AN RC OSCILLATOR

Thesis for the Degree of M. S.

MICHIGAN STATE COLLEGE

Roy John Smollett, Jr.

1950

This is to certify that the

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Roy John Smollett, Jr.

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J A Strelzoff
Major professor

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AN RC OSCILLATOR

By

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THESIS

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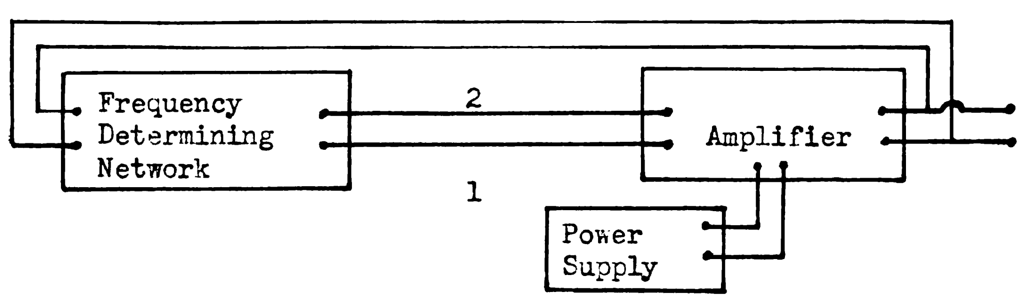
Nature of the problem:

The problem undertaken in this thesis is the design, construction, and testing of an electronic sine wave generator or oscillator to cover a frequency spectrum from sub-audio frequencies to radio frequencies. The oscillator is to produce sine waves with little distortion. The frequency calibration should be as nearly exact as can be obtained practically.

Method of solution:

The vast majority of electronic oscillators consists of the following component parts: a frequency determining network, an amplifier, and a method of limiting the amplitude of oscillations plus necessary associated equipment such as power supplies. This division into parts is somewhat artificial since each is definitely related to the others as indicated in Fig. 1.



Block diagram of an electronic oscillator.

Fig. 1

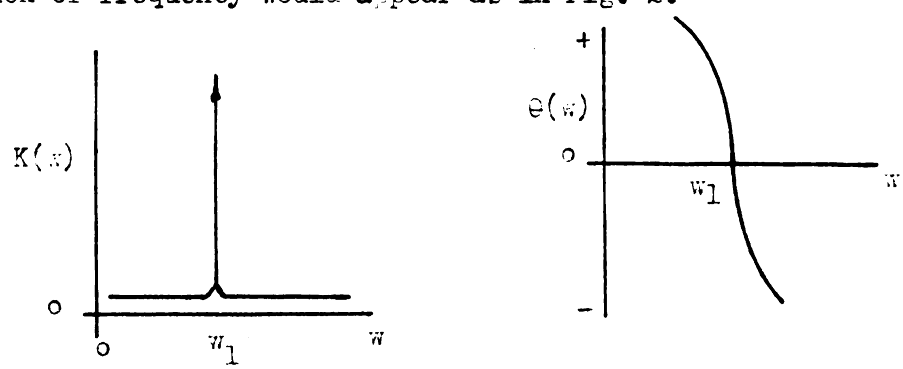
The method of designing such an oscillator is as follows: choose and design a frequency determining network; design the amplifier to provide the transfer characteristics required by the frequency determining network; design the amplitude control network to control the amplifier. And, then design the power supply and any other necessary equipment. Each of these phases of the problem will be treated under its separate heading.

The Frequency Determining Network:

The Barkhausen condition for oscillation (Ref. No. 1) referring to Fig. 1, requires that the net signal gain through the amplifier and through the frequency determining network will be exactly 1 + j0.

An ideal frequency determining network would maintain this condition at only one frequency, and that frequency would be independent of the transfer characteristics of the amplifier, temperature, and other circuit variables.

The transfer characteristics* of such a network plotted as a function of frequency would appear as in Fig. 2.



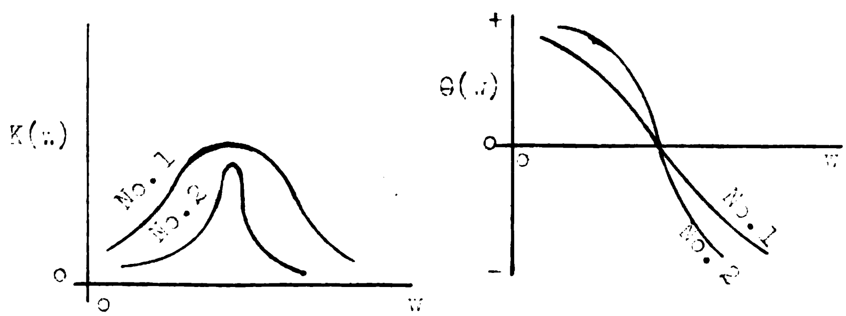
Transfer Characteristics of an Ideal Frequency Determining Network

Fig. 2

Such a network cannot exist for it would require no amplifier and would have no losses. However, the concept of an ideal frequency determining network has a value as a basis of judging networks that can be realized. For example, if the curves in Fig. 3 were the

* Transfer characteristics here refer to the ratio of output voltage to input voltage. $\frac{E\ out}{E\ in} = K(w)\ \varepsilon^{j\theta(w)}$

transfer characteristics of two different frequency determining networks, other things being equal, network No. 2 would be more desirable to use in an oscillator than network No. 1, because the frequency of oscillation would be more nearly independent of phase shift in the amplifier.



Typical Transfer Characteristics of Frequency
Determining Network

Fig. 3

The "Q" (Ref. No. 2) of a network is a much more analytical description of this property. An ideal network would have a Q of infinity. In Fig. 3 network No. 2 would have a greater Q than network No. 1. In general, the greater the Q of the frequency determining network, the more stable the oscillator.

