in parallel and are connected into the equipment through a step-up transformer. Signals from the input transformer are fed through a three-stage pre-amplifier, which has a flat response from 5 to 120 kc. The output of the pre-amplifier is fed into a low-pass filter with an acceptance band of 0 to 120 kc. Adjustments in gain are made in the pre-amplifier by controlling the amount of degeneration in the first two stages.

The signal output from the low-pass filter and the output of the local oscillator are fed into a linear mixer. In the linear mixer the signal voltage and the local oscillator voltage are added, but not heterodyned, and are sent through a phase inverter, and then into the double balanced converter or mixer of the receiver. The double balanced converter output consists of the beat between the local oscillator signal and the signal input and does not contain either the local oscillator frequency or the signal frequency. The output of the balanced converter is fed into an i-f stage operating at 270 kc and then into a conventional diode detector. If listening, rather than panoramic presentation, is desired, the output from a local beat frequency oscillator is mixed with the intermediate frequency in the diode detector. The output from the detector is then fed through a d-c restoration circuit and an amplitude limiting circuit to the vertical amplifier and then to the vertical plates of the cathode ray tube.

A voltage comparison marker circuit has been developed to show by means of a vertical step on the cathode ray tube trace the 10 kc band of frequencies the receiver will display if turned to the "narrow" position. The supply voltage for the X-1 Sonaramic Receiver is 115 volts, 60 cycles.

The equipment is contained in one cabinet 15" x 15" x 15" and is mounted by means of four shock mounts on the bottom of the cabinet and

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two on the top rear, as shown in Fig. 16. The cabinet is made of 1/16 inch steel except the bottom, which is of 1/16 inch Monel. Suitable louvers for ventilation and handles for carrying the equipment are provided on the sides of the cabinet. The components are mounted on two 1/16 inch Monel chasses, mounted in a horizontal plane one above the other. This type of construction facilitates maintenance and repair.

## 4.2 Pre-Amplifier

### 4.21 Input Circuits

Input jack J1 furnishes an all-purpose input which can be used with either a balanced or an unbalanced signal voltage. The attenuator between J1 and the grid of the first amplifier tube V1 is designed so that a voltage will appear at the grid of V1 no matter what type of input is used. Jacks J2 and J3 are connected in parallel, directly to the primary terminals of the input transformer.

T1 is a step-up input transformer with a flat frequency response from 5 to 120 kc. Plate 1 is a plot of its frequency characteristic. It is seen that there is less than 2 db variation from 2 to 120 kc when a resistor of 27 kc is connected across its secondary. By suitable choice of taps provided, the following nominal impedances are obtainable: 500 ohms balanced or unbalanced (taps 1 & 6), 125 ohms balanced or unbalanced (taps 1 & 3, 2 & 5, or 4 & 6), 31 ohms unbalanced (any two consecutive taps), and 281 ohms unbalanced (taps 1 & 5 or 2 & 6). A faraday cage type of shield is placed between primary and secondary windings and is grounded internally to the case. The primary may then be grounded elsewhere in any installation without introducing additional ground loops in existing circuits.

### 4.22 Band-pass Amplifier

The pre-amplifier contains three stages of amplification composed of two high gain tubes, V1 and V2, and one low gain, low-impedance tube, V3. Two ganged potentiometers, R9 and R19, in the cathode circuits of V1 and V2 furnish the amplifier gain control. Their operation is to vary the fraction of the cathode resistance which is by-passed. This method gives about 70 db of control with negligible effect upon the frequency characteristic. A calibration curve is shown on Plate 3. The 180 ohm cathode resistors, R10 and R20, in series with the ganged potentiometers, are present to insure good low-frequency response under minimum gain conditions. The voltage dividers in parallel with the cathode resistors are used to provide Class A bias for the tubes. An RC decoupling network (R6C2a for V1 and R16C6a for V2) is placed between the voltage divider and the grid return point to prevent signals from being fed back into the grid circuit. The amplifier stages are resistance coupled, the RC coupling networks being designed to give a dropping of the frequency characteristic of approximately 3 db at 5 kc. By attenuating all

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frequencies below 5 kc, microphonics and stray pick-up are minimized. An overall frequency response of the pre-amplifier, including the input transformer and low-pass filter, is shown on Plate 2.

The gain of V1 and V2 is about 100 each. The gain of V3 is approximately 4. The overall gain is such that a one microvolt signal in series with 500 ohms applied to the 500 ohm taps on the input, produces a visible deflection on the cathode ray tube screen.

### 4.23 Low Pass Filter

In the previous section it was pointed out that V3 has a low output impedance. This output impedance (3.9K ohms) matches the characteristic impedance of the low pass filter, T2, which has a 120 kilocycle cut-off frequency. The frequency characteristic and electrical details on this filter are shown on Plate 4. To preserve good frequency response, it is essential that the coupling condenser C12, be one of very low capacitance to ground - less than 25 mmfd. To complete the impedance matching, the low pass filter is terminated by 3.9K ohms. This filter is provided to prevent undesirable signals from creating spurious responses in the IF amplifier. To be more specific, the intermediate frequency is 270 kc. Any signal of approximately one-half the intermediate frequency can introduce a spurious signal either if there is an accompanying second harmonic of the signal or if there is a second harmonic produced in the balanced modulator, large enough to introduce a 4th order conversion into the IF amplifier. The latter is possible since the modulator does not balance out even orders. As shown by Plate 4, the signal level is down more than 50 db at 130 kilocycles.

### 4.3 Frequency Conversion

#### 4.31 Linear Mixer

The amplified signal voltage from the preamplifier and the local oscillator voltage from the tapped cathode follower V22B are mixed linearly in tube V8. The sum of these voltages appear in the common plate load of V8a and V8b. It is important to notice that there is no frequency conversion in this stage, but merely a voltage addition. That is if  $e_m = E_m \cos \omega_m t$  and  $e_c = E_c \cos \omega_c t$  represent the two signals as they enter V8, the output voltage,  $e_o$ , across the common plate load of V8 may be represented mathematically as  $e_o = e_m + e_c$ . Both V8a and V8b have degenerative cathodes. This property is to insure linear mixing (i.e. simple addition) even for very large signals. The coupling constants into the next stage are designed to discriminate against frequencies below 5 kilocycles. This tends to eliminate whatever hum might have been picked up in previous circuits.

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### 4.32 Phase Inverter

The phase inverter tube V9 is designed to give a push-pull output voltage between its plate and cathode. The linear mixed signals from V8 are introduced on the grid of the phase inverter. The effective cathode follower input of the phase inverter will permit large voltage swings on the grid without introducing distortion or frequency conversion. In order to obtain perfect converter operation in the balanced converter, it is necessary for the input signals to the converter from the inverter to be exactly equal in amplitude and  $180^\circ$  out of phase, assuming that there are identical pentodes in the balanced converter. Care was taken in the phase inverter circuit design to reduce phase shift. The signals from the cathode and plate of the converter are adjusted for the correct relative amplitude by means of a 5K ohm potentiometer, R49, in the cathode of the inverter. This amplitude control is a degenerative gain control similar to that used in the preamplifier in which the amount of cathode resistor which is bypassed may be varied. For minimum cathode degeneration, the ratio of plate output to cathode output is more than 2 to 1. A similar ratio of cathode output to plate output can be obtained when the gain control is set for maximum degeneration. Because the plate impedance of the tube is increased when degeneration is introduced in the cathode, the plate output contains considerably more power supply hum than the cathode. This results in 60 cycle unbalancing of the balanced mixer. The effect was minimized by reducing the ripple with an RC decoupling filter (R51 and C30).

### 4.33 Double Balanced Converter

The converter uses two tubes, V10 and V11, having a common (tuned) plate load impedance. Signals from the inverter are applied in balanced fashion to the two grids. There are reasons for using a converter of this type. First, the oscillator frequency sweeps down to the frequency of the I.F. amplifier and would produce a large apparent low frequency signal on the cathode ray tube trace if not balanced out. Second, the X-1 Sonaramic is essentially a superheterodyne receiver with an untuned preselector, and is subject to spurious responses. For example, a signal will generate harmonic components in the converter, since all such devices are non-linear. These harmonics may beat with the local oscillator producing spurious signals. In this mixer, any second harmonic so generated is balanced out, and subsequent harmonics are small because of the small curvature of the tube characteristics. Another type of spurious response occurs when the signal frequency is some sub-multiple of the intermediate frequency. This mixer discriminates against odd sub-multiples. The intermediate frequency is more than twice the high frequency cut-off of the preamplifier filter and therefore no submultiple of less order than the fourth can give appreciable spurious response. A theoretical analysis of the mixer performance is given in Appendix 2.

The tubes are operating at a very low  $g_m$  and at a frequency conversion gain of only unity. Whenever a large signal appears at the converter, a change in dynamic bias will automatically unbalance the modulator unless the  $g_m$  characteristics of the tubes are identical. However, at a grid bias of -30 volts, the  $g_m$  characteristics are very flat and any change in dynamic bias on the tube will change the  $g_m$  very little. This property permits the converter to remain balanced over a considerable range of signal voltages even though the tubes are not identical.

#### 4.4 Intermediate Frequency, Audio, and Video Circuits

##### 4.4.1 I. F. Amplifier

Tube V12 is the IF amplifier tube, and T5 and T6 are the IF transformers. The IF is 270kc. The frequency resolution obtainable on the X-1 Sonoram Receiver or any similar panoramic device is dependent upon the IF acceptance and upon the rate of scanning. If the bandwidth of the IF amplifier is too broad, the signals displayed on the cathode ray screen will be correspondingly broadened. On the other hand if the IF bandwidth is made too narrow, the displayed signal will again be too broad since the circuits "ring" for a longer part of the sweep. At the same time, the gain is reduced due to the insufficient time available for build-up of current in the tuned circuits. An optimum bandwidth gives the narrowest displayed pip. The condition for optimum bandwidth is roughly that the heterodyne signal sweep through the IF band in the time it takes to build up full currents. This time is  $\sim 1/\Delta F$  where  $\Delta F$  is the i-f bandwidth. If  $g$  is the number of sweep per second and  $F$  the total band swept, then  $\Delta F/gF$  is the time available to build up current and

$$\frac{1}{\Delta F} \sim \frac{\Delta F}{gF} \quad \text{or} \quad \Delta F \sim \sqrt{gF}$$

is the condition for optimum bandwidth. Measurements have shown that a useful design formula is

$$\Delta F = 1.25\sqrt{gF} \quad \text{Where } \Delta F$$

is measured 3db down. In Appendix 3 it is shown that bandwidths of 6.85 kc and 1.25kc are required on broad and narrow bands respectively.

The i.f. transformers use high Q, center-tapped, litz coils in stackpole G-1 powdered iron pots. The resonant impedance of each coil is about 0.5 megohms. Since this is comparable to the plate resistance of the tubes, the tap is used to reduce tube loading. Coupling is minimized by proper orientation of the coils in the cans. Required increase of coupling is then obtained by capacitive means. In the narrow band position this capacitance is 2 micromicrofarads between the coil taps of the first transformer. None is used in the second transformer. The broad bandwidth is obtained by damping and increased capacitive coupling. The amplifier is identical for broad and listen

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positions of the function switch. A certain amount of detuning takes place when the additional coupling is used, and this is compensated for by condensers C30 and C33 for T5 and C67 and C39 for T6. The gain in the narrow position would normally be considerably greater than that in the broad position. It is reduced to a value comparable to the gain in the broad position by switching R59 into cathode circuit of V12 in the narrow position.

Transformer T5 has a bandwidth of 1.5 kc (3 db down). Transformer T6 has a bandwidth of 3.5 kc (3 db down). When these two bandwidths are combined, the resultant overall bandwidth is 1.25 kc (3 db down). Overall response of the IF is shown in Plate 5.

#### 4.42 Second Detector

Tube V13a is a diode detector. The demodulated voltage appears across R150 and R63. R150 and C71 form a filter to reduce the amount of intermediate frequency in the output. C38 is an IF by-pass.

#### 4.43 Beat Frequency Oscillator

The beat frequency oscillator V13b is a Hartley oscillator tuned to the IF frequency and is coupled into the diode detector by a 2 mmfd condenser. The BFO can be used only in the Listen position. Since the IF bandwidth in the Listen position is 8 KC, the audio acceptance band is + or -4kc. The frequency of the B.F.O. is fixed at 270 kc. The beat note may be observed on the cathode ray tube as well as heard on the head phones.

#### 4.44 Audio Amplifier

The audio amplifier consists of a 6SN7 tube V26 one side of which is a degenerative voltage amplifier and the other a cathode follower feeding 600 ohm head phones through a 2 mfd. condenser. Plate 6 is a frequency response of the amplifier including the effect of the detector circuit feeding it. The output of the voltage amplifier V26A is used also for signal intensification of the cathode ray tube by coupling to the intensity grid of the latter.

#### 4.45 The Vertical Amplifier

The vertical amplifier consists of a "humdinger" type circuit and uses one type 6SN7 tube, V15a and V15b. The operation of this circuit is similar to the operation of the horizontal amplifier V23a and V23b, and reference should be made to Section 4.52 for the detailed operation. Plate 7, showing the characteristic of the circuit, applies equally well to both amplifiers.

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### 4.46 Diode Restorer

Tube V14A, one half of a type 6SL7 dual triode, is used as a diode to restore the signal voltage to a level determined by R69 and R70. Its action is as follows: negative signals are fed to it from the diode second detector through a condenser C40 whose function is to prevent any rectified noise voltage at the detector from shifting the trace in a vertical direction. Since the time average of the voltage on the output side of C40 would be zero without restoration, vertical position of the cathode ray tube base line would depend on the amplitude of a signal. This is especially unpleasant if the frequency of a pip is to be read from a calibration scale on the tube face, or if relative amplitude is to be determined. The diode insures that all signals are negative with respect to the potential of the junction of R69 and R70. In the absence of a signal the potential of this junction determines the base line position.

### 4.47 Vertical Amplitude Limiter

Referring to Plate 7 it is seen that in order for V15A to operate in the center of its characteristic curve, thereby giving the maximum linear voltage swing in its output, the grid must be operated at approximately +30 volts. Since the cathode of V15A operates at approximately +38 volts, about 8 volt bias is maintained on the tube. About 5 of the 30 volts is due to the drop in R129, and 25 in R69. The grid of V14B is connected to the ungrounded side of R69 and therefore the voltage drop in R129 is essentially independent of that in R69.

Tube V14B is a cathode follower used as a voltage limiter. It prevents large signals from deflecting the trace off the screen. Its action is simply explained if it is assumed first that the junction of R69 and R70 is fixed in potential. A negative signal on the grid of V14B cuts off the tube and reduces the drop in R129 from 5 volts to 0 volts. Thus the tube limits the output to 5 volts. However, the potential of the junction of R69 and R70 is slightly dependent on the current in V14B, thereby producing somewhat of a remote cut-off effect. With the shape of signals which appear at the detector output the result of this remote cut-off is a rounding of the tops of the limited pips on the cathode ray tube screen.

### 4.5 Sweep Circuits

#### 4.51 Saw-tooth Generator

The saw-tooth generator is used to supply a saw-tooth voltage to the reactance tube and to the horizontal amplifier, the output from the latter being used to sweep the electron beam across the screen of the cathode ray tube. The saw-tooth generator consists of V17, a type 2050 thyratron, and V18, a type 6SK7 pentode. V17 and V18 are connected in

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series. For the narrow band sweep rate (100 c.p.s.) condenser C46 is placed across the 2050.

Assume, for explanation, that the thyratron is non-conducting and that C46 is discharged. If this is true, the cathode of V17 is at approximately the plate potential of V17. The potential of the grid is fixed by a voltage divider from +350 volts to ground. The grid, therefore, is negative with respect to the cathode and V17 is held in the non-conducting state.

Condenser C46 is free to charge through V18, a constant current pentode.

As C46 charges, the 2050 cathode becomes less and less positive. When the cathode potential has dropped so that the grid to cathode voltage is proper for firing, as determined for the particular plate-cathode potential, V17 will fire and C46 will discharge through it.

Small changes in the sweep rate are made by adjusting the bias on the 6SK7 constant current pentode. The adjustment is made by two parallel potentiometers placed in the cathode circuit, one for broad and the other for narrow. The main function switch selects the appropriate bias. A large change in sweep rate is made by switching C47 in series with C46. The output from the sweep generator is a negative-going saw-tooth voltage which is AC coupled both to the frequency determining network and to the horizontal amplifier.

### 4.52 Horizontal Amplifier

The horizontal amplifier is a "humdinger" type circuit consisting of V23A and V23B, a type 6SN7. For convenience V23A will be referred to as the master tube, and V23b will be referred to as the slave tube. Plate 7 shows the variation in plate voltage of both the slave and master tube as the grid voltage on the master tube is varied. The bias on the slave tube is set at 30 volts. Approximately 30 volts of fixed bias is placed on the grid of V23a by means of a divider consisting of R120 and R122. From Plate 7 it may be seen that if the bias on both tubes is set to 30 volts, the plate voltage on each tube will be approximately 190 volts, and this voltage is appropriate for proper focusing of the cathode ray tube. The saw-tooth voltage of approximately 30 volts peak to peak is ac coupled to the master grid. Consider now the instantaneous value of the saw-tooth at the initial part of the sweep. In this condition the grid of the master tube will be approximately 45 volts positive. V23a will then conduct heavily, causing the master tube plate voltage to drop to approximately 65 volts. Since V23a and V23b have a common cathode resistor R123, the heavy current drawn by V23a will produce a bias across this common cathode resistor. This bias will oppose the 30 volts bias put on the grid of V23b by R126 and will tend to cut off V23b, the slave tube. The plate potential on V23b will then rise to approximately 350 volts.

