AUDIO-FREQUENCY
VECTOR-VOLTAGE INDICATOR

A THESIS

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Master of Science in Electrical Engineering

by

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AUDIO-FREQUENCY VECTOR-VOLTAGE INDICATOR

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I wish to express my sincere appreciation to Dr. F. O. Nottingham, Jr., of the Department of Electrical Engineering, not only for his suggestion of the problem, but also for his most valuable aid and guidance in its prosecution.

TABLE OF CONTENTS

LIST OF TABLES

LIST OF FIGURES

AUDIO-FREQUENCY VECTOR-VOLTAGE INDICATOR

CHAPTER I

INTRODUCTION

Throughout the fields of scientific research and engineering practice there is a great need for a fast and accurate method of measuring complex quantities such as alternating currents, alternating voltages, and impedances. Recent papers have described a device that performs this function but they are restricted to a small frequency range.[1,2]

The purpose of this thesis is to report the progress made to date on an adaptor which shows a complex quantity as a vector on the screen of an ordinary cathode-ray tube. Work on this thesis has established the method of achieving such results. The device herein described will function independently of the applied frequency. This report covers the design and construction of such a device and its associated frequency source covering the audio frequencies. Work is now being done on an extremely low-frequency source

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[1] E. A. Walker, A. H. Waynick and P. G. Sulzer, "Polar Vector Indicator", Electrical Engineering, Vol. 68 (June, 1949), p. 489.

[2] Peter G. Sulzer, "Vector Voltage Indicator", Electronics, Vol. 22 (June, 1949), pp. 107-9.

to excite this adaptor for use in servomechanism analysis.[3]

The device herein described is a compact unit with a self-contained frequency source continually tunable from 20 cycles to 20 kilocycles. By simply inserting a test network, a polar plot of the impedance diagram appears on the face of the cathode-ray tube as a function of frequency.

There are many uses for such a device. Applications include testing four-terminal networks for attenuation and phase shift, testing stability of feedback amplifiers, checking phase-shift networks, and plotting loci of complex transfer functions.

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[3]Frank O. Nottingham, Jr., "A Low Frequency Polar Vector Indicator". Unpublished conference paper presented at the A.I.E.E. Southern District Meeting, Miami Beach, Florida, April 11-13, 1951. 7 pp.

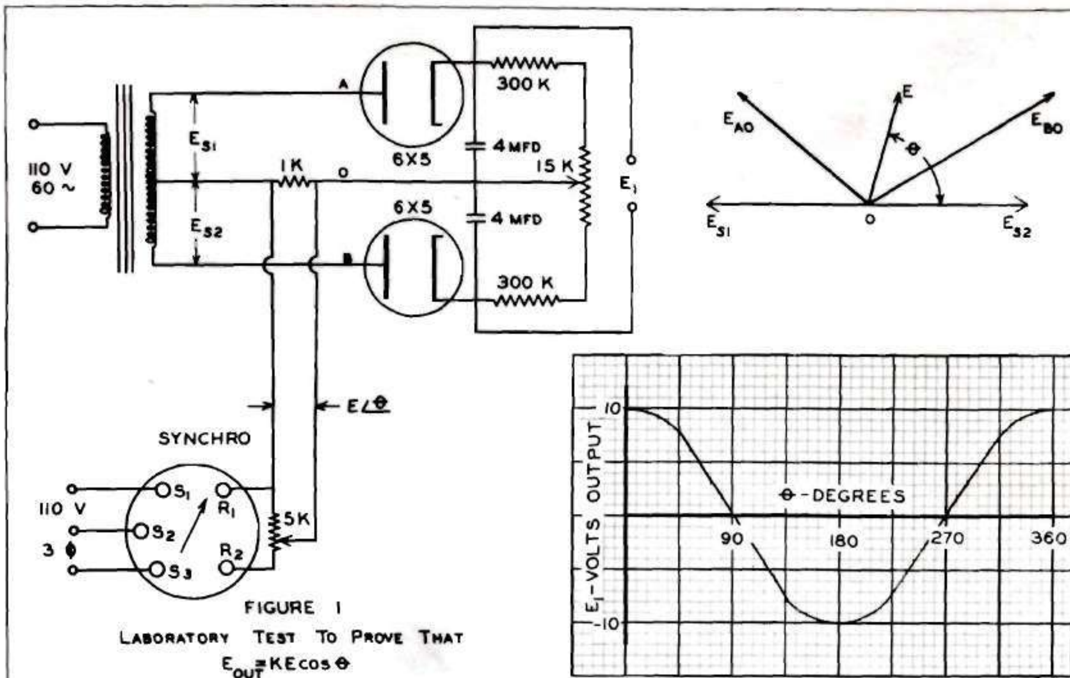

FIGURE 1

LABORATORY TEST TO PROVE THAT

$E_{OUT} = KE \cos \theta$

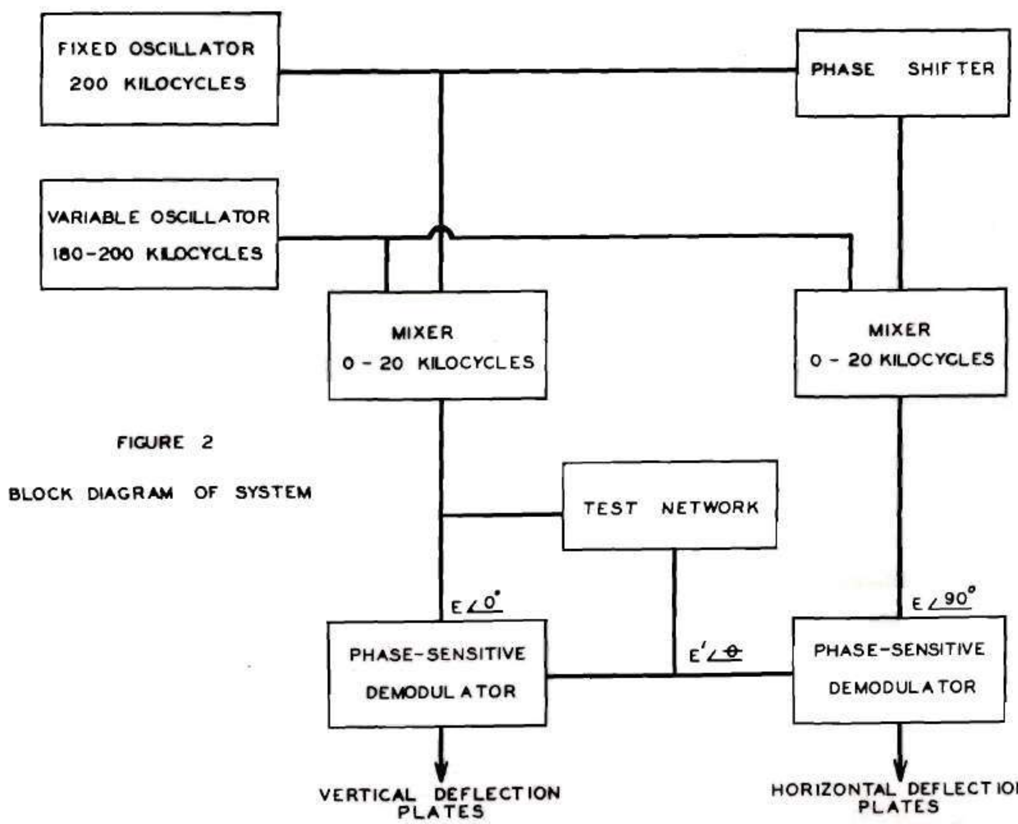

FIGURE 2

BLOCK DIAGRAM OF SYSTEM

CHAPTER II

HISTORY

The first work on this project was done in April, 1950 as a "Special Problems" project. The original idea was suggested by Dr. F. O. Nottingham, Jr.

Origin of Problem. It was suggested that with a circuit as shown in Fig. 1, a signal voltage in the return lead of a simple demodulator would cause the d-c voltage between cathodes to vary as KE cos θ. With no signal voltage applied, a balance occurs so that E_{out} is zero. The mathematics involved were checked and are shown in Appendix I. The circuit was constructed with the values shown and the following tests made.

1. With E constant at 50 volts and θ changed from zero to 360 degrees, E_{out} (at the terminals E_1) was found to vary as the cosine of θ.

2. With E varied from zero to 50 volts (constant phase angle), E_{out} was found to vary directly with the magnitude of E.

The results of these tests led to the conclusion that two such demodulators could be used to show a complex quantity on the screen of a cathode-ray tube. By shifting the phase on one demodulator 90 degrees its resulting output voltage will be KE sin θ. If KE cos θ is applied to the horizontal deflection plates of a cathode-ray tube and

KE sin θ is applied to the vertical deflection plates, the resulting deflection will represent $KE\underline{/\theta}$ and will appear as a spot at the tip of the vector.

Method of Attack. A solution was accomplished by using a system as shown in Fig. 2. Two demodulators are excited by two voltages with 90 degree phase difference. A sample of the reference voltage is used to drive a test network. The output from the network is then fed into the plate returns of the two demodulators. The resulting d-c voltage from each demodulator is applied to the horizontal and vertical deflection plates respectively of a cathode-ray tube. A spot will appear on the tube at the tip of the voltage vector, giving its magnitude and phase.

First System Considered. The first consideration was to use a system as shown in Fig. 3. A frequency of 200 kc was selected to excite the demodulators. The main reason for this choice was to make filtering of the demodulator output easier.

With such a system, there is a source of 0-20 kc voltage available for network testing. The output voltage from the test network is then fed into a mixer to return the test frequency to 200 kc. The injected signal to the demodulator must have the same frequency as the excitation voltage for the mathematics to be valid.

Second System Considered. This system could be simplified if the demodulators would function under a frequency range of 0-20 kc. One low-frequency oscillator could replace the two high-frequency oscillators and mixers. This led to the consideration of an oscillator with two outputs, both with constant voltage and 90 degree phase difference over a tuning range of 0-20 kc. With such a frequency source, a system as shown in Fig. 4 would produce the desired results.

Phase-Shift Oscillator. A simple phase-shift oscillator was designed with four 45 degree phase-shift branches as shown in Fig. 5. If R is made larger than R', the gain is independent of the frequency and the output will be constant.[4]

A study of the phase shift shows that when $R = X_c$ there is a 45 degree phase shift in each branch of the network. The individual branches react upon each other so some method of isolation must be employed to insure 45 degree phase shift in each branch. A cathode follower in each branch will isolate them and provide an output at the center branch that will be 90 degrees out of phase with the end branch. Such an oscillator was constructed and is shown in Fig. 6.

No combination of RC could be found that would preserve the 90 degree phase shift in a standard 10:1 frequency change. Tapering R would set the 90 degree

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[4] W. R. Hinton, "Phase-Shift Oscillator", Wireless Engineer, Vol. 27 (February, 1950), pp. 65-6.

shift for a particular frequency but there was still too much interaction to preserve the angle over a 10:1 frequency range.