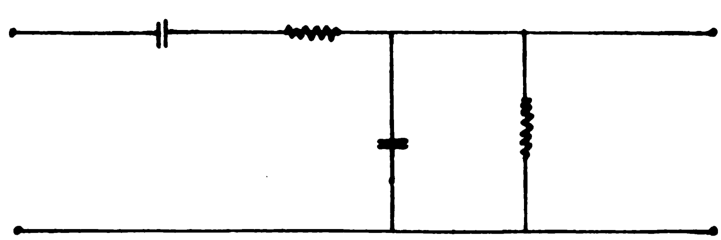
The Q of the network is not the only factor to be considered for this application. Some of the other factors are: the number of elements in the network, the time and temperature stability of the elements, the linearity of the elements, and the frequency range possible with the network.

Some of the more common frequency determining networks used in oscillators are: quartz crystals, magneto-strictive materials, inductance-capacitance resonant circuits, and resistance-capacitance circuits. These networks are listed in the order of magnitude of their Q. The magnitude of the Q for a quartz crystal is about 10,000 while a resistance-capacitance network might have a Q of 2. The common frequency range of a crystal controlled oscillator is 100 kc/s to 10 mc/s. A resistance-capacitance oscillator might have a range of frequencies from a fraction of a cycle per second to several megacycles per second.

The frequency determining network used in the oscillator to be described in this thesis is a resistance-capacitance network usually referred to as a Wien Bridge. (Ref. No. 3)

There are several reasons for this choice. It is one of the few practical networks for sub-audio frequencies. The frequency of oscillation is inversely proportional to resistance and capacitance rather than inversely proportional to the square root of these quantities. The oscillations produced are of exceptional purity when the network is used with a linear amplifier. The elements of the network are few and simple and are readily available in very stable forms.

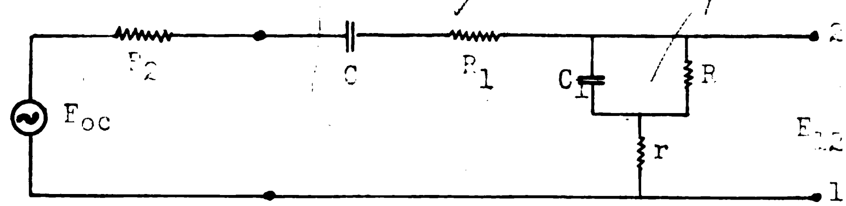
The network in its simplest form is shown in Fig. 4.



Wien Bridge Frequency Determining Network

Fig. 4

It is modified slightly when it is used in an oscillator. If the circuit of Fig. 1 is broken at the points 1-2 and the amplifier is replaced by the equivalent circuit of Thevinen's Theorem, (Ref. No. 4) the circuit becomes that of Fig. 5.



Equivalent Circuit of Amplifier and Frequency Determining Network

Fig. 5

The Barkhausen condition for oscillation requires that the ratio $\frac{E_{oc}}{E_{12}}$ be exactly equal to the gain of the amplifier. It is proposed to make the gain of the amplifier a pure, real number. An analysis of the network with this in view will give the gain requirements of the amplifier and the requirements of the network to produce any desired frequency.

Assume no output current and define Z_1 as the series impedance of R_1 and C. Define Z_2 as the parallel impedance of R and C_1.

Then $\frac{E_{oc}}{E_{12}} = \frac{r + R_2 + Z_1 + Z_2}{r + Z_2} = 1 + \frac{R_2 + Z_1}{r + Z_2}$

This will be a pure, real number if $\frac{r}{R_2} = \frac{Z_2}{Z_1}$

Since $\frac{R_2}{r}$ is a real number, $\frac{Z_1}{Z_2}$ must also be a real number.

$$Z_1 = \frac{1 + jwCR_1}{jwC} \qquad Z_2 = \frac{R}{1 + jRC_1w}$$

$$\frac{Z_1}{Z_2} = \frac{1 - w^2RR_1CC_1 + j(wR_1C + wRC_1)}{jRCw}$$

Then $\frac{E_{oc}}{E_{12}}$ will be a real number at the frequency w_o. Where $w_o^2RR_1CC_1 = 1$. And at this frequency $\frac{Z_1}{Z_2} = \frac{R_1}{R} + \frac{C_1}{C}$. Let $\frac{R_1}{R} = N$ and $\frac{C_1}{C} = M$. Then at w_o, $\frac{E_{oc}}{E_{12}} = 1 + N + M$ if $\frac{R_2}{r} = N + M$.

Or, if the oscillator is to produce the frequency w_o, the amplifier must have a gain of $1 + N + M$. The frequency determining network must be such that $RR_1CC_1 = \frac{1}{w_o^2}$ and $\frac{R_2}{r} = N + M$.

It is extremely difficult to design an amplifier whose gain is a real number over a wide range of frequencies. The effect of phase shift in the amplifier can be most easily understood if the transfer characteristics of the circuit of Fig. 5 are presented as a function of frequency.

$$\frac{E_{12}}{E_{oc}} = \frac{r + Z_2}{r + R_2 + Z_1 + Z_2}$$

$$Z_1 = NR\left(1 - j\frac{w_o}{w}\sqrt{\frac{M}{N}}\right) \qquad Z_2 = \frac{R}{1 + j\frac{w}{w_o}\sqrt{\frac{M}{N}}}$$

If $\frac{R_2}{r} = N + M$

$$\frac{E_{12}}{E_{oc}} = \frac{G(w)}{1 + N + M}\,E^{j\theta(w)}.$$ Where $G(w)$ and $\theta(w)$ are real.

