A HETERODYNE-TYPE BRIDGE DETECTOR FOR ULTRASONIC FREQUENCIES

by

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Abstract

This thesis describes a bridge detector designed to operate between 2 kilocycles and 100 kilocycles per second.

The detector is an electronic device which beats the incoming signal with a local oscillator signal. This effect is due to the heterodyne principle. The local oscillator is of the Wein Bridge type which utilizes a Wein Bridge C-R feedback circuit to produce stable harmonic-free oscillations.

The beat-frequency signal is filtered by a two-section tuned amplifier tuned to 1000 cycles per second. The amplifiers are of the parallel-T feedback type which use a C-R degenerative feedback network to produce selective amplification. The output signal is detected by means of earphones.

Good sensitivity is obtained by the use of high gain throughout the circuit.

Several capacitors were measured at frequencies up to 20 kilocycles on a Schering bridge using the instrument. Balances of five significant figures were easily obtained.

Section A

Introduction

I. Foreword

This thesis is a description of a bridge detector designed to operate in the region of the frequency spectrum between 2 kilocycles and 100 kilocycles. The detector amplifies the signal obtained from the bridge and converts it to a 1 kilocycle signal through a heterodyning process, thus enabling the bridge to be balanced by the use of earphones.

II. Some Reasons for the Operation of a Bridge at High Frequencies

There are many reasons why it might be desirable to operate a bridge at an ultrasonic frequency.

The effective resistance, capacitance, and inductance of circuit elements is a function of frequency, and therefore these quantities should be measured at the frequency at which they are to operate.[1,2]

In the investigation of such frequency-dependent effects, it would be advantageous to be able to make measurements at any frequency.

The frequency range chosen for this instrument is particularly important since the carrier frequencies used in telecommunication work lie in this region of the frequency spectrum. Therefore, it is important that ease of measurement of electrical quantities be achieved in this frequency range.

III. Available detectors for bridges

The simplest way to examine the types of detectors available for

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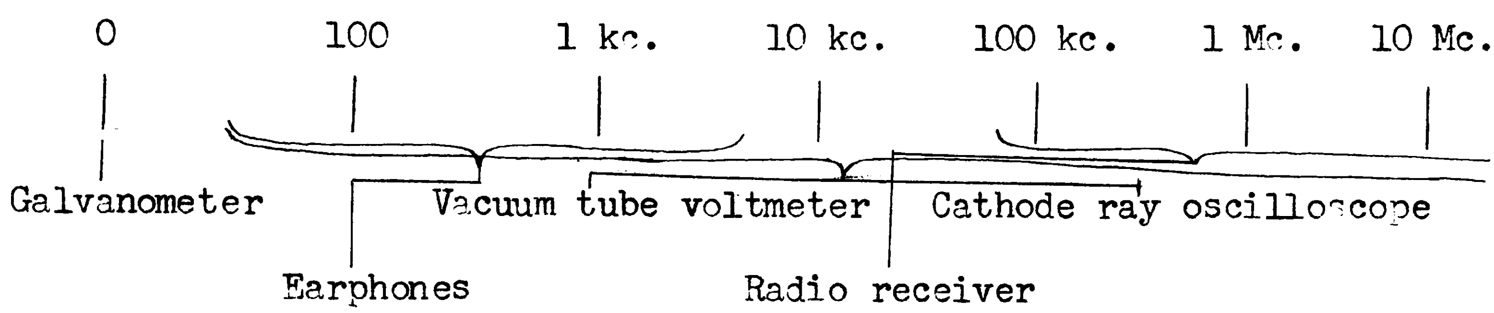
1 Harnwell, G.P. "Principles of Electricity and Electromagnetism" McGraw-Hill, p. 433

2 Morecroft, J.H. "Principles of Radio Communication", J. Wiley & Sons, 1944, p. 184

bridge measurements at any given frequency is to construct a chart of the frequency spectrum and available detectors as has been done in fig. 1.

fig. 1 Available Detectors Over the Frequency Spectrum

frequency in cycles per second



IV. The Evaluation of Detectors

In the evaluation of the quality of a detector the following considerations are of importance.

1. sensitivity
2. range of signals which give a null reading
3. frequency discrimination
4. cost and size

In the examination of the detectors on the above chart, some recommend themselves in all the points mentioned above, others do not.

V. An Investigation of Available Detectors

For example, let us consider the earphones. The sensitivity of the earphones may be increased almost indefinitely through the use of suitable amplifiers. By the same considerations, the range of signals which give a zero reading can be narrowed very greatly through amplification of the signals.

The operator may train his ear to distinguish very accurately between fundamental and harmonics in the signal. In addition, in the low frequency region, low pass filters may be used to partially eliminate harmonic content in the output of the bridge. Thus the bridge could easily be balanced at any one frequency, and therefore the frequency discrimination of the detector may be said to be good.

The last requirement, that of reasonable size and cost is well met by the earphones and accompanying amplifiers.

In the region above 100 kc. a radio receiver may be used as a detector. Since a radio is essentially a detector of low voltage signals, and is tuneable to any desired frequency, it is ideally suited to bridge signal detection. Good radio receivers are available at a reasonable cost.

However, upon examining the chart in fig. 1, it will be seen that there is a frequency range where only a limited number of types of detectors are available. The sensitivity of the human ear declines rapidly above 2 or 3 kc., while radio sets are not made which operate much below 100 kc. Thus between 2 kc. and 100 kc. there are only vacuum tube volt-meters and cathode ray oscilloscopes available as detectors. Both of these fail to meet fully all the qualifications set up for a good detector.

The sensitivity of a vacuum tube voltmeter is a function of the quality of the instrument. High quality instruments usually have low voltage scales where signal differences of 20×10^{-6} volts may be detected.

The vacuum tube voltmeter has the major disadvantage that it reads the total voltage of all the frequencies together which are applied to it. Filters may be used to partially eliminate unwanted frequencies, but complete elimination of harmonics in this way is difficult. A point will eventually be reached where the harmonic signals will mask the fundamental signal and prevent further sensitivity of reading from being obtained. Furthermore, if readings are to be made over a range of frequencies, the elimination of harmonics by the use of filters necessitates a filter with a variable cut-off frequency. This is a clumsy and difficult piece of apparatus to build and operate.

A vacuum tube voltmeter is an expensive instrument, and its cost increases as greater sensitivity is required.

In the cathode ray oscilloscope, the cathode ray tube is inherently an insensitive method of detecting voltages, due to the high voltage necessary on the deflecting plates of the tube to cause unit deflection of the beam. Thus great amplification of the signal, with all its accompanying difficulties must be employed to obtain sensitivity.

The cathode ray oscilloscope suffers from the same disadvantage as the vacuum tube voltmeter in that it detects all the applied frequencies. However, by the use of lissajous figures,- using the voltage input to the bridge as the horizontal sweep voltage on the cathode ray tube,- it is possible to distinguish quite well between fundamental and harmonic frequencies.

Here again the cost of the detecting instrument may become quite large as good sensitivity is demanded of it.

From the above considerations, it was felt that a detecting instrument to operate between 2 kc. and 100 kc. which would combine high sensitivity, good frequency discrimination, ease of operation, and low cost would be a decided asset in Electrical Measurement work.

Section B

The Bridge Detector

I. Form of the Proposed Instrument

It was proposed to build an electronic device to act as a bridge detector between 2 kc. and 100 kc. meeting the requirements set up in Section A.

In order to give maximum ease of operation of the instrument, as well as good sensitivity, it was decided to use earphones as the ultimate detecting device. Since the human ear has its maximum sensitivity in the region of 1000 cps, an output signal of this frequency was desired from the instrument.

The heterodyne principle was employed. The incoming signal is beat with the signal from a variable-frequency beat oscillator. The resulting signal is applied to a tuned amplifier which is tuned to 1000 cps. Thus, filtering action in the instrument is supplied by a single fixed tuned amplifier, rather than a variable frequency filter. This means that sharp filtering action may be obtained with a minimum of circuitry. At the same time, the tuneable portion of the circuit is transferred to the local oscillator circuit where wide variations of frequency are more easily obtained.

By these means the operator of the instrument is enabled to select and balance the fundamental of the applied signal, excluding all harmonics, with ease and accuracy.

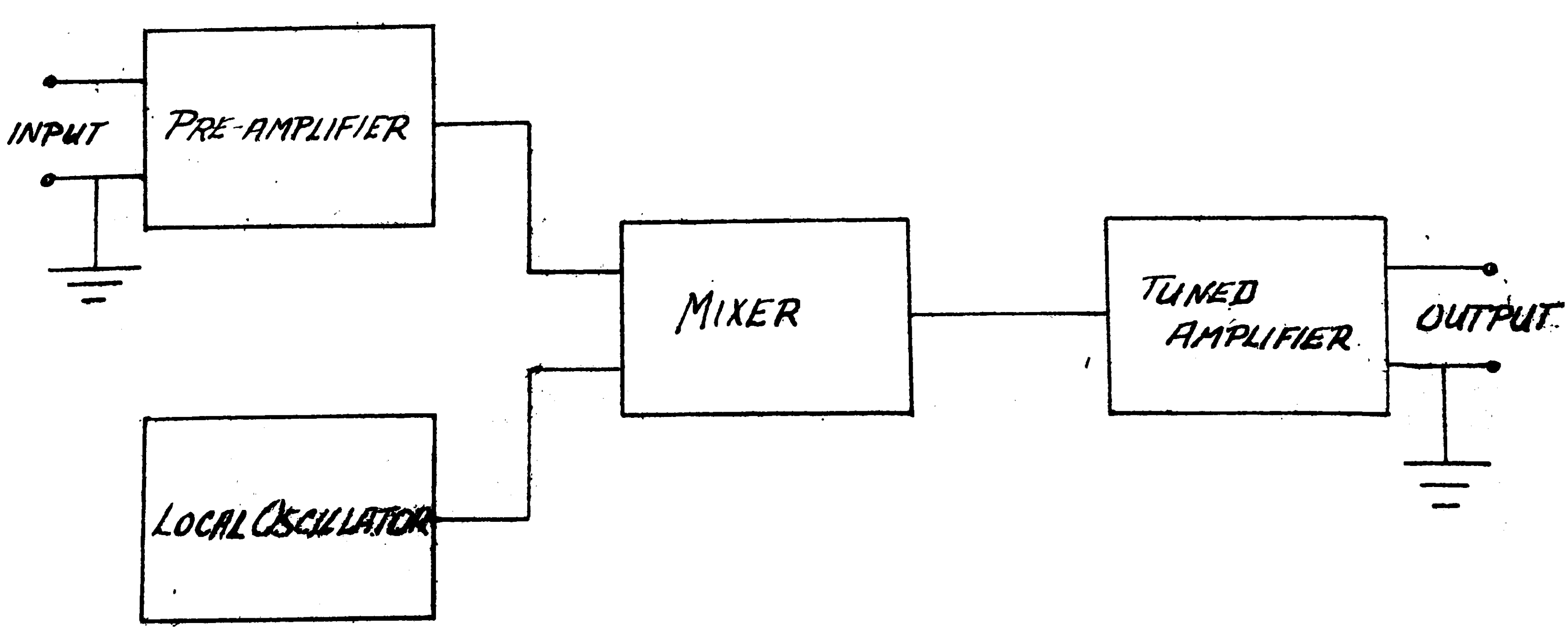
In order to obtain high sensitivity, high amplification of the incoming signal is employed. To do this a preamplifier is incorporated into the circuit.

In order to conserve space, miniature tubes are employed wherever possible, and all tuned circuits employed are of the R-C rather than the L-C type. This has the added advantage that there are no stray magnetic fields generated by inductors in the circuit which might conceivably affect the operation of some bridges employing inductances. The absence of coils also greatly reduces pickup due to stray fields from the bridge.

These methods enabled the entire detector including the power supply to be built on an 8" X 12" chassis. The cost of the apparatus was kept as low as possible by using components with wide tolerances wherever possible. (The entire apparatus could be built for approximately $50.00).

A block diagram of the apparatus is shown in fig. 2.

fig 2 BLOCK DIAGRAM OF THE APPARATUS



II. The Heterodyne Principle

It is desired to convert the incoming signal, which lies in the ultrasonic region of the frequency spectrum, into a 1000 cps. frequency which the ear will readily detect. To do this the incoming frequency ω_a will be beat with a local oscillator frequency ω_1.

The plate current of a vacuum tube may be represented by a power series[1]

$$i_p = i_o + K_1 e_g + K_2 e_g^2 \ldots\ldots \qquad \text{II-1}$$

where e_g is the applied grid signal

i_p is the plate current

K's are constants of the tube.

It is known that $|K_s|$ decreases as s increases.