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Consider next the instantaneous value of the saw-tooth wave at the end of the sweep. This will be the most negative value of the saw-tooth; hence, the grid voltage of the master tube will be at approximately +15 volts. With this voltage on the grid, the master tube will tend to conduct less heavily than formerly, thereby causing plate voltage of the master tube to rise to a value of about 350 volts. The reduced current drawn by V23A will cause a reduced bias voltage to be developed across the common cathode resistor; so that in effect V23b, will have less bias and will tend to conduct heavily. This change in bias will cause the plate voltage of the slave tube to drop to approximately 65 volts. Referring to Plate 7 again, if the voltage on the grid of the master tube varies in a saw-tooth manner, the voltage on the plates of V23a and V23b will likewise vary in a saw-tooth manner; and each will be 180° out of phase with the other. It is apparent now that with a 30 volt peak to peak saw-tooth on the master grid, a saw-tooth voltage of approximately 285 volts peak to peak will appear at the plate of both V23a and V23b. Since the saw-tooth voltage output of V23a and V23b are 180° out of phase, the output voltage from each of the two tubes can be applied to opposite deflection plates of the cathode ray tube; and the result will be a push-pull deflection of the electron beam. The effective deflection voltage is 570 volts.

Essentially the vertical amplifier, V15a and V15b, is the same type of circuit.

### 4.6 Electronic Scanning Circuits

#### 4.61 Local Oscillator and Buffer

V22a and V22b comprise the local oscillator and buffer, respectively. The local oscillator is tuned grid, untuned plate oscillator. The grid coil of T-8 is a three millihenry universal wound coil with a  $\frac{1}{2}$  in. inside diameter. The tickler or plate coil is a one millihenry universal wound coil with a  $\frac{1}{2}$  in. inside diameter. Both coils are mounted on  $\frac{1}{2}$  in. polystyrene hollow tubing containing a Stackpole S49 powdered iron slug. The grid coil is tuned by means of a 10-140 mmfd variable condenser in parallel with a 50 mmfd fixed condenser. R111, a 10K resistor across the plate coil, is used to decrease the feed back and improve the wave form output of the oscillator. The 60 millihenry choke in the plate of the oscillator prevents the oscillator voltage from leaking the ground through the power supply. The output voltage from the oscillator for various frequencies is shown in plate 8.

V22b is a conventional cathode follower. The signal output is taken from a tap point (R19 to ground) in the cathode circuit of the follower. The net average signal output from the buffer to the mixer is approximately 2 volts, peak to peak value.

4.62 Reactance Tube and Dummy Tube

The reactance tube V21, and the dummy tube V20 are type 6SK7 pentodes. The reactance tube operates across the grid coil of T8 and is thus in parallel with it. The reactance tube acts like an inductance, paralleling the inductance of the grid coil of T8. This may be seen in the following way.

A portion of the output voltage of T8 is made to lag in phase approximately  $90^\circ$  by means of R110 and a capacitor composed of C56 and the input capacity of V21. This voltage is then applied to the grid of the reactance tube where it produces a component of current in the plate circuit of V21 which lags the plate voltage by  $90^\circ$ , this being the property of an inductance.

In Appendix 1 it is shown that the effective reactance of a reactance tube is proportional to the transconductance. The saw-tooth applied to V21 through L4 and R108 changes the transconductance and thereby the frequency. R108 plus L4 was chosen roughly resonant in the band to insure high impedance. R108 eliminates the necessity for considering possible series resonances, but is chosen small enough to avoid certain errors which arise when grid current flows in V21.

Plate 9 is an idealized reactance tube-frequency characteristic.

A dummy tube, V20, is used in connection with the reactance tube. The two tubes share the same cathode resistor which is by-passed only for r-f. The grid of the V20 is taken to a positive voltage to bias the cathodes positive. This eliminates the necessity for a negative bias supply for the reactance tube. At the same time, the total current of the pair is almost constant as the reactance tube bias is altered by the saw-tooth. This places a constant load on the voltage regulator tubes V5 and V6 and permits their regulation range to be used almost exclusively in compensating for line voltage fluctuation.

The saw-tooth voltage used on the broad range is developed across the cathode resistance of V19A. That used on the narrow range is approximately one-tenth the broad range saw-tooth amplitude and is taken from an appropriate tap point on the same cathode resistor. The two magnitudes are approximately 100 volts and 10 volts respectively. To these saw-tooth voltages appropriate positive dc bias voltages must be added. DC voltage must be applied to the reactance tube in the broad position to insure that the reactance tube sweeps the frequency over the most linear portion of the characteristic as shown in Plate 10. On the narrow position, the dc bias applied to the reactance tube determines what 10kc portion of the band will be swept. The broad band bias is adjustable between 40 and 60 volts positive by the front panel screw-driver control, R146, and the narrow band bias is adjustable between 0 and 105 volts positive by the Manual Tuning knob, R102.

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After the addition of the bias voltages to the saw-tooth voltages, the resultants are divided down to approximately one tenth of their initial values across selected resistors R97, R98 for the broad range and R99 and R100 for the narrow range. The division ratios are equal within about 0.5%.

The necessity for adding large, magnitudes and then dividing down arises because the frequency marker circuit operates by comparing two voltages and the errors are greatly reduced by having the comparison at high level.

### 4.7 Frequency Marker Circuits

In normal use of the Sonaramic Receiver, the presence of a signal is first indicated by a pip on the broad-band sweep. For further analysis and identification it is desirable to switch to narrow sweep and/or listen position. In the interest of speed, it is necessary that the manual tuning knob be turned to the appropriate setting without the intermediate step of reading frequency from the display and setting it on the dial. This quick setting action is accomplished by means of a step in the trace on the broad sweep which is movable by means of the Manual Tuning knob. The step is moved to intersect the pip. Then, if the receiver is switched to narrow band or to listen position, it is correctly tuned. The step appears only on the broad sweep.

The step is generated in a voltage comparison circuit and is subsequently passed through shaping stages.

#### 4.71 Voltage Comparison

The essential element of the voltage comparison circuit is one section of a 6SN7, V19b, connected as a diode with a 2.2 megohm resistor, R96, in its plate circuit. This comparison device compares the voltage appearing across the two precision dividers consisting of R97, R98 and R99, R100. It will be remembered that a voltage of approximately 40 to 80 volts dc and a saw-tooth of approximately 105 volts appears across divider A. A dc voltage of from 0 to 105 volts and a saw-tooth signal of 10.5 volts appear across divider B. The diode comparison device, V19b, is connected between the tops of precision dividers A and B.

With the receiver in the broad position the voltage appearing at the cathode of V19b consists only of a dc potential variable from 0 to 105 volts since in this position no saw-tooth voltage is switched to the precision divider B through C51 and switch S1D1. The voltage appearing at the plate of V19b consists of a dc potential of between 40 and 80 volts and a saw-tooth voltage of approximately 105 volts, peak to peak amplitude. It will be remembered that the saw-tooth is negative-going and condenser coupled to the plate of V19b, the comparison diode. The bias appearing on the cathode of the comparison diode is fixed by the

Manual Tuning control R102. So long as the net amplitude of the saw-tooth voltage and dc bias applied to the plate and grid of the comparison diode is positive with respect to the bias appearing on the cathode, the diode will conduct; and a constant voltage will appear across R96. As the saw-tooth goes negative, however, it eventually comes to the point where the plate and grid become negative with respect to the cathode; and the diode ceases to conduct. Current will stop flowing through R96 at this instant, and the voltage appearing on the plate side of R96 will be an exact reproduction of the remainder of the saw-tooth wave shape appearing across precision divider A. It can be seen now that the output wave from the comparison diode is essentially a saw-tooth wave with the positive tip clipped off.

In actual operation difficulty is experienced in operating the marker at the extreme ends of the trace. Recommendations are listed in section 8.2 for improvement of the circuit.

#### 4.72 Shaping Circuits

The output wave form from V19b, the comparison diode, is direct coupled to V24a, a cathode follower, to prevent loading of the diode comparison circuit by shaping circuits. The first section of the marking step amplifier V24b, a section of a 6SL7 dual triode, is a cut-off type clipper with the cathode operating at ground potential. The grid and grounded cathode form a dc restoration circuit which will not allow the input voltage on the grid to become positive. The plate voltage is rather low as determined by R128 and R129; therefore the voltage at which the tube cuts off is fixed at a low value. The output of V24b is fed to the second stage of the marking step amplifier V25, a type 6J5 triode, which operates in a manner similar to V24b.

The output step is taken from a tap on the divider which lowers the plate voltage. It contains a dc level of about 10 volts and a step of about 4 volts and is direct coupled into that side of the vertical amplifier which is not used for signal. The centering voltage is applied at this same point. To permit independence of the two voltages, the centering potentiometer R72 and the divider tap point are separately connected to the same point on the vertical amplifier by 150k resistors R71 and R136, each voltage being halved thereby.

The marking step amplifier is switched on and off by grounding the grid of V25. This is done automatically when the receiver is switched either to the narrow or listen position and may also be manually accomplished by throwing S4, the marker "on-off" switch.

#### 4.8 Oscilloscope Circuit

The oscilloscope circuit consists of V16, a type 3BP1 cathode ray tube, and its associated focus, intensity, and intensity modulation circuits. The grid of V16 is maintained at about -1000 volt by the high

voltage supply. The cathode potential is variable from that of the grid to about 50 volt more positive by means of R33. The focus anode potential is variable by means of R35 and the second anode potential is kept at about +200 volts by R77 and R78. This potential is approximately the stand-by average potential of the deflection plates, a condition necessary for sharp focus.

#### 4.9 Power Supplies

##### 4.91 Low Voltage Supply

Intensification of pips is done by feeding a positive signal from the plate of the audio voltage amplifier, V26A, to the intensity grid. A 1 meg series resistance, R79, prevents driving the grid more positive than the cathode thereby avoiding "blooming".

The low voltage power supply uses a 5U4-G full wave rectifier tube (V4). A 115V 60 cycle transformer supplies a secondary terminal voltage of 450 volts each side of center-tap. When rectified and filtered this yields a net output of 350 volts at 150 milliamperes. One center-tapped 5V-3A secondary winding is used as a filament supply for the 5U4-G rectifier tube. Two 6.3V-5A secondary windings supply the filament voltages for all other tubes except the cathode ray tube. The primary and secondary windings are separated by an electrostatic shield to minimize the effect of r-f disturbances on the power lines. The reactance tube oscillator requires a very stable supply voltage since changes in bias or in plate supply introduce frequency shifts. In addition, the frequency marker network, the vertical video amplifier, and horizontal sweep amplifier require stable bias voltages. Consequently, double filtering is used in the power supply to reduce the a.c. ripple; and voltage regulator tubes, V5 and V6, are used to maintain a stable 210V positive supply for the reactance tube oscillator and a stable 105 volt positive supply for the critical bias networks.

##### 4.92 High Voltage Supply

The high voltage power supply employs a 2X2 half-wave rectifier. Because of the very low currents taken by the cathode ray oscilloscope, resistance-capacitance filtering is satisfactory. The transformer is designed to supply a filtered voltage of 1000V and a maximum current of 1 milliampere. One 6.3v-.7A secondary winding is used as the filament supply for the oscilloscope; and one 2.5V-4A secondary winding supplies the filament voltage for the rectifier tube.

#### 5. PROBLEM SPECIFICATIONS

The following paragraphs are references to enclosure A of reference (a), BuShips confidential letter, serial number C4201(940Ce-339-938) of 26 January 1945.

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Specification

The receiver shall give the following indications:

- (1) Visual indication of the spectrum from 0 to 120 kc.
- (2) Visual indication of any ten kc band within the 0 to 120 kc spectrum.
- (3) Audio output to a headset corresponding to extended ten kc band of reference (a) (A(2)). It shall be suitable to provide only one of the three above features at any one time with a switch provided to choose the type of indication desired in position (3) and oscilloscope presentation of audio wave form.

Performance

- (1) Visual range is from 0 to 100 kc; however, because of the low frequency water noise occurring in actual operation, the useful visual range is from 5 to 100 kc.
- (2) There is a visual indication of any ten kc band within the 5 to 100 kc spectrum.
- (3) The audio output to a headset corresponds to the center 8 kc portion of the ten kc band of reference (a), (A(2)).

The X-1 Sonaramic Receiver is provided with a three way switch to choose the type of indication desired. In position three, simultaneous listening and oscilloscope presentation of the audio wave form is shown.

Specification

Sensitivity shall be one microvolt for full deflection with built in sensitivity control.

Performance

One microvolt signal at the input jack yields a visible trace deflection on the oscilloscope. A built in sensitivity control provides a gain variation of about 70 db.

Specification

A vertical scale shall be compressed approximately logarithmically.

Performance

The Preamplifier has no automatic volume control. Considering the schemes of volume compression the following things become apparent:

- (a) If a weak pulse and a strong CW signal occur simultaneously, the weak pulse will be blanked out by the AVC action resulting from the strong CW signal.

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(b) In the amplifier logarithmic limiting of signals greater than predetermined level is undesirable because harmonics produced as a result of this limiting will produce spurious signals in the IF.

Specification The input impedance shall be at least one megohm.

Performance It has been recommended in references (h) and (i) that this specification be changed to read "approximately 100 ohms" and "500 ohms". The X-1 Sonaramic Receiver has provisions for a balanced or unbalanced input of 500 and 125 ohms and various other unbalanced impedances.

Specification Three paralleled input channels shall be provided.

Performance Two input jacks in parallel feed signals into the equipment through an impedance matching transformer. In addition, an all purpose input jack is feeding an attenuator so that large amplitude signals can be introduced into the receiver. The three input channels feed into the equipment simultaneously.

Specification The resolution on a ten kilocycle band presentation shall be 2 kc or better.

Performance Resolution in a ten kc band presentation is 2.5 kc.

Specification The equipment shall be able to indicate signals having pulse widths as narrow as three milliseconds and repetition rates as low as one per second.

Performance The sweep rate and band width of the X-1 Sonaramic Receiver were selected to meet these specifications using the mathematic analysis presented in Appendix 3 of this report. Preliminary tests at the Under Water Sound Laboratory in New London, reference (c), indicated satisfactory reception of 5 millisecond pulses occurring at a rate of one per second.

Specification A single three inch cathode ray tube shall be used.

Performance Presentation is on a type 3BP1 medium persistence cathode ray tube.

Specification Means shall be provided for attaching photographic equipment. Over all size shall be within the limits of 8" wide, 15" deep and 15" high.

Performance Because of the crash program under which the initial model of the X-1 Sonaramic Receiver was produced, time was not available to permit the provision for photographic equipment.

Early estimations of size requirement proved inadequate. The size of the X-1 Sonaramic Receiver is 15" by 15" by 15" less shock mounts, handles and knobs.

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## 6. PERFORMANCE TESTS (Ref. d)

Preliminary tests of the X-1 Sonaromic Receiver under actual operational conditions were made by engineers of the Communication Security Section, Radio Division, NRL, at the U. S. Naval Underwater Sound Laboratory on Saturday, 19 May 1945, aboard USS MACKEREL (SS204). The target vessel was the IX-97.

The tests were designed to see whether the Sonaromic Receiver would actually indicate the different frequencies of under water supersonic carriers sent out by the special Bell Laboratories driver on the target vessel. Of particular interest were the questions of minimum detectable signal strength and the signal strength attenuation over a wide range of frequencies. Frequencies of 18, 25, 50, 70, and 90 kilocycles were transmitted from IX-97. The pulse width varied from 5 to 70 milliseconds and the time interval of a separate frequency transmission was in most cases 20 seconds.

The data obtained yielded the following conclusions:

- a. The 25 kilocycle pulse was received every time it was transmitted over a range varying from 300 to 7000 yards.
- b. The 18 kilocycle pulse was received about 50% of the time during its transmission. The data taken gave no indication of any correlation between the range of the target and the probability of receiving the signal.
- c. The 50 kilocycle pulse was received only once during the test. The range was approximately 600 yards. During all other transmissions of the 50 kilocycle pulse, either no signal of any frequency was received or a spurious frequency near 25 kilocycles was picked up.
- d. For all 90 kilocycle pulse transmissions, no signal of that frequency was received; however, a strong signal of about 25 kilocycles was received.
- e. No 70 kilocycle pulse was received during its transmission. However, about 50% of the time a stray 25 kilocycle signal of medium strength was received.
- f. From all data obtained, there was no positive correlation between spurious signals received at 70 and 90 kilocycles and the range of the target vessel.

These results were confusing, in that the receiver indicated frequencies other than those which were supposedly being transmitted by the equipment in the IX-97. Further tests, however, indicated spurious

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transmissions from the target vessel, explaining the anomalies.

The QB head on the submarine was also tested by means of a monitor. The frequency response curve shown in Plate 12 was taken at 40 kc; the QB response was down almost 40 db below that at 25 kc; at 50 kc the response was down almost 20 db. This in addition to a spurious 25 kc signal from the driver may account for the poor reception obtained at 50 kc.

A final check of the X-1 Sonaromic Receiver with a standard signal generator indicated no difficulty in the receiver equipment itself. The frequency response of the receiver varied less than 3 db over the entire range from 5 to 100 kilocycles. In conclusion, the tests showed that the X-1 Sonaromic Receiver indicates in a satisfactory manner the frequencies of signals present in the water. However, because of the poor frequency response of the QB head on the submarine and the defective transmitter on the IX-97, the results were not as conclusive as had been anticipated.

## 7. CONCLUSIONS

Preliminary tests have shown that the X-1 Sonaromic Receiver is a reliable field instrument. With modifications as listed in Section 8, this receiver will perform with greater effectiveness and will be completely independent of any external instruments for alignment or calibration.

The X-1 Sonaromic Receiver indicates with certainty the presence of short duration signals occurring at a low repetition rate.

## 8. RECOMMENDATIONS

Since the original model was developed, additional work has been done on certain parts of the circuit to improve performance. This section is devoted to discussing various revisions that may profitably be included in future models. These recommendations are based on either development work completed after the model or are extrapolations from experience on the model and on similar gear.