It was then decided to return to the system shown in Fig. 3 and use this phase-shift oscillator to cover the 180-200 kc range.

The necessary RC changes were made, but when C was padded to cover the 180-200 kc range, R became so small that the output varied with the frequency. The best results were obtained with the output voltage varying 20 per cent over the tuning range. This circuit was abandoned and a straight four-branch network was used (the cathode followers were removed). A feedback circuit was added and AVC action employed to get a constant output voltage. With the output constant, the waveform was distorted.

The next system tried for a frequency source was one containing a crystal oscillator operating at 100 kc and a variable Wien-bridge oscillator operating from 90-100 kc.

<u>Crystal Oscillator</u>. The first circuit tried was a modified-Hartley oscillator. This circuit did not prove satisfactory due to the second harmonic component in the output voltage.

A circuit as shown in Fig. 7 was then constructed. The output voltage from this oscillator proved to be of good wave form and very stable under varying plate supply voltage.

Wien-Bridge Oscillator. A Wien-bridge oscillator was constructed as shown in Fig. 8. The output voltage was varied from 90-100 kc with constant magnitude and good wave form.

Mixer. A mixer was constructed as shown in Fig. 10. When the two oscillators were applied to this mixer there was too much attenuation in the output near the 10-kc point.

Tests. A phase shifter was installed in the fixed-frequency output. Initial tests gave an output with a phase difference variable from 80-100 degrees. The output increased sharply near the low-frequency end of the range. A careful check disclosed that the two oscillators were locking into synchronism as the variable oscillator neared 100 kc. All stages were well shielded from one another to prevent this interaction. It was then discovered that the two oscillators were drifting too much. The output was not useful under 1000 cycles. At frequencies below 1000 cycles the drift became very apparent with instability occurring near 200 cycles.

These oscillator tests proved that more work must be done on this section of the system.

Summary. The work to date has dealt primarily with the design and construction of a frequency source variable from 0-20 kc. It must have two outputs with 90 degree phase difference and constant magnitude over the entire tuning range.

Results. The results of the work to date are:

1. A system as shown in Fig. 2 is desirable.

2. Two different types of oscillators have too much frequency drift to be useful in the low-frequency range.

3. Interaction between the r-f units can be eliminated only by special care in wiring arrangements, proper shielding and proper decoupling networks for each section of the heterodyne oscillator.

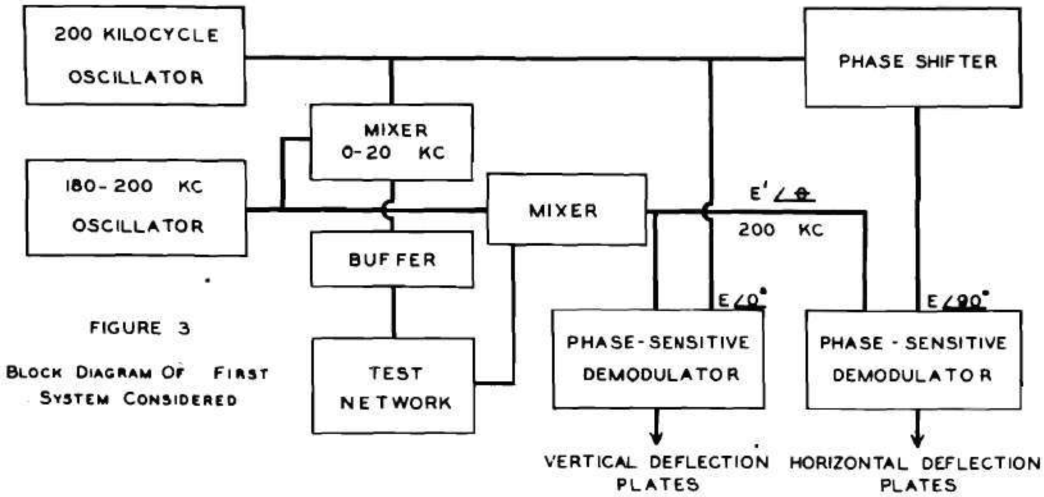


FIGURE 3

Block Diagram Of First System Considered

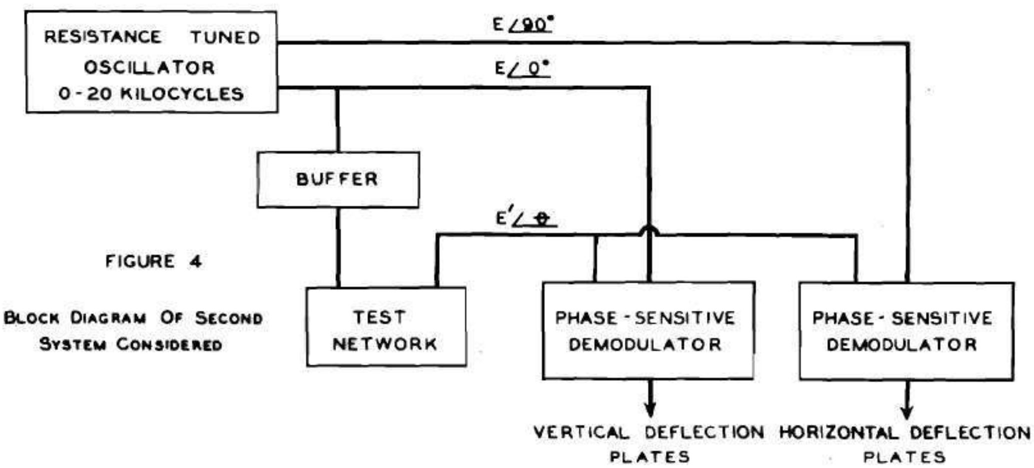
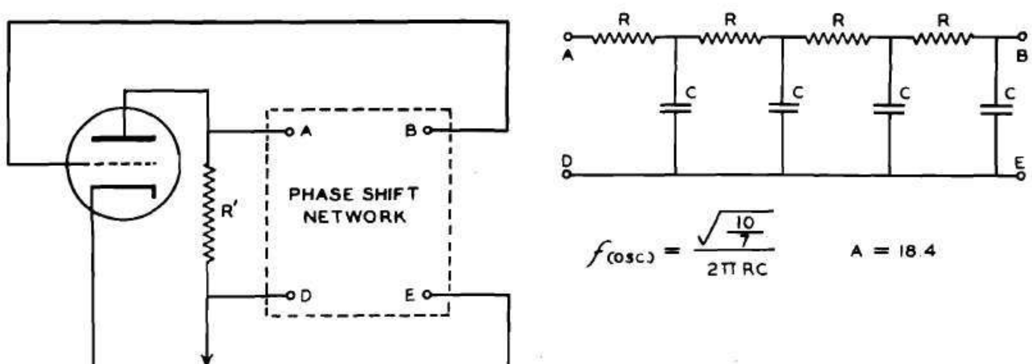


FIGURE 4

Block Diagram Of Second System Considered



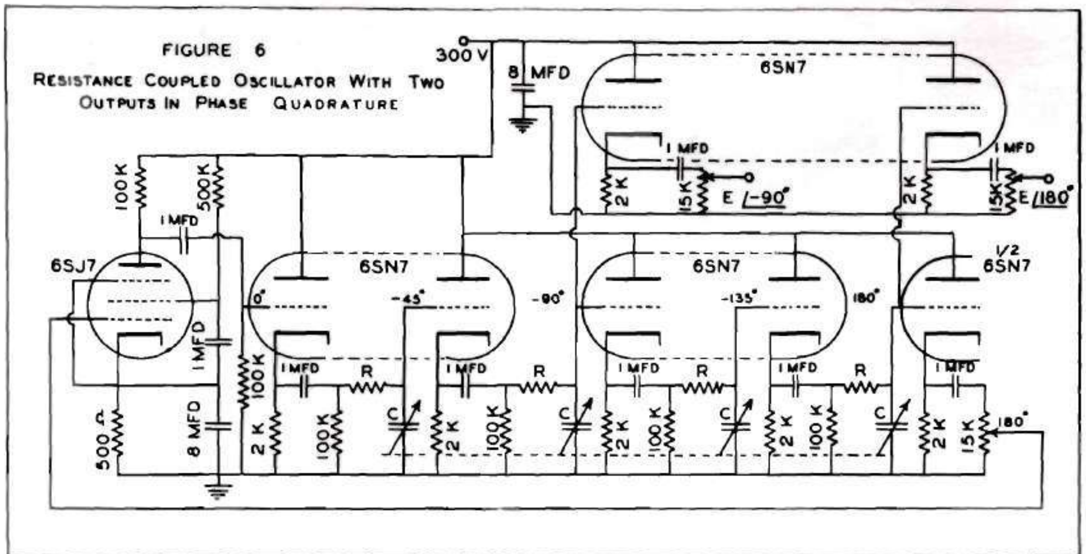
$$f_{(osc)} = \frac{\sqrt{\frac{10}{7}}}{2\pi RC} \qquad A = 18.4$$