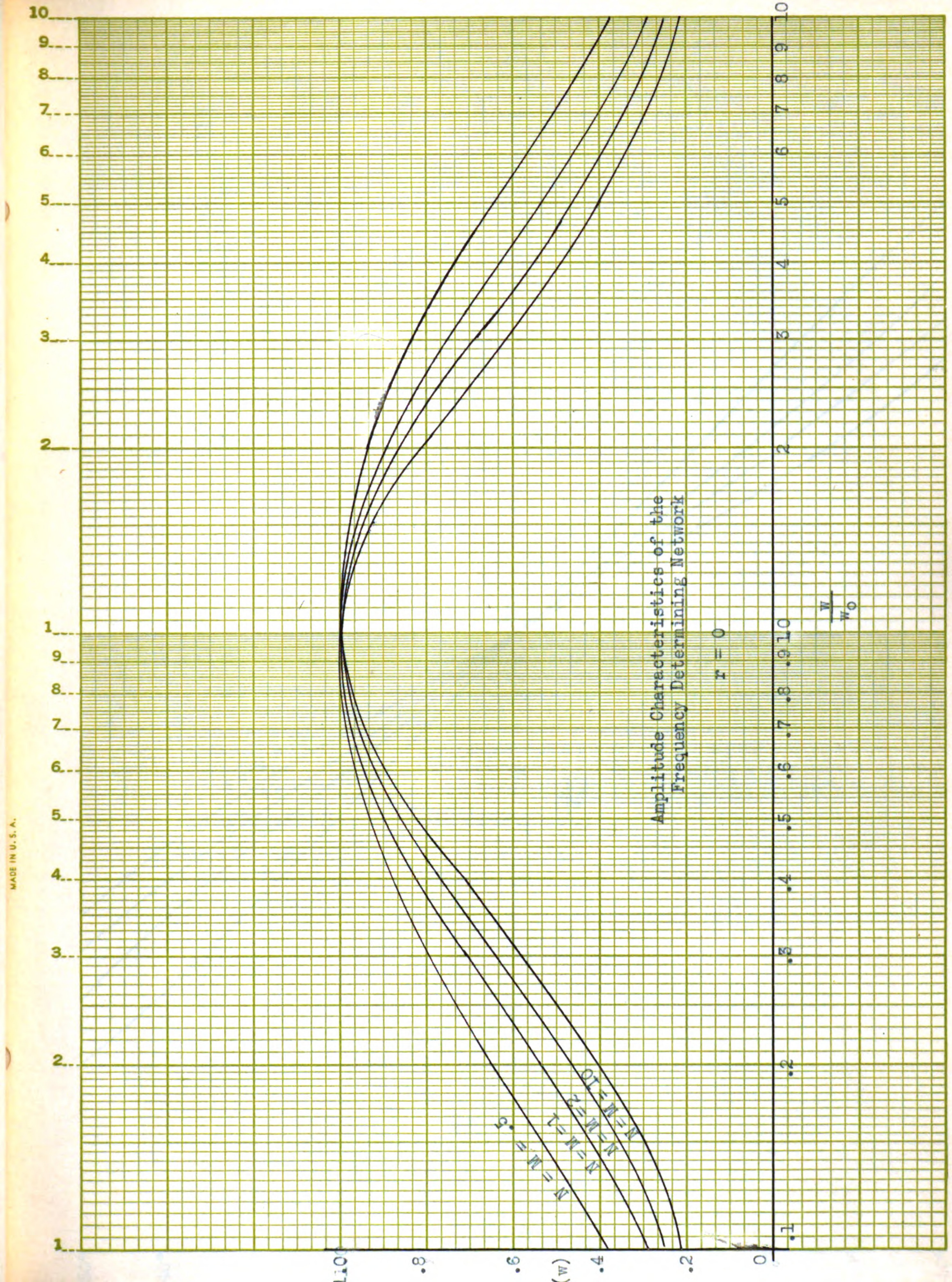
$$G(w) = \frac{\sqrt{1 + \dfrac{w^2}{w_o^2}\dfrac{M}{N}\dfrac{r^2}{(R + r)^2}}}{\sqrt{1 + \dfrac{w^2}{w_o^2}\dfrac{M}{N\,(R + r)^2}\left[r + \dfrac{RN}{1 + N + M}\left(1 - \dfrac{w_o^2}{w^2}\right)\right]^2}}$$

$$\tan\theta(w) = \frac{-\dfrac{w}{w_o}\,\dfrac{\sqrt{NM}}{1 + N + M}\,\dfrac{1}{1 + \frac{r}{R}}\left(1 - \dfrac{w_o^2}{w^2}\right)}{1 + \dfrac{w^2}{w_o^2}\dfrac{M}{N}\left[\left(\dfrac{1}{1 + \frac{R}{r}}\right)^2 + \dfrac{2}{2 + \frac{R}{r} + \frac{r}{R}}\,\dfrac{N}{1 + N + M}\left(1 - \dfrac{w_o^2}{w^2}\right)\right]}$$

When $w \cong w_o$, $\theta(w) \cong \tan\theta(w)$ and

$$\theta(w) \cong \frac{\dfrac{w}{w_o}\,\dfrac{\sqrt{NM}}{1 + N + M}\,\dfrac{2}{1 + \frac{r}{R}}\left(1 - \dfrac{w}{w_o}\right)}{1 + \dfrac{w^2}{w_o^2}\dfrac{M}{N}\left[\dfrac{1}{\left(1 + \frac{R}{r}\right)^2} - \dfrac{2}{2 + \frac{R}{r} + \frac{r}{R}}\,\dfrac{N}{1 + N + M}\left(1 - \dfrac{w}{w_o}\right)\right]}$$

Graphs of these functions are given on the following pages.