### 8.1 Reactance Tubes

Reference to Fig. 9 and Fig. 10 shows that the actual reactance oscillator characteristic differs considerably from the ideal reactance oscillator characteristic. Because the characteristic of the reactance oscillator is not perfectly linear, when the receiver is switched to the narrow position, a range of frequencies greater or less than 10 kc will be presented. This occurs because the same amplitude of the saw-tooth voltage is being applied to the reactance tube throughout the range of bias covered by the Manual Tuning.

One method of improving the linearity would be to change the reactance and dummy tubes from 6SK7's to 6AC7's. These tubes have a greater maximum  $G_m$  and, therefore, from reference to Appendix 1 it can be seen that not only the total frequency range will be increased but also the sweep linearity over any small portion of the band would be improved. Type 6AC7 tubes have a disadvantage, however, in that, because of their high  $G_m$ , the  $G_m$  characteristics may vary considerably from tube to tube. Thus the circuits become more susceptible to misalignment as tubes are changed. In addition, because of the close spacing of the elements in a 6AC7, such tubes experience large percentage changes in  $G_m$  under shock of vibration, thereby changing the characteristics of the reactance oscillator.

A second possible solution to the problem of increased linearity of the reactance oscillator characteristic of the X-1 Sonoram Receiver involves the use of push-pull reactance tubes as suggested by reference (j). In using a push-pull reactance tube set-up, it is desirable to have one tube act as an inductive reactance and the other tube act as a capacitive reactance. In the theoretical considerations of reactance tubes it can be shown that the equivalent capacitance of a reactance tube operating as a capacitive reactance is equal to the product of the mutual conductance by RC. The capacitive reactance circuit involves inverting the RC phasing network from the inductive reactance case. For the inductive case the effective reactance of the tube is inversely proportional to the mutual conductance and for the capacitive reactance tube the equivalent capacity is directly proportional to the mutual conductance. Since it is desirable to have the effective inductance of the inductive tube and the effective capacitance of the capacitive tube be a maximum simultaneously or minimum simultaneously, saw-tooth voltages  $180^\circ$  out of phase are placed on the capacitive tube and the inductive tube. The total frequency excursion, therefore, will be approximately the sum of the excursions of the individual reactance tubes. With this type of set-up it will be possible to use a smaller portion of the excursion of each of the tubes, thereby increasing the linearity of that portion of the characteristic which is used.

The marker circuit for use with push-pull reactance tubes is substantially unchanged provided there is a direct coupling scheme (such as the "Humdinger") for phase inversion between the two tubes. It is only necessary that the bias and sweep voltages be applied to one grid and this voltage, as before, be compared to the dc on the manual tuning potentiometer.

## 8.2 Frequency Marker

### Revision A

The three proposed marker circuits shown on Plate 15 differ in many respects from the one shown in Plate 14. The diode comparison network was retained in the proposed revisions to the marker circuit except that

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the values of the precision network resistors were increased as indicated in Plate 15. These changes were made necessary because the divider circuits were loading the diode to some extent and were rounding the peak of the sawtooth wave. This rounding effect prevented a voltage comparison at the extreme end of the saw-tooth. The output of the comparison diode is fed to a type 6SL7, one-half of which is maintained as a cathode follower similar to V24a. The new circuit differs in that the cathode resistor is split and the output taken from a tap on the cathode. The output from the cathode follower is fed into an amplifier in the other half of the 6SL7. This amplifier has a gain of approximately 5 and the inverted output is differentiated by means of a 250 mmfd condenser and a 68 K resistor. The output of the differentiating circuit will contain a high amplitude negative pulse in addition to the desired positive signal. In order to do any effective clipping, therefore, it becomes necessary to remove this high amplitude pip.

The bias on 6AC7 pentode clipper is set just below cut off so that the tube will conduct only during the positive portion of the cycle. Since the wave is ac coupled into the grid this occurs only during the square positive portion of the wave form. The output from the 6AC7 is fed into one-half of a 6SN7 acting as an amplifier. The output from this amplifier is fed into an electronic switch. The negative portion of the input wave holds the input tube of the switch at cut-off. When the wave form goes positive the switch is tripped. The output from the switch is taken from the plate of the input tube. In considering all the proposed revisions for frequency marking it is important that adequate output capacity be provided in the power supply filters. It might be desirable therefore, to increase C14 to 16 mfd.

#### Alternate Marker Circuit Revisions

Circuit B of Plate 15 uses several triode clipper stages to produce either a square marking step or a sharp marker pip. The circuit as shown will produce a marker pip. If a marking step similar to the one produced by the circuit now used in the X-1 is desired, the last clipper stage and the differentiating network in its grid circuit may be eliminated and the output taken from the plate of the preceding clipper.

Circuit C of Plate 15 is, in the first stages, similar to Circuit B. The output from point "Z" on circuit B is fed through an amplifier into an electronic switch. The negative pulse input to the switch causes the output tube of the switch to change abruptly from a cut-off to a fully conducting condition. After a time determined by the RC constant of the network from the plate of the input tube to the grid of the output tube, the switch will abruptly return to its initial condition, producing a notch. Thus, the output from this circuit will produce either a sharp marker pip or a marking notch which may be made any desired width by adjusting the time of the RC network.

In the circuit contained in the model the step position is adjusted to bisect a received signal pip. In the circuit which produces a marker pip, the latter is made to coincide with the center of the signal pip. In the circuit which produces a notch the leading edge is made to intersect the center of the signal.

#### Advantages and Disadvantages of Proposed Marker Circuits

Circuit A described in 8.2 above seems to be the more promising of the three circuits. The adjustment of the cathode resistance in the 6AC7 seems somewhat critical, however. Circuit B is an extremely simple circuit using only non-critical triode switches. Both circuit B and circuit C seem to have a tendency to multivibrate at the extreme end of the comparison range. It is anticipated that this could be eliminated by sufficient decoupling. It can be seen, however, that inasmuch as the sweep frequency at its lowest value is 334 cycles, this would involve using rather large decoupling condensers and resistors.

#### 8.3 Crystal Controlled Beat Frequency Oscillator

By replacing the present beat frequency oscillator by a crystal controlled oscillator, it will be possible to align the local oscillator of the receiver without using any external signal source. If this modification is made, the proposed alignment instructions set forth in Section 9 would be changed slightly.

#### 8.4 Crystal Calibrating Oscillator

The addition of a crystal calibrating oscillator would permit correct adjustment of the broad band frequency excursion without any external standard signal. The frequency of this oscillator should be set at the maximum supersonic frequency to be covered on the broad range. For the X-1 Sonaramic Receiver, the frequency would be set at 100kc.

This addition would facilitate alignment of the equipment in that a signal pip would appear at both ends of the horizontal trace - a zero pip at the low frequency limit as the local oscillator sweeps through the intermediate frequency and a pip from the crystal calibrating oscillator at the high frequency.

#### 8.5 Calibrating Voltmeter

If the recommendations in 8.3 and 8.4 are followed, and if, in addition, a suitable panel voltmeter that can be switched across the diode load resistance (R63) is incorporated, the receiver would be completely independent of external equipment for all alignment, including the I.F., the local oscillator, and the broad band frequency excursion.

### 8.6 Intermediate Frequency Changes

It is proposed that the intermediate frequency be increased from 270 kc to 375 kc. This would permit increased linearity of the reactance oscillator characteristic inasmuch as the ratio of the frequency excursion to the intermediate frequency would be decreased approximately 10%. It should still be possible with careful design to obtain the narrow band widths necessary.

### 8.7 Revised Chassis Design

Potentiometer R89, the frequency range potentiometer, should be incorporated as a front panel control near the "Wide Frequency Center" control, R146. This change would permit frequency calibration alignment of the receiver without its removal from the cabinet.

If space becomes a prime factor, it may be desirable to build the equipment in two units, one of which may be located remotely from the other. Space may be saved by using miniature tubes for all circuits except the vertical and horizontal amplifier. The maximum rated plate voltage of miniature tubes does not permit sufficient voltage swing in the vertical and horizontal amplifier circuits used to give full deflection of the cathode ray tube beam. Several indicator units may be used in remote locations. Each indicator should contain a "humdinger" and be fed by cathode followers in the receiver unit. Centering may be left in the indicators.

### 8.8 Fixed Bias on Balanced Converter

As mentioned in section 4.33, any considerable variation in signal strength into the converter produces a change in operating point and causes unbalance. This is observable as reappearance of the zero pip. The use of a VR tube to bias the cathodes of V10 and V11 positive is suggested. The grids should then be run to a suitable positive bias (common to both grids) to obtain proper operation.

## 9. PROPOSED ALIGNMENT SCHEME

Set band switch to L and manual tuning knob to zero, the maximum counter-clockwise position. Place a voltmeter with the negative terminal to the ungrounded junction of C38 and R63. Turn the B.F.O. switch to Off. Adjust the local oscillator condenser for a peak reading on the meter. Should the peak reading be low and indistinguishable, the magnitude of the reading on the meter may be increased by unbalancing the balanced converter by means of the "Balance Control" on the front panel.

Remove the sweep generator thyratron V17 and turn down the intensity control of the cathode ray tube to prevent burning the screen. Turn the band switch to N position and adjust C30, C33, C67 and C39 for a maximum

deflection of the meter. At this stage the oscillator has been adjusted so that its low frequency (the frequency of the beat frequency oscillator) is the center frequency of the IF band both on the Broad and Narrow positions.

Again throw the band switch to the Listen position, turn the BFO switch to On and adjust the BFO frequency for a zero beat in the phones. This insures that the BFO is oscillating at the IF frequency.

Reinsert the sweep thyratron V17 and turn up the CRT intensity until the trace is again visible.

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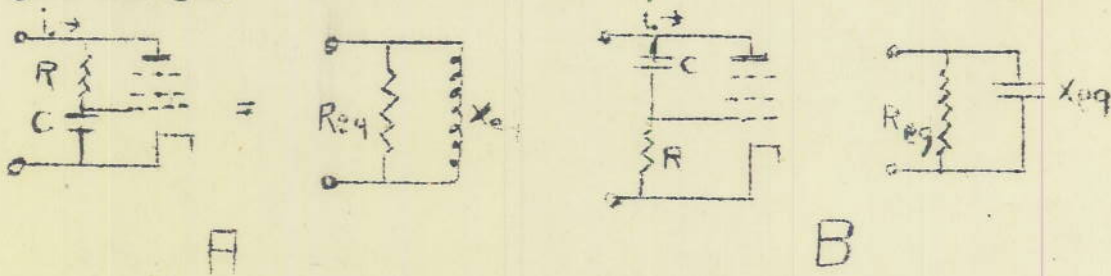
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APPENDIX I

MATHEMATICAL ANALYSIS OF INDUCTIVE AND CAPACITIVE REACTANCE TUBES

Circuits A and B below are examples of simple reactance tube circuits in basic form. A behaves like an inductance and B like a capacitance. Their basic principle is that the voltage applied across the input terminals produces a component of plate current 90° out of phase with the applied voltage by virtue of a phase shift network supplying the grid voltages.



The analysis of each circuit proceeds by a computation of the input admittance  $Y = G + jS$  from which the equivalent parallel reactance ( $X_{eq}$ ) and equivalent resistance ( $R_{eq}$ ) is obtained as

$$R_{eq} = 1/G, \quad X_{eq} = -1/S$$

For unit applied voltage in circuit A, a grid voltage  $\frac{jX_c}{R - jX_c}$  appears, where  $X_c = -1/\omega C$ . This produces a plate current,  $G_m$  times greater than the applied voltage,  $G_m$  being the transconductance. This current compared to the unit voltage yields the admittance

$$Y = \frac{G_m(X_c^2 - jRX_c)}{R^2 + X_c^2}$$

if the current through the phase network is neglected, which can be done whenever  $G_m R \gg 1$ . This is true in a practical case except when the tube is almost completely biased off, but the effect of the reactance tube is then almost nil. Using the dimensionless constant  $\lambda = \frac{-R}{X_c} = RC\omega$  we obtain:

Circuit A

- (1)  $R_{eq} = \frac{\lambda^2 + 1}{G_m}$
- (2)  $X_{eq} = \frac{\lambda^2 + 1}{\lambda G_m}$
- (3)  $Q = \frac{R_{eq}}{X_{eq}} = \lambda$

Circuit B

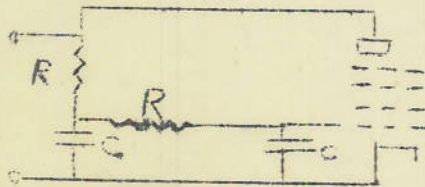
- $R_{eq} = \frac{\lambda^2 + 1}{\lambda^2 G_m}$
- $X_{eq} = \frac{\lambda^2 + 1}{\lambda G_m}$
- $Q = 1/\lambda$

The maximum reactive effect is obtained with that value of  $\lambda$  which makes  $X_{eq}$  a minimum. By differentiation of  $X_{eq}$  and equating to zero, we get an optimum  $\lambda = 1$ , and for both types of circuits

$$\begin{aligned} (4) \quad & R_{eq} = 2/G_m \\ (5) \quad & X_{eq} = 2/G_m \\ (6) \quad & Q = 1 \end{aligned}$$

This represents a great amount of damping which may be intolerable to any oscillator used in conjunction with the reactance tube. The Q may be raised at the expense of the magnitude of frequency deviation by choosing a more conservative mid-band value of  $\lambda$  as seen in equation (3). This implies a phase shift closer to  $90^\circ$ .  $\lambda = 1$  implies  $45^\circ$  of phase shift, 3db of attenuation at the grid, equal resistance and reactance in the phase network, and equal in-phase and quadrature currents in the reactance tube.

An intrinsically better circuit than either A or B (but more difficult to construct practically) employs a two stage R-C phase network and may be made either capacitive or inductive by proper arrangement of the components. The analysis of the inductive type is shown below.



The grid voltage for unit applied voltage to the terminals may be shown to be.

$$V = \frac{jX_c}{R+jX_c} \cdot \frac{jX_c}{R+jX_c + \frac{jRX_c}{R+jX_c}}$$

$$i = G_m V = Y = G+jS = \frac{1}{R_{eq}} - \frac{j}{X_{eq}}$$

$$\text{and for } \lambda \equiv -\frac{R}{X}$$

$$(7) \quad R_{eq} = \frac{\lambda^4 + 7\lambda^2 + 1}{G_m (1 - \lambda^2)}$$

$$(8) \quad X_{eq} = \frac{\lambda^4 + 7\lambda^2 + 1}{3G_m \lambda}$$

$$(9) \quad Q = \frac{3\lambda}{1 - \lambda^2}$$

(note that for  $\lambda > 1$ ,  $R_{eq} < 0$ , and the circuit becomes a phase-shift oscillator)

Optimum  $\lambda$  is obtained by differentiation and is  $\lambda_{optimum} = 0.368$  for which value  $R_{eq} = 2.3/G_m$  (10)

$$X_{eq} = 1.8/G_m \quad (11)$$

$$Q = 1.3 \quad (12)$$

Note that  $X_{eq}$  is lower (and therefore better) by  $2^{1.8}$  than the simpler case and  $Q$  is higher by  $1.3/1$ .

In practice, care must be taken to shield the components of the phase network from each other to prevent undesired phase-shift.

Improvement in reactance tube circuits is also possible by utilizing amplification between the phase network and the tube to permit shifts close to  $90^\circ$  without excessive attenuation.

## APPENDIX II

## THEORETICAL CONSIDERATIONS OF THE DOUBLE BALANCED CONVERTER

It is oftentimes desired to obtain the side band frequencies ( $\omega_c \pm \omega_m$ ) of two signals  $\omega_c$  and  $\omega_m$  and at the same time eliminate the fundamentals and all other nth order generated frequencies. Shown below is a simplified circuit of the double balanced modulators, V10 and V11. Triodes, operating in the non-linear portion of their mutual characteristic curves, will replace the 6SK7's actually used in the Sonaramic Receiver for this discussion.

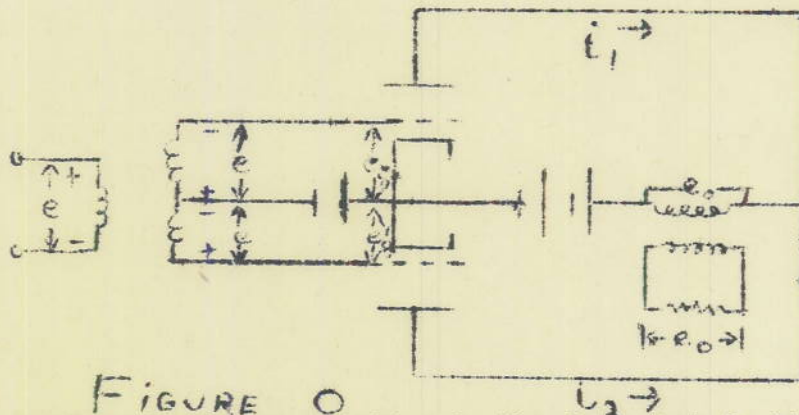


FIGURE 0

The modulator's importance lies in the fact that in it certain frequency components can be suppressed and do not appear in the output. It is similar to the push-pull amplifier; and the frequency suppression in the modulator is related to the fact that in the push-pull amplifier only the odd harmonics generated in the tubes appear in the output current, while the even harmonics appear in the plate-power supply current. The circuit as shown has one set of terminals for the introduction of the two voltages to be mixed, and a single set of terminals from which the output voltage,  $e_0$ , can be taken. It is assumed that the transformers are ideal and that a resistance load is connected across the output terminals. It is also convenient, though not essential, to assume that the turns ratios of the transformers are unity. The two tubes are to be considered identical. From the drawing, it is seen that

$$(1) \quad e_g = -e$$

and

$$(2) \quad e'_g = e$$

Since the transformers are ideal and the load is a pure resistance,

$e_0 \propto i_1 + i_2$ , where  $i_1$  and  $i_2$  are the a.c. components of the respective plate currents.

$$(3) \quad \therefore e_0 = K_0(i_1 + i_2)$$

The a.c. component of the plate current in a non-linear triode circuit may be given by a Taylor's series approximation as

$$(4) \quad i_1 = -a_1 e + a_2 e^2 - a_3 e^3 + a_4 e^4 - \dots \text{etc.}$$

$$(5) \quad \text{similarly, } i_2 = a_1 e + a_2 e^2 + a_3 e^3 + a_4 e^4 + \dots \text{etc.}$$

Substituting (4) and (5) into (3),  
gives

$$(6) \quad e_o = K_o (2a_2 e^2 + 2a_4 e^4 + \dots \text{etc.})$$

This result shows that only even order terms will appear in the output of the balanced modulator, provided the modulator is balanced.