FIGURE 5

Resistance Tuned Oscillator Using A Phase Shift Network

FIGURE 6

RESISTANCE COUPLED OSCILLATOR WITH TWO OUTPUTS IN PHASE QUADRATURE



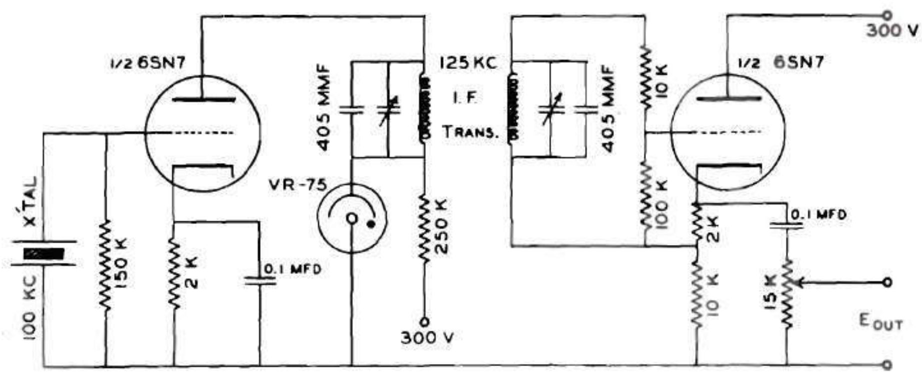
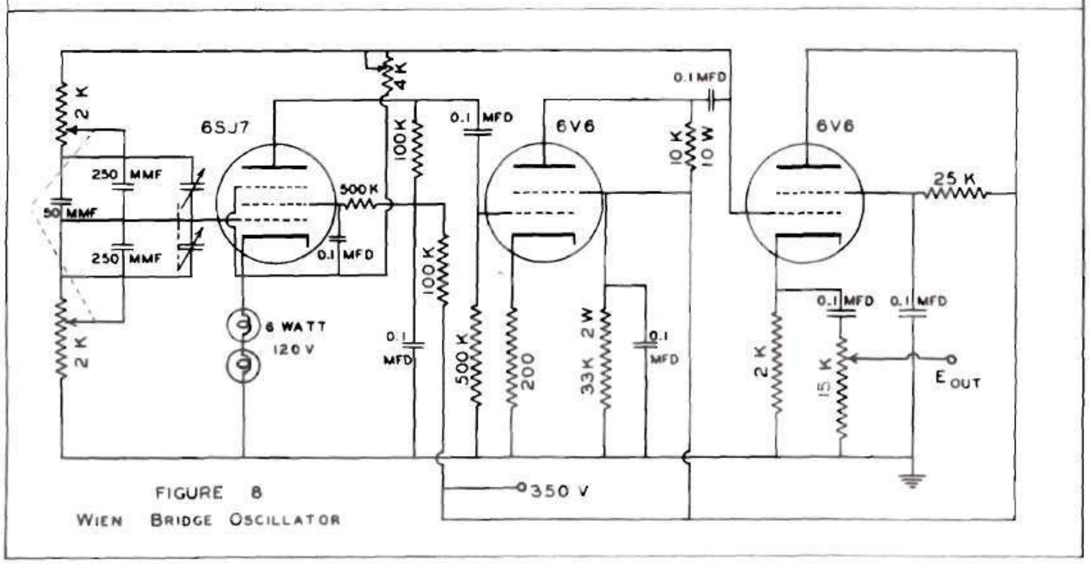


FIGURE 7

CRYSTAL OSCILLATOR

FIGURE 8

WIEN BRIDGE OSCILLATOR



CHAPTER III

HETERODYNE OSCILLATOR DESIGN

The design of a heterodyne oscillator involves a number of special problems,[5] among which are the prevention of interaction between the oscillators, elimination of harmonics and other undesired frequencies, improvement of frequency stability, and the prevention of variation in output level. Interaction of the two r-f oscillators causes them to pull into synchronism when their frequency difference is small and thereby prevents the production of low-audio frequency. Interaction may be prevented by adequate shielding, use of chokes or decoupling resistors, and by correct methods of coupling the outputs of the r-f oscillators to the mixer tube. The coupling method used here is a pentagrid-mixer tube, the oscillator outputs being applied to grids 1 and 3 which are shielded from each other.

The output of a heterodyne oscillator cannot be free of harmonics unless the output of at least one r-f oscillator is also free of harmonics and the higher-order distortion of the mixer is negligible. Distortion will result if the two r-f voltages contain harmonics of the

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[5] Herbert J. Reich, "Heterodyne Oscillators", Theory and Applications of Electron Tubes, New York: McGraw-Hill Book Company, Inc., 1939, p. 348.

same order. This can be explained by reference to fixed and variable oscillators whose frequencies are ω_r and $\omega_r - \omega_a$, respectively, where ω_a is the desired audio frequency. The $\underline{n}^{th}$ harmonic of the fixed frequency is $\underline{n}\omega_r$ and that of the variable frequency $\underline{n}(\omega_r - \omega_a)$. The output of the detector will contain not only the difference of the fundamental frequencies, $\omega_r - (\omega_r - \omega_a) = \omega_a$, but also the difference of the harmonics, $\underline{n}\omega_r - \underline{n}(\omega_r - \omega_a) = \underline{n}\omega_a$. Thus, if the two r-f voltages applied to the mixer contain harmonics of the same order, the beat-frequency output of the heterodyne oscillator will also contain a harmonic of the same order. This type of distortion may be prevented by filtering out the harmonics of at least one of the r-f oscillators.

The frequency stability of a heterodyne oscillator depends on the stability of the individual r-f oscillators. If the r-f oscillators tend to drift, then the output of the heterodyne oscillator will not be suitable for low-frequency work. Any drift will be especially noticeable in the neighborhood of 20-200 cycles.

To minimize the frequency instability of the heterodyne oscillator it is best to use the two r-f oscillators that are identical in components and structural details. If the two oscillators have the same drift characteristics, then the error in the heterodyne-oscillator output will be minimized and the low-audio frequencies may be realized.

<u>R-F Oscillator Section</u>. The results evidenced from

the previous oscillators led to the construction of two identical r-f oscillators, as shown in Fig. 9.[6]

Each oscillator was constructed on an individual chassis to eliminate as far as possible any chance of interaction between the two r-f units. The chassis consists of a piece of one-eighth inch aluminum measuring 3.5 x 8 inches.

The frequency determining elements of the oscillator are R_1, R_2, and C_1. These elements were padded and trimmed until an optimum operating frequency range was reached.

R_3 and C_2 were made adjustable so that the amount of regenerative feedback could be controlled for optimum operation. These two elements determine the harmonic content and magnitude of the output voltage. The controls for these adjustments are reached through the tops of the shield cans.

The components are laid out in such a manner as to keep all leads as short as possible. The oscillators are not self-contained in that they depend on an outside source for filament and plate voltages. One side of the filament supply is grounded where it enters the chassis. The other side is by-passed with a 0.5-mfd. capacitor. The plate supply is fed through a decoupling pi network consisting of a series 60-mh. r-f choke shunted to ground by 0.1-mfd. capacitors.

All ground connections are made to one bus that is

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[6] Peter G. Sulzer, "Wide-Range RC Oscillators", Radio and Television News, Vol. 44 (September, 1950), p. 43.

connected to the chassis at one point. The r-f unit is mounted on a chassis measuring 11 x 17 x 3 inches as shown by the layout photograph in Appendix III. The r-f chassis is supported by four 1.5-inch chassis spacers and is completely covered by an aluminum shield.

Both oscillators were made adjustable over the same frequency range. The tuning capacitor control is brought through one end of the shield where it is controlled by a 5:1 vernier. The two oscillators are mounted side by side on the main chassis as shown in Appendix III.

The r-f unit to be used as the variable-frequency control is centered on the main chassis with its control accessible from the front panel. The vernier for this unit is fitted with a 6.5-inch aluminum disk that is calibrated in kilocycles. The other r-f unit is reversed and mounted on the right-hand side of the control unit. The vernier for this last unit is then available from the back for calibration purposes.

Mixer Construction. Two mixers were constructed as shown in Fig. 10 and were placed as shown in Appendix III. The input and output leads were shielded as shown in the circuit diagram. All the components were arranged so as to keep the leads as short as possible. One filament lead is grounded at the tube socket and the other is by-passed with a 0.1-mfd. capacitor to ground. Care was taken to make all ground connections to one bus and then run this bus to ground at only one point. The decoupling network for each mixer consists of a 10,000-ohm resistor by-passed to ground through a 1-mfd. capacitor.

Two outputs are obtained from each mixer. One output goes directly to the grid of a phase inverter while the other output is made available on the front panel through a cathode follower. The phase inverter is the first stage of the demodulator section and is described in detail in a later chapter.

A 6SA7 pentagrid-converter type tube was selected for the mixer circuit. The variable-frequency voltage is applied to the control grid through a 0.1-mfd. coupling capacitor while the second grid is driven by the fixed-frequency voltage.

The filtering network for the plate circuit was designed by ordinary low-pass filter design principles; however, it was found necessary to place a trimming capacitor across the r-f choke to minimize the 200-kc. component in the output voltage. Slight modifications of the design had to be made in order to get minimum attenuation over the tuning range of the oscillator and still maintain a constant voltage output. The final design is shown in the circuit diagram.

Phase Shifter. A phase-shift network was installed in the fixed-frequency oscillator output. This network consists of a circuit as shown in Fig. 9. The potentiometer control is brought through the top of the main chassis and is located between the two mixer tubes as shown in Appendix III.

In order that the two audio-frequency outputs may be made equal, two potentiometers were inserted in the r-f oscillator outputs. This system is shown in Fig. 11 and it consists of a potentiometer across each r-f oscillator output. The full output of the variable r-f unit goes to the second grid of the quardrature mixer while the full output of the fixed-frequency r-f unit goes through the phase shifter to the second grid of the in-phase mixer. The tapped output of the fixed r-f unit goes to the control grid of the quadrature mixer and the tapped voltage of the variable r-f unit goes to the control grid of the reference mixer. The two output voltages from the heterodyne oscillators can be made equal in magnitude by adjustment of these two potentiometers. There is no need for further adjustments once these voltages are made equal so the controls were placed in a small aluminum box below the chassis where they were well shielded.

Summary. The heterodyne oscillator assembly is complete and constitutes the frequency source for the adapter. After the assembly was completed, it was tested for the following:

1. Frequency range of each of the two r-f units.

2. Magnitude and wave form of the output of each r-f oscillator.

3. Frequency range and wave form of the mixer output.

4. Phase difference of the mixer outputs.

Results. The results of the above tests are listed below.