Amplitude Characteristics of the
Frequency Determining Network
r = 0
$\frac{w}{w_o}$
N = M = .5
N = M = 1
N = M = 2
N = M = 10
(w)
1.00
.8
.6
.4
.2
0
.1
.2
.3
.4
.5
.6
.7
.8
.9
1.0
2
3
4
5
6
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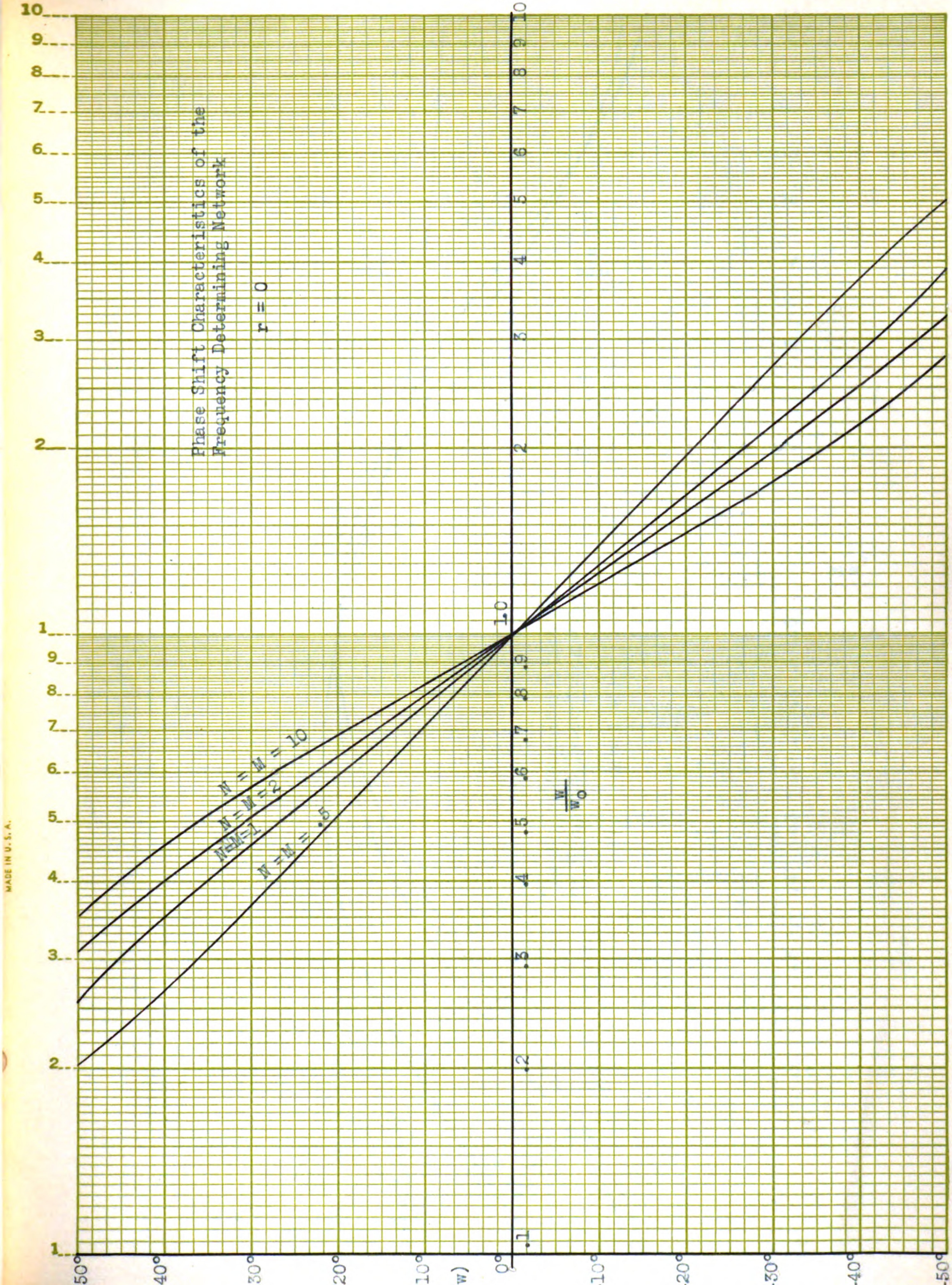
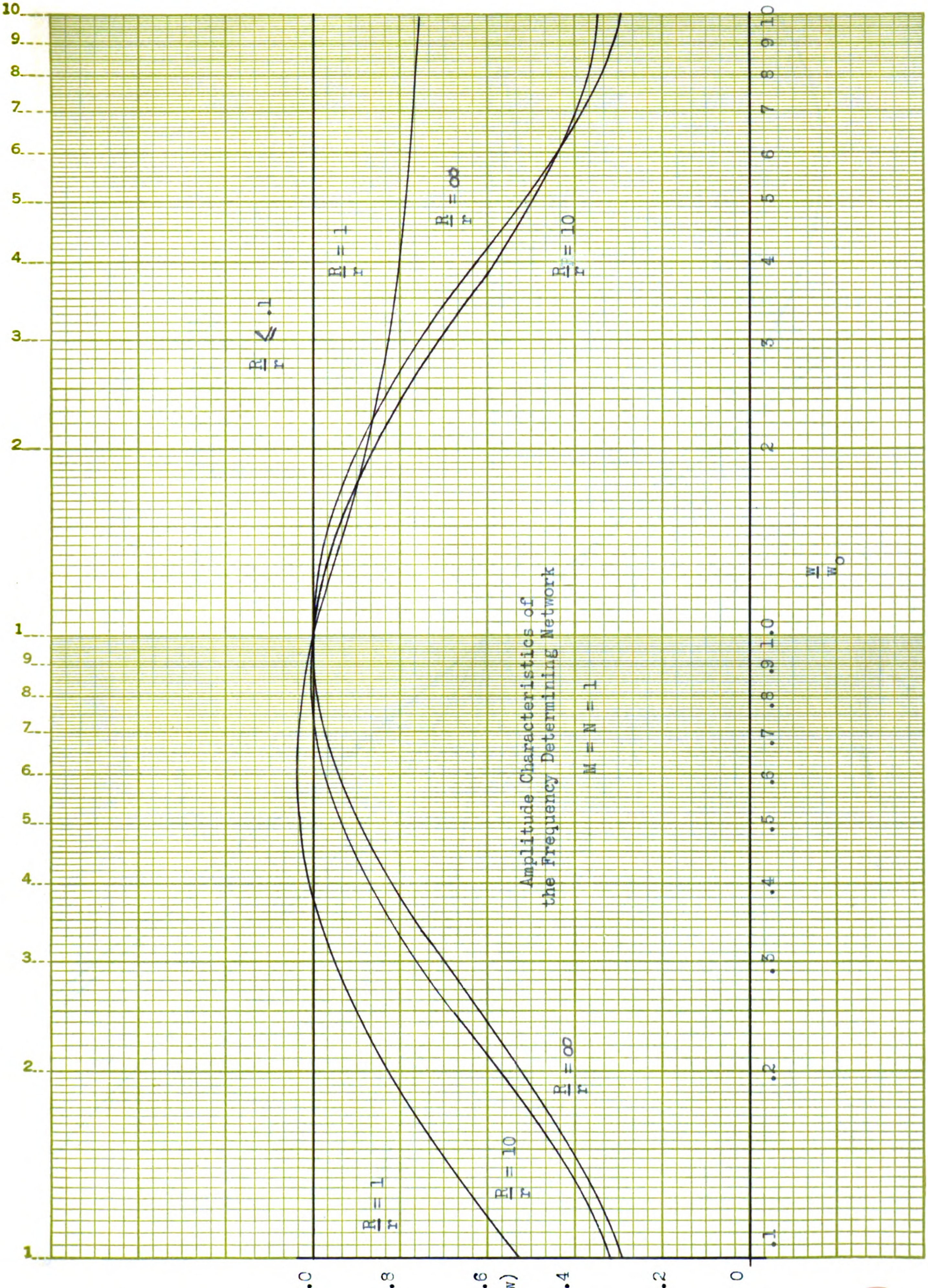
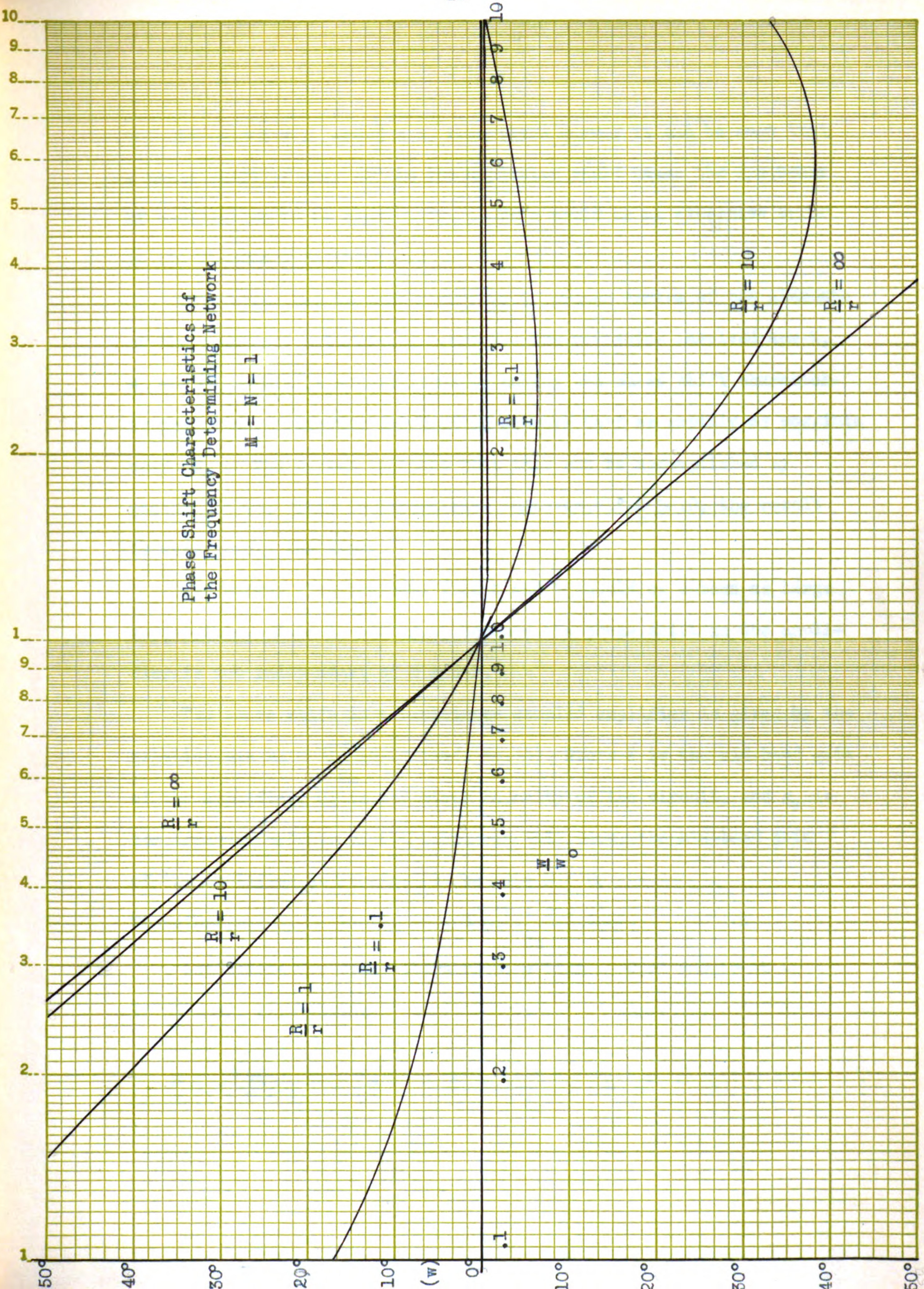