If both ω_a and ω_1 are applied to the grid (or grids) of the tube

$$e_g = a \text{ Cos } (\omega_a t) + b \text{ Cos } (\omega_1 t).$$

Taking the first three terms of II-1[2]

$$i_p = i_o + K_1 (a \text{ Cos } \omega_a t + b \text{ Cos } \omega_1 t)$$

$$+ K_2 (a^2/2)(1 + \text{Cos } 2\omega_a t) + (b^2/2)(1 + \text{Cos } 2\omega_1 t)$$

$$+ ab \text{ Cos } (\omega_a - \omega_1)t + ab \text{ Cos } (\omega_a + \omega_1)t \qquad \text{II-2}$$

Thus the frequencies present due to the first two terms of the expansion are

$$0,\ \omega_a,\ \omega_1,\ 2\omega_a,\ 2\omega_1,\ (\omega_a + \omega_1),\ |\omega_a - \omega_1|\ .$$

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1 Appleton, E.V. "Thermionic Vacuum Tubes" Methuen & Co., 1932, p.81

2 Morecroft, J.H. "Principles of Radio Communication", J. Wiley & Son, 1944 p. 585.

If the local oscillator is tuned in such a manner that $|\omega_a - \omega_1|$ = 1000 cps, then all of the other frequencies present will be in the ultrasonic region and be inaudible (assuming ω_1 is in the ultrasonic region). Furthermore, if the output of this tube is filtered with a sharp band-pass filter in some manner, (in this case a tuned amplifier tuned to 1000 cps is used) the only frequency on the output side of the filter will be $|\omega_a - \omega_1|$ (= 1000 cps).

The inclusion of further terms of the series II-1 will produce more different frequencies[1], but most of them will be outside of the audible range. All these signals due to higher terms will be small, since K_s decreases in size as s increases. Therefore, those extra signals which lie in the audible range will be eliminated by the following frequency filter.

Assume that the input signal contains harmonics $n\omega_a$ where n is any interger, and that the local oscillator is tuned so that $|\omega_a-\omega_1|$=1000 cps.

The output of the mixer tube will now contain frequencies

$$0,\ \omega_1,\ \sum_{n=1}^{\infty} n\omega_a,\ \sum_{n=1}^{\infty} |n\omega_a \pm \omega_1|$$

Of these the following amplifier will only pass $|\omega_a - \omega_1|$ (= 1000 cps).

Assume that even the local oscillator signal contains harmonics $m\omega_1$ (where m is any interger), and that the local oscillator is tuned so that $|\omega_a - \omega_1|$ = 1000 cps. Then the output of the mixer tube will contain frequencies

$$0,\ \sum_{m=1}^{\infty} m\omega_1,\ \sum_{n=1}^{\infty} n\omega_a,\ \sum_{m=1}^{\infty}\sum_{n=1}^{\infty} |n\omega_a \pm m\omega_1|.$$

1 Appleton, E.V. "Thermionic Vacuum Tubes", Methuen & Co., 1932, p.81.

Of these the following tuned amplifier will still pass $|\omega_a - \omega_1|$ (=1000 cps). However it is also possible that $|n\omega_a - m\omega_1|$ = 1000 cps at this point of tuning. This would introduce unwanted frequencies in the measurement. This condition is extremely unlikely for if

$$|\omega_a - \omega_1| = 1000 \text{ cps}$$

and $$|n\omega_a - m\omega_1| = 1000 \text{ cps}$$

then $$n = \frac{1000 + m\omega_1}{1000 + \omega_1} .$$

Thus this condition is possible only for a very few values of ω_1.

This unwanted condition may be entirely eliminated if there are no harmonics in the local oscillator output.

Thus the heterodyne process can be employed to produce an audible note from a high frequency signal in such a manner that only the fundamental of the input signal is detected.

It is customary to use a multigrid tube as a mixer tube, thus enabling the impression of signals ω_a and ω_1 on the tube output while still isolating the signal inputs from each other.

III The Wein Bridge Oscillator

1. The Characteristics of the Desired Oscillator

An oscillator was required which would have a continuously variable frequency range from 1 kc. to 100 kc. It was necessary that the output be as pure a sine-wave as possible, since harmonics in the oscillator would make tuning the apparatus more difficult as well as introducing possible unwanted frequencies in the output. In keeping with the rest of the apparatus, it was desired to keep the physical size of the components used to a minimum.

The Wein bridge oscillator was chosen because it best fulfilled the above conditions.

2. The Wein Bridge

Consider the bridge circuit in fig. 3. The balance conditions for the bridge are[1]

$$\frac{R_2}{R_1} = \frac{R_s}{R_p} + \frac{C_p}{C_s} \qquad \text{III-1}$$

$$\omega_o^2\, C_s C_p R_s R_p = 1 \qquad \text{III-2}$$

It is customary to operate this bridge with $R_s = R_p = R$

$C_s = C_p = C.$

Then III-1 becomes

$$\frac{R_2}{R_1} = 2 \qquad \text{III-3}$$

III-2 becomes

$$\omega_o CR = 1 \qquad \text{III-4}$$

3. The Wein Bridge Oscillator

In fig. 4 the components R_1, R_2, C_s, C_p, R_s, R_p, represent a Wein Bridge which is used to cause regenerative feedback from tube 2 to tube 1 of one particular frequency, ω_o, and degenerative feedback of all others. The frequency at which the circuit oscillates is determined by the bridge network in the following manner. A negative feedback voltage is fed to the cathode of the tube by the resistive network R_1, R_2. This voltage is independent of frequency. The network composed of $C_s R_s C_p R_p$ applies a positive feedback voltage to the grid which is a maximum at the frequency

1 Terman, F.E. "Radio Engineers Handbook", McGraw-Hill, 1943, p. 905.

fig 3

THE WIEN BRIDGE

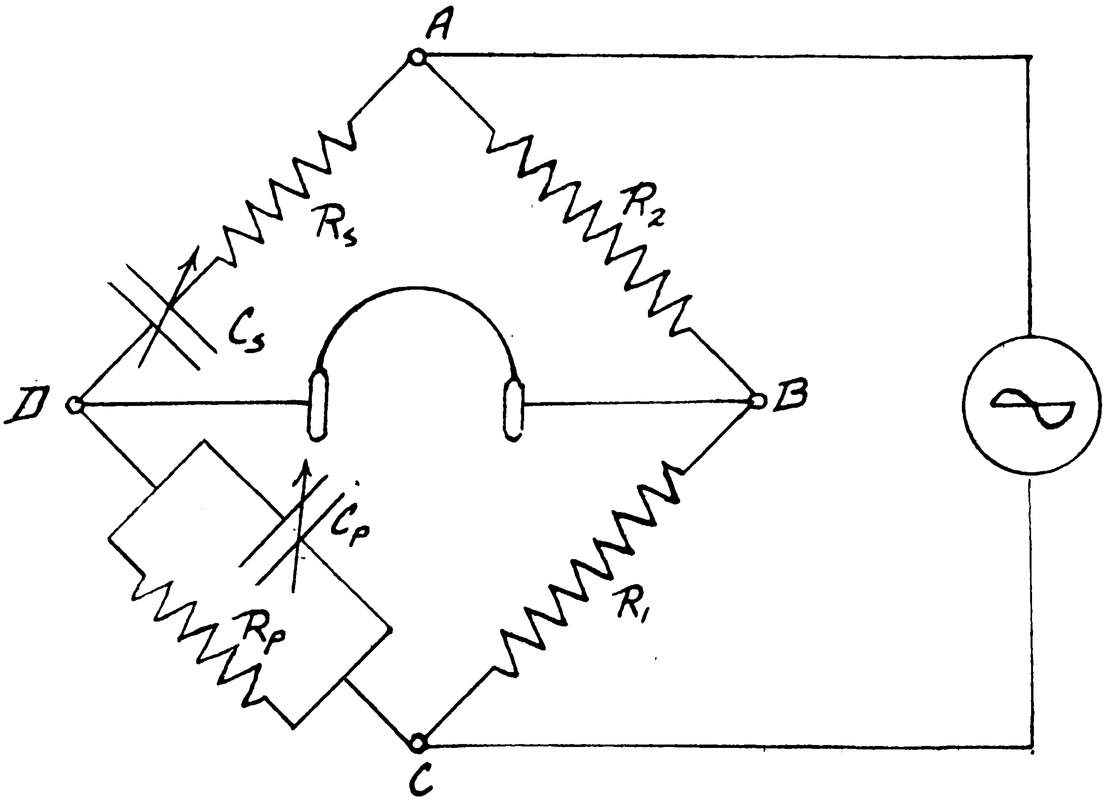
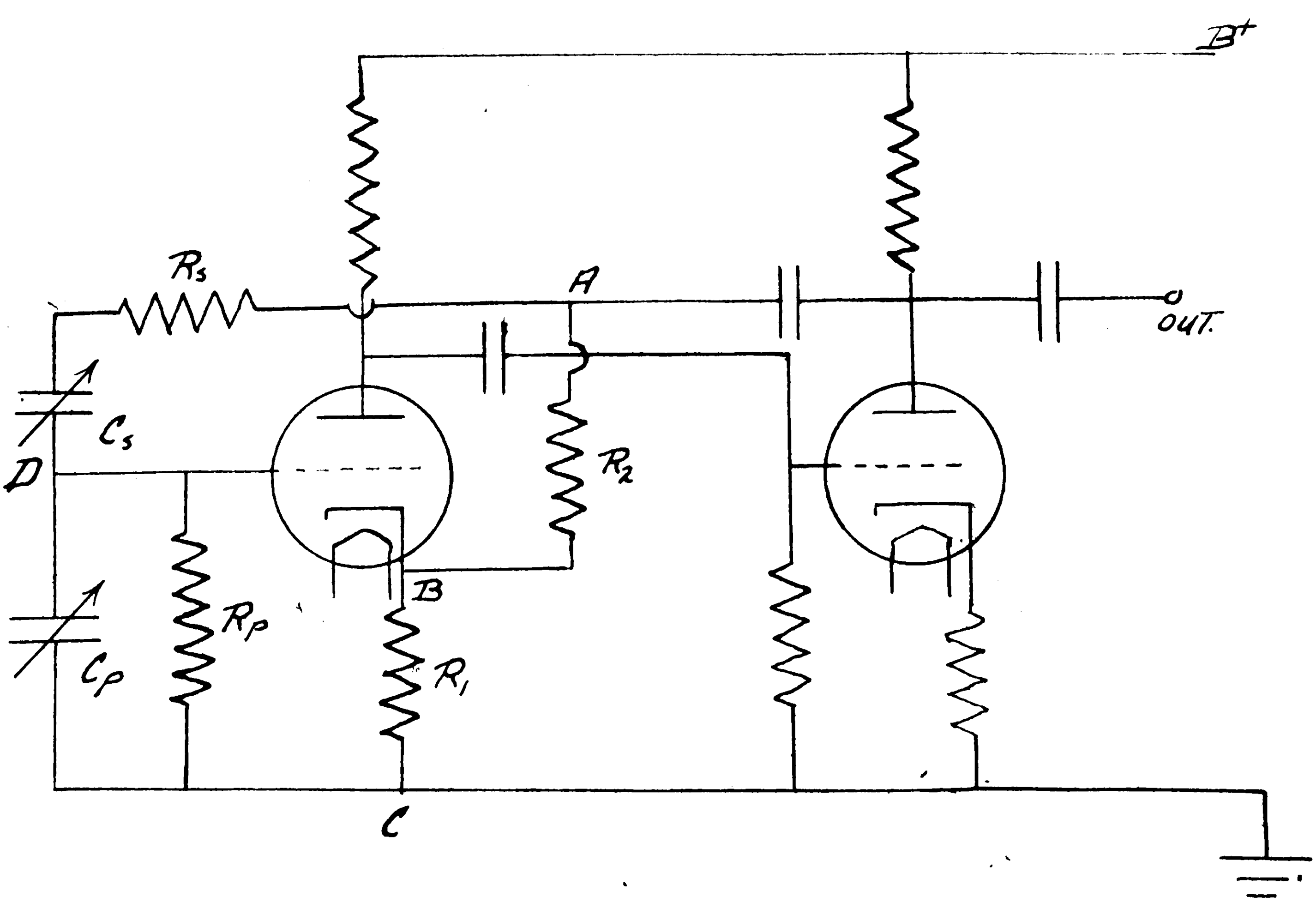


fig 4

THE WIEN BRIDGE OSCILLATOR



NOTE:- TRIODES ARE SHOWN FOR SIMPLICITY. ORDINARILY, PENTODES ARE USED FOR HIGHER GAIN.