1. The frequency range of both r-f units is 165-180 kilocycles.

2. The r-f output of the units is of good wave form and is 10 volts in magnitude.

3. The mixer output is stable and constant with good form from 60 cycles to 13.5 kilocycles.

4. The two mixers have 90 degree phase difference over the entire low-frequency range.

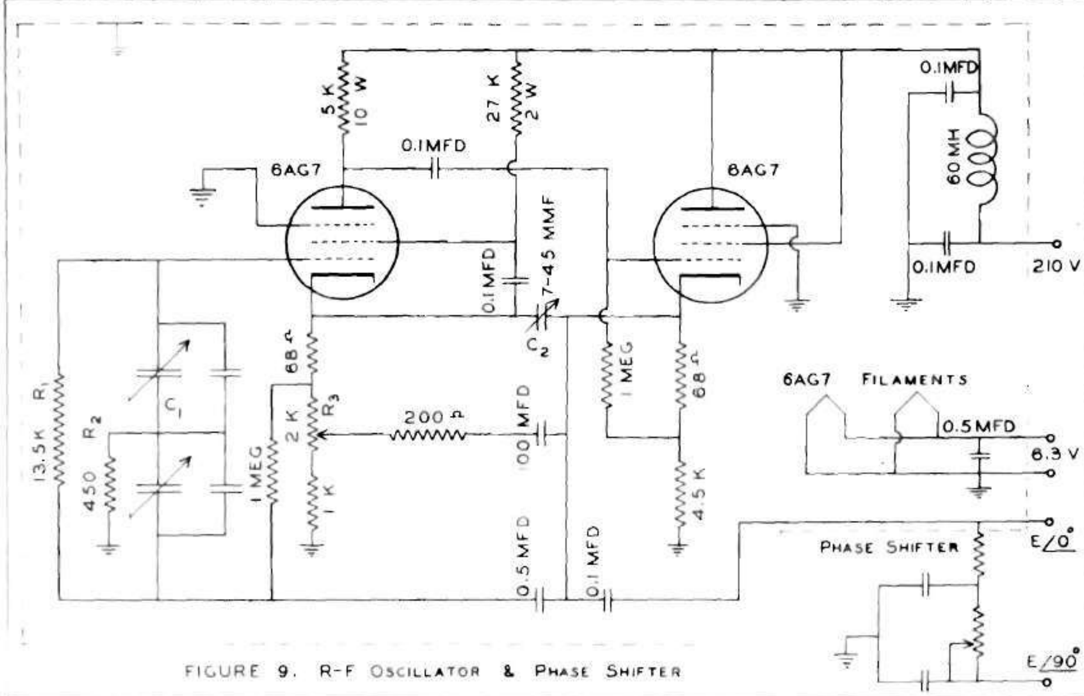

FIGURE 9. R-F OSCILLATOR & PHASE SHIFTER

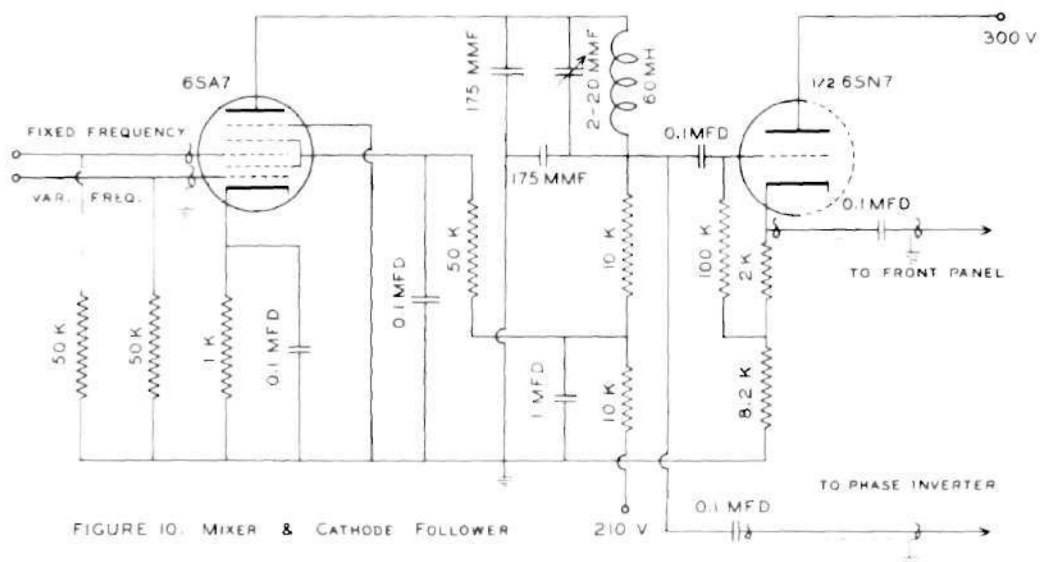

FIGURE 10. MIXER & CATHODE FOLLOWER

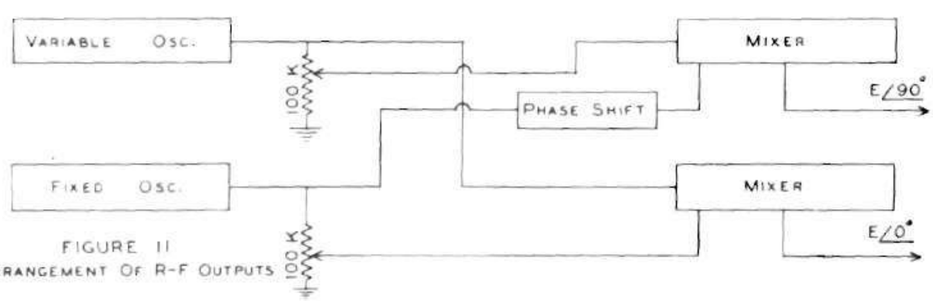

FIGURE 11
ARRANGEMENT OF R-F OUTPUTS

CHAPTER IV

DEMODULATOR ASSEMBLY

The heterodyne oscillator has two constant output voltages that have a 90 degree phase difference. The sources of these output voltages are grounded on one side and are not suitable for use with the demodulator circuit. The demodulators require two input voltages that have 180 degree phase difference.

The heterodyne oscillator outputs are fed into two individual channels, each containing a phase inverter, two cathode followers and a demodulator as shown in Fig. 12.

Phase Inverter. Conversion from an unbalanced source to a balanced output requires a phase inverter. This inversion requires a network with two output voltages equal in magnitude but opposite in phase, and proportional to the driving voltage. While a transformer fulfills this requirement, its frequency response is not uniform.

A vacuum tube is more economical and more uniform in frequency response to invert the phase in a circuit that will automatically equalize the two output voltages. A circuit was constructed as shown in Fig. 12. By using a circuit such as this, we sacrifice the automatic-equalizer section for circuit gain. By using an equalizer circuit, the best gain available is unity.[7] This circuit as constructed

[7] Myron S. Wheeler, "An Analysis of Three Self-Balancing Phase Inverters", Proceedings of the I.R.E., Vol. 34 (February, 1946), pp. 67-9.

gives a gain of thirteen.

The potentiometer in the plate circuit of the first tube provides an easy means of equalizing the output signals. The output phase difference was found to be very near 180 degrees over the entire 0-20 kc tuning range. There is negligible distortion of the input signal as it is inverted and amplified.

Cathode Follower. The output voltages from the phase inverters cannot be coupled directly to the demodulators because of the impedance differences. The demodulators have an impedance to ground of 100,000 ohms on the negative half-cycle and 50,000 ohms on the positive half-cycle. Such a varying load cannot be coupled to the high-impedance plate circuit of the phase inverters. The impedance of the phase inverters can be lowered to a usable value by use of a cathode follower.

Demodulators. Two demodulator units were constructed according to the schematic shown in Fig. 12. Each unit contains a 6H6 type tube arranged as two half-wave rectifiers. The plate input voltages are taken from the cathode followers through 1-mfd. capacitors. The plate circuits are then returned to ground through 100,000-ohm resistors. The load network of each 6H6 channel consists of a 100,000-ohm resistor with a 0.5-mfd. capacitor shunted across it. The common junction of the four load circuits is returned to ground through a 15,000-ohm resistor. Because the test voltage is introduced at A, this point is carried to an appropriate terminal on the front panel of the chassis.

The circuit arrangement for the whole demodulator assembly is shown in Appendix III.

Summary. The demodulator assembly now contains two channels, each one consisting of a phase inverter, two cathode followers and two half-wave demodulator units.

The phase inverters are driven by unbalanced voltages with 90 degree phase differences. Each channel then inverts the phase and provides a balanced output to the demodulators through a set of cathode followers. The plate voltages on the demodulators are 90 degrees apart.

The demodulator plate voltages are adjusted so that the two channels are balanced. This balance is achieved by adjusting the mixer inputs until the plate voltages of the through-channels in the demodulators are equal. Adjustment of the phase-inverter circuits then will equalize the other two plate voltages individually.

There must be some means of fine adjustment in the demodulator load circuits so that an exact balance can be made with no signal applied. This is accomplished by using a potentiometer arrangement as shown in Fig. 12.

The following tests were made after the demodulators were balanced.

A. With no test signal applied to the assembly, the following measurements were made.

(1) The plate-to-cathode voltage.

(2) The plate-to-ground voltage.

(3) The phase difference from plate-to-plate of each demodulator.

(4) The phase difference between the two demodulator plates.

(5) The output voltages to ground.

(6) The output voltages between cathodes.

B. With an in-phase test signal of variable magnitude applied, measure the output voltage of each demodulator as a function of the applied voltage.

C. With a quadrature test signal of variable magnitude applied, measure the output voltage of each demodulator, as a function of the applied voltage.

D. With a test signal of 45 degree phase displacement applied, measure the output voltage of each demodulator as a function of the applied voltage.

Results. The results of the above tests were:

A. (1) The plate-to-cathode voltages are 13.2 volts.

(2) The plate-to-ground voltages are 13.2 volts.

(3) The phase difference between plates of each demodulator is 180 degrees.

(4) The corresponding plate voltages of the two demodulators are 90 degrees apart.

(5) The d-c voltages from cathode-to-ground are 10.8 volts.

(6) The output voltages between cathodes are zero.

B. The d-c output voltage of the in-phase channel varies linearly with the a-c input when the applied voltage is in-phase and remains zero when the applied voltage is in phase quadrature with the reference.

C. The d-c output voltage of the quadrature channel varies linearly with the a-c input when the applied voltage is in phase quadrature and remains zero when the applied voltage is in-phase with the reference.

D. With a variable a-c voltage of 45 degree phase displacement applied, both channels varied linearly and equally with the applied voltage.

The results of B, C, and D are shown in Table I and Fig. 16.