Phase Shift Characteristics of the
Frequency Determining Network
r = 0
N = M = 10
N = M = 2
N=M=1
N = M = .5
$\frac{w}{w_0}$
50° 40° 30° 20° 10° 0° 10° 20° 30° 40° 50°
.1 .2 .3 .4 .5 .6 .7 .8 .9 1.0 2 3 4 5 6 7 8 9 10
MADE IN U. S. A.


Amplitude Characteristics of
the Frequency Determining Network
M = N = 1
R/r ≤ .1
R/r = 1
R/r = 10
R/r = ∞
R/r = 1
R/r = 10
R/r = ∞
w/w_o
.1 .2 .3 .4 .5 .6 .7 .8 .9 1.0 2 3 4 5 6 7 8 9 10
0 .2 .4 .6 .8 1.0


Phase Shift Characteristics of
the Frequency Determining Network
M = N = 1
R/r = ∞
R/r = 10
R/r = 1
R/r = .1
w/w_o
(w)
50° 40° 30° 20° 10° 0° 10° 20° 30° 40° 50°
.1 .2 .3 .4 .5 .6 .7 .8 .9 1.0 2 3 4 5 6 7 8 9 10

If $N = M = 1$ and $\frac{R}{r} = 1$, from these curves it can be seen that a phase shift of only -5° in the amplifier would cause the frequency of oscillation to be about $0.7w_o$. This would be an excessive error in the frequency calibration.