ω_o given by the bridge balance conditions (III-4). If the bridge is correctly balanced the positive feedback is just sufficient to overcome the negative feedback at the frequency ω_o and regenerative oscillations take place.[1,2] At frequencies not near ω_o, the positive feedback is small compared to the negative feedback,- thus keeping the harmonic content of the output to very small proportions.[3]

A more detailed examination of the circuit will show that while this is substantially what happens, the oscillator actually oscillates at a frequency slightly below that given by $\omega_o^2 = 1/C_pC_sR_pR_s$. (See Appendix I).

4. Automatic Amplitude Control[2]

In order that the circuit may remain stable from the viewpoint of amplitude of output oscillations, the bridge must automatically unbalance itself in such a way as to enhance the oscillations if they become too weak and to damp the oscillations if they become too great. This is done by varying the amount of negative feedback by varying the ratio R_1/R_2. This has the effect of varying the amplification, and is done by placing a thermally sensitive resistor at either R_1 or R_2.

Assume that R_1 is the thermally sensitive resistor. If the oscillations are too strong, too large a current flows in R_1, and therefore R_1 should increase, thus increasing the negative feedback voltage and damping the oscillations. If the oscillations are too

1 Chance, B., Hughes, V., MacNichol, E.F., Sayre, D., and Williams, F.C. "Waveforms", (M.I.T. Rad.Lab.Ser.), McGraw-Hill 1949, p. 118.
2 Terman, F.E., Buss, R.R., Hewlett, W.R., Cahill, F.C. "Some Applications of Negative Feedback with Particular Reference to Laboratory Equipment" Proc. Inst. Radio Engrs., N.Y. 27, 10, 649,
3 Scott, H.H. "A New Type of Selective Circuit and Some Applications" Proc. Inst. Radio Engrs., N.Y. 26, 2, 226.

weak, too small a current flows in R_1 and therefore R_1 should decrease in order to enhance the oscillations. This effect may be achieved by using an incandescent light bulb for R_1 - for as I increases in the lamp - R increases, thus producing the desired result of keeping the circuit in equilibrium.

5. Characteristics of the Wein Bridge Oscillator

The oscillator discussed above has easily variable frequency of good stability - the stability being governed by the bridge network.[1] As mentioned before, low frequencies may be obtained with small sized components, because of the use of the R-C network. In addition, there are no inductances present, and therefore no stray magnetic fields are generated to interfere with bridge operation. Similarly, the absence of coils reduces pickup from fields caused by outside circuits. The bridge network automatically discourages harmonics, so that harmonic content should be low if the amplifier produces little distortion.[1,2] Therefore, high gain may be employed provided the amplifier operates in the distortion free region of its characteristic. The resulting high output signal is advantageous, as may be seen from equation 2 of Section B-II on the heterodyne principle, which shows that the amplitude of the desired signal $|\omega_a - \omega_1|$ is proportional to ab where b is the amplitude of the local oscillator signal. However, a point will be reached where further increase of b will result in a decrease of the amplitude of $|\omega_a - \omega_1|$. This comes about through a consideration of further terms of the output series II-1.[3]

1 Chance, et al (See page 12)
2 Scott (See page 12, #3)
3 Appleton, E.V. "Thermionic Vacuum Tubes", Methuen & Co. 1932 p. 81.

IV. The Tuned Amplifier

1. Negative feedback amplifiers

It was decided to use a tuned amplifier of the negative feedback type to obtain frequency selectivity.

Consider the amplifier circuit in fig. 5. Without feedback

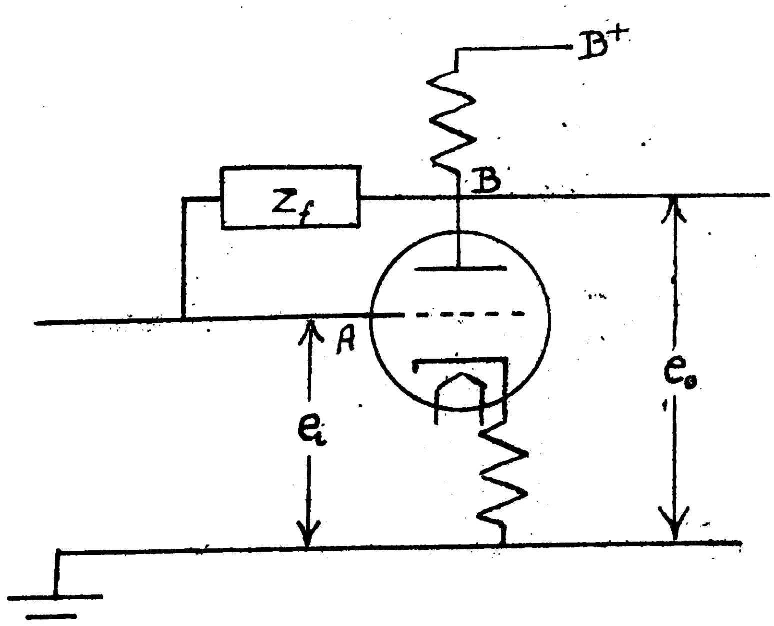
$$e_o = Ae$$

where A = amplification

e = applied signal.

fig 5

FEEDBACK AMPLIFIER



Suppose a fraction β of e_o is fed back from B to A in such a way that the fed-back portion βe_o tends to cancel e, i.e. βe_o is 180^o out of phase with e, and therefore β has a (-) sign. Then[1]

$$e_i = e - \beta e_o$$

$$e_i = - \frac{e_o}{A}$$

$$\therefore \quad e_o = - \frac{Ae}{1-A\beta} \qquad \text{IV-1}$$

1 Millman, J., Seely, S. "Electronics", McGraw-Hill, 1941, p. 607.

If a circuit can be devised with an impedance Z_f such that at one frequency, ω_o, $Z_f = \infty$ and at neighbouring frequencies Z_f has a finite value, then at ω_o, $\beta = 0$, and therefore $e_o = -Ae$ from IV-1 whereas at neighbouring frequencies $\beta \neq 0$ and thus $e_o = -Ae/(1-\beta A)$ which is smaller than e_o at ω_o. A frequency selective amplifier which amplifies frequency ω_o to a greater extent than all others has thus been devised.

Since the voltage at B is already 180° out of phase with that of A, the desired network should ideally have a zero phase shift characteristic. However, if the phase shift caused by the network is small at frequencies not near ω_o, the input voltage, e will still tend to be cancelled by the fed-back voltage βe_o, so as to produce attenuation of the output signal at these frequencies. Near ω_o, the quantity βe_o is small, so that as far as equation IV-1 is concerned, the phase shift caused by Z_f is of little importance, from the viewpoint of frequency selection. The possibility of regenerative feedback due to too large a phase shift in Z_f will be dealt with later.

2. The Parallel-T network

Consider the general parallel-T network in fig. 6. This network has an infinite impedance when the following relations are satisfied.[1,2]

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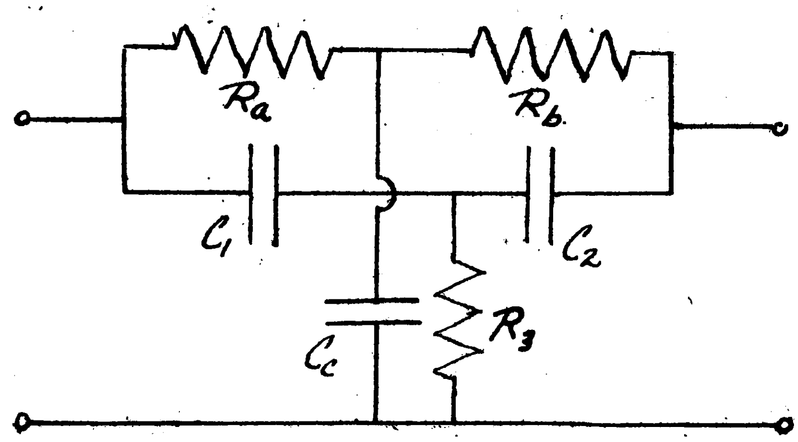
1 Hillan, A.B. "The Parallel-T Bridge Amplifier", J. Inst. Elec. Engrs, Pt. III 94,27,42.

2 Valley, G.E., Wallman, H. "Vacuum Tube Amplifiers" (M.I.T. Rad.Lab.Ser.) McGraw-Hill, 1948, p. 387.

$$\omega_o^2 = \frac{1}{C_1 C_2 R_3 (R_a + R_b)} \qquad \text{IV-2}$$

$$\omega_o^2 = \frac{C_1 + C_2}{C_1 C_2 C_c R_a R_b} \qquad \text{IV-3}$$

fig. 6 GENERAL PARALLEL-T NETWORK.



Thus, in order to achieve balance at any frequency, ω_d, two components, R_3 and C_c must be adjusted.

It is common practice in the use of this network to make

$$R_a = R_b = R \qquad \text{IV-4}$$

$$C_1 = C_2 = C \qquad \text{IV-5}$$

This is the form used in this apparatus, and hence for simplicity, these equalities will be used in all succeeding discussion.

Eliminating ω_o in IV-2 and IV-3

$$\frac{R_a R_b}{R_3(R_a + R_b)} = \frac{C_1 + C_2}{C_c} = n$$

$$\text{or} \quad R_3 = \frac{R}{2n} \quad C_c = \frac{2C}{n} \qquad \text{IV-6}$$

For the case where the load impedance of the network is infinite, the network may be treated as a potential divider. If β is the transmission parameter of the network, then[1]

$$\beta = \frac{(\omega/\omega_o - \omega_o/\omega)}{(\omega/\omega_o - \omega_o/\omega) + 2j\ (n+1)/\sqrt{n}} \qquad \text{IV-7}$$

To find the value of n which would give the sharpest response characteristic

deg? of p?

$$\left.\frac{d\beta}{dp}\right|_{p=1} = \frac{\sqrt{n}}{n+1}$$

$$\frac{d}{dn}\left(\left.\frac{d\beta}{dp}\right|_{p=1}\right) = 1 - n = 0$$

where $p = \omega/\omega_o$

Therefore, $n = 1$ is the optimum value. However, the slope of $\left.\frac{d\beta}{dp}\right|_{p=1}$ at $n = 1$ is a slowly varying function, and thus if n lies in the neighbourhood of 1, the response characteristic will still retain its sharp peak.[1]

With reference to the method of balancing the network by varying R_3 and C_c, suggested by IV-2 and IV-3, it will be seen that such a variation will also have an effect on the shape of the response characteristic. If the value of n set by R_3 and C_c is not near 1, then the response peak of the network will be relatively broad.

It will be seen that equation IV-7 is logarithmically symmetric in ω about ω_o; that is, the interchange of ω_o/ω for ω/ω_o (where $\omega_o/\omega = \omega/\omega_o$) makes no difference. Thus the response characteristic of the network is symmetrical about ω_o when plotted on a logarithmic base.

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1 Valley, G.E., Wallman, H. "Vacuum Tube Amplifiers" (M.I.T. Rad.Lab.Ser.) McGraw-Hill, 1948, p. 387.

Also, from equation IV-7, $\beta \rightarrow 1$ as $\omega \rightarrow 0$ or $\omega \rightarrow \infty$

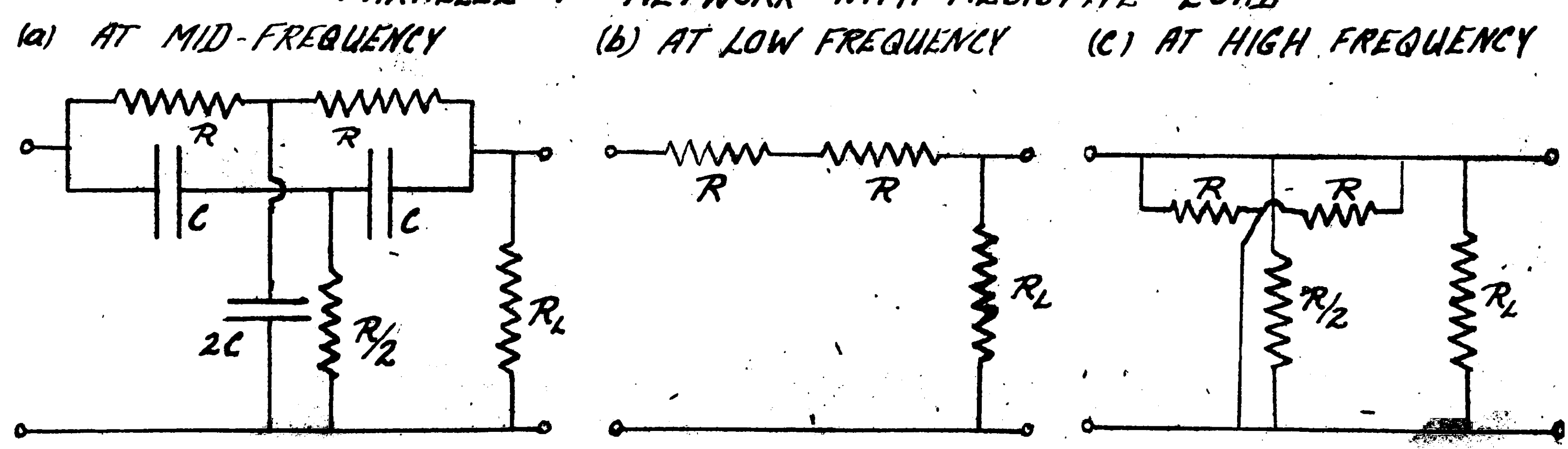
However, the network will always be operating into a finite impedance. If this load impedance is a pure resistance, R_L as in fig. 7(a), the circuit no longer has a symmetrical selectivity curve. For the circuit in fig. 7(b) is the low frequency equivalent of the circuit in 7(a) and in this circuit

$$\beta = R_L/(R_L + 2R) \neq 1$$

The circuit in fig. 7(c) is the high frequency equivalent of fig. 7(a) and in this circuit $\beta = 1$.