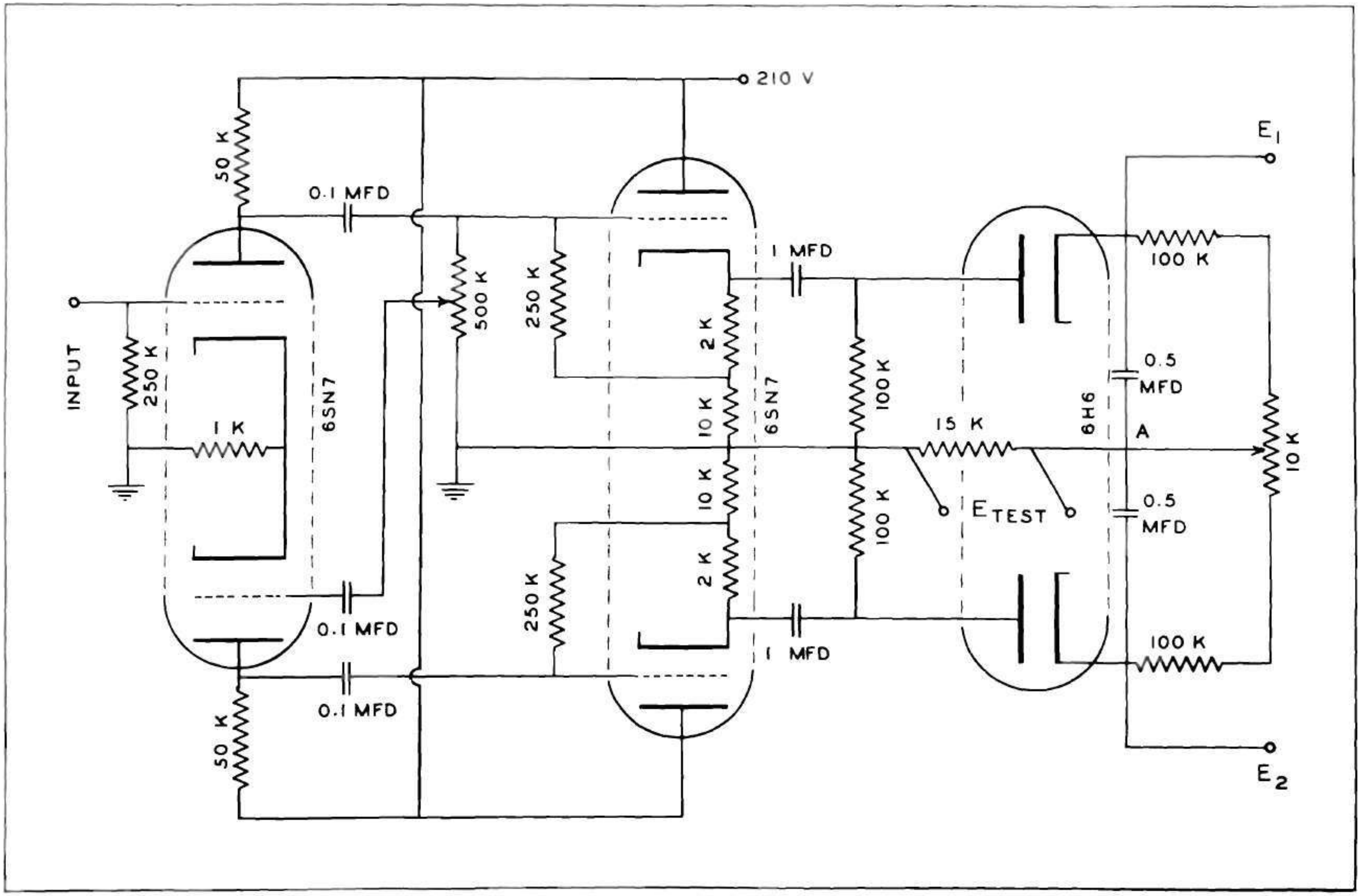

FIGURE 12. DEMODULATOR ASSEMBLY

CHAPTER V

DIRECT-COUPLED AMPLIFIERS

The output from the demodulators is direct current and is of such low level that it must be amplified before it is placed on the plates of the cathode-ray tube.

A circuit that will amplify direct current must be directly coupled. Direct-coupled amplifiers possess certain disadvantages, and their use is limited to slowly varying inputs.[8]

Because of the fact that the grid of one stage is directly coupled to the plate of the previous stage, it is necessary to include d-c sources in the grid circuits in order that the quiescent conditions be those of Class-A operation. This is a disadvantage in that the necessary high-voltage grid-bias batteries are expensive. A second disadvantage of the direct-coupled amplifier is the inherent instability associated with the direct coupling. The characteristics of the tubes in the circuit change with time, or the output voltage of a-c line-operated power supplies, likewise change with time. Since such changes are amplified, some precautions must be taken to overcome this instability. For this reason, balanced circuits with

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[8] Samuel Seely, "Direct-Coupled Amplifiers", Electronic-Tube Circuits, New York: McGraw-Hill Book Company, Inc., 1950, p.111-13.

degenerative feedback are used.

<u>The Difference Amplifier</u>. To avoid the disadvantages of a direct-coupled amplifier, the difference amplifier shown in Fig. 13 was constructed. If one stage of the amplifier is considered, the output is found to be proportional to the difference of the input voltages. The output from one stage is

$$(e_{o1} - e_{o2}) = \frac{-\mu R_L}{R_L + r_p}(e_1 - e_2).$$

The difference amplifier can be used to indicate the exact point of balance between two d-c potentials. In this circuit, when there is no input signal and when the tubes are properly matched, the voltage output of the amplifier is zero. When a signal voltage is applied, an amplified signal voltage results.

Throughout the amplifier the circuits must be balanced when the input signal is balanced or is zero. At the output of the first stage there are two potentiometers in a balancing circuit. One is placed in the plate circuit of the first stage and another in the grid circuit of the following stage. A balance is maintained in the third stage by degenerative coupling to the second stage. The final stage of amplification is balanced by slight adjustment of one plate-load resistor. A potentiometer in the grid circuit of the third stage provides a gain control for the magnitude of the output voltage.

Construction. Two difference amplifiers are needed, one for each demodulator output. Each one was constructed on an individual chassis measuring 12 x 5 x 2 inches. The four potentiometer controls were brought through the top of the chassis for easy access. The layout details are shown in Appendix III.

Summary. The outputs of the demodulators are direct current and are difference voltages on the order of ±2 volts. This output voltage must be on the order of 125 volts to get full deflection on a standard five-inch cathode-ray tube. Two difference amplifiers were used to amplify the signals to a useable level.

Each amplifier was tested for amplification and linearity.

Results. The gain of the amplifiers is approximately eighty and is linear for values of input voltage up to one volt. The measurements are shown in Table II and curves are shown in Fig. 17.

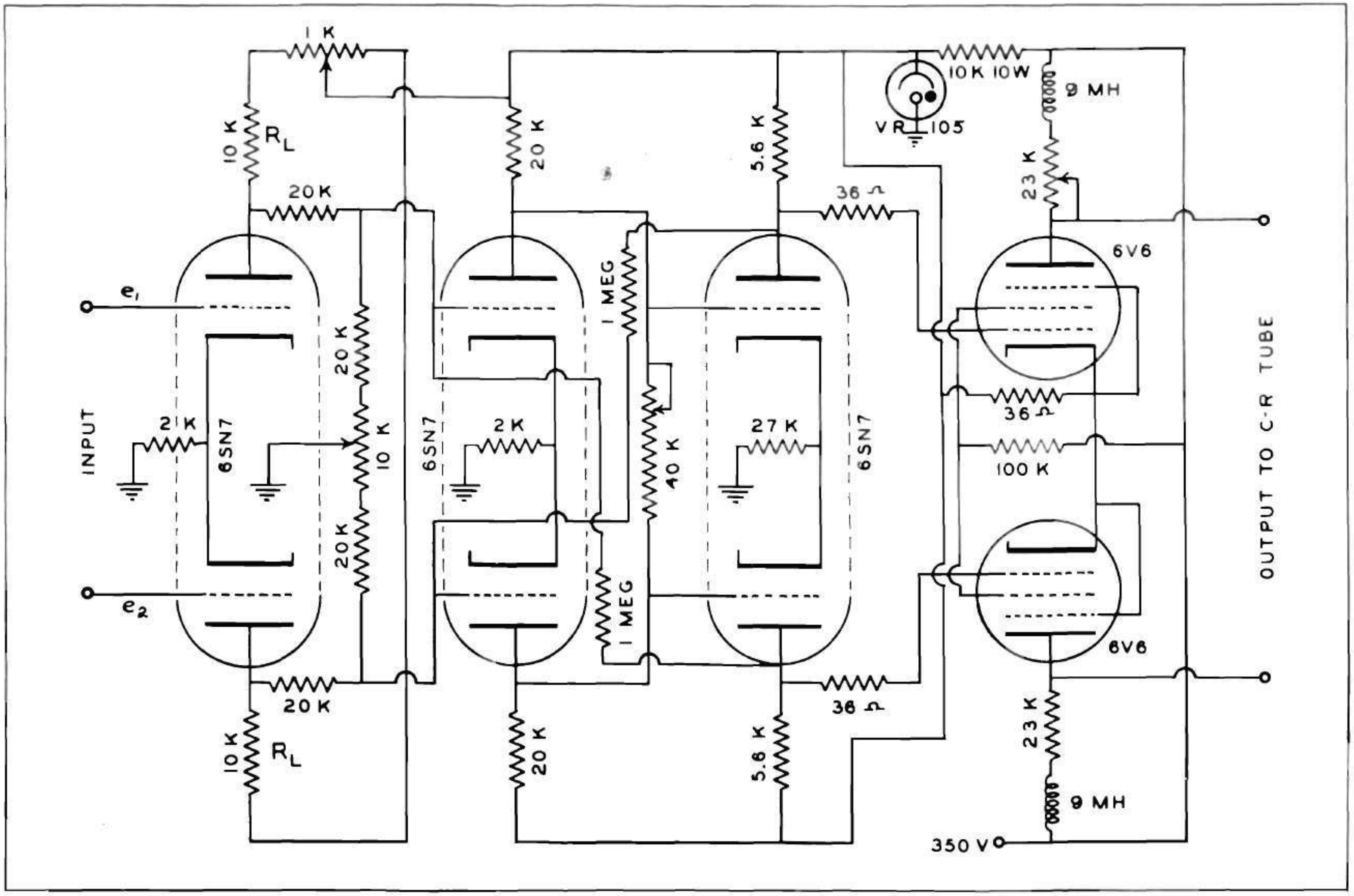

FIGURE 13. DIRECT-COUPLED AMPLIFIER

CHAPTER VI

SUMMARY AND CONCLUSIONS

The completed adaptor contains the following elements.

A. Main chassis

(1) Heterodyne oscillator

a. Two r-f oscillators

b. Two mixers

c. Two cathode followers

(2) Demodulator assembly

a. Two phase inverters

b. Two cathode followers

c. Two half-wave demodulators

B. Separate chassis

(1) Two direct-coupled amplifiers

(2) Power supply

<u>Alignment</u>. The elements of the adaptor were aligned and balanced according to the instructions in their respective chapters.

Hereafter, the in-phase channel will be referred to as the cosine channel and the quadrature channel as the sine channel.

The two outputs from the direct-coupled amplifiers are applied to the deflection plates of a cathode-ray tube, care being taken to place the output of the cosine amplifier on the horizontal plates and the output of the sine amplifier on the vertical plates. The spot on the cathode-ray tube is centered when all the circuits are balanced and there is no signal applied.

A jumper is then placed from the in-phase test signal source to the demodulator input terminal. The gain control of the cosine amplifier is adjusted until the spot on the cathode-ray tube is deflected full scale to the right. A jumper is then placed from the quadrature test signal source to the demodulator input terminal. Next, the gain control of the sine amplifier is adjusted until the spot on the cathode-ray tube is deflected full scale upward.

A test network can now be inserted and the vector of the output voltage will appear on the face of the cathode-ray tube as a function of frequency.

Tests. A transmission line consisting of twelve pi sections was constructed. Each section contains a 10-mh. coil (5 ohms) and two 0.0125-mfd. capacitors. The line was tested with an audio-frequency generator to determine the proper termination. The line appeared to have fewer reflections over the audio-frequency range with a resistive terminal load of 500 ohms.

This line was used as a test network for the adaptor. To get a true representation of the voltages at different points along the line, the line input terminals were connected directly to the cathode of the cathode follower in the test signal source. This connection was used in order to eliminate the coupling capacitor at the output of the cathode follower.

The voltages at different points along the transmission line were fed to the demodulators and observed on the cathode-ray tube as the frequency applied to the line was varied over the audio spectrum.