As has been stated in Part 4.31,  $e = e_m + e_c$ , where  $e_c$  is the signal voltage and  $e_m$  is the local oscillator voltage.

$$\therefore (7) \quad e_o = K_o [2a_2 (e_m^2 + 2e_m e_c + e_c^2) + \dots \text{etc.}]$$

Trigonometric identities will show that  $e_c^2$  represents the second harmonic of the signal voltage, while  $e_m e_c$  represents the side band frequencies,  $\omega_m \pm \omega_c$ . With the proper selection of a low pass filter in the preamplifier and a suitable intermediate frequency, as discussed in Part 4.23, side band frequency ( $\omega_m - \omega_c$ ) will be the only signal below the 4th order of conversion entering the IF amplifier.

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## APPENDIX III

### DETERMINATION OF OPTIMUM BAND WIDTHS AND SWEEP RATES

The X-1 Sonaromic Receiver includes both a wide band and a narrow band presentation, only one display being used at any one time. It is specified that the wide band sweep cover a total of 100 kc with a repetition rate sufficiently rapid to insure the detection of all incoming pulse signals of duration exceeding 3 milliseconds. For the narrow band display, however, it is not required that the probability of intercepting each incoming signal be unity. The narrow band sweep must be chosen at least as great as the acceptance band of the broad band display; and both acceptance bands must afford good resolution of displayed frequencies.

Let the following notation be used:

- F = sweep band width
- $\Delta F$  = acceptance band width
- g = sweep repetition rate
- T = pulse period, (repetitive pulses being considered)
- $\Delta T$  = pulse duration
- n = number of sweeps per pulse
- p = probability that a pulse will be intercepted during a given pulse period.

The first condition imposed upon the wide band display, namely, that all pulses of duration greater than 3 milliseconds be intercepted, is satisfied if the sweep frequency g exceeds

$$\frac{1}{\Delta T} = \frac{1}{3 \times 10^{-3}} \approx 334 \text{ cycles per second}$$

It is assumed that optimum resolution is obtained when the acceptance band  $\Delta F$ , the sweep band F, and the sweep frequency g, are related by the equation

$$\Delta F = k \sqrt{gF}, \quad (3\text{db down})$$

A rough derivation of this is given in section 4.41. Experimental investigations conducted at the Naval Research Laboratory indicate that k should be 1.25, and that the accompanying resolution is  $2\Delta F$  (6db down).

Using  $g = 300$  sweeps per second (somewhat better than 334 from the practical standpoint of setting the sweep rate by reference to the 60 cycle line) and  $F = 100$  kc, one finds that  $\Delta F$  should be 6.85 kc (3db down) for optimum resolution, and that the resolution (6db down) is roughly 13.7kc.

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In the discussion of the narrow band display, it becomes necessary to know the probability of intercepting a specific incoming pulse signal. To determine this probability, the acceptance band can be thought of as a small interval moving in the direction of increasing frequencies over the larger frequency interval comprising the sweep band. Neglecting end effects, the pulse is intercepted if, when it comes on, the leading edge of the acceptance band lies in the interval from  $f - g\Delta T$  to  $f + \Delta F$ , where  $f$  is the frequency at which the pulse appears on the display. The total time favorable to interception of the pulse during a single sweep is

$$\frac{g\Delta T + \Delta F}{g}$$

$$= \Delta T + \frac{\Delta F}{g},$$

in the sense that if the pulse comes on at any time during a certain interval of the length given, then it will be intercepted. There being  $n$  sweeps per pulse, the total time out of a period  $T$ , favorable to seeing the pulse is

$$n\Delta T + \frac{n\Delta F}{g}$$

Hence the probability that a given pulse will be intercepted is

$$p = n \frac{\Delta T}{T} + \frac{n\Delta F}{gT}$$

or

$$p = g\Delta T + \frac{\Delta F}{F}.$$

Since  $p$  increases with  $g$  and  $\Delta F$ , it is best to take  $\Delta F$  as large as permissible. If it be specified that for the narrow band display, the sweep be 10kc, and that the acceptance band be 1.25kc, then optimum resolution of 2.5kc is obtained by using a sweep repetition rate of 100 sweeps per second.

With these values for the quantities  $g$ ,  $\Delta F$ , and  $F$ , the probability of seeing a 3 millisecond repetitive pulse in a single pulse period is

$$p = .425$$

Considering the viewing of separate pulses as independent events, one can draw up the following table:

<u>Number of pulse periods</u>	<u>Probability of seeing at least one pulse</u>	<u>Probability of seeing at least two pulses</u>
1	.425	0
2	.669	.181
3	.810	.388
4	.891	.567
5	.937	.707
6	.964	.802
7	.979	.874
8	.988	.916
9	.993	.948
10	.996	.966

For a 5 millisecond pulse:

$$p = .625$$

1	.625	0
2	.859	.391
3	.947	.683
4	.980	.848
5	.992	.927
6	.997	.967
7	.999	.986
8	1	.995
9	1	.998
10	1	.999

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## REFERENCES

### Problem References:

- (a) BuShips ltr Ser. No. C-4201(940Ce-339-938) of 26 Jan 1945 to NRL
- (b) CNO Secret ltr to BuShips and BuOrd S-S75-1/100 dated 4 Dec 1944 (940 Ser. No. 120415)
- (c) U. S. Naval Underwater Sound Laboratory Memorandum - NP24(A16)(WWS) Ser. 0412 of 24 May 1945
- (d) NRL Memorandum C-S68/46(476-10), C-471-186/45 dated 10 March 1945 to Supt. Sound Division
- (e) OBH Sonar Conference - C-F42-1/84/RCM(320:JRZ) Ser C320-172/45 of 6 March 1945
- (f) Telephone Conference Report - C-F42-1/84/RCM(320:JRZ) Ser C320-173/45 of 6 March 1945
- (g) Letter to BuShips Ser. No. C-471-239/45, C-S68/46(476/10).

### Technical References:

- (h) RCA Review - Volume 5, No. 1 (1940-1941) Pages 8-9 "Reactance Tube Frequency Modulators" by M. G. Crosby
- (i) NDRC Report, Div. 15 dated 10-23-44 "Notes on the Selection of Values for Use in the Phase Network of Reactance Tubes".
- (j) Electronics - August 1939, Page 14 "Trigger Circuits" by H. J. Reich.

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PRIMARY: 70 TURNS  
#36 SCF CT  
1-16 MM

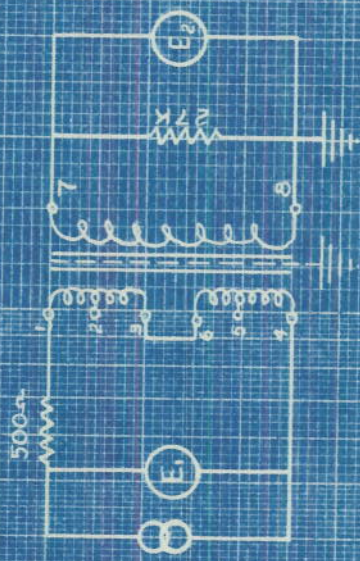
SECONDARY 1300 TURNS  
#40 SSE

UNIVERSAL WOUND  
CORE FF 26-27 "4750"  
100% INTERLEAVED  
SQUARE STACK

INSULATION .003"K ON  
EACH SIDE OF .003"  
COPPER SHIELD

POTTED IN WAX IN 4046  
CASE

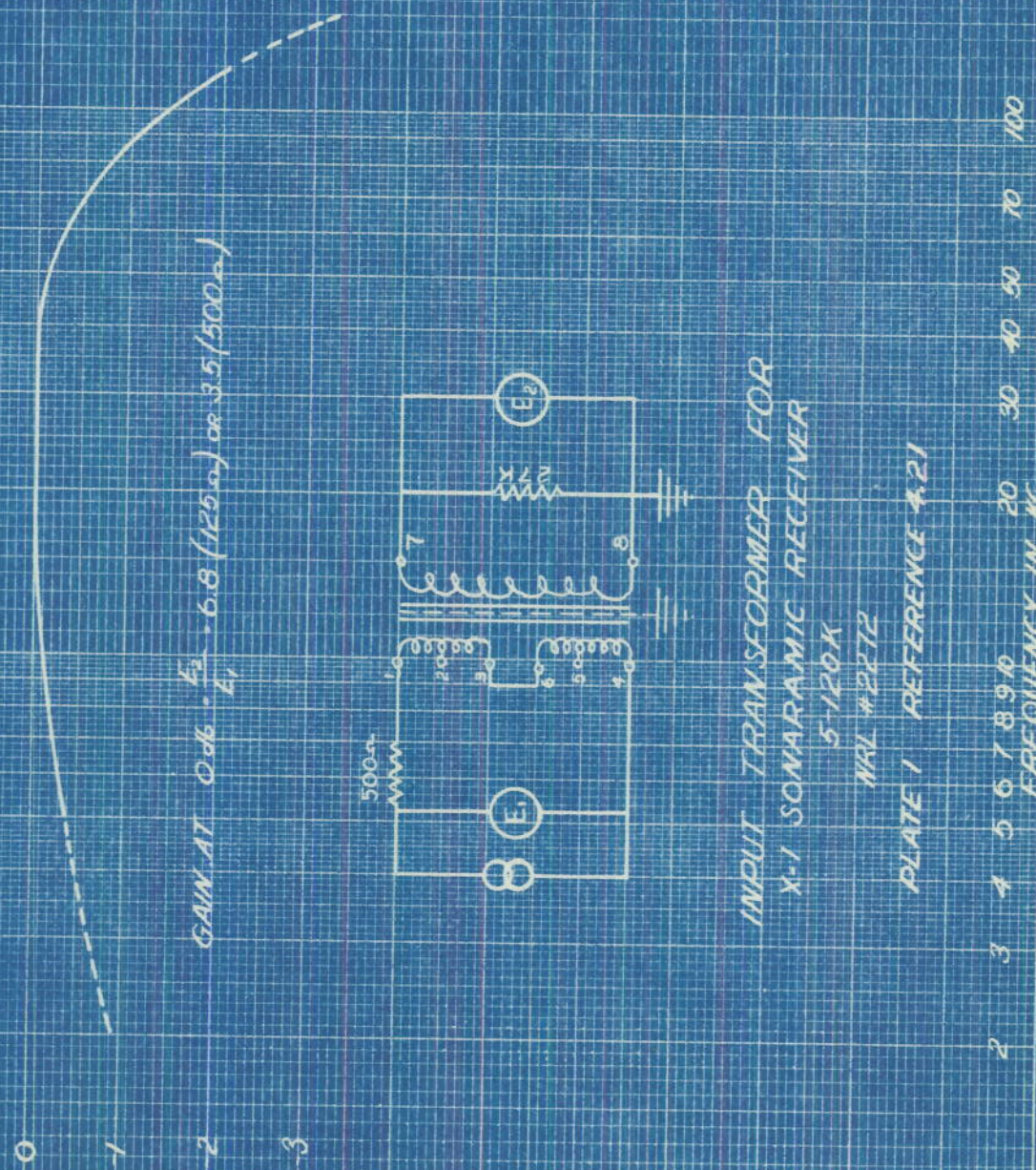
GAIN AT  $0.46 \cdot \frac{E_2}{E_1} = 6.0$  (125 $\mu$ ) OR 3.5 (500 $\mu$ )