This error can be made a minimum by proper choice of the elements of the frequency determining network. The variables involved are N, M, and $\frac{R}{r}$. This is accomplished by making the phase shift characteristics of the frequency determining network vary as rapidly as possible near w_o. The curves indicate that this can be accomplished by making $\frac{R}{r} = \infty$. This is impossible. However, if $\frac{R}{r} \gg 10$, the result is very nearly the same.

The curves also indicate that M and N should be chosen as large as possible. This is not true, because the phase shift in the amplifier is not independent of M and N.

The phase shift in a resistance coupled amplifier is given by the relation $\tan \theta_a = \frac{-wKC_s}{g_m}$. (Ref. No. 5) Where K is the mid-band gain of the amplifier, C_s is the stray capacity of the circuit, and g_m is the transconductance of the tube. In this case, the mid-band gain of the amplifier is required to be $1 + N + M$.

Therefore, $$\tan \theta_a = - \frac{wC_s(1 + N + M)}{g_m} .$$

Conditions for oscillation require that $\theta(w) + \theta_a = 0$.

If $\frac{R}{r} = \infty$, this yields

$$- \frac{w \sqrt{NM}}{w_o(1 + N + M)}\left(1 - \frac{w_o^2}{w^2}\right) - \frac{w\, C_s\,(1 + N + M)}{g_m} = 0$$

Which gives:

$$\left(1 - \frac{w_0^2}{w^2}\right) = -\frac{C_s w_0 (1 + N + M)^2}{g_m \sqrt{NM}} .$$

This relation is a measure of the theoretical dial calibration error. It is a minimum when $N = M = 0.5$. If $N = M = 1$; the ratio of the value of this function to its minimum value is $\frac{9}{8}$. This is no great increase in calibration error, and a choice of $M = N = 1$ would simplify construction considerably by reducing the number of components of different values.

For these reasons choices of $N = M = 1$ and $\frac{R}{r} \geqslant 10$ are nearly optimum.

The approximation that results when considering $r = 0$ is not an unreal one. In commercially available oscillators r is not present in most circuits. However, R_2 is present in the form of the impedance seen looking back into the output of the amplifier. This is ordinarily small compared to R and does not introduce a serious error in the dial calibration. (A serious error here means an error of the order of magnitude of the precision of the dial setting. The circuits that neglect R_2 are usually tuned by a variable air condenser. The precision of the dial setting is about 1%. If greater precision is required, it is usually necessary to consider R_2 .)

If the frequency of oscillation is to be varied by a variable condenser, R_2 can be neglected for approximate tuning oscillators. R_2 can be made a part of R_1 for more exact tuning oscillators. In the latter case $N = \frac{R_1 + R_2}{R}$.

If the frequency is to be varied by a decade conductance for R_1 and R, r should exist physically for greatest ease of construction and minimum dial calibration error. R_2 can be made quite small if the last stage of the amplifier is a cathode follower; this will help keep $\frac{R}{r}$ large.

There are advantages and disadvantages to each of these methods of tuning the oscillator.

A variable air condenser will give a continuous frequency spectrum over the range of the condenser. The tuning condenser ordinarily can give values of C over a ten to one range. It is possible to get a continuous frequency spectrum over a wide range of frequencies by including a band switch to change the values of resistance. The precision of dial setting can be made as great as practical by increasing the dial size and/or adding a worm gear drive to the condenser. The physical size of the variable condenser limits the lowest frequency of such an oscillator to about 20 c.p.s. The phase shift in the amplifier limits the upper frequency of such an oscillator.

If the tuning unit consists of a decade conductance, the frequency spectrum of the oscillator is not continuous. However, if the decade conductance is a four dial conductance, the points of the frequency spectrum will be separated by 100 parts in a million. This separation is small enough to be considered continuous for frequencies up to several megacycles. This figure is also the limit of dial calibration of the best variable air condenser. By using a four dial decade conductance, it is possible to set the frequency over four decade bands with a single condenser. This is not possible with the variable condenser. The lower frequency limit of oscillation with a decade conductance tuning is a fraction of a cycle per second. It is limited by the maximum resistance value that may be placed in the grid circuit of the amplifier input, and the physical size of satisfactory fixed condensers. The upper frequency limit is determined by the phase shift in the amplifier and the impedance of the frequency determining network. The decade conductance does offer great ease and precision of the frequency setting.

The upper frequency limit is about the same regardless of how the tuning is accomplished. Any phase shift in the amplifier would cause the oscillator to oscillate at a frequency lower than the calibrated frequency. From the curves it is evident that a phase shift of but one degree will cause the oscillator to oscillate at a frequency approximately 95% of the dial setting. The phase shift in the amplifier is a direct function of the frequency.

The smallest value of the tuning condenser, either fixed or variable, should be from 5 to 10 times the stray capacity of the circuit, or 50 to 100 micro-micro-farads.

The smallest value of the tuning resistor, either fixed or variable, is determined by the output impedance of the amplifier. When the output stage of the amplifier is a cathode follower, a low output impedance stage, the smallest permissible value of tuning resistor is about $\frac{5}{g_m}$ of the cathode follower if the distortion is to be kept small.

These limitations on the circuit at higher frequencies place the highest frequency of satisfactory operation at about one megacycle.