A low-pass filter section was used as another test network. This network contains three 50-mh. (60 ohms) coils in series. The two center junction points are by-passed to ground through 0.025-mfd. capacitors. This network is also connected directly to the cathode-follower source and it is terminated with a 5,000-ohm resistor.

Results. With the transmission line used as a test network, the following results were noted.

1. The input voltage remains constant with a change in frequency but has a very slight phase shift. This indicates a mis-match in the input impedance and the presence of reflections in the line.

2. A photograph of the vector voltage at the receiver end of the transmission line is shown in Fig. 14A and a photograph of the vector voltage at the end of the fourth section is shown in Fig. 14B. Both voltages are shown as a function of frequency.

3. The frequency was set at a value that made the transmission line one wave length long. The voltage at each station along the line was observed and found to be nearly equal in magnitude. A composite photograph of the vector

voltage at each station was made and is shown in Fig. 14C.

The low-pass filter was then used as a test network and the folloeing results recorded.

1. The input voltage remained constant over the entire frequency range.

2. The output voltage with a variable-frequency input is shown in Fig. 15. As the frequency increased, the vector rotated clockwise, indicating a phase lag and shortened, indicating attenuation. The total phase shift was a little over 360 degrees, which is to be expected with a three-section filter of this type.

Conclusions. Although no measurements were made to determine the accuracy of the adaptor, it is believed that so long as the demodulators and amplifiers are not driven beyond their respective linear ranges, the accuracy should be excellent.

The results of the tests made so far show that the adaptor will be very useful for many other experiments, such as testing the stability of feedback amplifiers, checking phase-shift networks and filter sections, and analysis of transmission lines and servomechanisms.

This adaptor is by no means limited just to the audio frequencies. A modification of the frequency source would make this adaptor work just as well for lower and higher frequencies.

Recommendations. As a result of the experiments run to date, the following recommendations are made.

1. The test signal source should be modified such that an output voltage on the order of 10-volts rms. with a very low output impedance would be available.

2. The linearity of the direct-coupled amplifiers should be improved by devising a method of providing successively higher plate voltages for each stage of amplification.

3. A series of tests should be made to determine the accuracy of the adaptor.

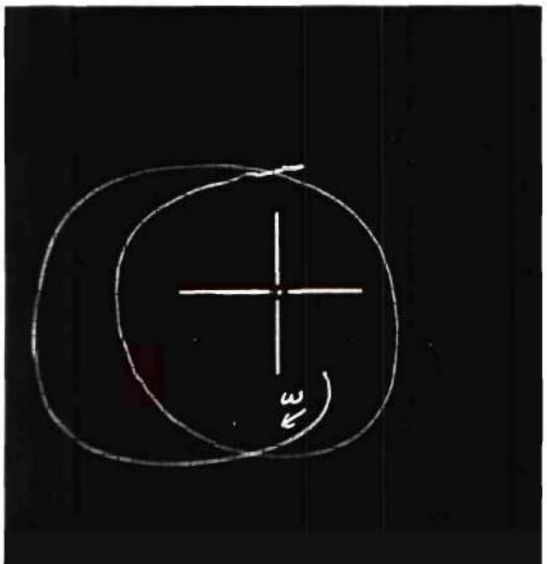

FIGURE 14 A

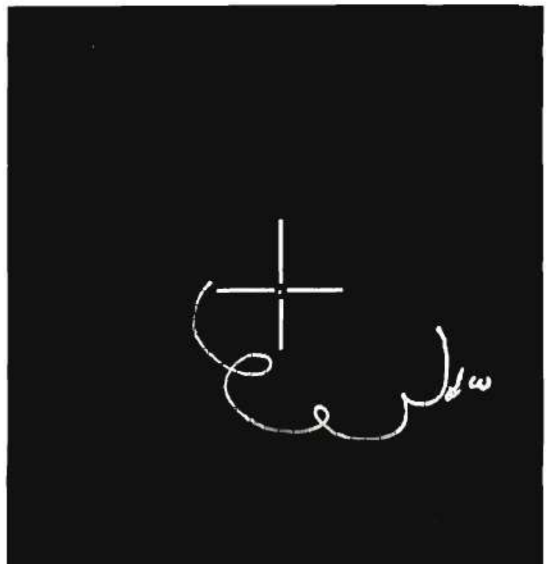

FIGURE 14 B

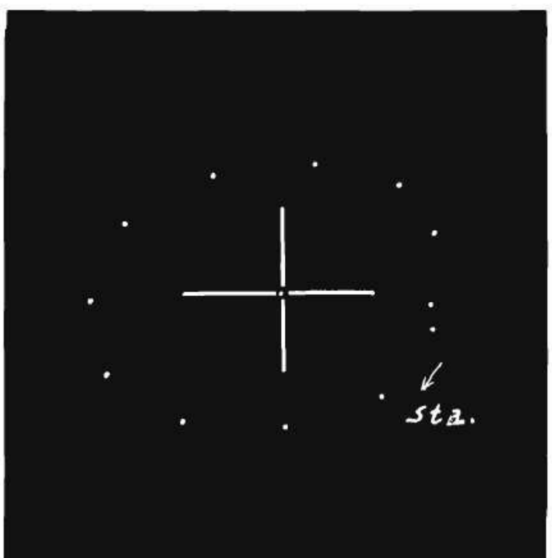

FIGURE 14 C

FIGURE 15

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BIBLIOGRAPHY

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APPENDIX

APPENDIX I

DEMODULATOR CHARACTERISTICS

If two diodes are connected as shown in Fig. 1, it can be shown that for small signal voltages, $E\underline{/\theta}$, E_{out} (at terminals E_1) is proportional to the magnitude of E and to the cosine of θ.

$$E_{ab} = E_{ao} - E_{bo}$$

$$E_{ao} = (E_{s1}^2 + E^2 - 2E_{s1}E \cos \theta)^{1/2} = (A - B)^{1/2}$$

$$E_{bo} = (E_{s2}^2 + E^2 + 2E_{s2}E \cos \theta)^{1/2} = (A + B)^{1/2}$$

$$(A - B)^{1/2} = A^{1/2} - 1/2\, A^{-1/2}B - 1/8\, A^{-3/2}B^2 - 1/16\, A^{-5/2}B^3 + \ldots$$

$$(A + B)^{1/2} = A^{1/2} + 1/2\, A^{-1/2}B - 1/8\, A^{-3/2}B^2 + 1/16\, A^{-5/2}B^3 + \ldots$$

$$E_{ab} = -\,(A^{-1/2}B + 1/8\, A^{-5/2}B^3 + \ldots)$$

$$E_{out} = KE_{ab}$$

K is the ratio of the d-c output voltage to the effective a-c input voltage of each diode; therefore, neglecting the negative sign

$$E_{out} = K\left[\frac{2E_sE \cos \theta}{(E_s^2 + E^2)^{1/2}} + \frac{(E_sE \cos \theta)^3}{(E_s^2 + E^2)^{5/2}} + \ldots\right]$$

and if $E \leqq 0.1E_s$, the higher-order terms can be neglected. Then

$$E_{out} = K'E \cos \theta.$$

If θ is replaced by $(90 + \theta)$, the output voltage then becomes

$$E_{out} = K'E \sin \theta.$$

APPENDIX II

TABLES AND GRAPHS

TABLE I: Test of the Demodulator Assembly.

Curves of E_{in} vs. E_{out} are shown in Fig. 16.

TABLE II: Test for Linearity of the Direct-Coupled Amplifiers.

Curves of E_{in} vs. E_{out} are shown in Fig. 17.

APPENDIX II

TABLE I: Test of the Demodulator Assembly.

A-C Volts Input (rms.)			D-C Volts Output Cosine			Sine		
0°	45°	90°	0°	45°	90°	0°	45°	90°
	0.0		0.00	0.00	0	0	0.00	0.00
	0.1		0.20	0.10	"	"	0.08	0.17
	0.2		0.38	0.24	"	"	0.20	0.35
	0.3		0.56	0.40	"	"	0.35	0.54
	0.4		0.75	0.55	"	"	0.48	0.71
	0.5		0.95	0.70	"	"	0.60	0.92
	0.6		1.12	0.84	"	"	0.75	1.21
	0.7		1.33	1.00	"	"	0.90	1.32
	0.8		1.53	1.10	"	"	1.00	1.52
	0.9		1.72	1.24	"	"	1.21	1.72
	1.0		1.90	1.35	"	"	1.35	1.90

Curves of E_{in} vs. E_{out} are shown in Fig. 16.

APPENDIX II

TABLE II: Test for Linearity of the Direct-Coupled Amplifiers.

Input Volts (d-c)

	Cosine Channel		Sine Channel	
±	+	-	+	-
0.0	0	0	0	0
0.1	7	10	10	10
0.2	16	20	18	18
0.3	27	28	26	28
0.4	37	38	34	38
0.5	46	47	41	46
0.6	54	55	48	56
0.7	61	64	56	64
0.8	67	71	62	70
0.9	71	80	68	74
1.0	75	87	74	79
1.1	79	93	79	81
1.2	83	98	83	84
1.3	86	100	87	88
1.4	89	105	90	91
1.5	92	110	93	94
1.6	94	112	95	96
1.7	96	115	97	98
1.8	98	118	98	99
1.9	99	120	99	100
2.0	100	122	100	101

Curves of E_{in} vs. E_{out} are shown in Fig. 17.