INPUT TRANSFORMER FOR  
X-1 SONAR AM RECEIVER

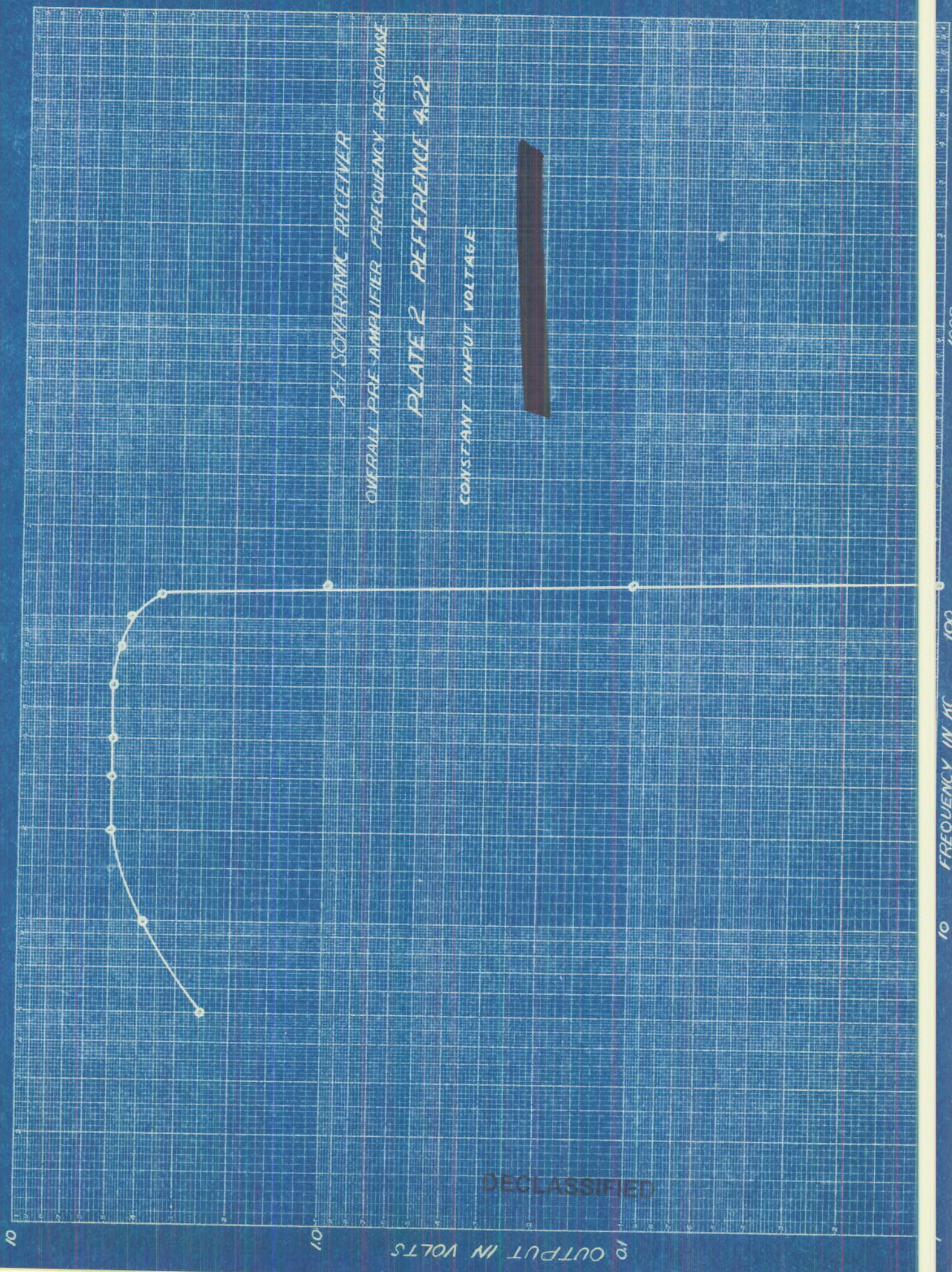
5-120K  
MIL #2272

PLATE 1 REFERENCE #21



FREQUENCY IN KC

4 JUN 60



X-1 SONAROMATIC RECEIVER

OVERALL PRE-AMPLIFIER FREQUENCY RESPONSE

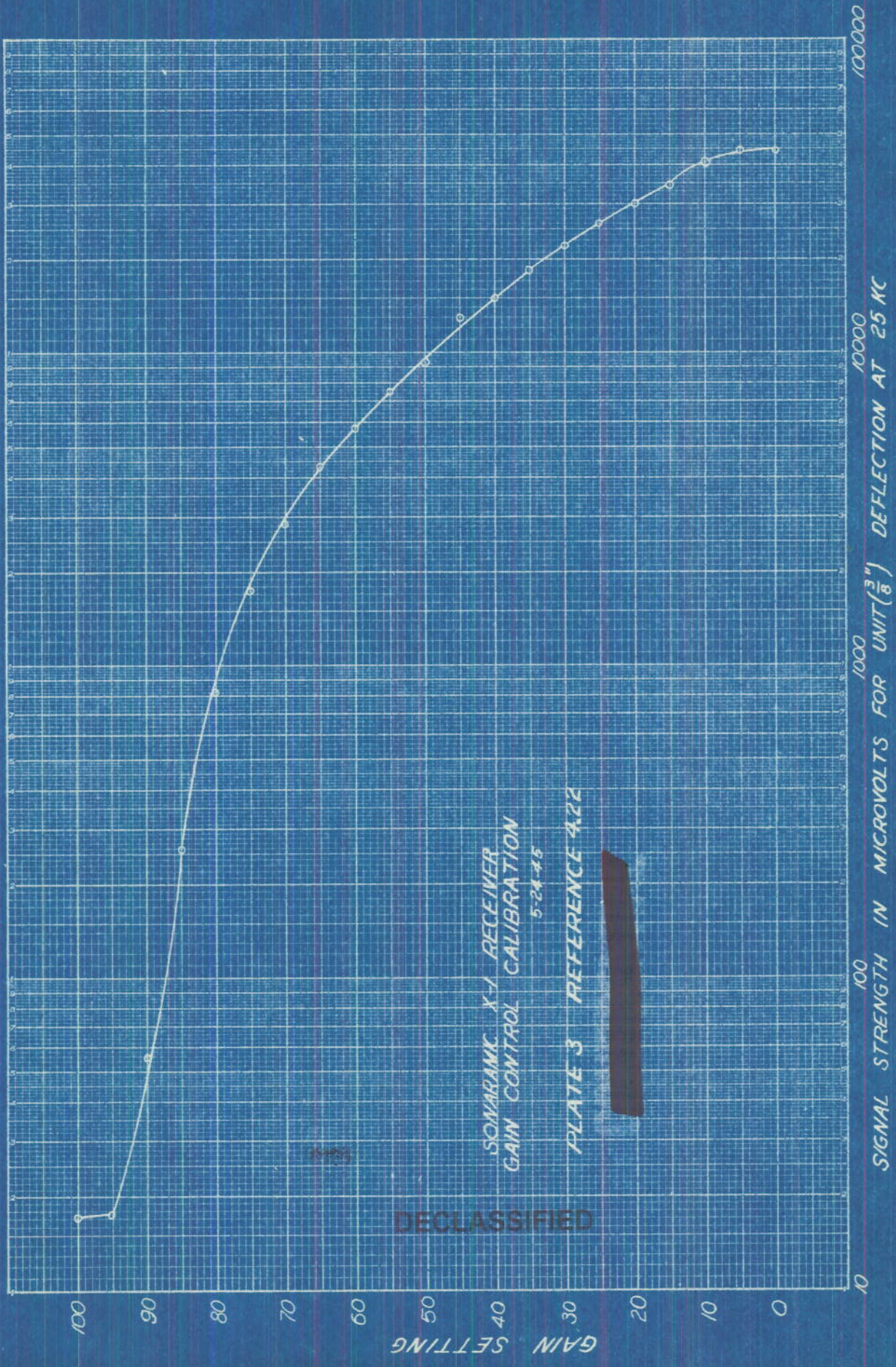
PLATE 2 REFERENCE 422

CONSTANT INPUT VOLTAGE



DECLASSIFIED

NO. 31726, 20 DIVISIONS PER INCH (120 DIVISIONS) BY FOUR CIRCLES PER INCH



SOMARMC X-1 RECEIVER  
GAIN CONTROL CALIBRATION  
5-24-46

PLATE 3 REFERENCE 422



DECLASSIFIED

SIGNAL STRENGTH IN MICROVOLTS FOR UNIT  $(\frac{3}{8})$  DEFLECTION AT 25 KC

10.39 mH COIL - 4 PIES

OTHERS 2 PIES

EACH PIE 150 TURNS #38G

UNIVERSAL WIND

1/2" I.D. 1/8" CAM

CORES - W.F. MOLYBDENUM

PERMALLOY RINGS

0.8" O.D. x 0.5" I.D. x 0.25" H

M60

CHICAGO CAN #1046



INSERTION LOSS at 24 KC

$$= 20 \log \frac{1}{2} = -0.4 \text{ db}$$

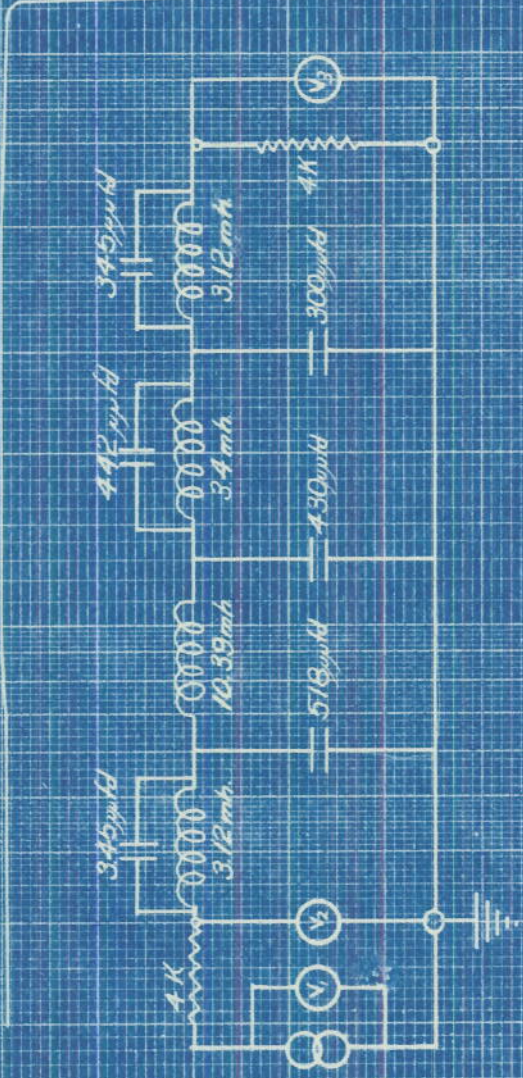
f<sub>c</sub> - 120

R - 4000

476-X-427/2

4-14-46

1000



X-1 SONARAMIC RECEIVER  
LOW PASS FILTER  
PLATE 4 REFERENCE 4.23

FREQUENCY IN KC

100

10

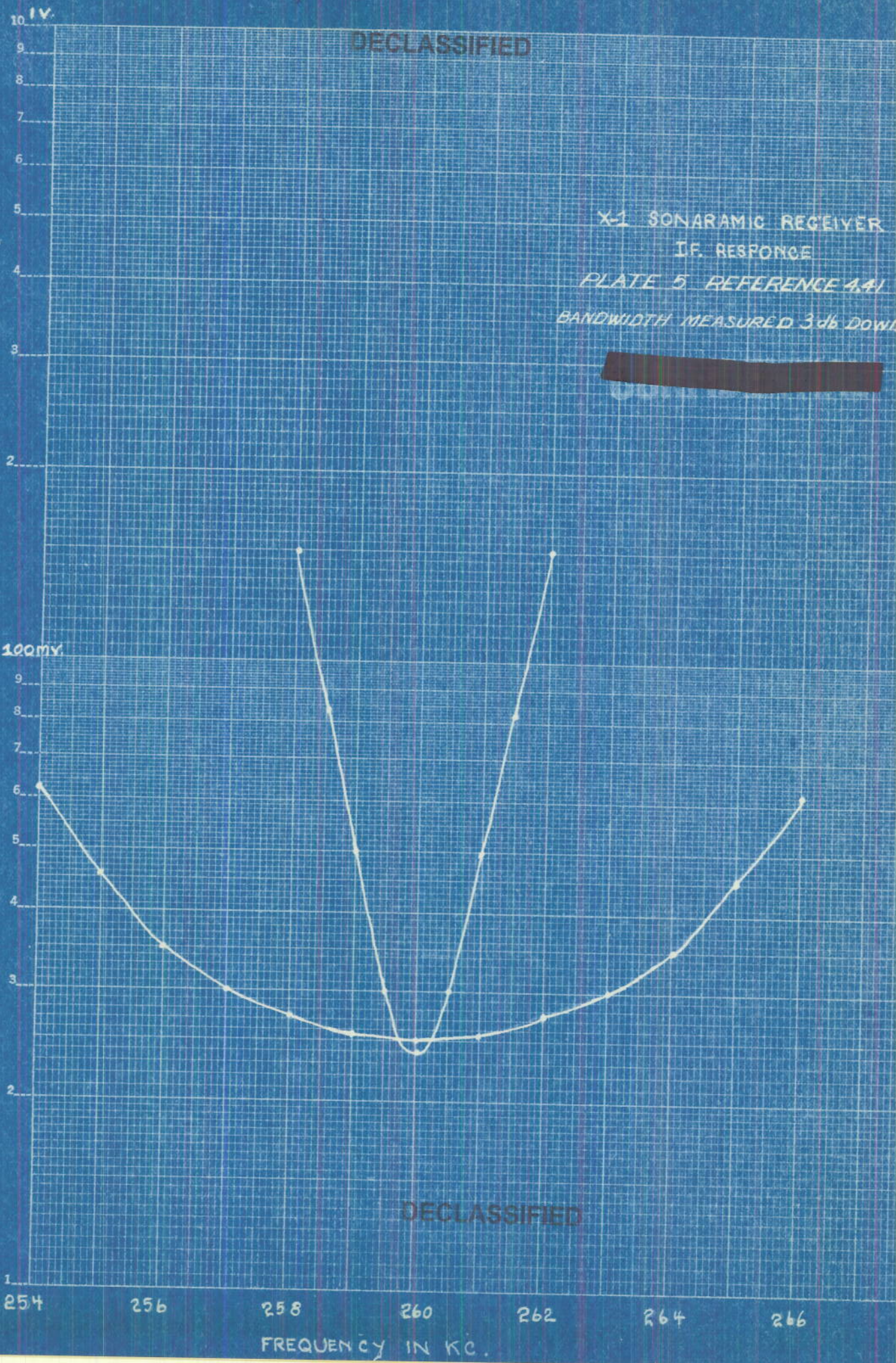
1

DECLASSIFIED

X-1 SONARAMIC RECEIVER  
I.F. RESPONSE  
PLATE 5 REFERENCE 4.41  
BANDWIDTH MEASURED 3db DOWN



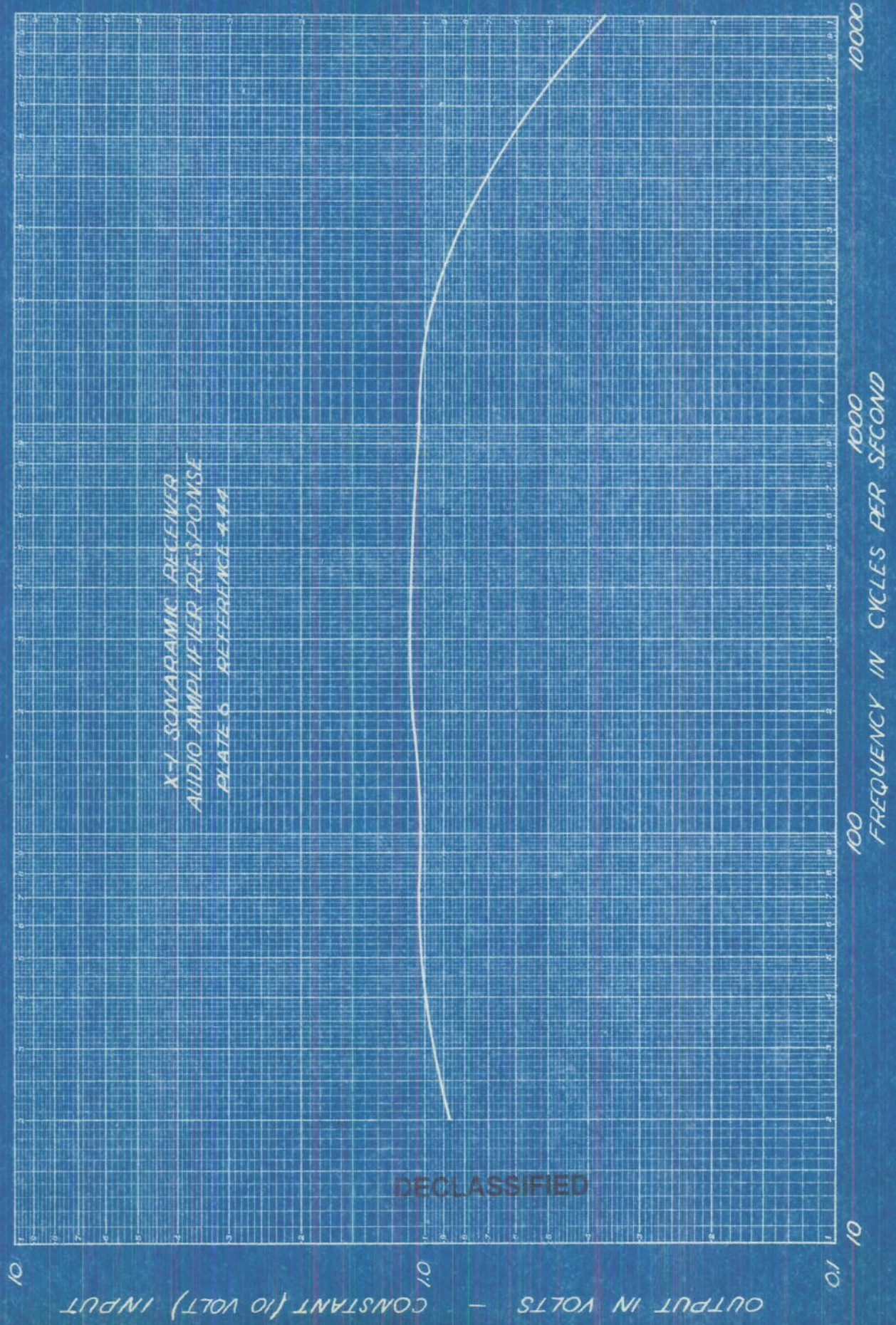
INPUT TO I.F. FOR CONSTANT 1.0V. OUTPUT FROM DETECTOR



KEUFFEL & ESSER CO., N. Y. NO. 359-83  
Semi-Logarithmic, 2 Cycles, X 30 to the Inch,  
MADE IN U. S. A.