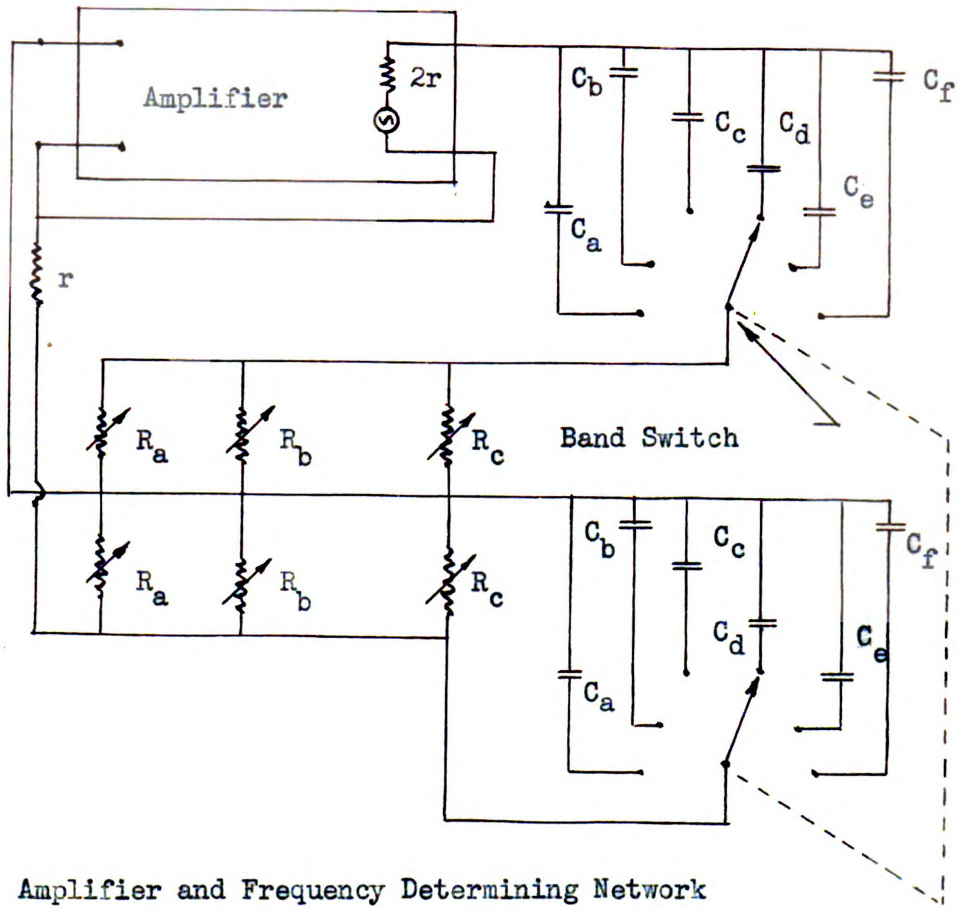
In view of the previous discussion, a design for a frequency determining network using a three dial conductance is given in Fig. 6. R_a, R_b, and R_c are all constructed as shown in Fig. 7. A conductance G, appears for R_a between terminals 1-1'. A conductance 2G, appears for R_a between terminals 2-2'. A conductance 3G, appears between terminals 3-3' for R_a .

A conductance $\frac{G}{10}$ appears between terminals 1-1' for R_b, a conductance $\frac{2G}{10}$ appears between terminals 2-2' for R_b , etc. R_c is similar except they have nominal values over 100. The combination of R_a , R_b , and R_c in parallel thus form a three dial decade conductance.

Reasonable values for these components are:
$\frac{10^5}{\pi}$ ohms between terminals 1-1' on R_a . Remaining values for R_a , R_b , and R_c can be determined as outlined above. C_a = 5uf, C_b = .5uf, C_c = .05uf, C_d = .005uf, C_e = .0005uf, and C_f = .00005uf.

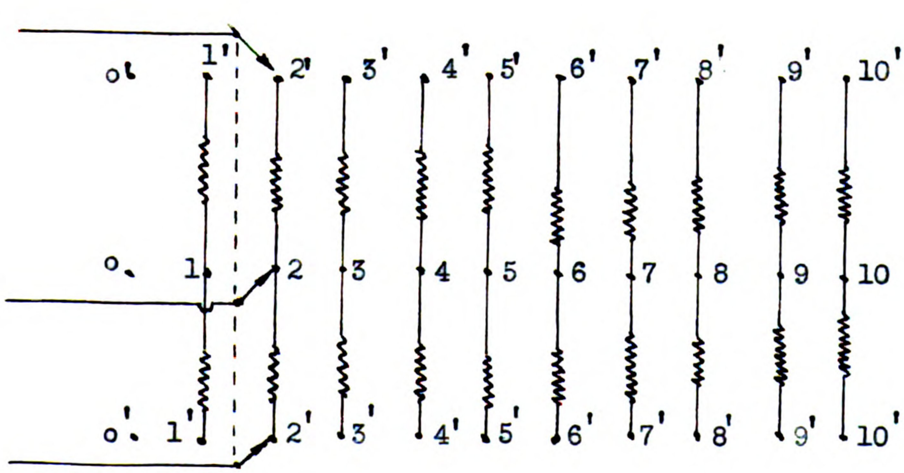
This network produces a range of frequencies as listed below:

Band	Frequency Range	Smallest Incremental Frequency	Resistors	Condensers
1	.01-.1 cps	.01 cps	R_c	C_a
1	.1 - 1 cps	.01	R_b, R_c	C_a
1	1 - 10 cps	.01	R_a, R_b, R_c	C_a
2	10 -100 cps	.1	"	C_b
3	100 - 1 kcs	1	"	C_c
4	1 - 10 kcs	10	"	C_d
5	10 -100 kcs	100	"	C_e
6	100 -1000 kcs	1000	"	C_f



Amplifier and Frequency Determining Network

Fig. 6



Circuit of R_a

Fig. 7

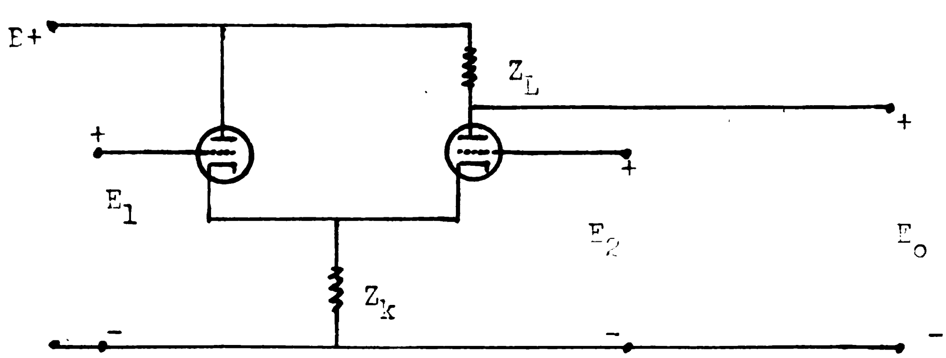
This circuit produces a truly decade oscillator. The minimum resistance is 3103 ohms excluding r. The minimum capacity is 50 uuf.

