Armed Services Technical Information Agency

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U N R E S T R I C T E D

INTERIM DEVELOPMENT REPORT FOR AN/URM-41
RADIO INTERFERENCE-FIELD INTENSITY
MEASURING EQUIPMENT

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Contract NObsr-63057 31 December 1953

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INTERIM DEVELOPMENT REPORT #4

FOR

AN/URM-41

RADIO INTERFERENCE - FIELD INTENSITY
MEASURING EQUIPMENT

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This Report Covers the Period 22 September 1953 through 31 December 1953

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STODDART AIRCRAFT RADIO COMPANY, INC.

6644 SANTA MONICA BOULEVARD - HOLLYWOOD 38, CALIFORNIA

NAVY DEPARTMENT - BUREAU OF SHIPS - ELECTRONICS DIVISION

Contract Number NObsr-63057 31 December 1953

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ABSTRACT

1.1 This report covers developmental work accomplished during the period 22 September 1953 through 31 December 1953 on Navy Contract NObsr-63057.

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PURPOSE

2.1 Radio Test Set AN/URM-41 is to be a Very-Low-Frequency Radio Interference and Field Intensity Measuring equipment covering the frequency range of 30 to 15,000 cycles-per-second. The equipment to be built on this contract, NObsr-63057, is to be an alternate approach to the problem of making field intensity and noise measurements at these very low frequencies.

2.2 The equipment is to be transportable, suitable for use aboard ship, ashore, and in the laboratory.

2.3 The specification applicable to the AN/URM-41, SHIPS-R-734, recommends that the test set consists of the following units:

(a) A selective amplifier for harmonic analysis.
(b) A broad-band audio amplifier for pulse analysis and observation.
(c) A vacuum tube voltmeter at the output of (a) and (b).
(d) Suitable circuits for calibrating the test set.
(e) A suitable power supply.

2.4 Pick-up devices and coupling units as well as all other devices necessary for efficient and dependable operation of the equipment are to be included.

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GENERAL FACTUAL DATA

3.1 ENGINEERING PERSONNEL ASSIGNED TO THIS WORK.

1.	Al Parker	Chief Engineer
2.	Donald Radmacher	Assistant Chief Engineer
3.	Gerald Essex	Project Engineer
4.	Vernon Moore	Project Engineer
5.	Hollice Favors	Assistant Project Engineer
6.	William Glasspool	Assistant Project Engineer
7.	Carlyle Moore	Assistant Project Engineer
8.	William Humphries	Test Engineer
9.	Richard Williams	Test Engineer
10.	Richard Van Wickle	Technical Writer

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DETAIL FACTUAL DATA

4.1 The work summarized by this report includes development of an improved modulator, low-pass filter, preamplifier and oscillator. Temperature and voltage supply stabilization of the selective amplifier and local oscillator has been achieved. Further work on the broad-band amplifier, detector and metering circuits with particular emphasis on switching problems was carried out but not completed. Balancing problems associated with the electric field probe have been investigated.

4.2 The following are included with this report:

Figure 1. Preamplifier, Modulator, and Low-Pass Filter, Schematic Diagram.

4.3 BALANCED MODULATOR:

Previous work on the balanced modulator culminating in the schematic diagram, figure 5 of the previous report, indicated that the circuit had serious limitations for our purpose. Among these limitations are:

a. Poor and variable balance of local oscillator residual.

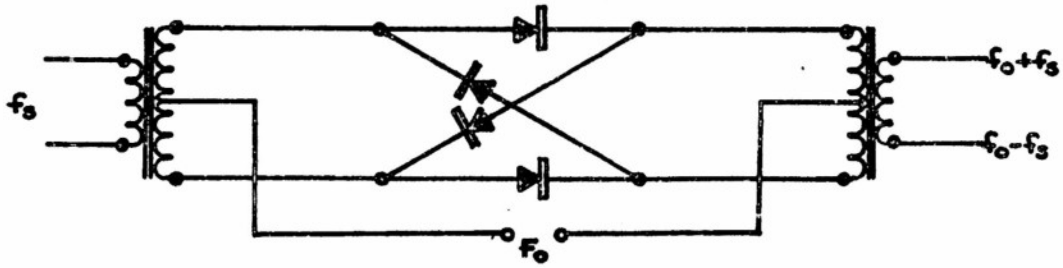
b. Extreme sensitivity to supply voltage variations.

c. High 2nd and 3rd harmonic responses.

d. Undesired responses, although generally better than 60 db down, were large in number and could easily be confused for signal components due to their not being in easily identifiable harmonic relationships to signal or oscillator frequencies.