DECLASSIFIED

DECLASSIFIED



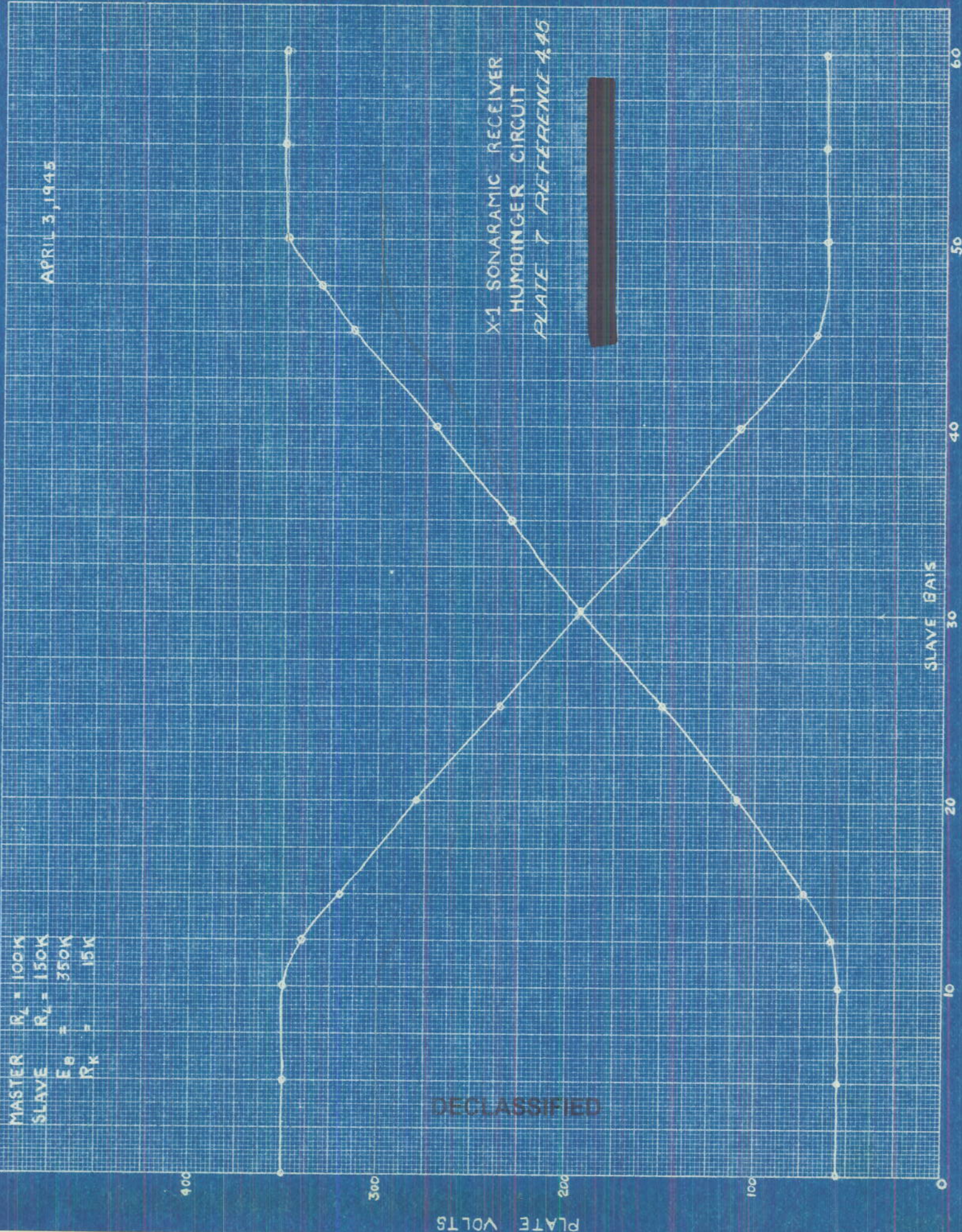
APRIL 3, 1945

X-1 SONAR MIC RECEIVER  
HUMBINGER CIRCUIT  
PLATE 7 REFERENCE 4.46



MASTER  $R_L = 100K$   
SLAVE  $R_L = 150K$   
 $E_B = 350K$   
 $P_K = 15K$

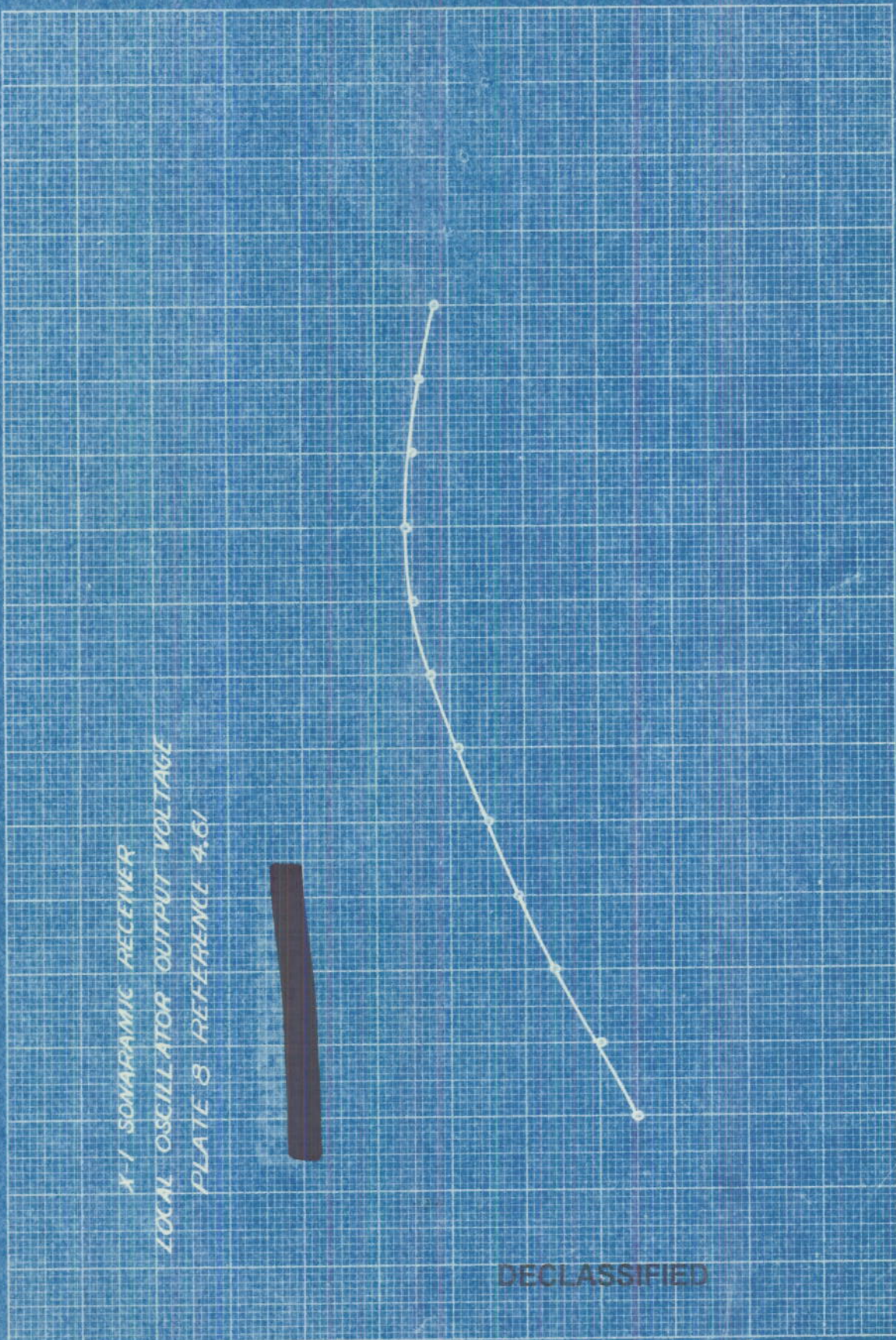
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X-1 SONAROMATIC RECEIVER  
LOCAL OSCILLATOR OUTPUT VOLTAGE  
PLATE 8 REFERENCE 4.61

OUTPUT VOLTS (PEAK-PEAK)

FREQUENCY IN KC.



DECLASSIFIED

SONARAMIC X-1 RECEIVER  
IDEAL REACTANCE OSCILLATOR CHARACTERISTIC  
PLATE 9 REFERENCE 462



NO. 319A. MILLIMETERS, 208 BY 250, DIVISIONS  
COLEX ELECT. COMP. CO., INC., WORWOOD, MASSACHUSETTS.  
PH. U. S. 410  
400  
390  
380  
370  
360  
350  
340  
330  
320  
310  
300  
290  
280  
270  
260  
250

FREQUENCY IN KC

BROAD SWEEP  
RANGE  
120 KC

NARROW SWEEP  
RANGE  
10 KC

0.19 VOLTS  
SWEEP

MANUAL TUNING  
SWEEP

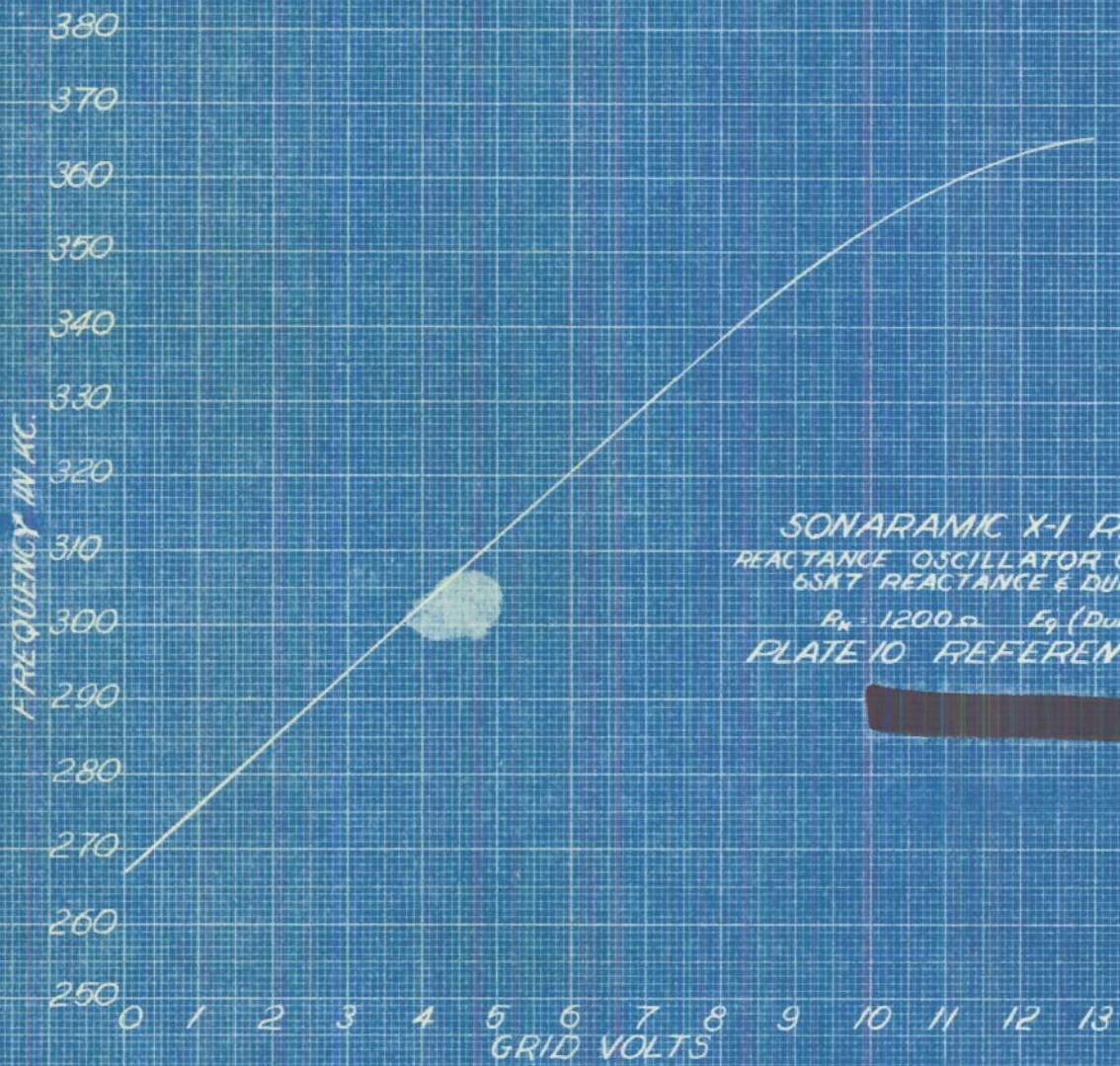
FREQUENCY CENTER  
WIDE  
SETTING

2.20 VOLTS  
SWEEP

GRID VOLTS

DECI 4 ASSIFIED

DECLASSIFIED



SONAROMIC X-1 RECEIVER  
REACTANCE OSCILLATOR CHARACTERISTIC  
6SK7 REACTANCE & DUMMY TUBE  
R<sub>k</sub> = 1200 Ω E<sub>g</sub> (Dummy) = +3 VOLTS  
PLATE 10 REFERENCE 4.62



DECLASSIFIED

6 MAY 57

110

105

100

95

90

85

80

75

70

65

60

55

50

45

40

35

30

25

20

15

10

5

5

10

15

20

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45

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55

60

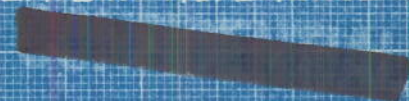
65

70

75

DIAL SETTING

SONARAMIC XI RECEIVER  
MANUAL TUNING CALIBRATION  
PLATE II REFERENCE 463

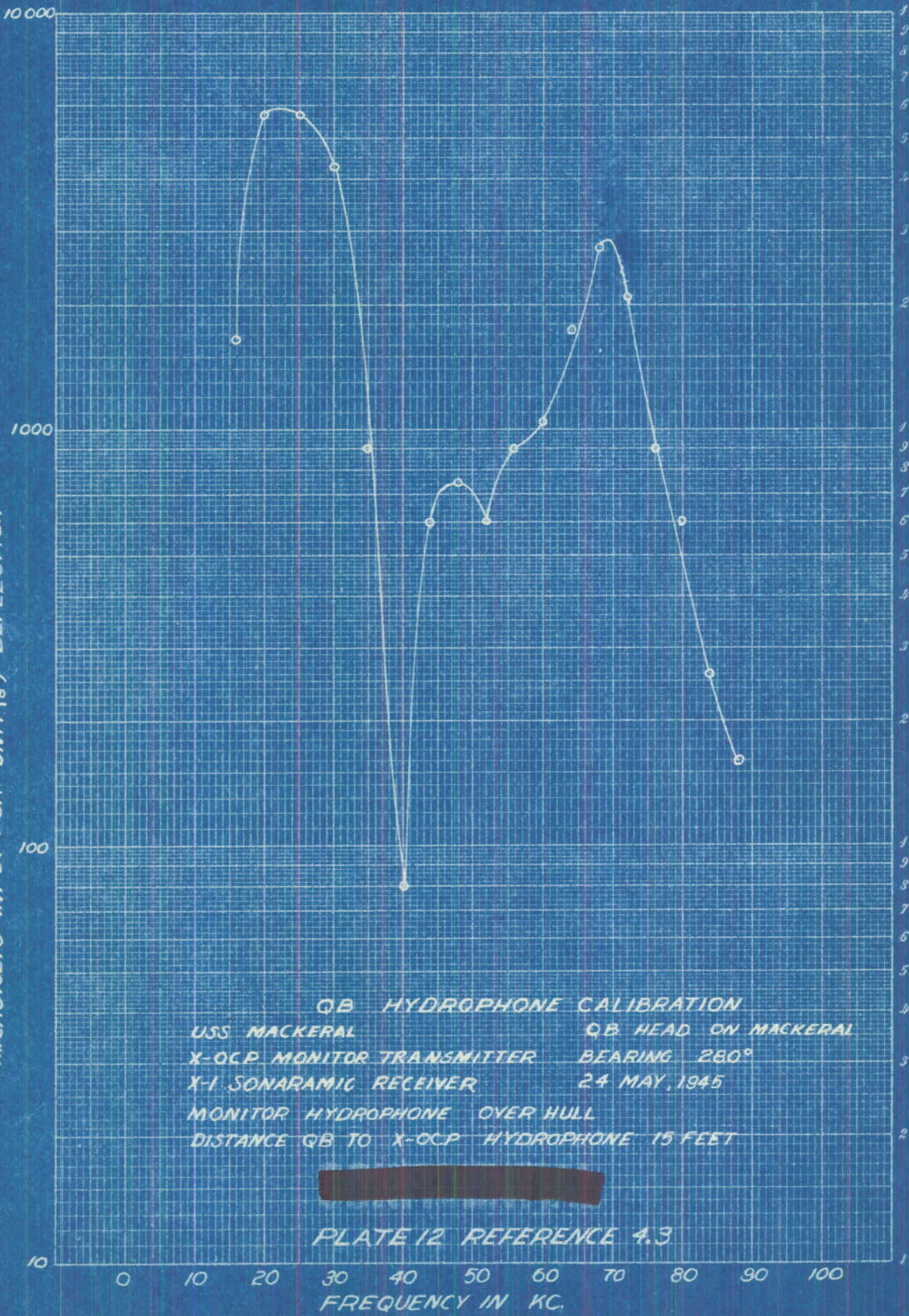


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12 MAY 15



MICROVOLTS INPUT FOR UNIT (2/8") DEFLECTION



QB HYDROPHONE CALIBRATION  
 USS MACKERAL QB HEAD ON MACKERAL  
 X-OCP MONITOR TRANSMITTER BEARING 280°  
 X-1 SONARAMIC RECEIVER 24 MAY, 1945  
 MONITOR HYDROPHONE OVER HULL  
 DISTANCE QB TO X-OCP HYDROPHONE 15 FEET