fig 7

PARALLEL-T NETWORK WITH RESISTIVE LOAD



The symmetry may be restored to the selectivity curve by using a load composed of a resistance R_l and a capacitance C_l in series,[1] where

$$R_L = mR \qquad \text{IV-8}$$

$$C_L = C/m \qquad \text{IV-9}$$

1 Hillan, A.B. "The Parallel-T Bridge Amplifier", J. Inst. Elec. Engrs, Pt. III, 94, 27, 42.

It is found that as m gets larger, the selectivity curve of the network gets sharper. However, after $m > 5$, the additional improvement obtained by increasing m further is small.[1]

In IV-7, if $n = 1$ and $p = \omega/\omega_o$, then

$$\beta = \frac{1}{1+4jp/(p^2-1)} \qquad \text{IV-10}$$

If this equation is plotted in polar form

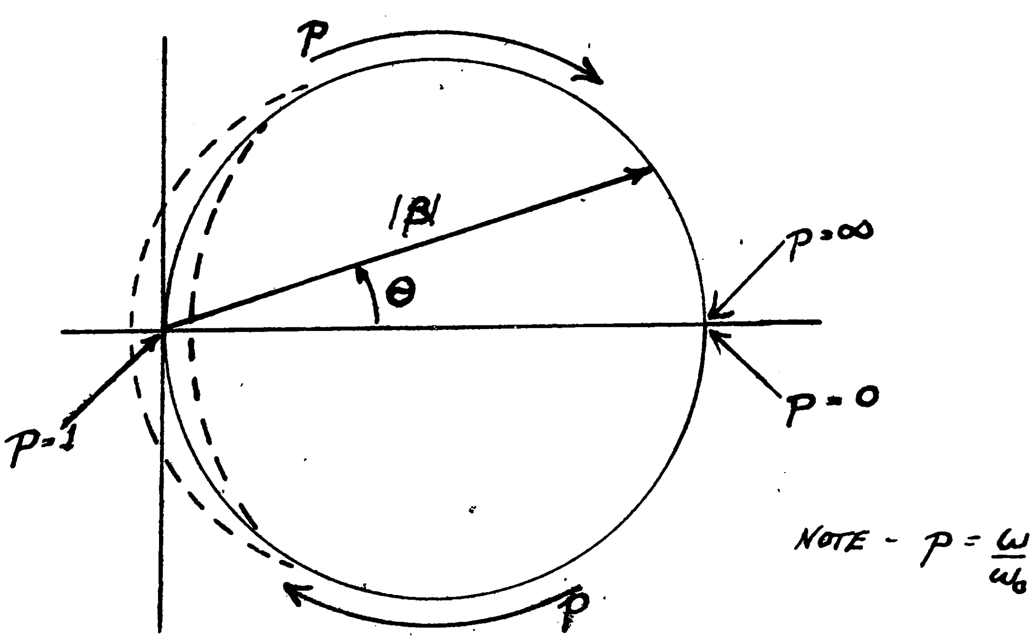
$$\beta = |\beta| \exp^{j\theta}$$

where $$|\beta| = \frac{1}{1+16p^2/(1-p^2)^2}$$

and $$\theta = \tan^{-1} \frac{4p}{p^2-1}$$ from equation IV-10.

Then a polar plot such as that shown in fig. 8 results.[1]

fig 8 POLAR PLOT OF THE FEEDBACK PARAMETER β



1 Hillan (See page 18)

It will be seen from fig. 8 that the phase change caused by the circuit is small when the frequency is not near ω_0. Thus this circuit satisfies the phase change conditions set up in the discussion of negative feedback amplifiers.

However, if the network is incorrectly balanced, the locus of β does not pass through the origin, but may follow a locus similar to one of those shown dotted in fig. 8. Thus, positive feedback may occur in the region of nominal balance frequency. This condition will enhance the output signal obtained but will also cause a loss of stability in the circuit. If the effect is too great, regenerative oscillations will result.

A circuit has now been devised which, if used as Z_f, in fig. 5 will have the proper frequency selective characteristics and phase characteristics to make the circuit shown in fig. 5 a frequency selective amplifier.

If this circuit is analyzed it will be found that the peak amplification varies inversely as m + 1. Thus the selectivity obtained and the peak amplification represent a compromise on the value of m.[1]

Section C

The Construction of the Apparatus

I. General Construction Details

It was desired to produce a finished, self-contained instrument of compact size and moderate cost. Simplicity of operation was of prime importance.

1 Hillan (See page 18)

The entire instrument including the power supply was built on a standard 8 X 12" Hammond chassis. A front panel of Arborite was placed on the 8-inch side of the chassis. This permits the completed instrument to be placed on a shelf with other apparatus with the use of a minimum of shelf length.

All controls necessary for the operation of the instrument are mounted on the front panel, as are the input and output terminals. The input and output terminals are banana plug binding posts mounted on 3/4" centres. This permits their use with a standard twin banana plug. One of each pair of terminals is connected directly to the ground. This permits the instrument to be used with a shielded transformer in the bridge circuit.

For ease in tuning a Vernier adjustment is provided on the tuning condenser. This is achieved through a simple pulley system which results in a slowing of the direct drive by a factor of four.

The A-C input enters the chassis at the back. There is also a ground terminal at the back of the chassis.

A circuit diagram of the detector circuit is given in fig. 9 with the adjoining Table I containing a components parts list. In addition photographs of the apparatus are given in fig. 13.

II. The Wein Bridge Oscillator

A Wein Bridge Oscillator, such as that described in Section B-III was used as the local oscillator in the circuit. A miniature 6AU6 voltage amplifier pentode was used as the oscillator tube, while a miniature 6AK6 power amplifier pentode was used as the output tube.

Output frequencies of from 1.2 kc. to 130 kc. at an RMS voltage of 35 volts are obtained from this oscillator. This frequency range is obtained in two overlapping steps - from 1.2 kc. to 13 kc. and from 12 kc. to 130 kc. These frequency ranges approximate those predicted from equation III-4, which are 1.2 kc. to 13.25 kc. and 12.1 kc. to 132.5 kc., where the first range is calculated using R = 100 kilohms and C varying from 1320 picofarads to 120 picofarads. The high range uses the same condenser and an R of 10.0 kilohms. The change-over from low to high frequency is obtained by substituting R_{36} and R_{39} (fig. 9) for R_{37} and R_{38} by means of a switch mounted in the front panel. The frequency is varied by varying the size of C_{32} and C_{33} (in fig. 9) simultaneously by means of a knob on the front panel. The overlapping of the ranges is caused by the fact that the condensers actually vary in size from minimum to maximum by a factor of 11. It was felt that this overlapping contributed to the ease of operation of the instrument. The variable condenser is mounted on top of the chassis, and is completely shielded to eliminate variable stray capacitances.

The resistors R_{36}, R_{37}, R_{38}, R_{39} (fig. 9), which are all in the frequency determining network, are all wirewound for greater frequency stability.

The amount of negative feedback to the cathode of tube 5 is controlled by a rheostat, P_2. This rheostat is also wirewound for greater stability. The rheostat was adjusted until the output of the oscillator was a maximum without noticeable distortion of the waveform. This point was reached when the RMS value of the output was 35 volts.

For higher outputs, noticeable distortion of the waveform could be detected on a CRO screen. The rheostat P_2 was adjusted to the optimum value and locked into position. In the operation of the instrument afterwards, it was found that the harmonic content of the output was so low that it was difficult to detect its effects.

From the analysis in Section B-III, it will be seen that maximum output is obtained from this oscillator when the associated Wein Bridge frequency determining circuit is just balanced. The condition imposed on the bridge was that $R_s = R_p$ and $C_s = C_p$. If R_s and R_p are out of balance, then equation III-1 will be out of balance. But the ratio R_s/R_p is still constant, since R_s and R_p are constant.

Therefore, the amount of unbalance caused will be constant and will not vary over the frequency range. But if C_s and C_p are unbalanced the ratio C_s/C_p will not be constant, since C_s and C_p are variable. The percentage unbalance of the bridge will thus be greatest when the condensers are at their lowest value, i.e. at high frequency, than when they are at their highest value. Therefore the output of the oscillator will be lower at high frequencies than at lower frequencies. It was found that a small unbalance of the condensers would cause the output to be reduced by as much as 50 percent at high frequencies.

In discussing the incandescent lamp used as an automatic amplitude control in Section B-III, it was shown that as the current through the lamp increased, the lamp heated up and its resistance increased, thus rebalancing the current. However, this process naturally takes a finite time, during which the oscillator must operate in some state between the initial and final state. It was found that the system operated best when the lamp was kept quite hot,- in fact if the output of the oscillator was turned down below 20 volts RMS, sudden changes in frequency, and consequent altering of balance conditions would cause sharp increases or decreases of output voltage which would gradually disappear as the circuit returned to normal.

This effect is explained by the fact that when the lamp is operated at a high temperature, the resistance varies more rapidly with temperature. In addition, if the lamp is operated at a high temperature in relation to its surroundings, heat is radiated and conducted away from it more rapidly, and the lamp thus becomes more sensitive to instantaneous changes of heat generated by the current in it. Thus the lamp should be operated as hot as possible if the automatic amplitude control is to have a short time lag.

III. The Mixer Tube

A miniature type 6BE6 pentagrid convertor was used to heterodyne the local oscillator signal with the incoming signal. Grid #1 was used as the control grid for the local oscillator signal, while grid #3 was used for the incoming signal. Grids #2 and #4, which are joined together within the tube, were used as screen grids. This arrangement serves to reduce the interelectrode capacities involving the two control grids to a minimum. In particular, the capacity between grid #1 and grid #3 is greatly reduced, thus effectively isolating the input signal from the local oscillator signal.

IV. The Tuned Amplifier

In view of the fact that the tuned amplifier was to be required to select a 1000 cps signal obtained by tuning a local oscillator signal to beat with a signal which may have a frequency as high as 100 kc., it was felt that the amplifier should have a broad maximum of response. For this reason, two cascaded tuned amplifiers of the parallel-T feedback type described in Section B-IV, and tuned 70 cps apart, were used.

Since it was desired that the response characteristic have a broad flat maximum, the response curves of the individual amplifiers could not be too sharp. This was accomplished by not tuning the parallel-T feedback circuits to their maximum selectivity. By making the value of n (Part B-IV 6,7) slightly greater than 1, and by making the value of m (Part B-IV 8,9) approximately 2, the selectivity curves of the individual amplifiers were made quite broad. Care was taken however not to make them too broad, lest beats from harmonics of the signal also be amplified, thus spoiling the selectivity of the instrument.

The broadening of the peaks also had the effect of increasing the effective amplification factor of the amplifier at the maximum. In the first place it was remarked in Section B-IV that the peak amplification of the individual amplifier varies inversely as $m + 1$. Thus the low value of m employed gave high peak amplification. Also, since the second amplifier operates with the other's output signal as its input signal, the broadness of the peaks means that when one amplifier is operating at maximum response, the other is amplifying the same signal more than would be the case if the peaks were much narrower.

The parallel-T's were balanced by cut and try methods rather than the more tedious matching of components. This method enabled the use of wide tolerance ($\pm$ 10%) components. By this method, response peaks were obtained at 1000 cps and 1070 cps. The adjustment of P_3 and C_{20}, P_4 and C_{31}, not only tuned the networks to their balance points, but enabled slight variations of the position of these balance points in the frequency spectrum. For larger variations (of more than 20 cps) it was found that adjusting the values of R and C in the "cross" of the T was the easiest method (equations IV-2 and IV-3). In order to obtain two peaks of equal height on the overall response curve, it was necessary to make the peak amplification of tube #4 alone, whose peak is at 1070 cps higher than that of tube #3 alone, whose peak is at 1000 cps. This was done by slightly detuning the parallel-T network of tube #3. The above effect is contrary to what would be expected from a consideration of IV-7 which shows that the response curve is logarithmically symmetrical about its maximum. From this, one would expect that the low frequency tube should have the higher amplification since at maximum response it would be operating on a lower portion of the response curve of the other tube than the high frequency tube in similar circumstances.