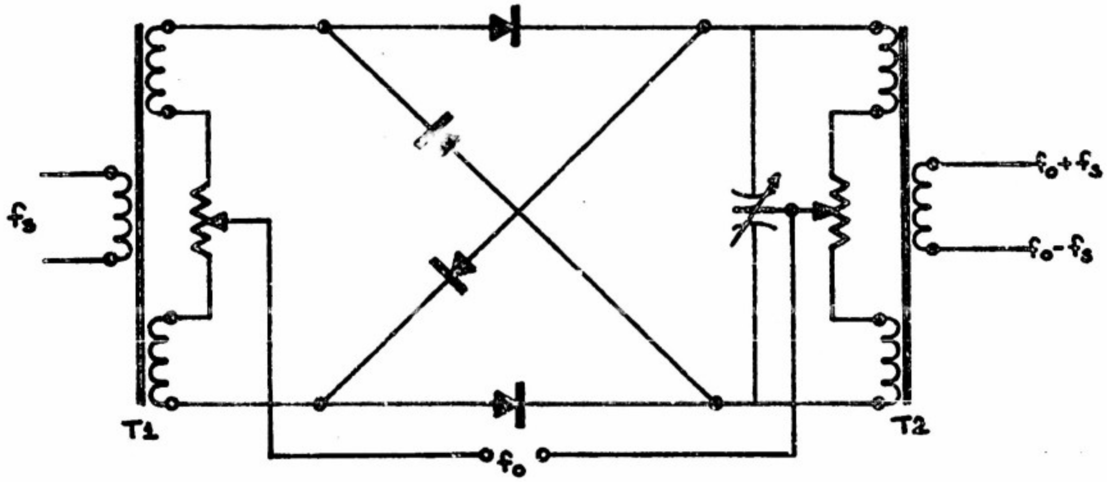
4.4 RING OR LATTICE MODULATOR:

In view of the balanced modulator limitations, it was decided to make an alternate approach. Other modulators balancing out the local oscillator include the series ring, shunt bridge, double balanced, and ring modulators. Of these, the ring type seemed most suitable because the output consists only of sidebands with no oscillator or signal components when correctly balanced and the residual modulation products are at a minimum.



The optimum operating conditions, methods of balance and requirements of such a modulator were investigated in the absence of references of a useful nature. Manual balance controls were added to the circuit and germanium crystal diodes were used for compactness and stability. These crystals were obtained in suitable matched forms from both General Electric and Sylvania.

The addition of balance controls to the ring modulator is believed to be original and took the following form. Other variations of circuitry are possible.



The transformers used were of the hybrid type used in carrier telephony and are obtainable commercially from more than one manufacturer. After setting the circuit up with a suitable preamplifier and low-pass filter it gave better results than the balanced modulator in most respects. Probably the circuit's most advantageous characteristic is the stability it experiences with respect to variations in supply voltages, temperature and oscillator balance. The most important requirement of the circuit is stable driving impedances. Frequency response and balance stability are dependent to a great extent on the source impedance of f_s, while capacitive balance is dependent to a large extent on the source impedance of f_o. Harmonic responses of the less obviously harmonically related types are of negligible amplitude. The most troublesome are the two types $f_o = 2 f_s$ and $f_o = 3 f_s$, in which case the amplitudes are of the same order as with the balanced modulator except that the $f_o = 2 f_s$ response is slightly greater than $f_o = 3 f_s$, which is reverse of that with the balanced modulator.

Noise level of the germanium diodes is higher than that of the 6SJ7's used in the balanced modulator but is of no nuisance value at the levels employed.

4.5 COMPARISON OF MODULATORS:

	Balanced Modulator	Ring Modulator
Frequency Response	Flat	Flat
Oscillator balance	Unusable between 30 cycles and 100 cycles except with limitations in harmonic response and dynamic range.	Usable at all sensitivities down to 25 cycles.
$fo=2fs$	66 db down (minimum)	58 db (minimum)
$fo=3fs$	55 db " "	69 db "
fs where $fs=\frac{fo}{2}$	56 db " "	70 db "
fs where $fs=\frac{fo}{5}$	67 db " "	Negligible
Various other residual modulation products	Many in number varying from 59.0 db down and averaging 65 db down.	Better than 75 db down and few in number.
Sensitivity with preamp.	.3 microvolt down to 1 kilocycle. Minimum 2.0 microvolts at 30 cycles, 600 ohm input.	Better than .05 microvolt down to 1 kilocycle. Minimum .75 microvolt at 65 cycles, 600 ohm input.
Heat	3 extra relatively high current tubes compared with ring type.	Negligible.
Economy		Probably more expensive due to cost of transformers and germanium diode quads.

	Balanced Modulator	Ring Modulator
Shielding		More difficult to shield due to transformers, although no trouble was experienced with the transformers in use.
Bulk		Bulkier in its present form but with possibility of making smaller than the balanced modulator.
Supply requirements.	Very expensive and bulky A and B voltage regulation necessary	No regulation necessary
Oscillator requirement	Oscillator voltage must be constant with frequency. Low impedance supply.	Oscillator voltage can vary ± 50% with frequency with no change in operation.

4.6 SPURIOUS RESPONSES:

Considerable time was spent investigating the cause and cure of spurious responses including harmonics. The known causes of such responses were:

a. Preamplifier.

b. Ring modulator.

c. Transformer distortion.

d. Spurious signals from generators in use.

e. Power line harmonics.