PLATE 12 REFERENCE 4.3

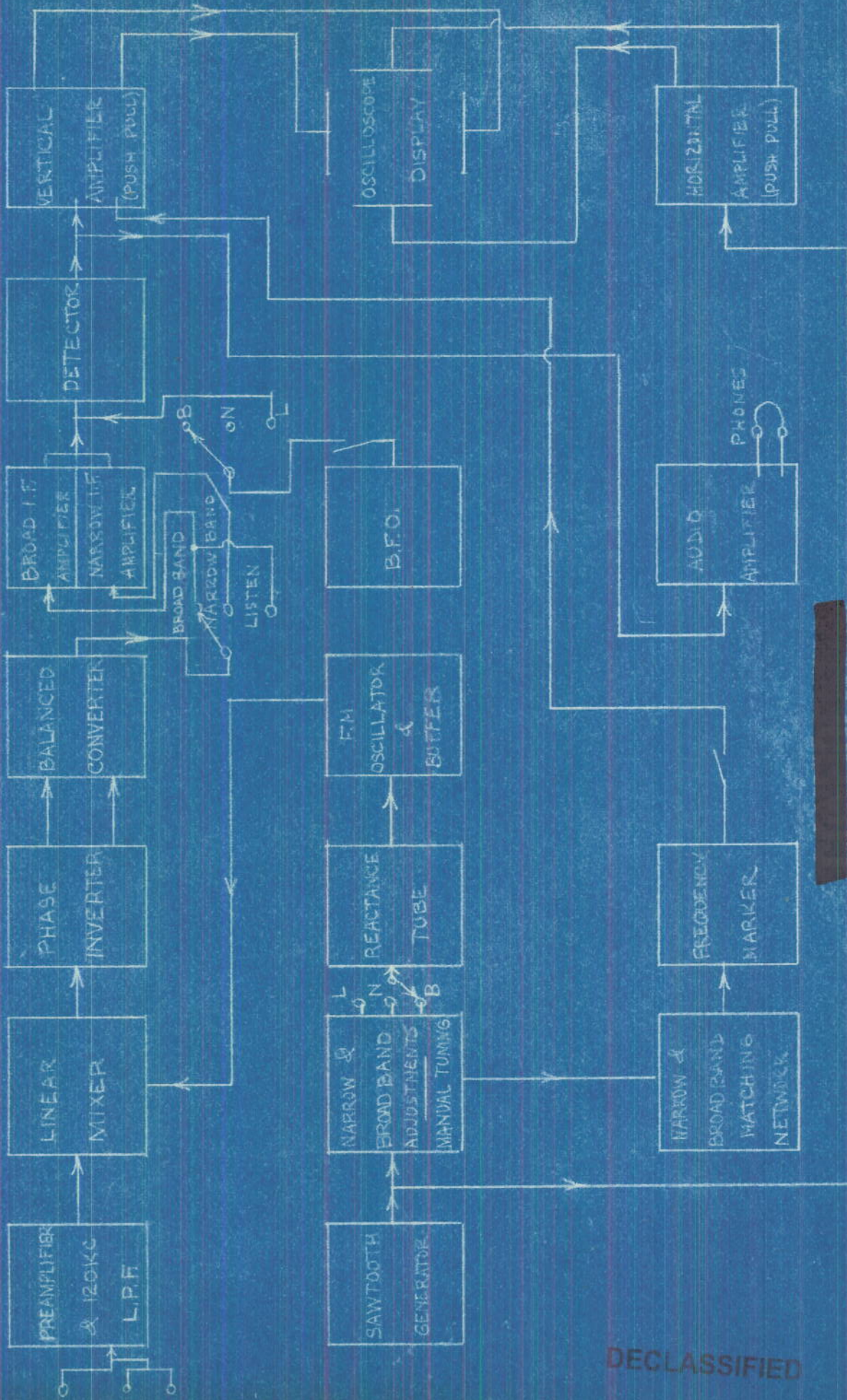


PLATE 13

BLOCK DIAGRAM OF

SCALE

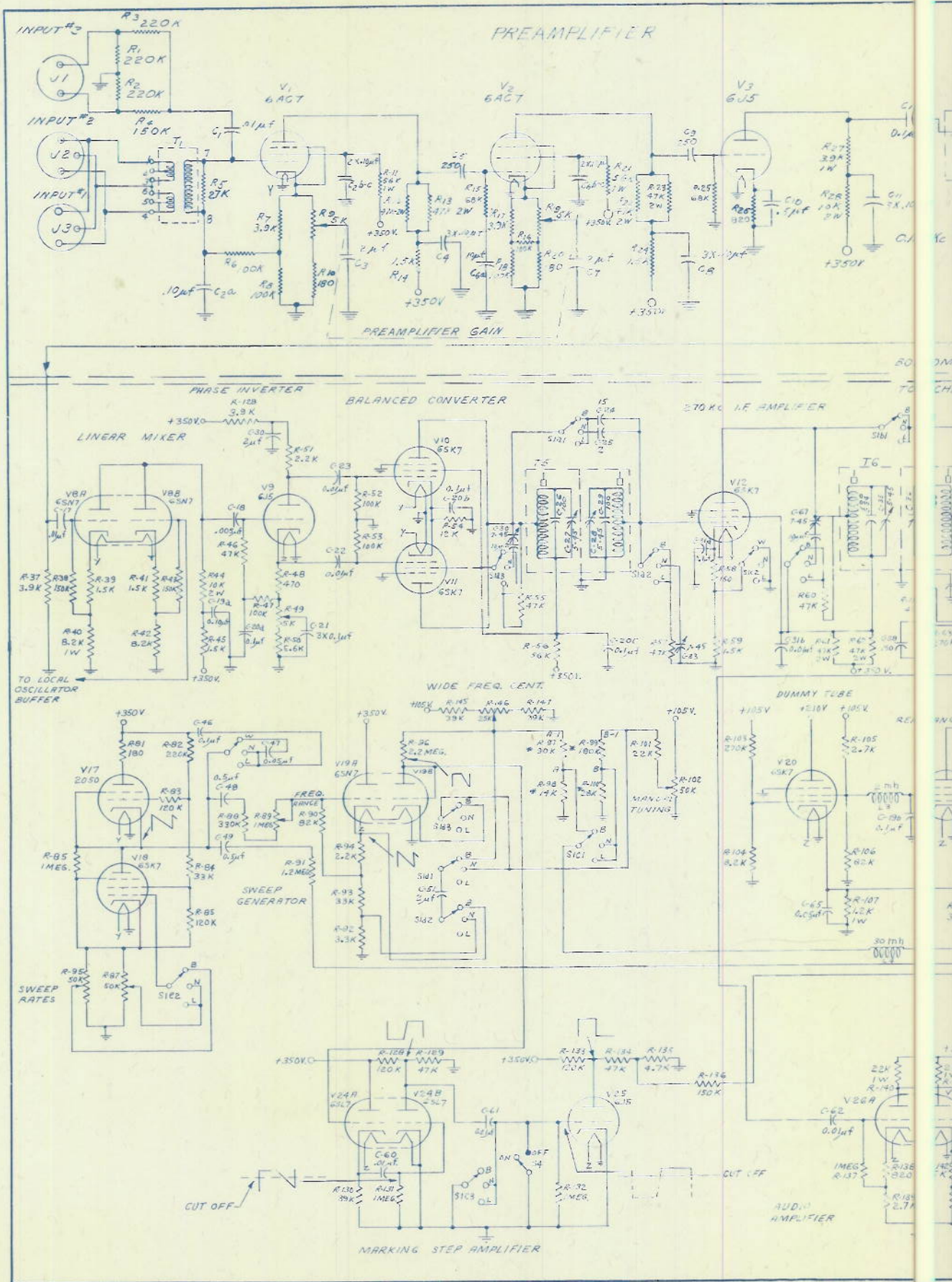
U. S. NAVAL RESEARCH LABORATORY WASHINGTON 20, D. C.

UNLESS OTHERWISE SPECIFIED, TOLERANCES ARE ±

APPR'D.

DATE 4/28/45

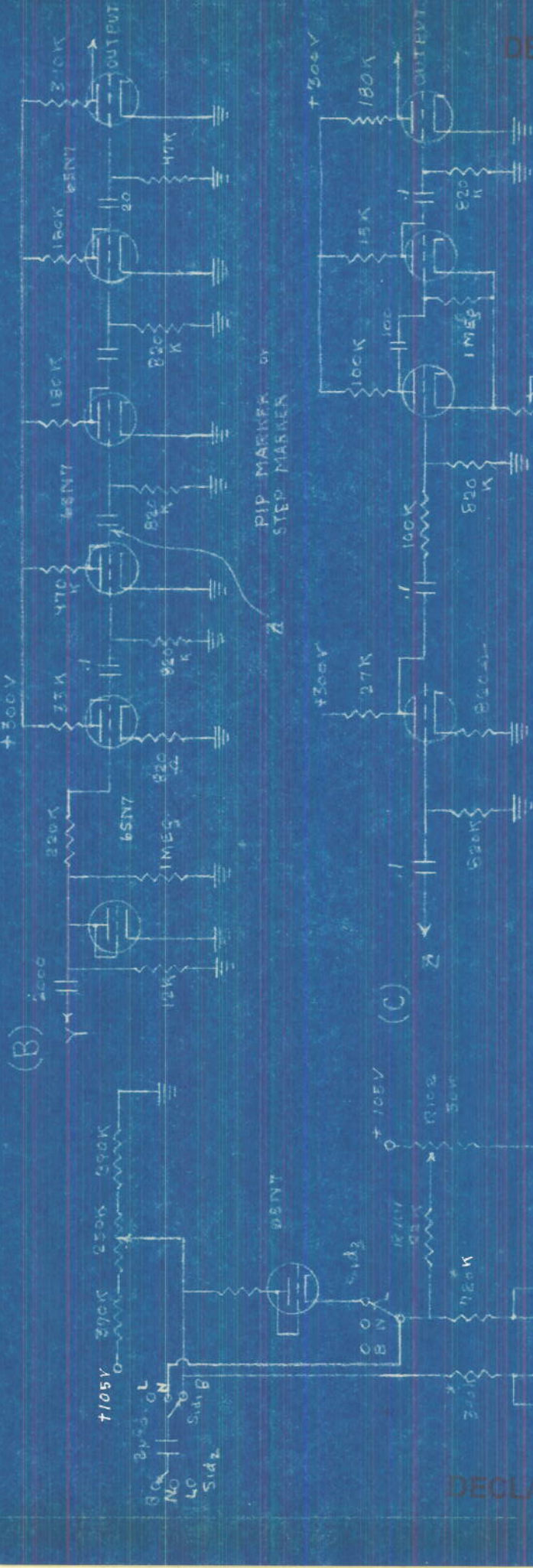
ROOM 308



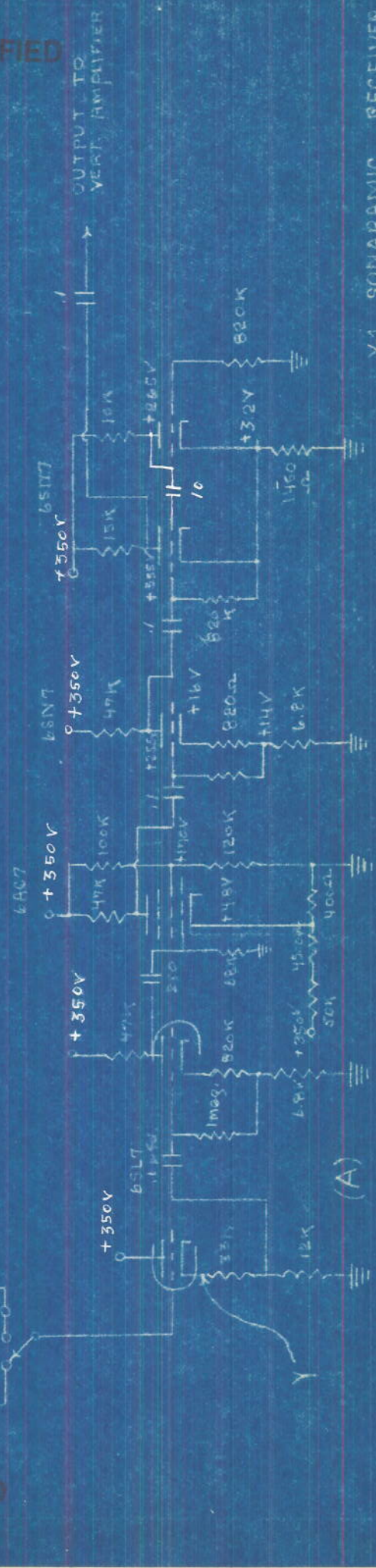
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DECLASSIFIED



NOTCH MARKER



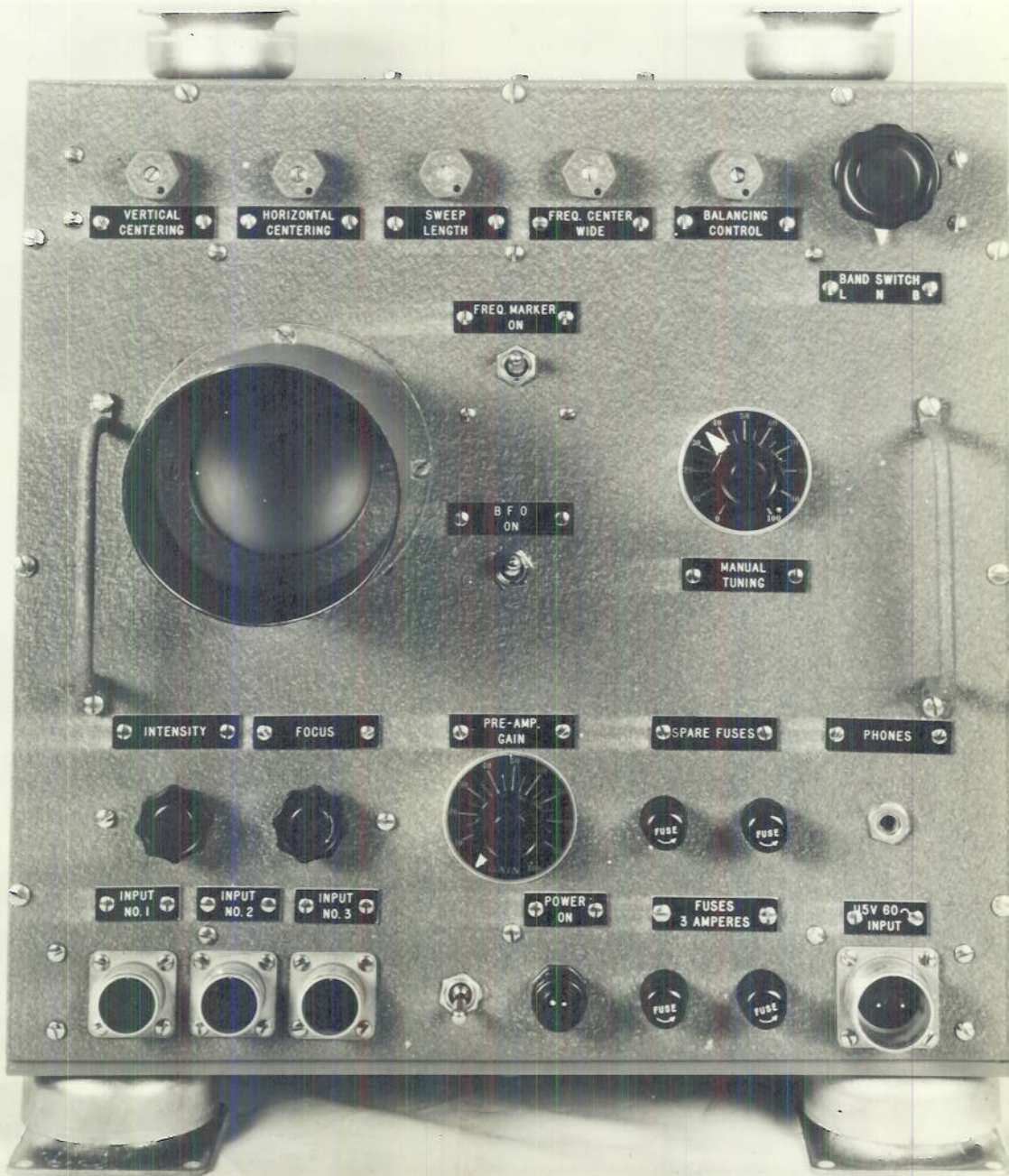
X-1 SONAROMATIC RECEIVER  
REVISED FREQUENCY  
MARKER CIRCUITS

STEP MARKER

PLATE 14 857 0-2

DECLASSIFIED

DECLASSIFIED



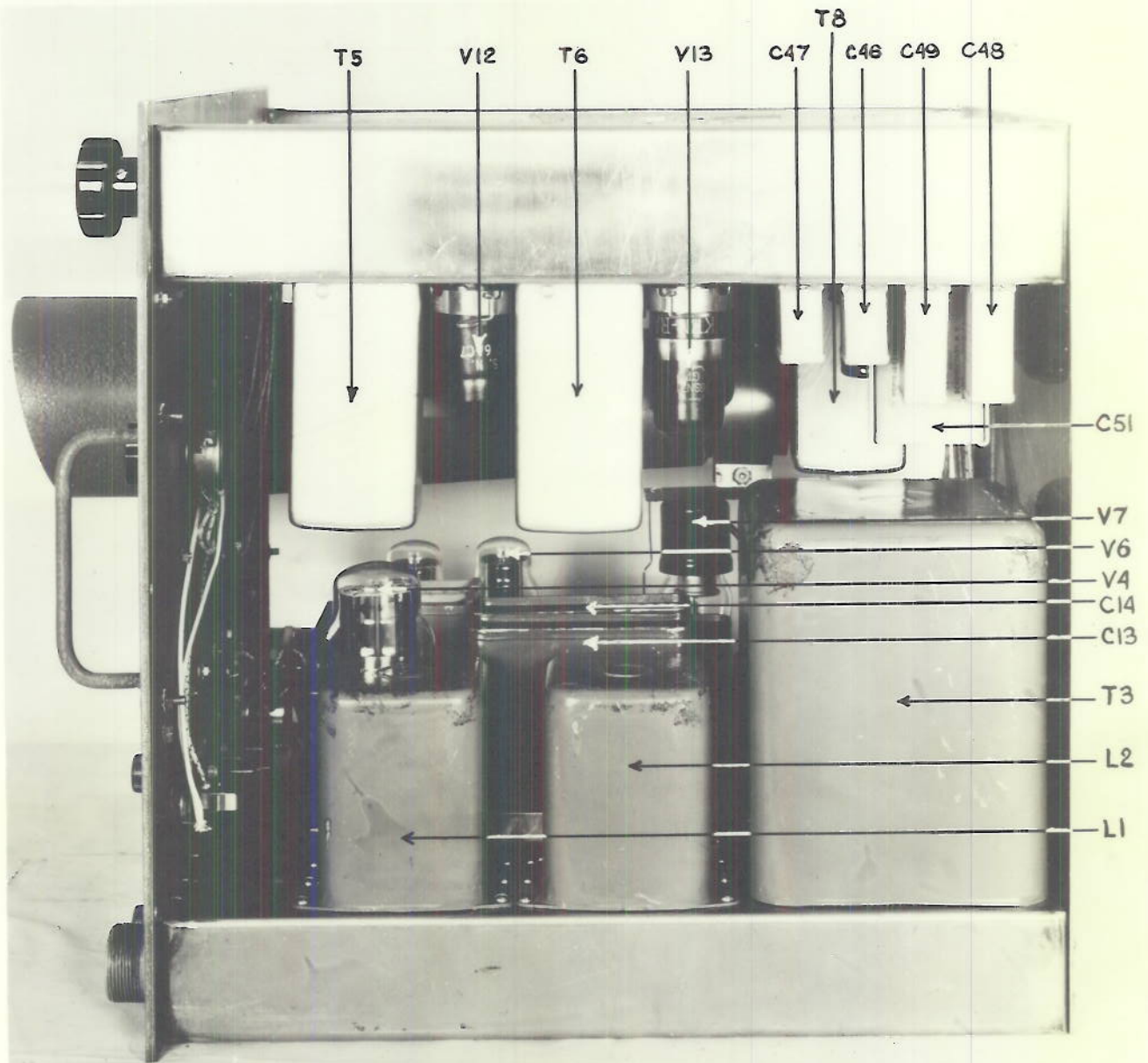
FRONT PANEL  
SONARAMIC X-1 RECEIVER

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PLATE 16

DECLASSIFIED



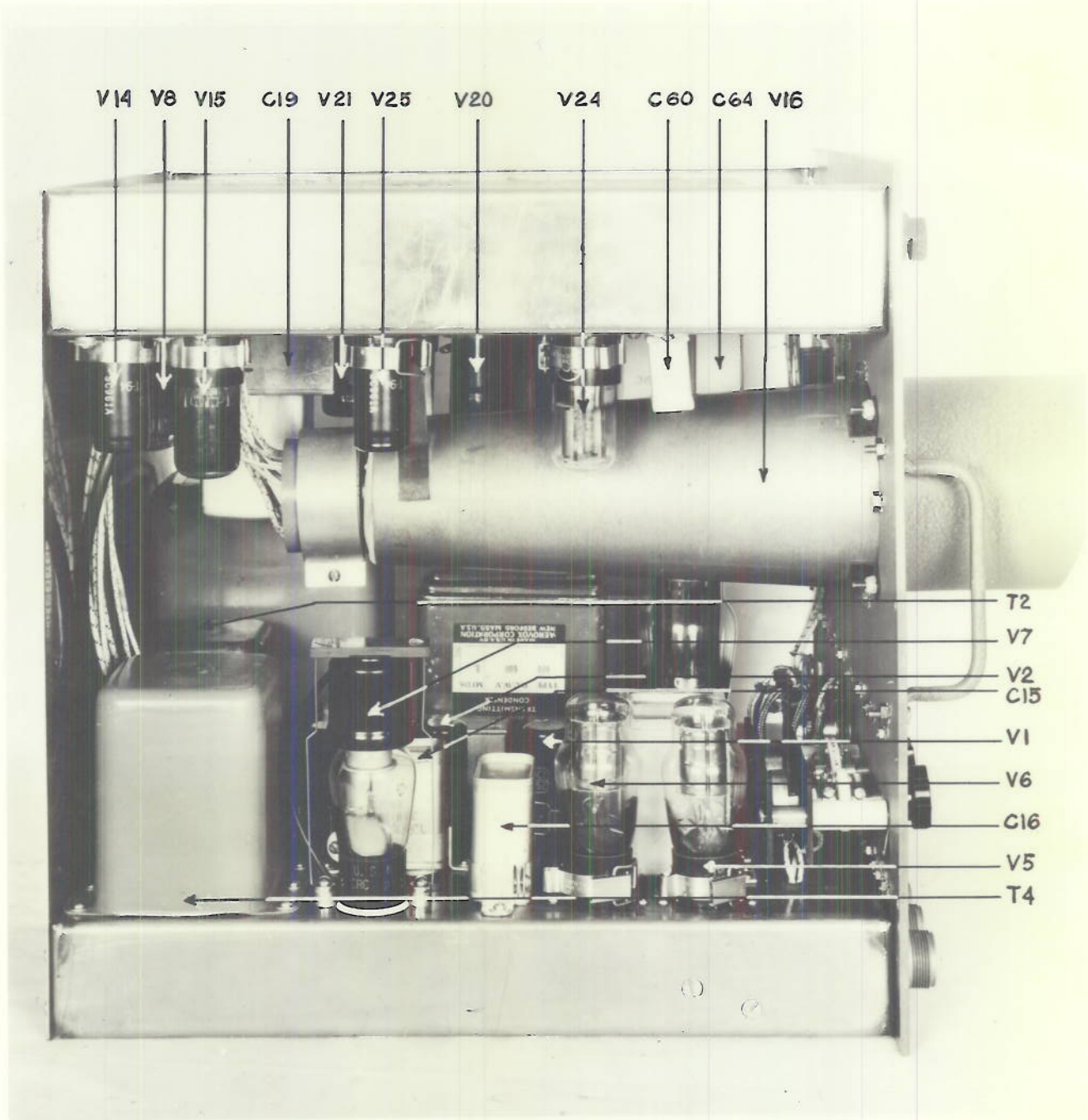
RIGHT SIDE VIEW  
SONARAMIC X-1 RECEIVER