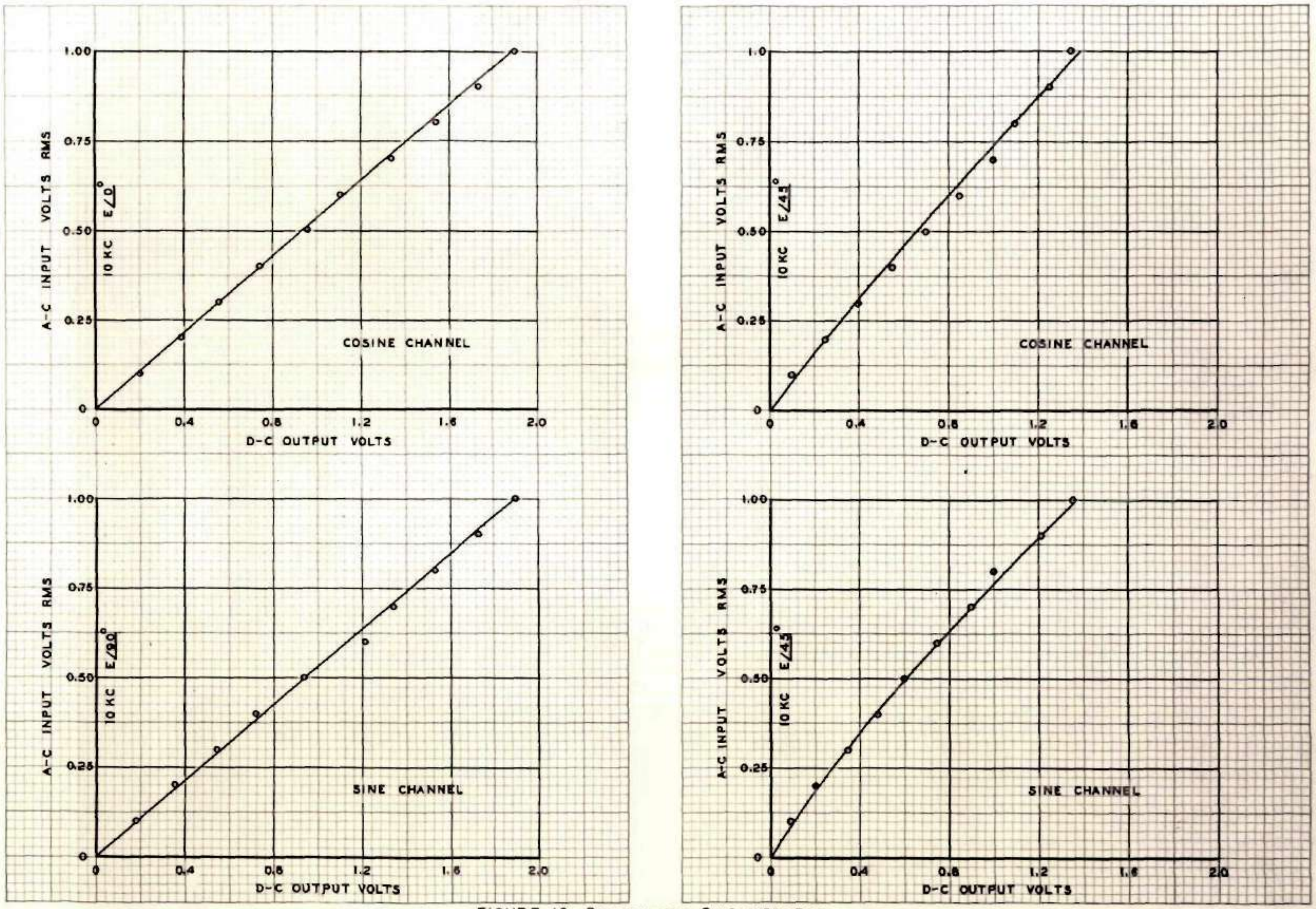
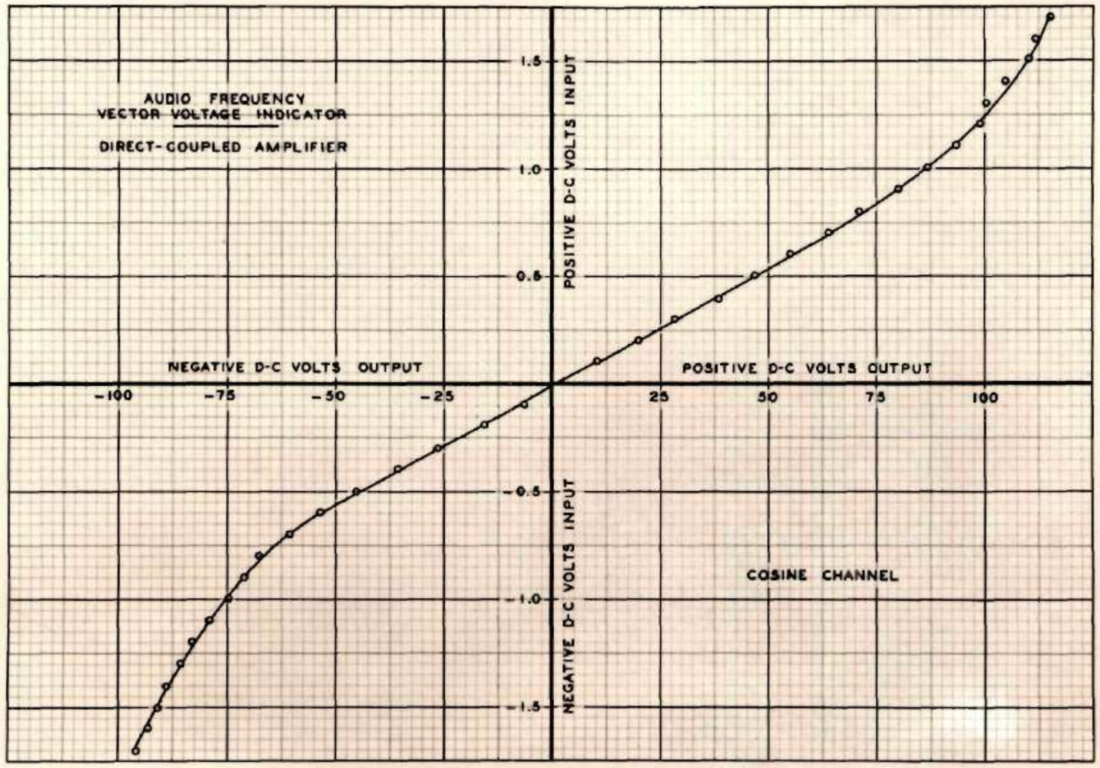


FIGURE 16. DEMODULATOR CHARACTERISTICS



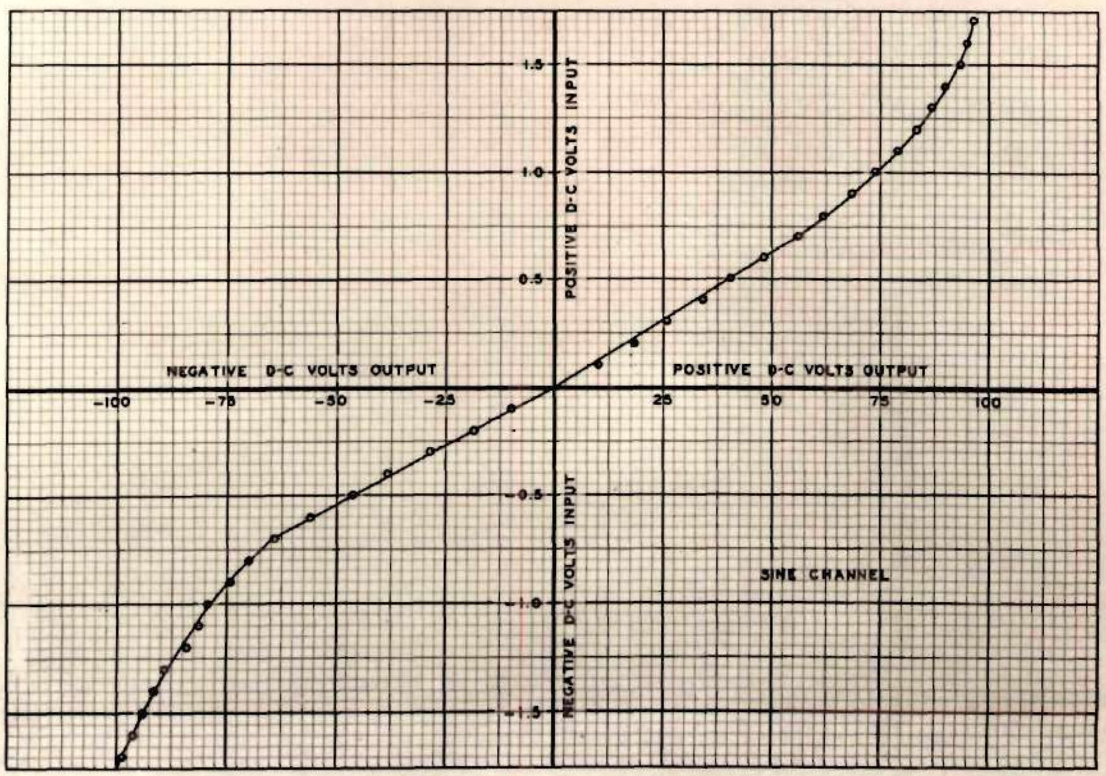


FIGURE 17. AMPLIFIER CHARACTERISTICS

APPENDIX III

CHASSIS LAYOUT

A composite photograph of the adaptor and its power supply is shown in Fig. 18.

Arrangement of the components in both the amplifier and the main chassis is shown in Fig. 19.