Measurements indicated that the preamplifier if designed with a large overall value of negative feedback would be free from harmonics provided levels in use were sufficiently low to restrict working to the linear portions of the tube characteristics. Ring modulator distortions proved to be the limiting factor and, in particular, the 2nd harmonic was the worst. Various methods of balancing the modulator for minimum

2nd harmonic proved fruitless and the only parameter affecting this harmonic substantially was oscillator voltage. This was adjusted for optimum signal-to-harmonic ratio. Transformer distortion was not troublesome except insofar as it contributes to ring modulator spurious responses. It was found that the Hewlett-Packard Signal Generator HP206A was adding 6 db 3rd harmonic distortion to any measurements made at approximately 60 db down from the input level. The 2nd harmonic content of the generator output was negligible.

Frequency response of the ring modulator was not found to be flat until the source impedance of the signal was adjusted for optimum. The source impedance of the oscillator supply was also found to be critical. The impedance of T2 balance potentiometer to ground has to be high in order to keep the range of the capacity balance within practical values.

A special transformer was designed to replace the buffer tube V4 but changes in the resistive balance caused frequency pulling due to loading effects on the oscillator cathode. This was undesirable so the original setup of buffer and transformer as in figure 1 was retained.

The effect of temperature on the ring modulator and, in particular, on the germanium diodes was investigated and it was found to be insensitive to ambient temperature changes through a large range. For a 55° C change in temperature, a change of only 3/4 db in gain was recorded. Balance controls required only a small readjustment to balance at thirty cycles.

B+ and A supply variations leave the ring modulator unaffected as to balance, gain, and harmonic responses.

Since better results were obtainable with the ring modulator it was decided to adopt this in preference to the simple balanced type used previously. The preamplifier and filter designs were therefore modified to fit the ring modulator requirements.

4.7 LOW PASS FILTER:

It is desirable to have as flat a frequency response as possible in the type of noise meter being designed and previous work had indicated that the low-pass filter would be the limiting factor. In order to obtain a flat response the low-pass filter should be flat in amplitude with frequency from 30 cps to 15 kcs ± 1/2 db from the calibrating source. With the IF at 20 kcs and requiring the response to signal input at 20 kcs to be 60 db down or better, the filter design became formidable. In all practical filters built to this

specification the cut off started slightly before 15 kcs resulting in a deviation from flat response. Compensation for this proved difficult. It was therefore decided to shift the I-F frequency to 25 kcs.

A mathematical investigation around these figures was made and an m-derived design arrived at using a three-section pi type filter. Practical values dictated an image impedance of 1,000 ohms. A filter was then constructed and proved satisfactory for our purpose; frequency response measured less than 1/2 db change throughout the pass band. Stray wiring capacitance and noise pick-up caused less trouble than in previous designs due to the lower image impedance involved.

4.8 Preamplifier requirements were altered by the new filter and redesign work was carried out. Several type preamplifiers were fabricated using power tubes and various amounts of feedback. In order to make gain and harmonic responses independent of supply changes and temperature over a wide limit, negative feedback had to be employed to a large extent. This made a two tube preamplifier a necessity. Considerable investigation has been made into the choice of tubes for this purpose and the RCA 5879 proved to be the best compromise, low noise and microphonics being the main consideration. With an overall feedback of 50 db and a total gain of 18 db, the frequency response was essentially flat over the required range. A low impedance output of 1000 ohms is required in order to feed the low pass filter and any variation from this value causes mismatch and consequential poor frequency response within the pass band of the filter. A cathode follower was designed to provide this 1000 ohm match and to provide negligible distortion. A pentode type cathode follower was employed to make input and output voltage as nearly similar as possible. The output impedance of the cathode follower was calculated to be 234 ohms and was confirmed by measurement. A series resistor then converted the output impedance to 1000 ohms.

Some residual 60 cycle voltage was found in the preamplifier despite dc-supplied heaters and elaborately filtered B+. This was traced to the 5879 front end tube being sensitive to stray magnetic fields. A simple iron shield eliminated this while a non-magnetic alloy type was ineffective. The sixty cycle heater supply to the preamplifier produced 60, 120, 180 and 240 cycle responses; these could only be removed by a dc heater supply. Even careful heater balance and a low harmonic regulated supply transformer proved useless. With dc heater supply and magnetic shielding, 60-cycle pickup could be reduced below noise level with a 600-ohm input and measured only 1.5 microvolts at 100,000 ohm input impedance.

In narrow-band operation, the noise level with a 600-ohm assymetrical input was less than 0.1 microvolt over 90% of the tuning