Response curves for each of the tubes individually and of the two together are plotted in fig. 11. These curves were obtained by putting a 0.1 volt signal on the points A and B (fig. 9) outside the tuned grid loads of the individual amplifiers, and measuring the output of the tubes with a vacuum tube voltmeter. Data obtained in this manner is contained in Table II.

It should be pointed out that while the curves so obtained correspond to the conditions under which tube #3, and the whole amplifier must operate, they do not correspond to the conditions under which tube #4 must operate. For while tube #3 has a constant input, and thus operates over a flat portion of the tube characteristic (assuming a small input signal), tube #4 in ordinary operation has a variable input depending on frequency, which is much larger than the constant signal input of tube #3, in the region of 1000 cps, and may thus force the tube to operate for part of each cycle on a non-linear portion of the tube characteristics. This would have the effect of reducing the effective amplification of the tube under operating conditions, thus necessitating a higher peak amplification of the tube under the conditions which were in effect when the curves were taken.

This condition is shown by the dotted curve for tube #4 in fig. 11 which is obtained by ~~subs~~tracting the decibel gain of tube #3 from the total decibel gain of the amplifier. This gives the decibel gain of tube #4. The values so obtained are tabulated under calculated gain in the last column of Table II.

It will be noted that for input signals above 15 volts the output of tube #4 is markedly different from that at low signal input. The major effect is to lower the peak amplification to a value of 28 decibels under operating conditions.

In the construction of this amplifier, difficulty was encountered with regenerative oscillations (as discussed in section B-IV). If the resistance R3 in the "upright" of the T was made too small, or the capacitance, C_c, too big, conditions were obtained where regenerative oscillations occurred. This difficulty was overcome (1) by careful tuning of the feedback networks so as not to go past their balance points and reach this condition for positive feedback; (2) by reducing the gain of the amplifiers themselves until they were moderate gain amplifiers. This is to satisfy the Nyquist criterion[1] which states that oscillations will occur if the product $A\beta = 1$ (where A is the amplification and β is the feedback factor and both are complex quantities.) In this case β is small, so that if A is reduced, $A\beta$ will be less than 1 and the oscillations will cease. (3) A small amount of negative feedback to the cathode of each tube was employed. This was achieved by using a relatively small cathode bypass condenser.

Care must be taken in designing an amplifier of this type that the components of the feedback network will have a high enough impedance that the tube impedance is not shunted by the network.[2] In addition, the capacitances of the network should be large enough that the interelectrode capacitances will not unbalance the bridge unduly.

From the curve in fig. 11 it will be seen that the amplifier so built has the desired characteristics. A fairly large pass band with a broad square top approximately 100 cycles broad is obtained. The high amplification obtained,- 43 decibels in the peak,- is a favourable factor, while the attenuation caused at low and high frequency is also of great assistance.

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1 Terman, F.E. "Radio Engrs Handbook", McGraw-Hill, 1943, p. 398

2 Hillan, A.B. "The Parallel-T Bridge Amplifier", J. Inst. Elec. Engrs, Pt. III, <u>94</u>,27,42.

V. The Preamplifier

An experimental setup consisting of the mixer, local oscillator and tuned amplifier was built and tested on a bridge. It was found that a preamplifier of 40 decibels gain was necessary in order to balance bridges to five significant figures, which was the accuracy of balance desired.

Since the instrument is to be used over a range of frequencies from 2 kc. to 100 kc., the preamplifier must have a 40 decibel amplification over this frequency range. It was originally decided to use two moderate gain R-C coupled amplifiers as the preamplifier. However, when the circuit was built, super-regenerative oscillations at a frequency of 1000 cps occurred with the main feedback loop being directly from the output to the input terminals. These oscillations occurred when there was no input signal, being apparently started by random noise in the circuit. A small input signal served to squelch these oscillations. However since it was desired to balance bridges to a null signal, the oscillations had to be eliminated.

These oscillations were due to the high gain of the instrument, coupled with proper phase relations between output and input, as well as some coupling of the various stages through the power supply. Heavy decoupling of the whole amplifier chain was resorted to, as well as shielding of the input circuit. At this point oscillations could be eliminated by the use of a shielded input terminal. However, it was felt that this would complicate the use of the apparatus needlessly. Therefore one stage of preamplification was eliminated. This not only reduced the gain but it altered the phase relations between the output and input and thus effectively halted oscillations.

By making the preamplifier a fairly high gain amplifier, it was still possible to obtain a 40 decibel gain over the desired frequency range. This is illustrated in fig. 12, which is a plot on a logarithmic frequency scale of the response of the amplifier. This plot was obtained by applying a signal of 0.1 volts at various frequencies to the input of the tube and measuring the output with a vacuum tube voltmeter. The data so obtained is contained in Table III.

A gain control on the preamplifier was desired as a means of controlling the output of the instrument and thus enabling preliminary balancing of the bridge when relatively high signals are obtained from it. When a 1 megohm carbon potentiometer was placed across the grid circuit of the input tube as a gain control a high noise level resulted in the instrument. This noise was due partly to thermal noise and partly to the noise common to all carbon type resistors caused by granular contacts changing.[1] This noise voltage varies as the resistance, and is a voltage generated directly across the terminals of the resistor.

This noise was very largely eliminated by placing the 1 megohm potentiometer P_1 (fig. 9) in series with the grid, which then had a 68 kilohm resistor as a grid leak to ground. This arrangement effectively removed a large part of the noise voltage from the circuit. At all times the noise voltage was now only that developed in the 68 kilohm grid leak, rather than that developed in the 1 megohm potentiometer which was much higher. This circuit has the disadvantage, however, that it is not possible to turn the input signal down to zero. If a bridge oscillator with a variable input is used, this difficulty is easily overcome.

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1 Terman, F.E. "Radio Engineers Handbook", McGraw-Hill, 1943, p. 477

It was found in using the instrument, that oscillations tended to occur due to interaction between the output (earphone) leads and the input leads. However, the use of a shielded input lead effectively eliminated this difficulty.

VI. The Power Supply

A power supply designed to give 300 volts at 100 milliamperes and 6.3 volts and 4 amperes was built into the instrument. The supply utilized a conventional 2-stage choke input filter to give a high degree of filtering. The supply may be switched on and off by a switch in the front panel. A circuit diagram of the power supply is given in fig. 10.

Section D

The Finished Instrument

From an examination of the data presented, and allowing for a 5 decibel loss in the converter tube, it will be seen that the instrument has an 80 decibel gain over the operating frequency range when correctly tuned. This means that the instrument has a 5-volt output (which can easily be detected by earphones) for a 5×10^{-4} volt input.

However, an easier way to see the effectiveness of the instrument is to consider readings made with it under actual operating conditions.

The Schering bridge shown in fig. 13 was used in this trial. The balance conditions for this bridge are given by[1]

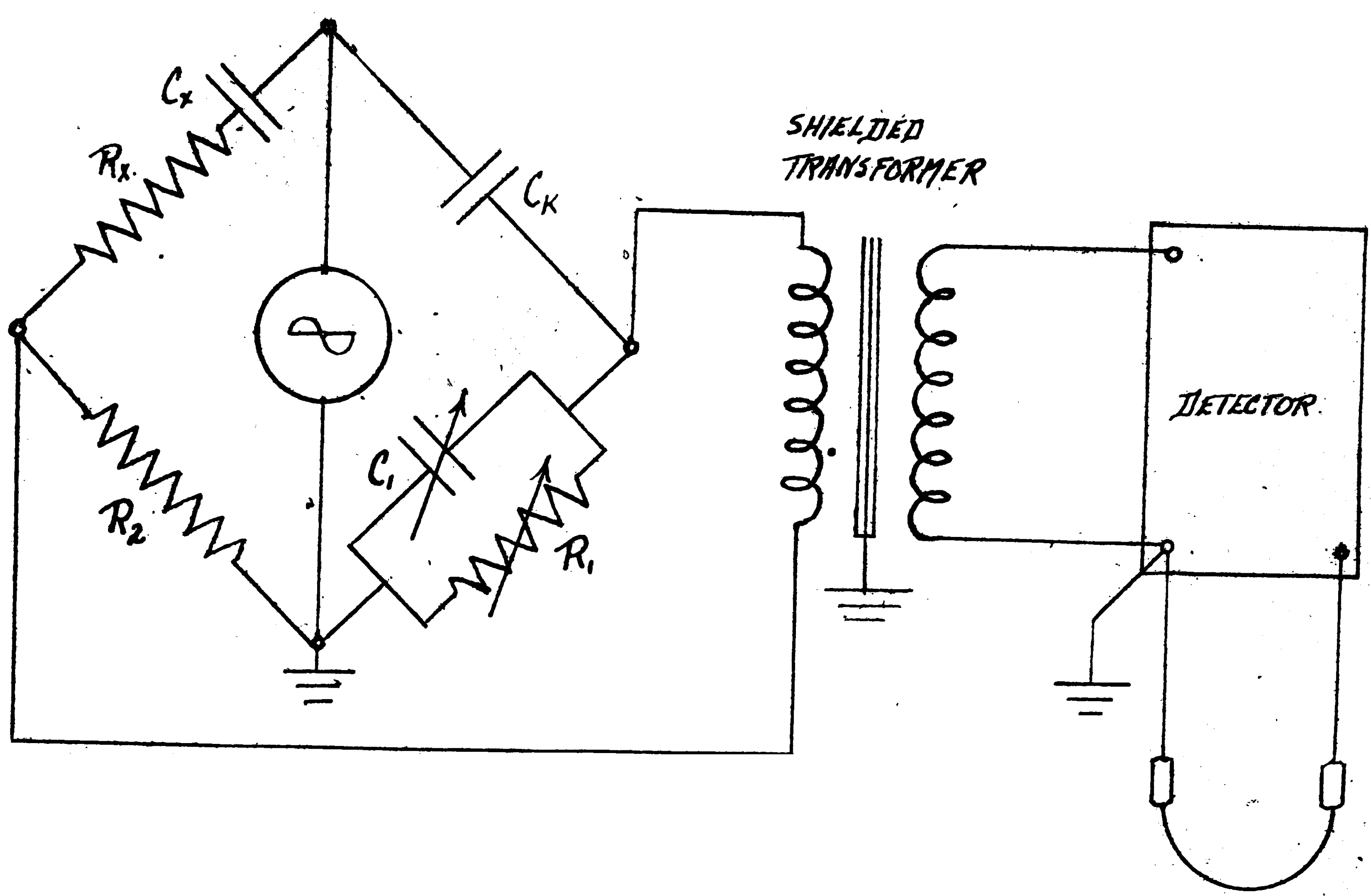
$$C_x = R_1 C_k / R_2 \qquad \text{D-1}$$

$$R_x = R_2 C_1 / C_k \qquad \text{D-2}$$

1 Terman, F.E. "Radio Engineers Handbook", McGraw-Hill, 1943, p. 905.

fig 13.

THE SCHERING BRIDGE



R_x and C_x are the resistance and capacitance, respectively, of the unknown, C_k is a known standard capacitance, R_2 is a fixed resistance and R_1 and C_1 are the variable resistance and capacitance respectively. The bridge used was a Sullivan bridge, with components accurate to 0.1 percent.

Readings were taken on two condensers separately and then of the two in series and in parallel at various frequencies. The readings were then checked from the formulae

$$C_p \doteq C_a + C_b \qquad \text{D-3}$$

$$C_s = \frac{C_a C_b}{C_a + C_b} \qquad \text{D-4}$$

$$R_s = R_a + R_b \qquad \text{D-5}$$

where the subscript s refers to series and p to parallel.

The exact formula for D-3 is given by[1]

$$C_a' = \frac{C_a}{1+\omega^2 C_a^2 R_a{}^2}$$

$$C_b' = \frac{C_b}{1+\omega^2 C_b{}^2 R_b{}^2}$$

For good quality condensers the value R_a or R_b is small, and therefore

$$\omega_o{}^2 C_a{}^2 R_a{}^2 \ll 1.$$

Thus D-3

$$C_p = C_a + C_b$$

holds approximately.

The bridge was first balanced at 1 kc., using the ordinary audio amplifier and earphones provided. Then, the detector was inserted into the circuit and the bridge was again balanced at intervals of 5 kc. until 20 kc., which was the maximum frequency of the bridge oscillator.

The results of these measurements are given in Table IV. In all cases the results obtained have errors which are of the same order as those obtained at 1 kc. with standard equipment, and thus must be errors inherent to the bridge equipment.

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1. Hague, B. "A.C. Bridge Methods", Pitman & Sons, 1946, p. 190.