**CONFIDENTIAL**

DECLASSIFIED

PLATE 17

DECLASSIFIED

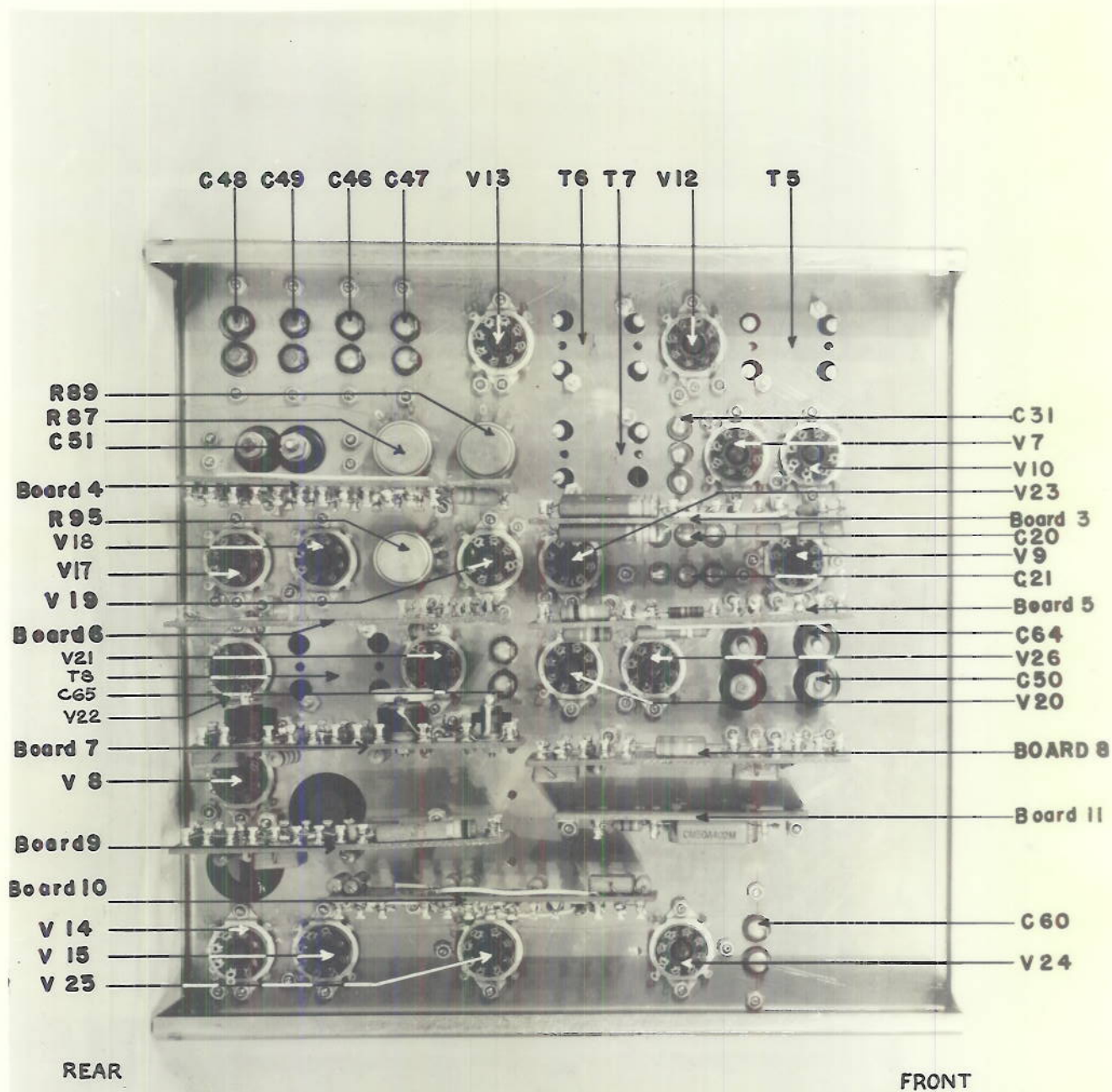


LEFT SIDE VIEW  
SONARAMIC X-1 RECEIVER

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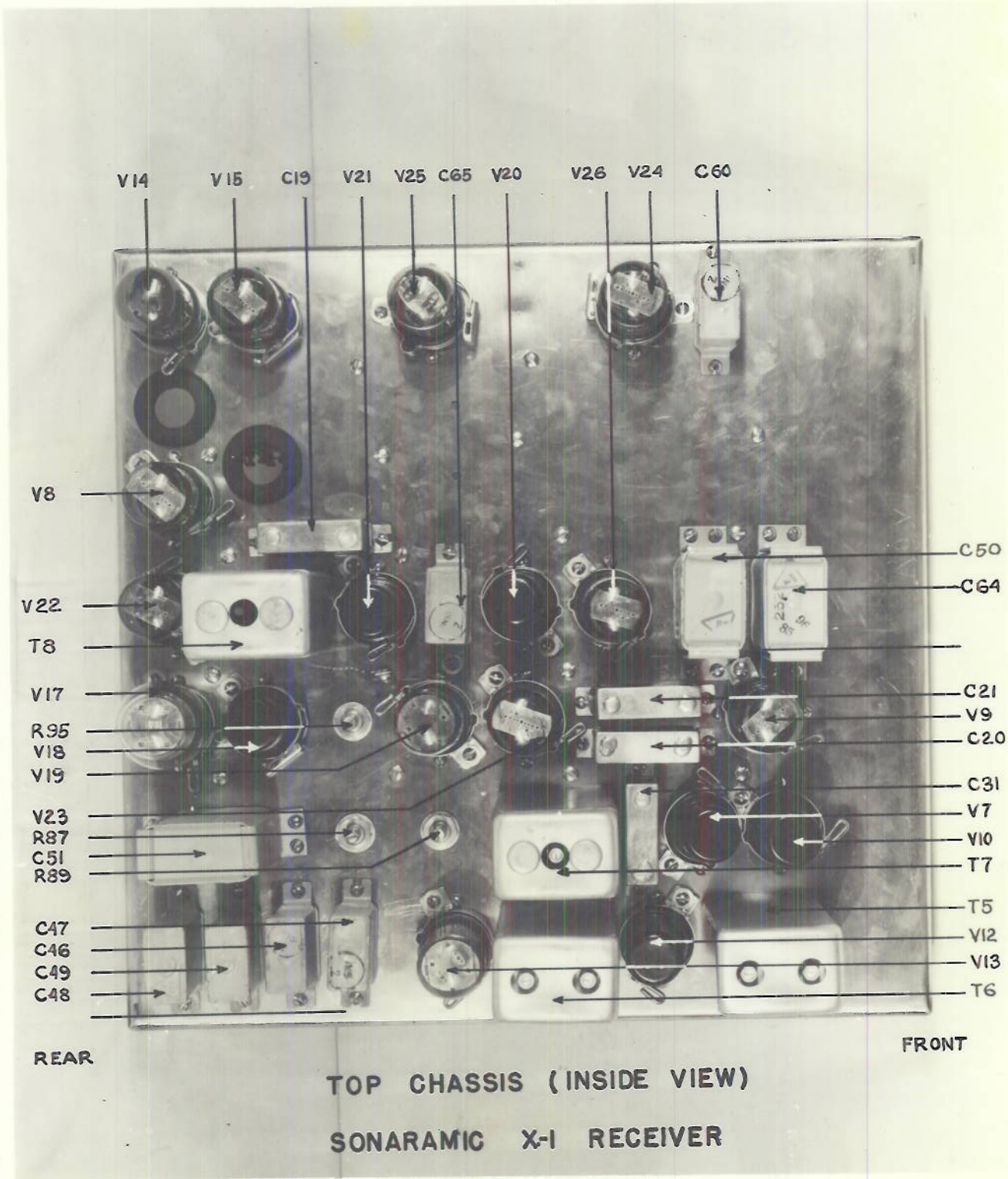
TOP CHASSIS (OUTSIDE VIEW)

SONARAMIC X-1 RECEIVER

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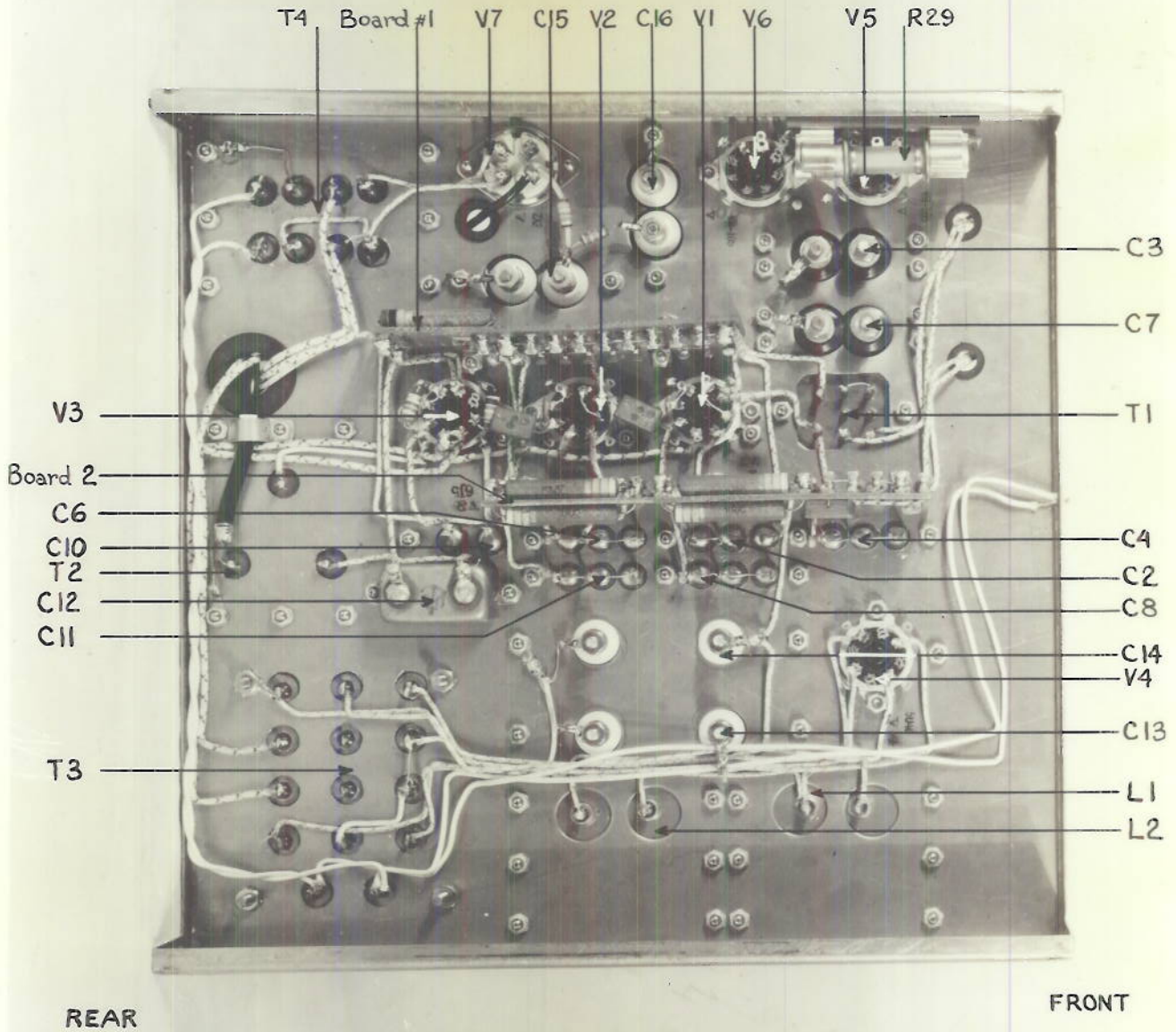
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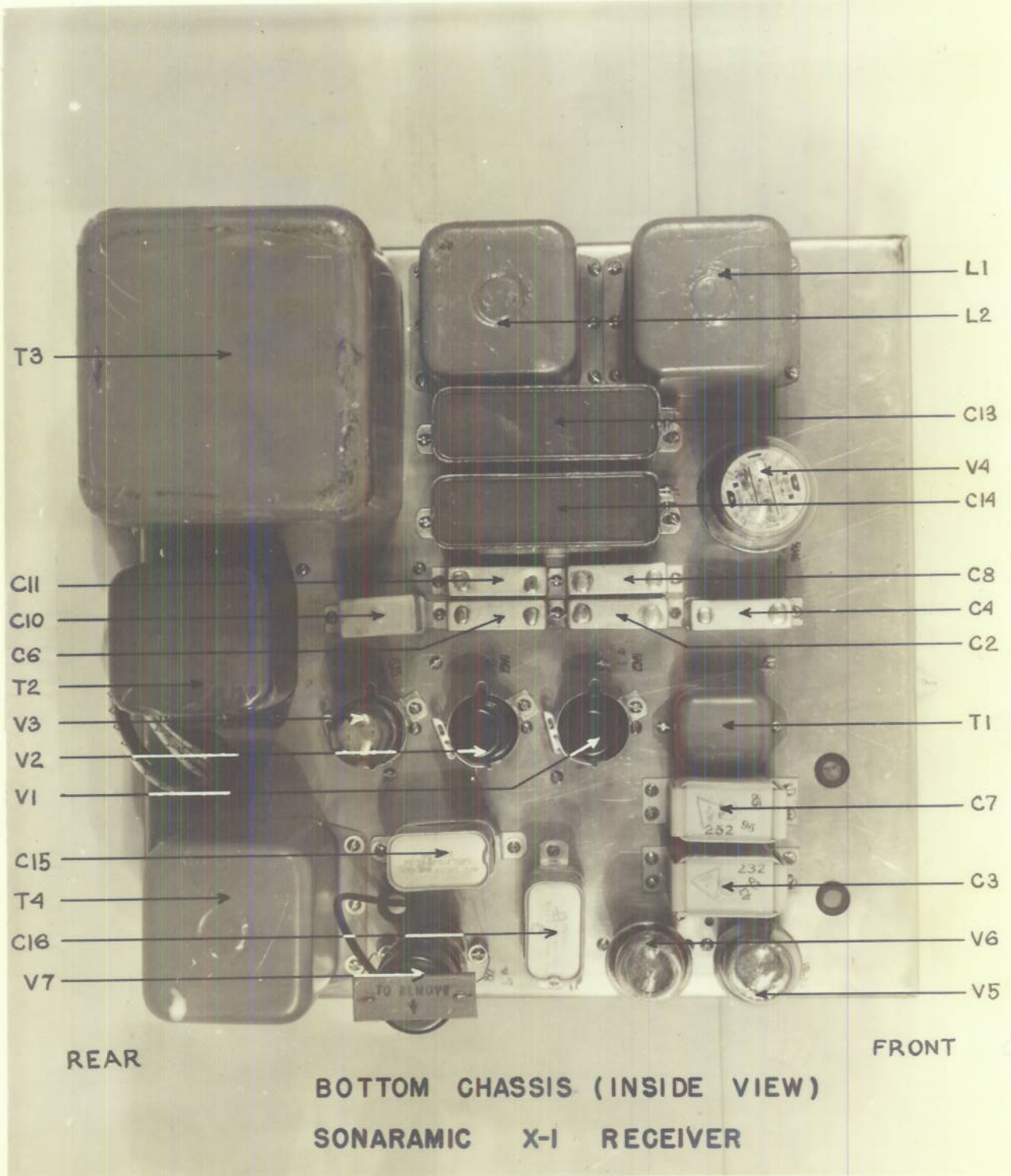


BOTTOM CHASSIS (OUTSIDE VIEW)

SONARAMIC X-1 RECEIVER

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