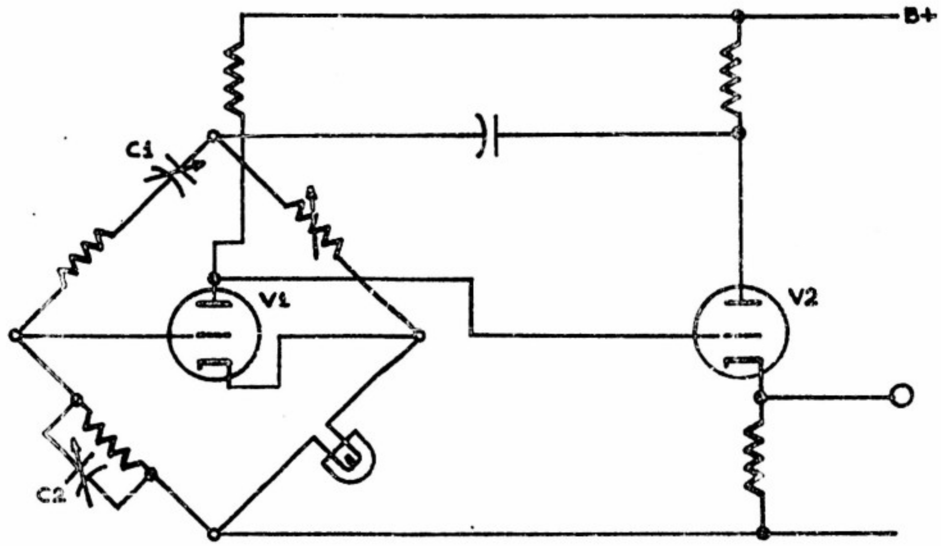
range and amounted to less than 0.05 microvolt at the higher frequencies. Below 1000 cps noise gradually increased to 0.7 microvolt at 30 cps. With 100,000-ohm assymetrical input the figures were only slightly worse, ranging from approximately 0.1 microvolt at the highest down to 0.8 microvolt at the lowest frequencies. The gradual increase in noise below 1000 cps is concluded to be due to "flicker effect". Little information is supplied by tube manufacturers regarding this effect. It is not proposed to investigate "flicker effect" at this time unless it proves desirable to obtain higher sensitivities at very low frequencies.

4.9 LOCAL OSCILLATOR:

Local oscillator design has been explored to some extent. Our original standard of comparison, the Hewlett-Packard Wave Analyzer, drifts in frequency considerably during operation and requires constant readjustment for the first hour. The temperature range of the URM-41 must be much wider than the above instruments. When the URM-41 receiver is tuned to the low end, thirty cycles, the local oscillator is operating at 25,030 cps and should be held to within a few cycles of its operating frequency. This would be easy for fixed frequency operation but the oscillator must be variable. Fortunately a small amount of zero setting is permissible, provided the dial calibration remains valid. This would mean using the I-F amplifier as a frequency standard for zero setting. This is within practical limitations as work on the stabilization of the IF indicates good frequency stability.

The primary oscillator requirements are resetability and freedom from drift with temperature and supply voltages. Output voltage stability is no longer a prime consideration although some work along these lines has been carried out for use with the alternative balanced modulator.

The oscillator dial calibration was studied and, for convenience of operation, an arbitrary scale was devised which provided dial expansion for frequencies below 1000 cps approximating a square law. Mechanical variable ratio drives to the tuning capacity were studied but eventually discarded in favor of especially shaping the stator plates and using a high quality fixed ratio drive. As a Wien bridge oscillator has been decided upon as being most suitable for our purpose, a two gang capacitor is required. In order to eliminate end and side thrust, a standard straight line capacitor was modified by installing ball bearing assemblies in place of simple sleeve bearings. The stator plates were shaped as necessary. Unfortunately a standard mid-line capacitor provides for minimum capacity change at minimum capacity while we require minimum capacity change at maximum capacity.



A basic circuit of the oscillator used is given above. From inspection it can be seen that two feedback paths exist from V2 plate to V1, one via the grid is positive and the other via the cathode is negative feedback. For optimum stability with changes in supply voltages and amplifier characteristics, as much negative feedback as possible for a given output should be employed. This means the total amplifier gain before feedback should be as high as possible, consequently high gain pentodes were used under optimum conditions for V1 and V2. The screen to cathode voltage was stabilized with bypass capacitors in each case to increase gain. Great care had to be employed with these capacitors to avoid low frequency phase shift and consequent spurious oscillation. An oscillator resulted which is virtually independent of changes in non-bridge circuit elements and tubes. Also the change in frequency with B+ changes was approximately 1 part in 100,000 per volt change. Changes due to heater voltage variations were negligible.

The choice of bridge components for the oscillator depends primarily on temperature coefficients and is still under investigation. Thus far the best results have been obtained with wire-wound resistors and special silver-mica capacitors having low-temperature coefficients.

The electric field probe has been a subject of further study since the last report. A linear field was set up and balance of the probe checked. It was evident that some form of input capacity balance would be necessary. This was done and the probe could then be used to plot lines of electric fields and for "homing" on interference, etc. Work was commenced on the elimination of the probe transformer, incorporation of a reversing switch and general mechanical design of the probe.

4.10 SELECTIVE AMPLIFIER:

As mentioned in connection with the low-pass filter, it was found desirable to raise the intermediate frequency to 25 kcs. This relatively small increase simplified the design of the low-pass filter considerably with a degree of sacrifice in I-F selectivity (or inherent stability) for a given set of components. However, the I-F amplifier required improvement in design for other reasons, chiefly that of excessive bulk.