It is interesting to note the variation of effective capacity and resistance of the unknown with frequency. The value of capacity decreases by 2% in 20 kc., thus giving ample proof of the necessity of measuring the values of circuit components at the frequency at which they are to be used. However, it should be realized that the values obtained are not corrected for frequency-dependent effects in the bridge components, and are thus not true values.

In the operation of the bridge using the detector, it was found that the bridge could be balanced with ease to five significant figures. This is about the limit of accuracy of most bridge components. Sharply defined null points were observed,- in every case it was possible to adjust R_1 until there was no doubt as to the correct figure on the fifth dial.

Good stability of tuning was observed, since the initial tuning appeared to remain stable throughout any series of readings.

Care must be taken in the operation of the instrument that it is correctly tuned with the fundamental of the local oscillator signal beating with the fundamental of the input signal. This point is easily identified, since subsiduary, or harmonic, beats are necessarily very sharp tuning points, whereas the fundamental tuning point is broader. The fundamental beat, being the strongest signal, can be heard starting at a fairly high pitch, decreasing in frequency and increasing in intensity to 1000 cps as the instrument is tuned, then decreasing in both pitch and intensity to zero. The process is then reversed as the beat $|\omega_a - \omega_1|$ passes through zero and begins to increase in frequency again. The instrument may be operated with ω_1 either above or below ω_a

equally well except at low frequencies where it is best to operate with ω_1 above ω_a in order to eliminate the local oscillator frequency from the output.

Section E

Conclusions

It has been shown that the instrument constructed had the following points to recommend it

1) ease of operation
2) high sensitivity
3) good frequency discrimination
4) small size and low cost.

The heterodyne method employed enabled separation of the tuning and frequency filtering circuits into two separate parts of the circuit. This enabled sharp frequency discrimination while still permitting ease of tuning to a wide range of frequencies. The high gain employed and the use of earphones as the ultimate detecting instrument resulted in high sensitivity.

Throughout the instrument, circuits were chosen with regard to compact size and possible wide tolerances of components. This factor kept size and cost to a minimum.

The operator of the instrument must be careful to tune it to the fundamental of the input signal. This is easily done, as already described and provided proper instructions are given, no difficulty should arise here

It is therefore felt that the instrument is ideally suited for bridge detection in the ultrasonic frequency range.

fig 9

CIRCUIT DIAGRAM OF THE APPARATUS.

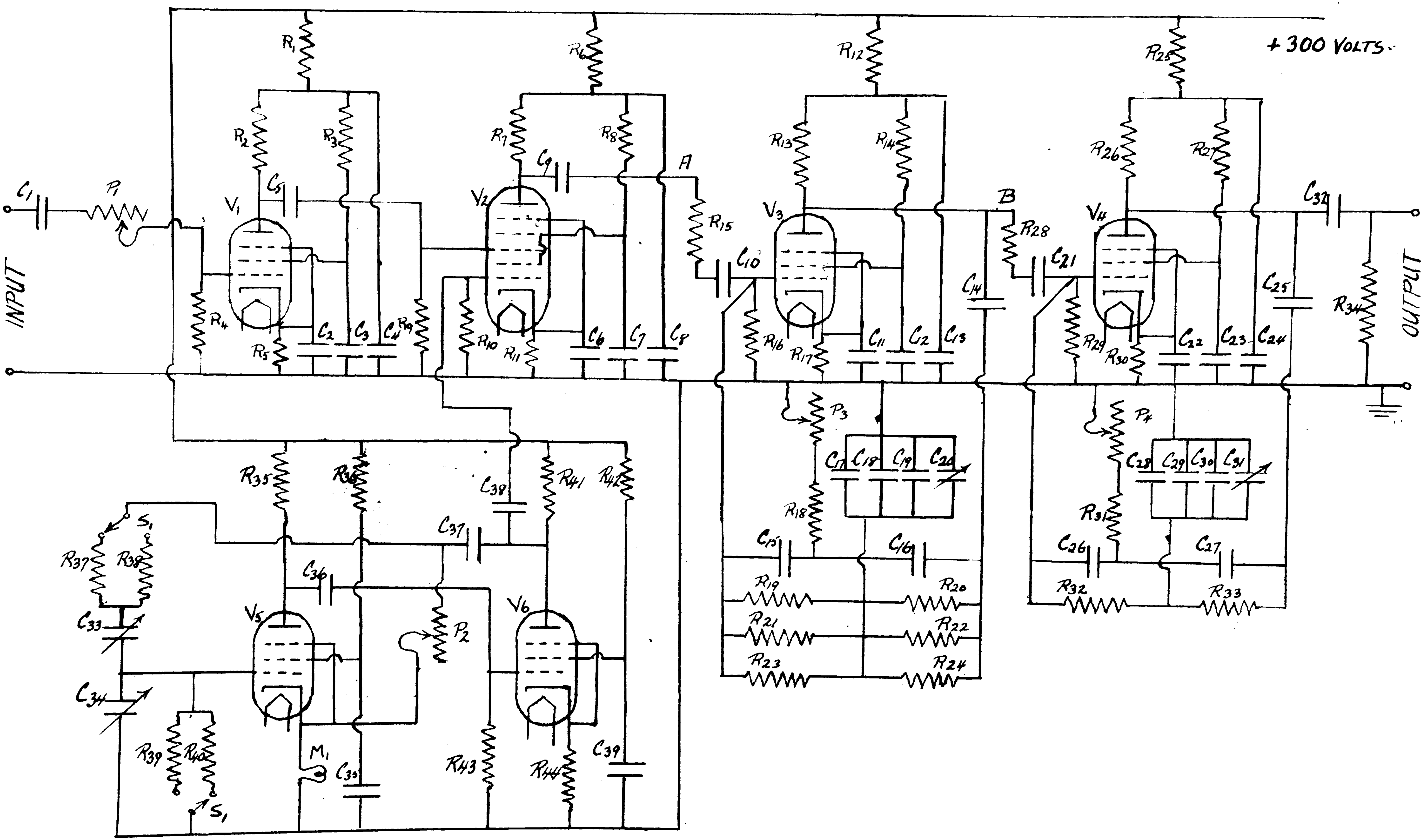


Table I. - Components List for Circuit in fig. 9

Vacuum Tubes

V_1, V_3, V_4, V_5	6AU6 miniature
V_2	6BE6 miniature
V_6	6AK6 miniature

Resistors - all resistors are composition type, with tolerance $\pm$ 10%, except where otherwise stated.

R_1	10,000 ohms	1 watt	R_{21}, R_{22}	1,000,000 ohms	1/2 watt
R_2	100,000 ohms	1/2 watt	R_{23}, R_{24}	3,300,000 ohms	1/2 watt
R_3	470,000 ohms	1/2 watt	R_{25},	10,000 ohms	1 watt
R_4	68,000 ohms	1/2 watt	R_{26}	100,000 ohms	1/2 watt
R_5	470 ohms	1 watt	R_{27}	470,000 ohms	1/2 watt
R_6	10,000 ohms	1 watt	R_{28}	470,000 ohms	1/2 watt
R_7	15,000 ohms	1 watt	R_{29}	220,000 ohms	1/2 watt
R_8	27,000 ohms	1 watt	R_{30}	470 ohms	1 watt
R_9	470,000 ohms	1/2 watt	R_{31}	47,000 ohms	1/2 watt
R_{10}	470,000 ohms	1/2 watt	R_{32}, R_{33}	270,000 ohms	1/2 watt
R_{11}	120 ohms	1 watt	R_{34}	68,000 ohms	1/2 watt
R_{12}	10,000 ohms	1 watt	R_{35}	220,000 ohms	1/2 watt
R_{13}	100,000 ohms	1/2 watt	R_{36}	470,000 ohms	1/2 watt
R_{14}	470,000 ohms	1/2 watt	R_{37}, R_{40}	100,000 ohms	wirewound
R_{15}	470,000 ohms	1/2 watt	R_{38}, R_{39}	10,000 ohms	wirewound
R_{16}	220,000 ohms	1/2 watt	R_{41}	10,000 ohms	1 watt
R_{17}	470 ohms	1 watt	R_{42}	68,000 ohms	1/2 watt
R_{18}	47,000 ohms	1/2 watt	R_{43}	470,000 ohms	1/2 watt
R_{19}, R_{20}	270,000 ohms	1/2 watt	R_{44}	560 ohms	1 watt

Table I. (cont'd)

Capacitors - all components ± 10% except where otherwise stated.

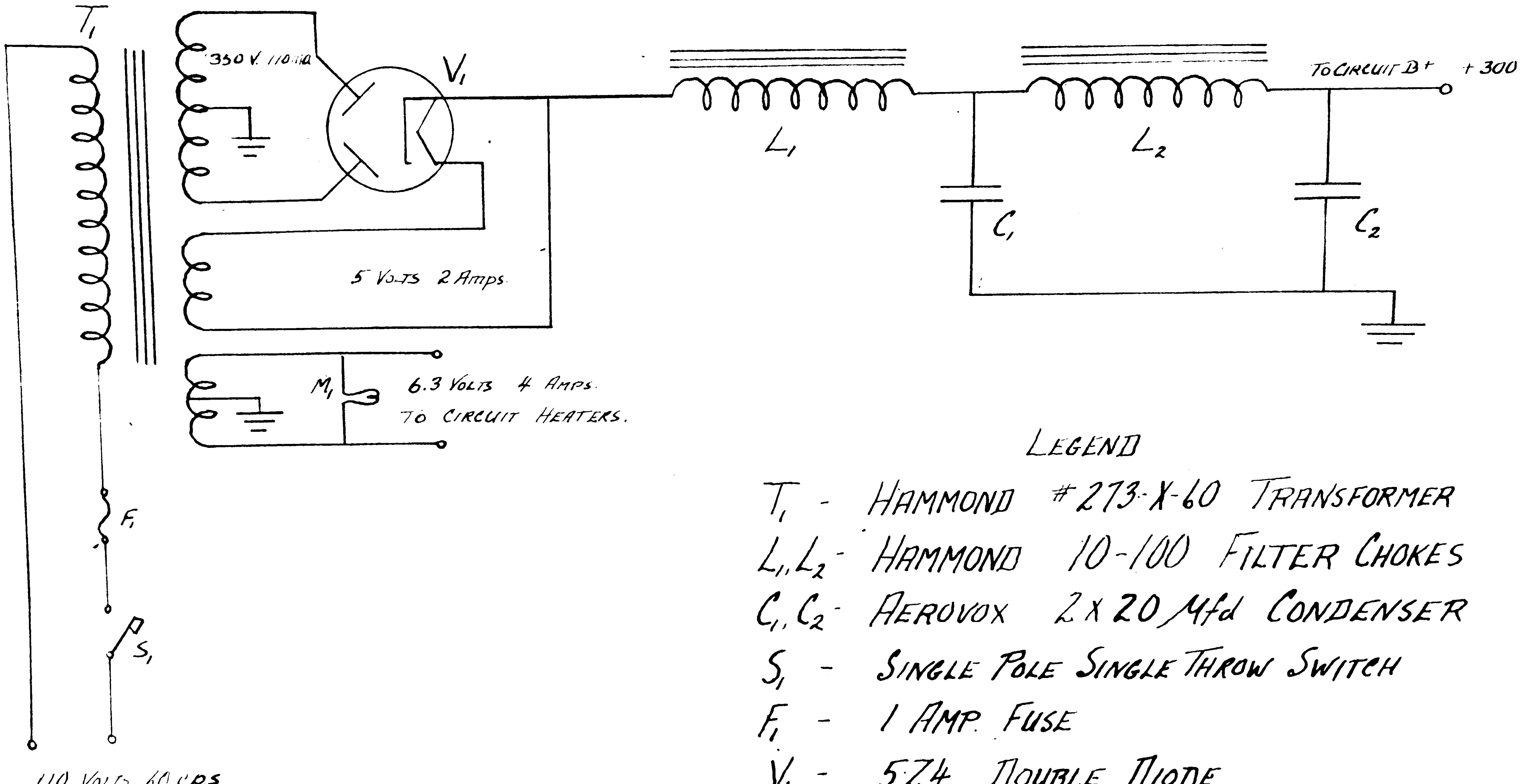
C_1	0.01 microfarads	600 volts	paper
C_2	12 microfarads	50 volts	dry electrolytic
C_3	0.05 microfarads	600 volts	paper
C_4	0.5 microfarads	600 volts	paper
C_5	400 picofarads		mica
C_6	0.1 microfarads	600 volts	paper
C_7	0.5 microfarads	600 volts	paper
C_8	0.5 microfarads	600 volts	paper
C_9	0.1 microfarads	600 volts	paper
C_{10}	400 picofarads		mica
C_{11}	0.01 microfarads	600 volts	paper
C_{12}	0.03 microfarads	600 volts	paper
C_{13}	0.5 microfarads	600 volts	paper
C_{14}	0.01 microfarads	600 volts	paper
C_{15}, C_{16}	700 picofarads		mica
C_{17}	0.001 microfarads		mica
C_{18}	400 picofarads		mica
C_{19}	50 picofarads		mica
C_{20}	150 picofarads		trimmer
C_{21}	400 picofarads		mica
C_{22}	0.01 microfarads	600 volts	paper
C_{23}	0.03 microfarads	600 volts	paper
C_{24}	0.5 microfarads	600 volts	paper
C_{25}	0.01 microfarads	600 volts	paper
C_{26}, C_{27}	600 picofarads		mica
C_{28}	0.001 microfarads		mica
C_{29}	500 picofarads		mica
C_{30}	50 picofarads		mica
C_{31}	150 picofarads		trimmer
C_{32}	0.1 microfarads	600 volts	paper
C_{33}, C_{34}	120 to 1320 picofarads		balanced variable, ganged
C_{35}	0.03 microfarads	600 volts	paper
C_{36}	0.01 microfarads	600 volts	paper
C_{37}	0.05 microfarads	600 volts	paper
C_{38}	0.05 microfarads	600 volts	paper
C_{39}	0.03 microfarads	600 volts	paper