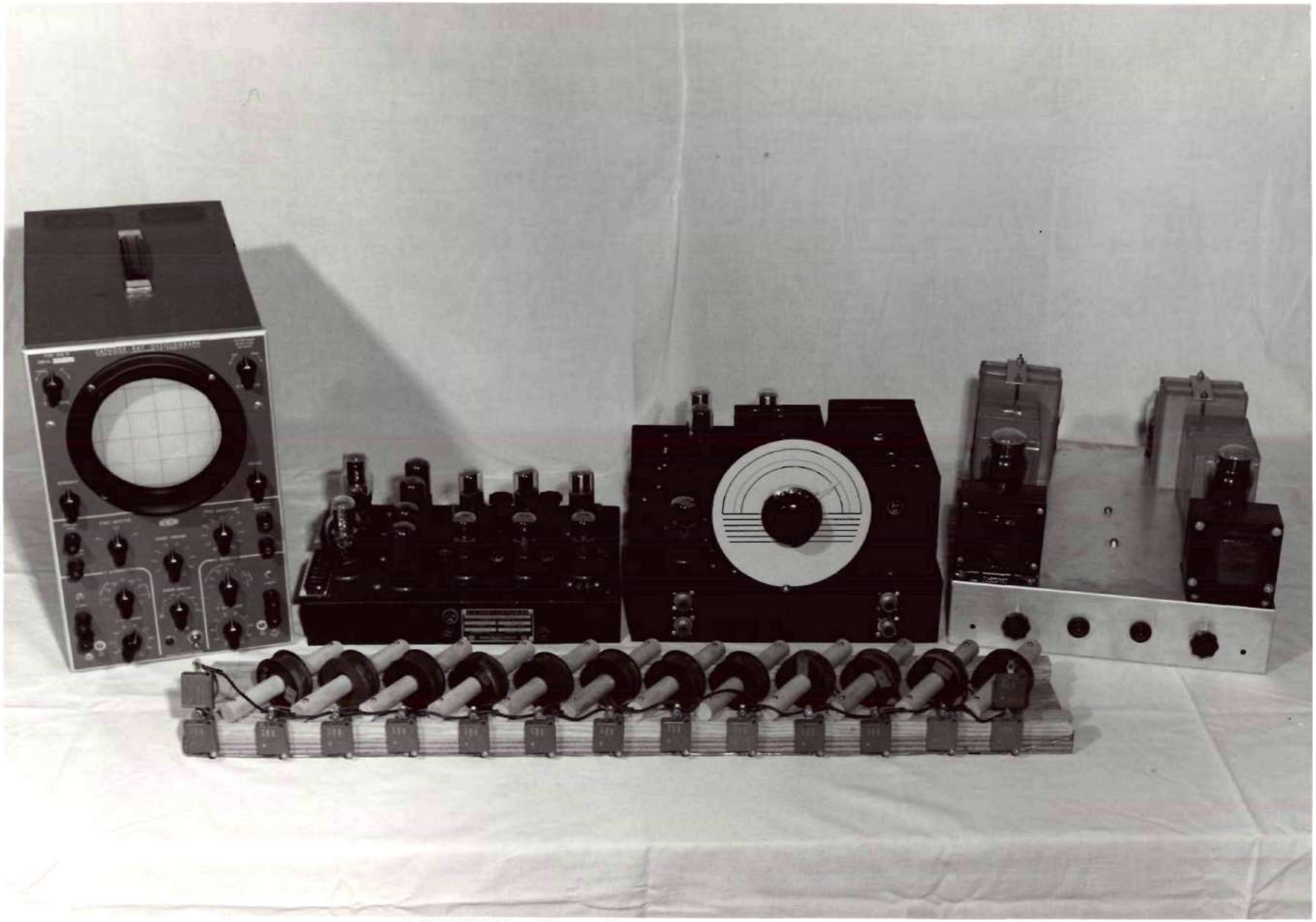
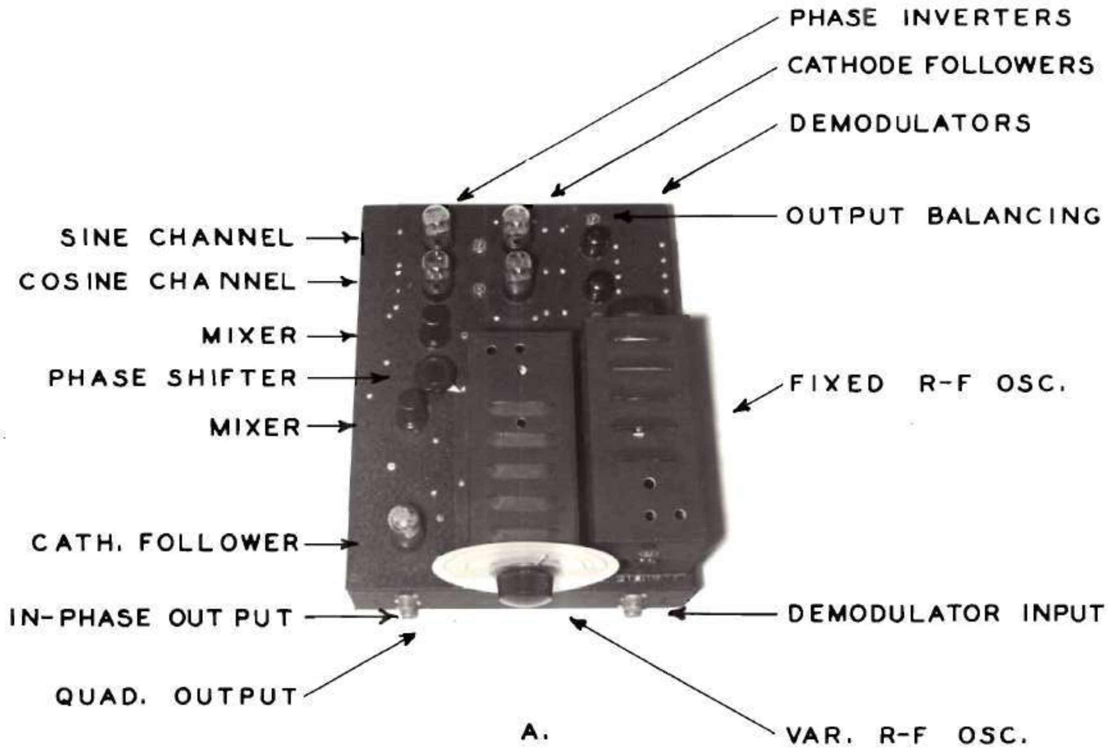
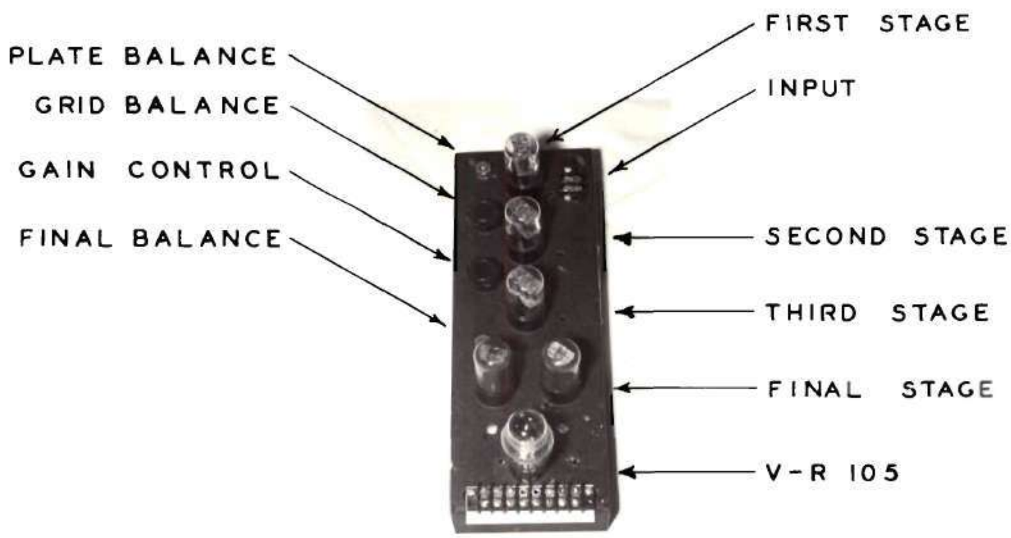


FIGURE 18. COMPLETE ADAPTOR



A.

MAIN CHASSIS



B.

D-C AMPLIFIER

FIGURE 19. CHASSIS LAYOUT

APPENDIX IV

POWER SUPPLY FOR THE ADAPTOR

The power requirements for the adaptor are as follows:

Stage/ Tubes	Plate Current Amperes	Filament Current Amperes
Heterodyne Oscillator		
4--6AG7	0.160	2.60
2--6SA7	0.024	0.60
1--6SN7	0.020	0.60
Demodulator Assembly		
4--6SN7	0.080	2.40
2--6H6	0.000	0.60
Amplifiers		
6--6SN7	0.120	3.60
4--6V6	0.200	1.80
Totals:	0.604	12.20

Two power supplies were constructed on a 17 x 13 x 3 inch chassis according to the schematic diagram shown in Fig. 20.

The voltage supplied to the heterodyne oscillator is regulated by voltage-regulator tubes. The oscillators are supplied by 210 volts and the mixer circuits are supplied by 300 volts.

The amplifiers are supplied from laboratory equipment.

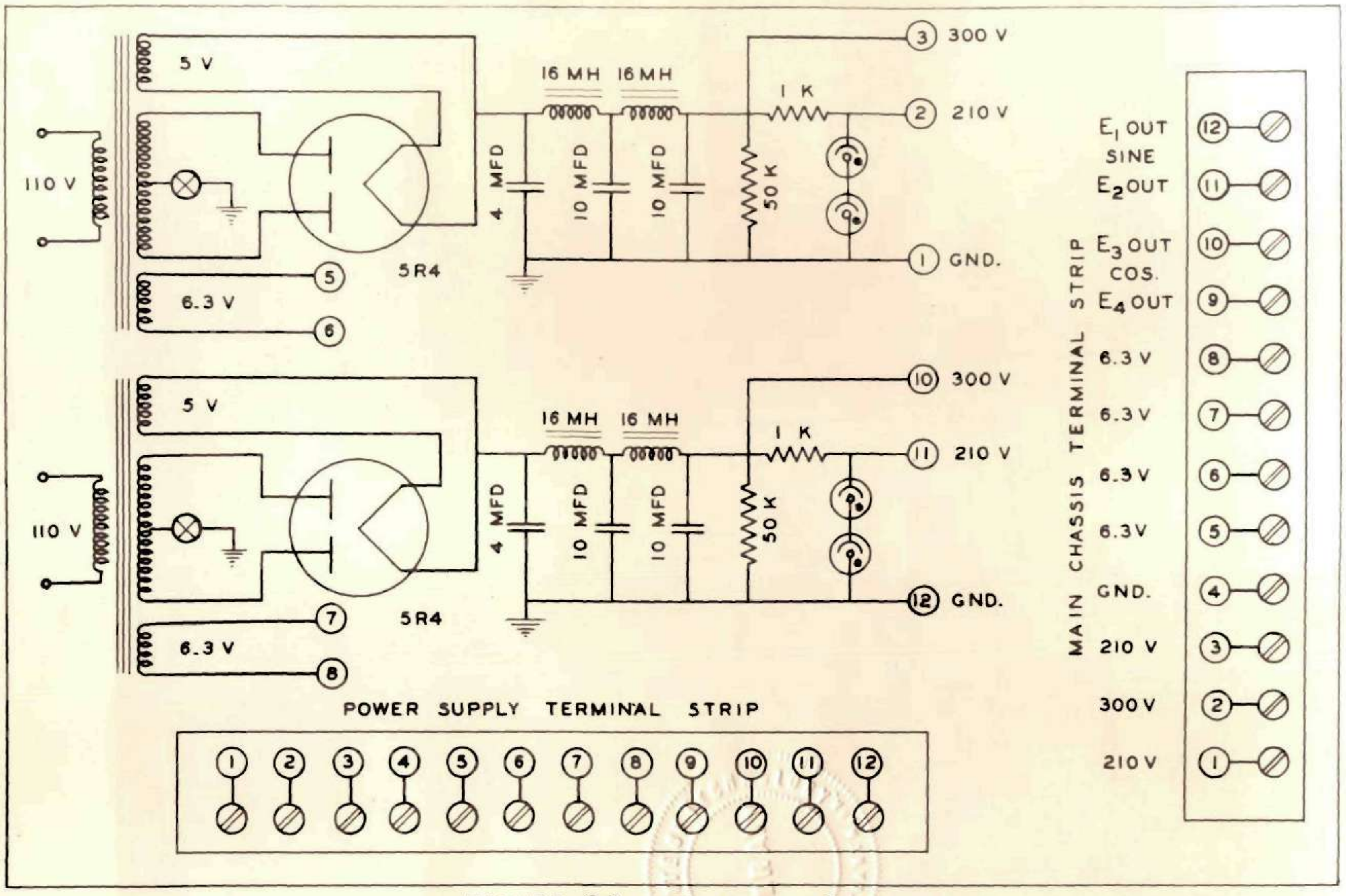


FIGURE 20. POWER SUPPLY