With an amplifier of the type we are using, the design, in general, involves a compromise between gain and selectivity if a given degree of stability is to be maintained. Stability is, of course, greatest when the maximum amount of negative feedback is employed. With high feedback, the selectivity and gain are controlled almost entirely by the tuned circuit. If this circuit has relatively high Q, the performance of the amplifier will be relatively unaffected by normal changes in plate and heater voltages and small variations in tube characteristics. Stabilization with respect to variations in temperature is, of course, still controlled by the temperature stability of the various detail components. The coil and capacitor of the tuned circuit, as well as all the resistors in the positive and negative feedback paths, must have the best possible temperature characteristics.

All preliminary work with the selective amplifier has been done with toroidal coils having the highest possible Q. These coils are approximately 2½ inches in diameter and weigh approximately 9 ounces. With an inductance of approximately 4 millihenries, a capacity of 0.016 mfd is required to tune to 20 kcs. The large size of the coil and capacitor discourages attempts at effective miniaturization. In addition, with such a large tuning capacitance, there is quite a problem in providing the actual adjustment for resonance. For reasons of stability, only an air trimmer or good ceramic trimmer can be employed. Assuming the use of a 100 mmfd air trimmer (which is relatively large physically), the combined variation of the fixed capacity and the inductance would have to be less than 0.5%. Obviously, the bulk of tuning adjustment consists of selection of several capacitors which would combine to bring the circuit within range of the 100 mmfd trimmer.

After due consideration, it appears that too great a premium is being paid to obtain a coil with a Q as high as 360. It is relatively easy to obtain a Q of 250 with a coil having a diameter of 1¼ inches and a weight of less than two ounces. The reduction in bulk is over six-to-one. In addition, it will probably be advantageous to go to higher inductance in order to obtain greater initial gain and permit a greater proportion of negative feedback to be used.

Work done thus far indicates that as miniaturization of the design progresses, the problem of isolation of the individual

stages becomes more acute. In fact, it appears definite that these stages should be made up as individually-shielded assemblies. It is better too that the negative feedback and bandwidth control potentiometers be located within their respective shielded stages rather than as a part of a gang potentiometer assembly to which shielded wiring must be run from each stage. Such an arrangement will, of course, require rack and pinion or bevel gearing to operate all controls in unison.

Extensive investigation into the problem of stabilization of the selective amplifier has been made on a single stage. Stability of gain and frequency is desired for variations in ambient temperature, plate and heater voltages, and for different positions of the bandwidth control. Frequency stability was achieved by the use of stabilized molybdenum permalloy cores, the most stable mica capacitors available supplemented by a compensating ceramic capacitor. The desired stabilisation of gain vs. temperature was obtained by the use of the highest quality deposited carbon resistors in all except plate and screen supply circuits. Nearly perfect stabilization could be realized by the use of a thermistor; however, this required selection of some of the other resistors for their exact temperature characteristics, making each stage a custom design, which was considered impractical. Stability of frequency and gain for variation of the bandwidth control was improved by using a relatively high resistance bandwidth potentiometer and arranging the layout within the stage to keep the output circuit isolated from the input tube grid and tuned circuit. Interstage feedback by way of coupling from input or output terminals of stages was, fortunately, no problem due to the low effective stage-to-stage gain and the attenuation afforded by the high input grid coupling resistors.

4.11 DETECTOR:

During this period, some thought has been given to the type of detector to be used and the manner in which the output meter scale is to be calibrated. Normal design procedure for a Radio Interference-Field Intensity meter second detector calls for a response time, in the field intensity function, fast enough to completely follow the highest frequency modulation permitted by the bandpass of the I-F amplifier. This response is still slow with respect to the I-F carrier and, in the absence of modulation, the second detector, in the field intensity function, is a peak-type detector. From this, it can readily be seen why the field intensity and quasi-peak circuits should indicate the same output on an unmodulated signal.

Tentatively, the output meter scale used for the selective amplifier portion of the URM-41 will be calibrated in terms of RMS value of a sine wave. Shaping by the narrow bandpass of the selective amplifier generally supplies a sine wave to the second detector regardless of the original waveform of the signal.

There are possibilities of low-frequency modulation appearing on the signal and of its riding through the selective amplifier and appering across the second detector load resistor. This situation would be much more likely to occur when using the wider bandpass portion of the selective amplifier. Ability to identify and observe such modulation would be desirable. Two second detector functions are therefore proposed:

a. A detector circuit with a response time just fast enough to follow the maximum modulation permitted by the selective amplifier bandpass. To the signal carrier, this detector circuit would be of the peak-reading type discussed above.

b. A detector circuit with a response time slow with respect to even the modulation that might be present. This detector circuit would measure the peak of the modulation. It would be equivalent to the usual Quasi-Peak circuit but would have a longer discharge time.

The two detector functions just described would provide the same indication on an unmodulated signal since both are peak indicating circuits at the intermediate frequency.

4.12 BROADBAND AMPLIFIER:

In the broadband amplifier portion of the URM-41, the second detector becomes a rectifier since there is no carrier frequency involved. To provide an accurate measurement of the input, the bandpass of the rectifier circuit (field intensity) should be equal to or greater than that of the preceding broadband amplifier. Thus, we have an averaging rectifier circuit.