Other Components

S_1 - double-pole double-throw switch
M_1 - 3 watt, 115 volt incandescent lamp.
P_1 - 1,000,000 ohm carbon rheostat
P_2 - 5,000 ohm wirewound potentiometer
P_3, P_4 - 50,000 ohm carbon potentiometers

fig 10

THE POWER SUPPLY



LEGEND

- T_1 - HAMMOND #273-X-60 TRANSFORMER
- L_1, L_2 - HAMMOND 10-100 FILTER CHOKES
- C_1, C_2 - AEROVOX 2 X 20 Mfd CONDENSER
- S_1 - SINGLE POLE SINGLE THROW SWITCH
- F_1 - 1 AMP. FUSE
- V_1 - 5Z4 DOUBLE DIODE
- M_1 - 6.3 VOLT PILOT LIGHT

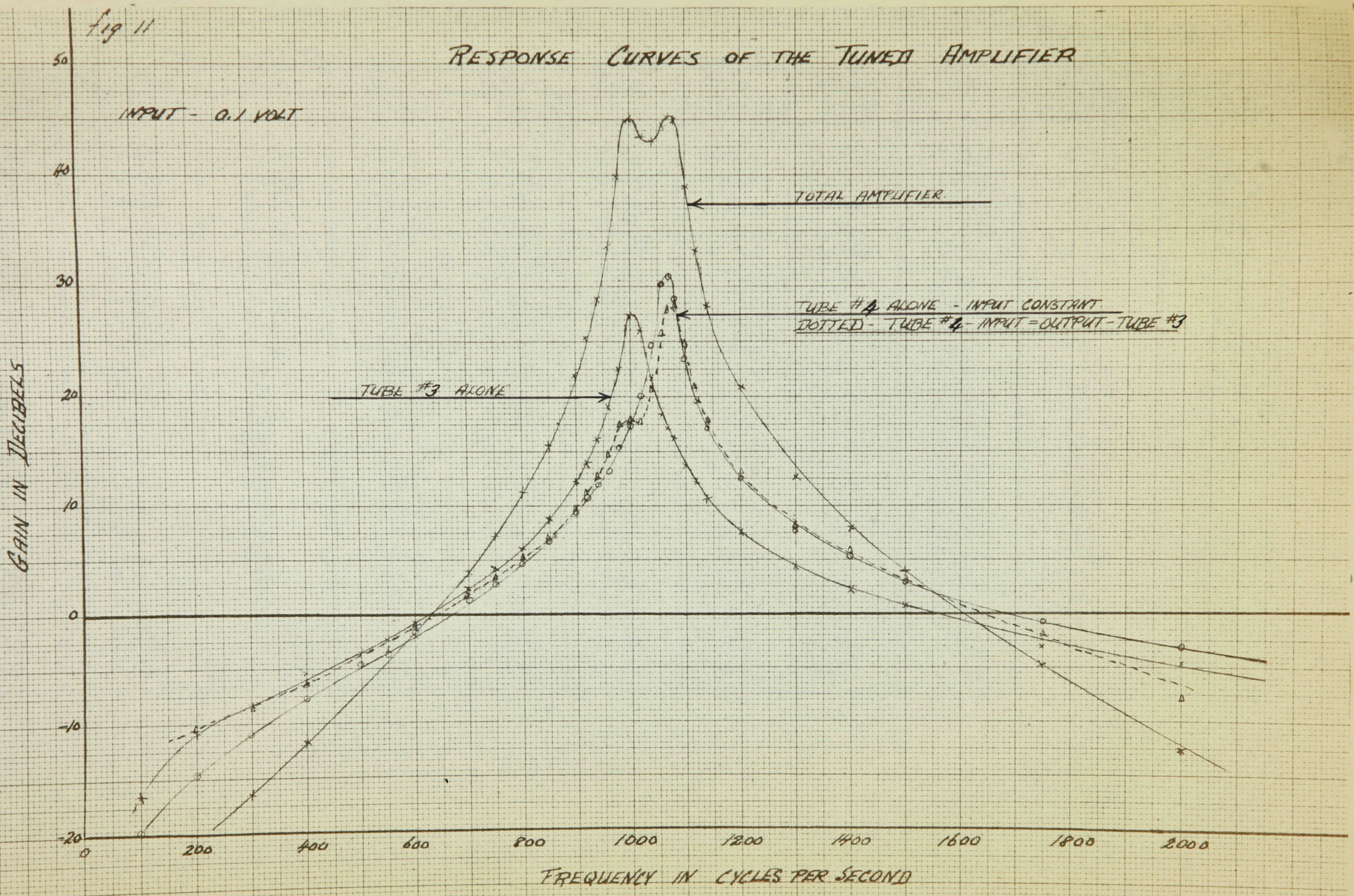


fig 11
RESPONSE CURVES OF THE TUNED AMPLIFIER
INPUT - 0.1 VOLT
TOTAL AMPLIFIER
TUBE #4 ALONE - INPUT CONSTANT
DOTTED - TUBE #4 - INPUT = OUTPUT - TUBE #3
TUBE #3 ALONE
GAIN IN DECIBELS
50
40
30
20
10
0
-10
-20
0
200
400
600
800
1000
1200
1400
1600
1800
2000
FREQUENCY IN CYCLES PER SECOND

Table II. Data for Response Characteristics of the Tuned Amplifier

(input 0.1 Volts)

Frequency	Total Amplifier			Tube number 3			Tube number 4			
	Output	Gain.	Gain	Output	Gain	Gain	Output	Gain	Gain	Gain (calc.)
cps	Volts		decibels	Volts		decibels	Volts		decibels	decibels
100	0.007	0.07	-23.1	0.015	0.15	-16.5	0.010	0.10	-20.0	-7.6
200	0.009	0.09	-20.9	0.026	0.26	-10.7	0.019	0.19	-14.4	-10.2
300	0.015	0.15	-16.5	0.038	0.38	- 8.4	0.028	0.28	-11.0	-8.0
400	0.026	0.26	-11.7	0.054	0.54	- 5.4	0.041	0.41	- 7.7	-6.3
500	0.100	1.0	0	0.067	0.67	- 3.5	0.059	0.59	- 4.6	3.5
600	0.078	0.78	- 2.2	0.091	0.91	- 0.8	0.08	0.8	- 1.9	-1.3
700	0.15	1.5	3.8	0.13	1.3	2.2	0.11	1.1	1.1	1.6
800	0.36	3.6	11.1	0.20	2.0	6.0	0.17	1.7	4.5	5.1
850	0.62	6.2	15.8	0.28	2.8	8.8	0.22	2.2	6.7	7.0
900	1.3	13	22.0	0.40	4.0	12.1	0.29	2.9	9.3	9.9
940	2.8	28	28.8	0.64	6.4	16.1	0.39	3.9	11.7	12.7
960	4.8	48	33.6	0.90	9.0	19.1	0.45	4.5	13.1	14.5
980	10.0	100	40.0	13.4	13.4	22.5	0.58	5.8	15.3	17.5
1000	17.9	179	45.0	2.35	23.5	27.4	0.74	7.4	17.3	17.6
1020	14.9	149	43.4	2.01	20.1	26.1	1.00	10.0	20.0	17.3
1040	13.7	137	42.6	1.26	12.6	22.0	1.70	17.0	24.6	20.6
1060	16.4	164	44.3	0.84	8.4	18.5	3.30	33.0	30.4	25.8
1080	17.8	178	45.0	0.67	6.7	16.5	2.75	27.5	28.8	28.5
1100	8.7	87	38.8	0.50	5.0	13.9	1.44	14.4	23.3	24.9
1200	1.1	11	20.8	0.24	2.4	7.6	0.41	4.1	12.3	13.2
1500	0.16	1.6	4.0	0.11	1.1	0.8	0.14	1.4	2.9	3.2
1750	0.057	0.57	-4.9	0.072	0.72	-2.9	0.093	0.93	-0.6	-2.0
2000	0.037	0.37	-12.6	0.061	0.61	-4.4	0.070	0.70	-3.1	-8.0