The output meter scale for the broadband portion can be calibrated (1) in RMS microvolts with respect to a sine wave or (2) in terms of microvolts average. If the RMS (.707 x peak of sine wave) method of calibration is used, the reading will be true only if the waveform applied to the rectifier is a sine wave. If the average method of calibration is used (.636 x peak of sine wave), the average reading will be true of any waveform. Since all types of waveforms will be encountered in broadband measurements, the average calibration appears the most suitable.

The broadband rectifier would also contain functions of quasi-peak and peak. Behavior of these two functions would be similar to their counterparts in the AN/PRM-1A equipment since the bandwidths are of the same order.

Broadband measurements can encounter signals in which the positive and negative portions of the waveform are dissimilar. Simple half-wave rectification of these signals would not be the proper way

to obtain a reliable measurement. Inaccuracies incurred by this method are commonly called "turn-over" discrepancies. Full-wave rectification of the signal to obtain average, quasi-peak, and peak readings involves rather complicated amplifier and detector circuitry, but would give more reliable measurements. However, even full-wave rectification does not supply sufficient data to analyze a signal. Separate half-wave rectification of the positive and negative portions of an assymetrical signal would provide the desired data. This method retains the less complex circuits required of half-wave rectification. The one disadvantage is that two readings, positive and negative, would have to be taken and then added to provide peak-to-peak voltage readings in quasi-peak and peak functions, and true average voltage in the field intensity function. Positive and negative voltage data enables analysis of signal as to symmetry and type of waveform. The ratio of average to peak reading can be used to identify the waveform; for instance: if the ratio of peak to average reading is .636 and the + and - readings are equal, the signal is a sine wave.

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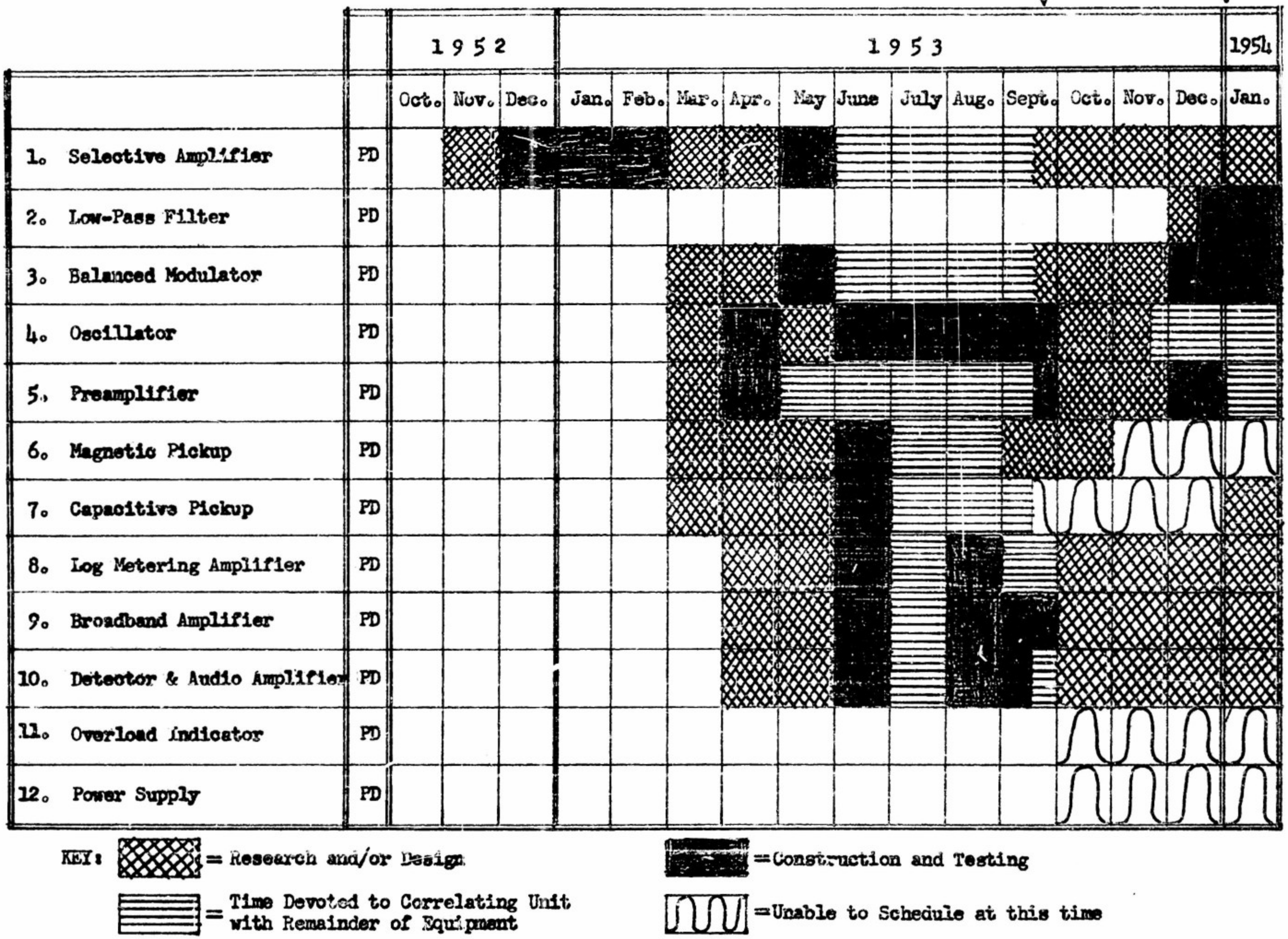
Contract NObsr-63057

PROJECT PERFORMANCE & SCHEDULE

Date: 31 Dec., 1953
Period Covered by Report: 9-22-53 - 12-31-53



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5.1 CONCLUSIONS:

Performance of the ring modulator has been highly satisfactory and no further time will be spent on the older balanced modulator. A satisfactory design for the new low-pass filter has been achieved. Improvement of the preamplifier and oscillator performances are almost adequate for use with the next model. Requirements for redesign of the selective amplifier are fairly well known as a result of the recent investigations. Continued investigation of the broadband amplifier and the logarithmic metering amplifier is required. More work will be required on the capacitive input probe.

6.1 PROGRAM FOR NEXT INTERVAL:

Effort will be directed toward completing all details of electrical design of the selective amplifier, ring modulator, oscillator, preamplifier, attenuator system, calibrating system, beat-frequency oscillator, overload indicator, etc., so that formal mechanical design for the next model can proceed. Electrical design of the wide band amplifier and metering circuitry will be pushed as rapidly as possible.

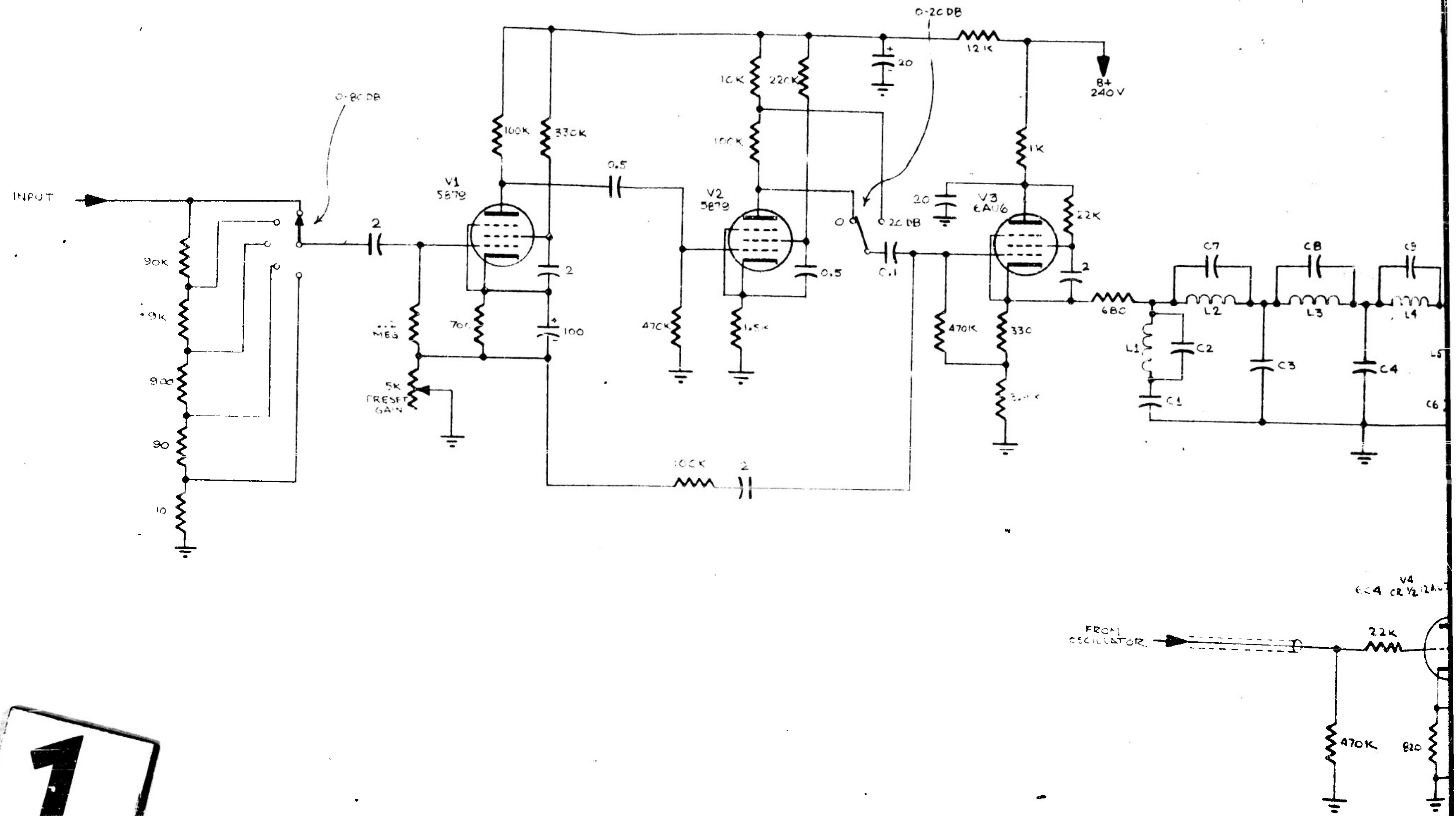

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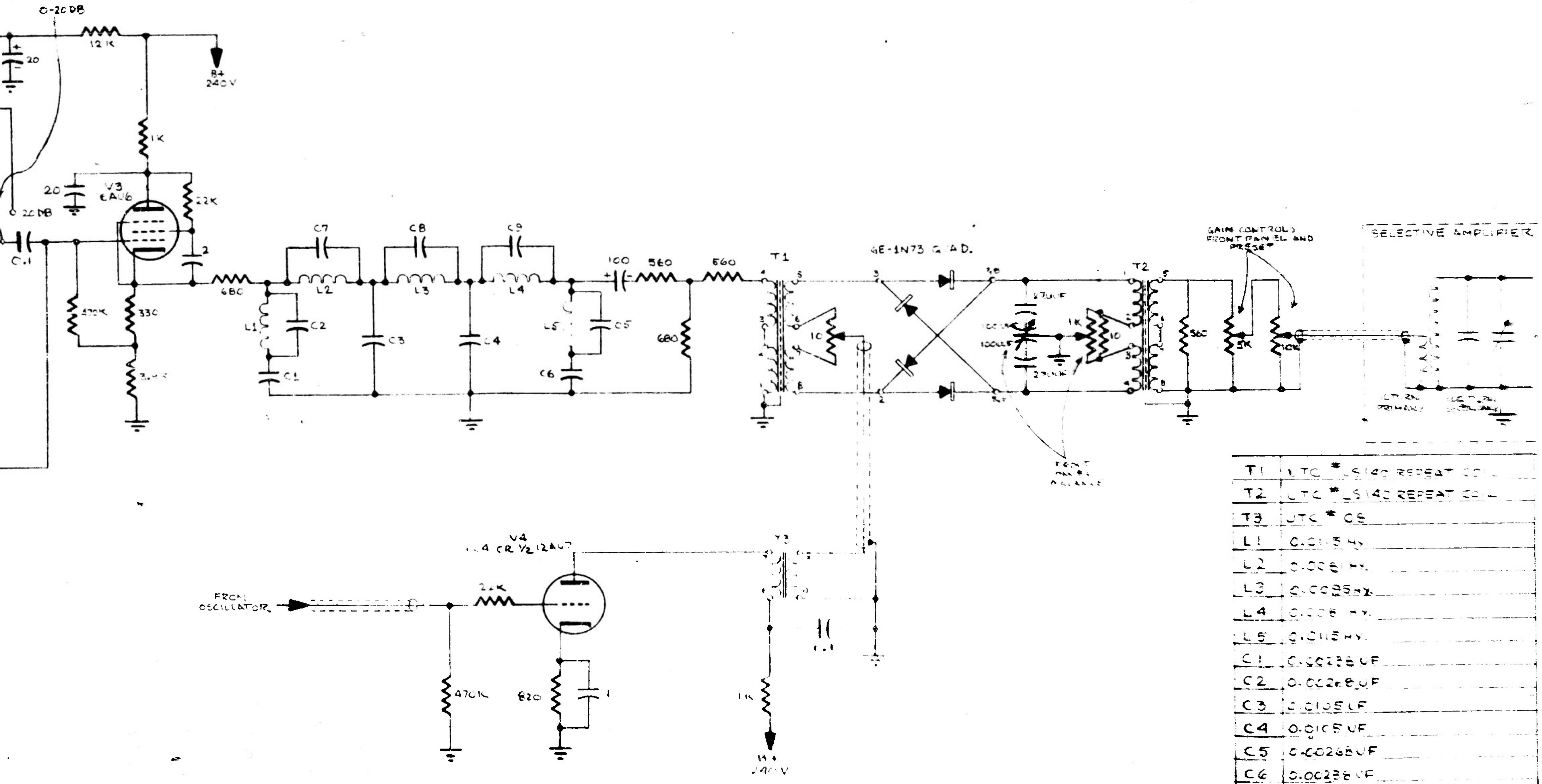
Figure 1. PREAMPLI



T1	UTC # LS140 REPEAT COIL
T2	UTC # LS140 REPEAT COIL
T3	UTC # CS
L1	0.0115 HY
L2	0.0061 HY.
L3	0.0085 HY.
L4	0.008 HY.
L5	0.0115 HY.
C1	0.00238 UF
C2	0.00268 UF
C3	0.0105 UF
C4	0.0105 UF
C5	0.00265 UF
C6	0.00238 UF
C7	0.0037 UF
C8	0.0042 UF
C9	0.0037 UF

Figure 1. PREAMPLIFIER, MODULATOR, AND LOW PASS FILTER, SCHEMATIC DIAGRAM

2