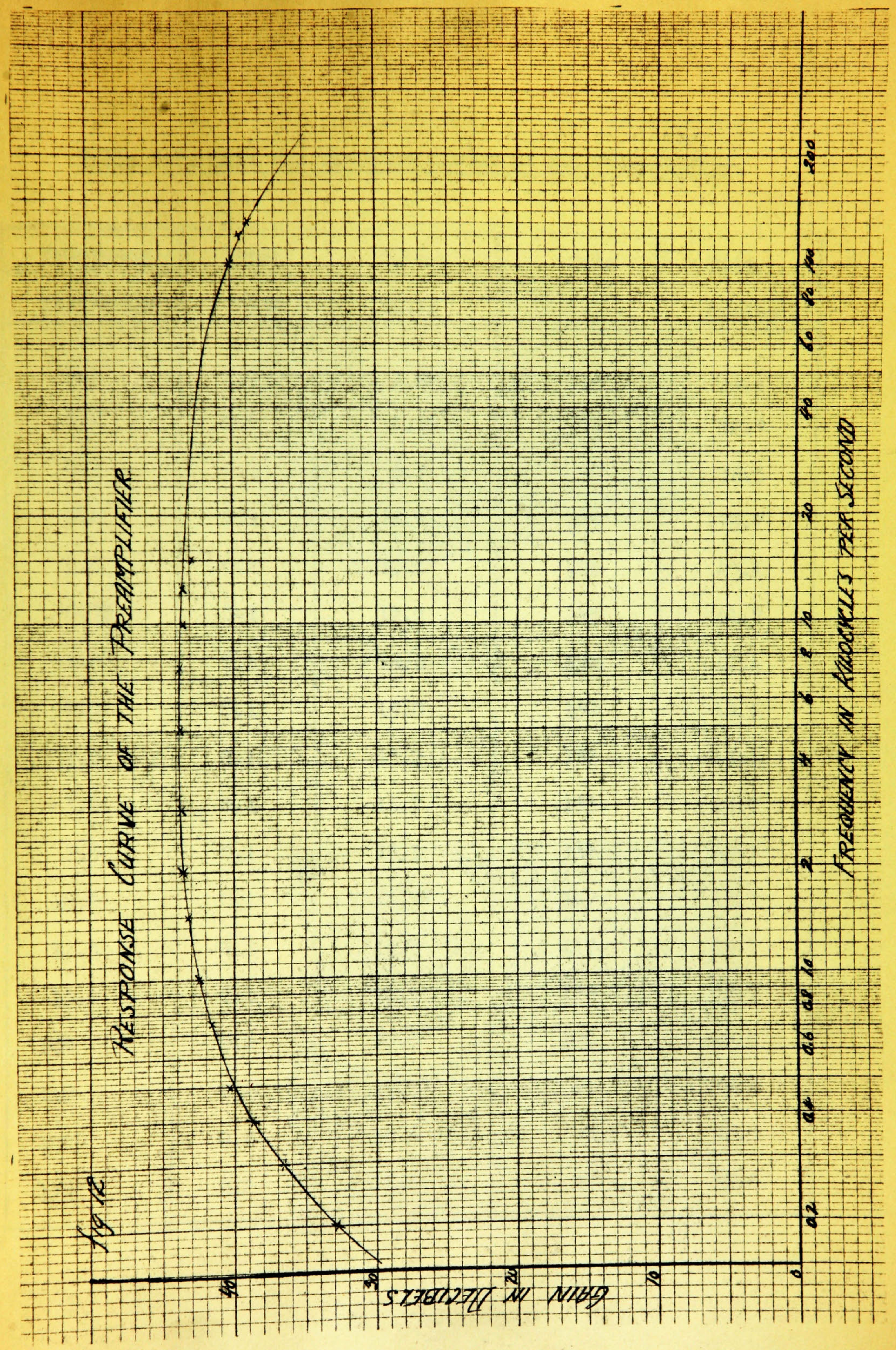


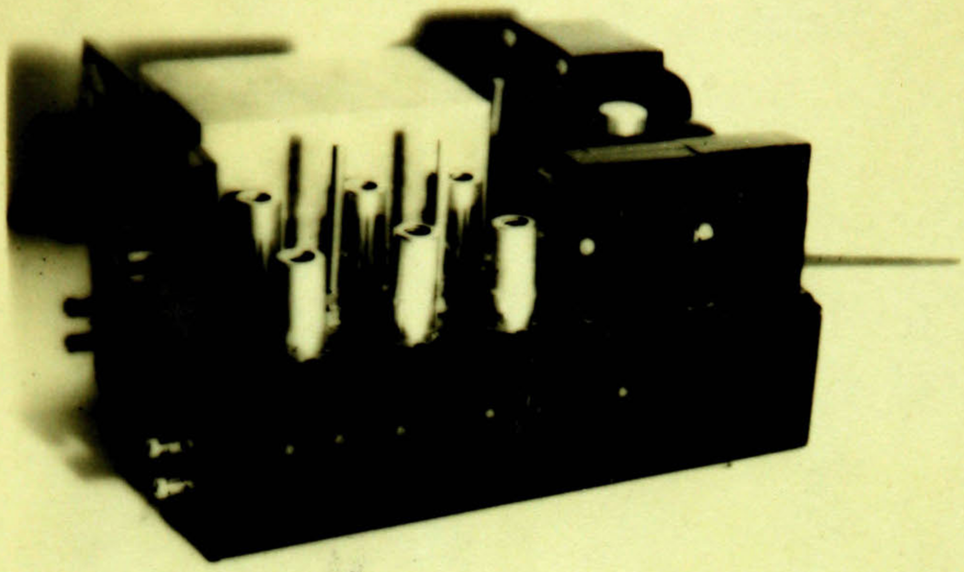
Fig 12 Response Curve of the Preamplifier

Table III. Data for Response Curves of the Preamplifier.

Frequency Kilocycles	Input Volts	Output Volts	Gain	Gain Decibels
0.1	0.1	1.80	18	25.1
0.2	0.1	4.40	44	32.9
0.3	0.1	6.7	67	36.5
0.4	0.1	8.6	86	38.7
0.5	0.1	10.1	101	40.1
0.75	0.1	12.2	122	41.7
1.0	0.1	13.2	132	42.4
2.0	0.1	15.0	150	43.5
3.0	0.1	15.2	152	43.6
5.0	0.1	15.3	153	43.7
7.5	0.1	15.1	151	43.5
10.0	0.1	14.7	147	43.3
12.5	0.1	14.7	147	43.3
15.0	0.1	13.6	136	42.7
100	0.2	20.0	100	40.0
110	0.2	19.0	95	39.5
120	0.2	18.6	93.0	39.3
130	0.2	17.5	87.5	38.8

fig 13

PHOTOGRAPHS OF THE DETECTOR



a) A GENERAL VIEW

b) THE FRONT PANEL

Table IV. Data from Measurements on the Schering Bridge

R_2 = 200 ohms C_k = 0.2 microfarads

	R_1 ohms	C_x μf	$C_{calc.}$ μf	%error	C_1 μf	R_x ohms	$R_{calc.}$ ohms	%error
1000 cps								
C_a	234.6_0	2.346_0			0.0044_5	4.4_5		
C_b	94.4_9	0.944_9			0.0090_0	9.0_0		
$C_{a+b}{}^p$	329.3_4	3.293_4	3.290_9	0.07_5	0.0030_4	3.0_4		
$C_{a+b}{}^s$	67.2_0	0.672_0	0.673_6	0.2_4	0.0138_2	13.8_2	13.4_5	2.9
5000 cps								
C_a	233.0_5	2.330_5			0.0013_7	1.3_7		
C_b	93.9_0	0.939_0			0.0025_6	2.5_6		
$C_{a+b}{}^p$	327.2_7	3.272_7	3.269_5	0.1_0	0.0008_7	0.8_7		
$C_{a+b}{}^s$	66.7_9	0.667_9	0.669_3	0.2_1	0.0041_5	4.1_5	3.9_3	5.3
10000 cps								
C_a	232.1_1	2.321_1			0.0008_0	0.8_0		
C_b	93.6_0	0.936_0			0.0016_4	1.6_4		
$C_{a+b}{}^p$	325.9_9	3.259_9	3.257_1	0.0_8	0.0007_6	0.7_6		
$C_{a+b}{}^s$	66.5_8	0.665_8	0.667_0	0.1_8	0.0026_2	2.6_2	2.4_4	6.9
15000 cps								
C_a	231.3_8	2.313_8			0.0006_7	0.6_7		
C_b	93.2_3	0.932_3			0.0013_1	1.3_1		
$C_{a+b}{}^p$	325.1_5	3.251_5	3.246_1	0.1_7	0.0006_0	0.6_0		
$C_{a+b}{}^s$	66.5_3	0.665_3	0.664_2	0.1_0	0.0021_5	2.1_5	1.9_8	7.9
20000 cps								
C_a	230.59	2.305_9			0.00056	0.56		
C_b	92.9_2	0.929_2			0.0011_5	1.1_5		
$C_{a+b}{}^p$	324.0_7	3.240_7	3.235_1	0.1_7	0.0003_8	0.3_8		
$C_{a+b}{}^s$	66.1_1	0.661_1	0.660_8	0.0_4	0.0018_8	1.8_8	1.7_1	9.0

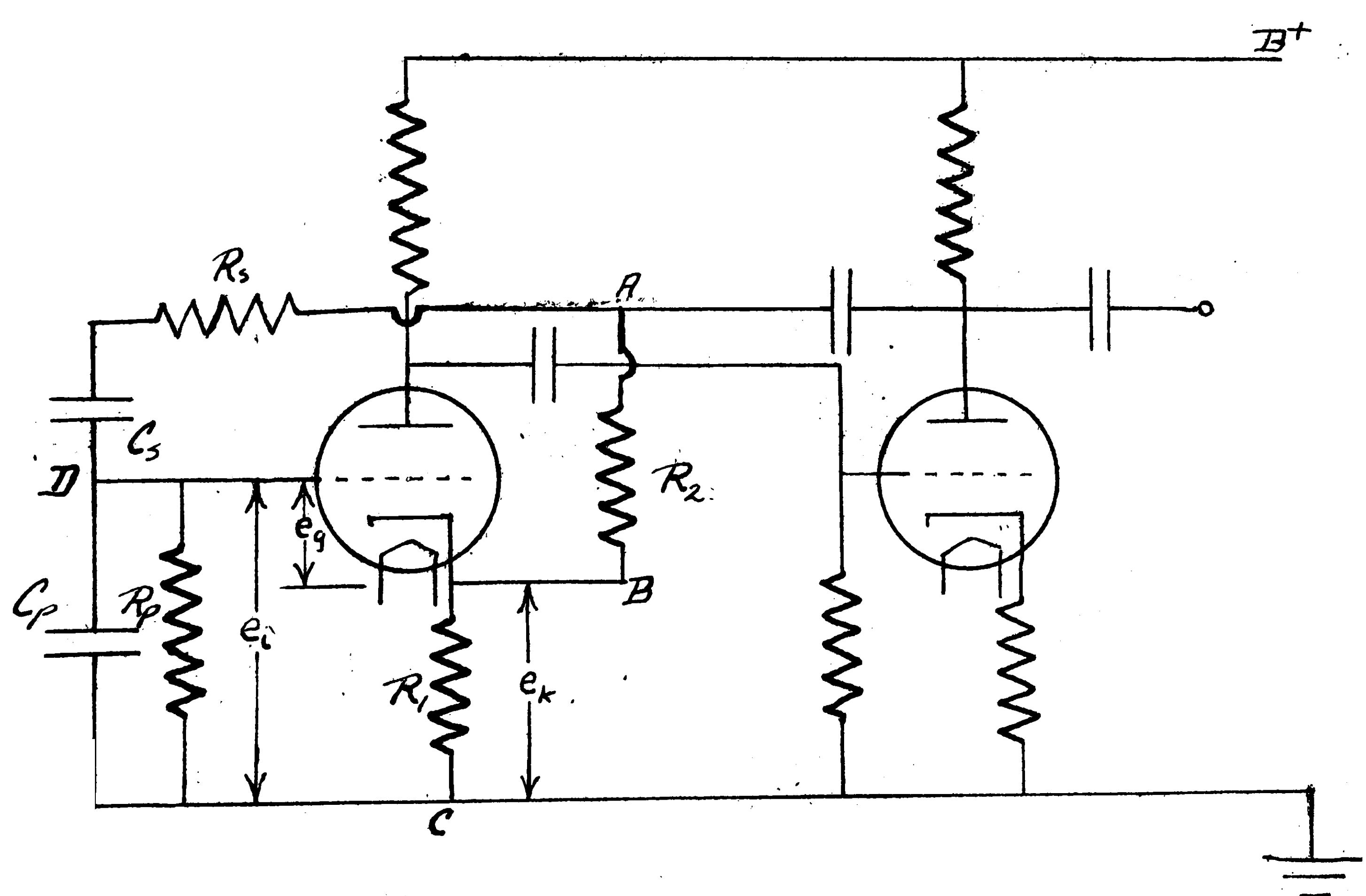
Note: All readings of R_1, above, are to ± 0.01 ohms.
All readings of C_1, above, are to ± 0.00005 microfarads.
All components in the bridge used are accurate to 0.1 percent.
s refers to series readings.
p refers to parallel readings.

Appendix I

To show that the Wein Bridge oscillator actually oscillates at a frequency slightly below that given by

$$\omega_o^2 = 1/C_pR_pC_sR_s \qquad \text{A-1}$$

fig 14 THE WIEN BRIDGE OSCILLATOR



NOTE - TRIODES ARE SHOWN FOR SIMPLICITY
ORDINARILY, PENTODES ARE USED

In fig. 14 when conditions are steady (i.e. oscillations are occurring)

$$e_g = e_i - e_k \qquad \text{A-2}$$

$$e_k = \frac{e_o R_1}{R_1+R_2} = \frac{e_o}{K}, \quad K = \frac{R_1 + R_2}{R_1} \doteq 3 \text{ from III-3} \qquad \text{A-3}$$

$$e_i = \frac{e_o z_p}{z_p+z_s}$$

$$= \frac{e_o}{1+\frac{R_s}{R_p}+\frac{C_p}{C_s}+j\omega C_p R_s+\frac{1}{j\omega C_s R_p}}$$

$$= \frac{e_o}{r + jx} \qquad \text{A-4}$$

where

$$r = 1 + R_s/R_p + C_p/C_s \doteq 3 \qquad \text{A-5}$$

and

$$x = j\omega C_p R_s - 1/j\omega C_s R_p \qquad \text{A-6}$$

For an amplifier,

$$e_o = e_g (A + jB) \text{ where } B \ll A$$

from A-2, 3, and 4

$$e_o = e_o(1/(r + jx) - 1/K)(A + jB)$$

$$rK + jxK = (K-r) A + xB + j[(K-r) B-xA]$$

If B is small, as it should be for a good amplifier; equating in phase and quadrature terms to 0

$$rK \doteq (K-r) A \qquad \text{A-7}$$

$$xK \doteq (K-r) B-xA \qquad \text{A-8}$$

from A-6

$$A = rK/K-r$$

But from A-3 and 5,

$$r = 3, K = 3.$$

This would mean that A approaches ∞. This is impossible. Therefore $K > r = 3$.

Now substituting in A-8 from A-7

$$(K-r)^2 B = K^2 x \qquad \text{or} \qquad x = \delta^2 B/K^2 \qquad \text{A-9}$$

where δ and B are both small quantities.

From A-6

$$\omega C_p R_s - 1/\omega C_s R_p \doteq \delta^2 B/K^2 \doteq 0$$

Therefore

$$\omega_o^2 C_p C_s R_p R_s = 1 + \omega C_s R_p \, B \, \delta^2/K^2.$$

or the oscillation frequency is slightly different from the value given by

$$\omega_o^2 C_p C_s R_p R_s = 1.$$

Since ω, C_s, R_p, δ^2, and K^2 are inherently positive, the nature of the correction will depend upon the sign of B. If B is negative, ω will be below ω_o; if B is positive, ω will be above ω_o.

It will be seen that the higher A and the smaller B, the closer the oscillator comes to fulfilling equation A-1.

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