<u>The Amplifier</u>:

The amplifier requirements as implied or stated in the previous sections are as follows. It should have a gain of 3 + j0 over the frequency range .01cps to 1000kcs. The output impedance should be less than 300 ohms. The input impedance should be very large; at least 10 megohm. The distortion should be small.

The above frequency requirements make it necessary to use a direct-coupled amplifier. (Ref. No. 6) The cathode-coupled amplifier is such a circuit, and it has recently achieved considerable attention for this application. (Ref. No. 7,8,9,10,11, and 12)

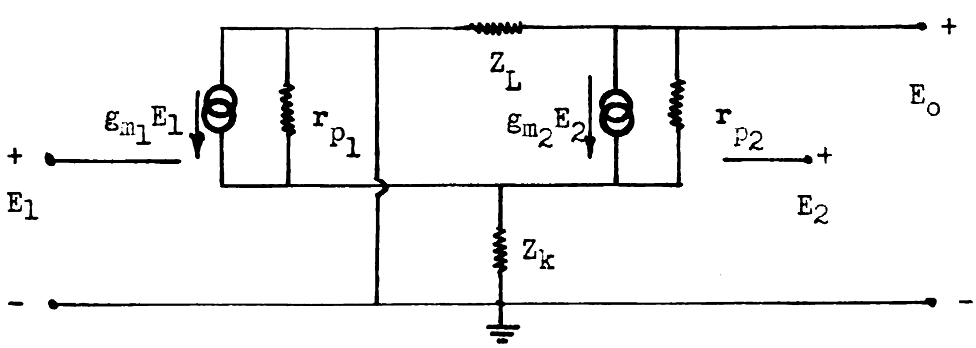
Fig. 8 illustrates the circuit for this amplifier.



Actual Circuit of a Cathode-Coupled Amplifier

Fig. 8

Applying the equivalent plate circuit theorem, (Ref. No. 13) this becomes Fig. 9



Equivalent Circuit of a Cathode-Coupled Amplifier

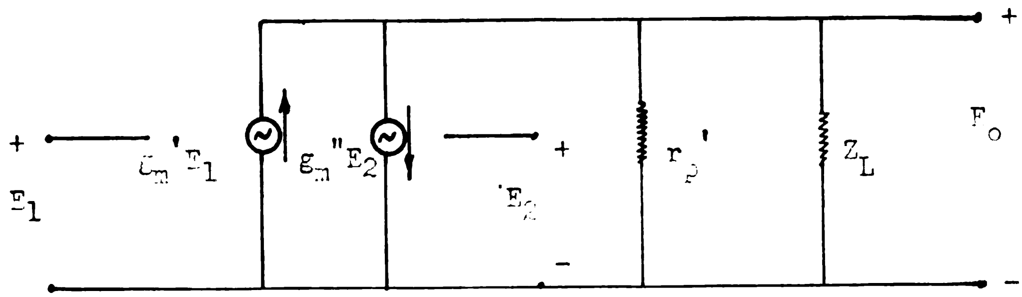
Fig. 9

An analysis of this circuit gives the output voltage, E_o , in terms of the input voltages E_1 , and E_2 , and the circuit parameters.

$$E_o = \frac{E_1\, g_{m_1} g_{m_2}\left(1 + \dfrac{1}{g_{m_2} r_{p_2}}\right)}{\left[g_{m_1} + g_{m_2} + \dfrac{1}{r_{p_1}} + \dfrac{1}{r_{p_2}} + \dfrac{1}{Z_k}\right]\left[\dfrac{1}{Z_L} + \dfrac{1}{r_{p_2}}\,\dfrac{\left(g_{m_1} + \dfrac{1}{r_{p_1}} + \dfrac{1}{Z_k}\right)}{\left(g_{m_1} + g_{m_2} + \dfrac{1}{r_{p_1}} + \dfrac{1}{r_{p_2}} + \dfrac{1}{Z_k}\right)}\right]}$$

$$- \frac{E_2 g_{m_2}}{\left[1 + \dfrac{g_{m_2} + \dfrac{1}{r_{p_2}}}{g_{m_1} + \dfrac{1}{r_{p_1}} + \dfrac{1}{Z_k}}\right]\left[\dfrac{1}{Z_L} + \dfrac{1\left(g_{m_1} + \dfrac{1}{r_{p_1}} + \dfrac{1}{Z_k}\right)}{r_{p_2}\left(g_{m_1} + g_{m_2} + \dfrac{1}{r_{p_1}} + \dfrac{1}{r_{p_2}} + \dfrac{1}{Z_k}\right)}\right]}$$

This equation would apply equally well to the simpler circuit of Fig. 10.



A Second Equivalent Circuit for a Cathode-Coupled Amplifier

Fig. 10

In this circuit

$$g_m'' = \frac{g_{m_2}\left[1 + \dfrac{1}{g_{m_1}(1 + \frac{1}{u_1})Z_k}\right]}{\dfrac{g_{m_2}(1 + \frac{1}{u_1})}{g_{m_1}(1 + \frac{1}{u_2})} + 1 + \dfrac{1}{g_{m_1}(1 + \frac{1}{u_1})Z_k}}$$

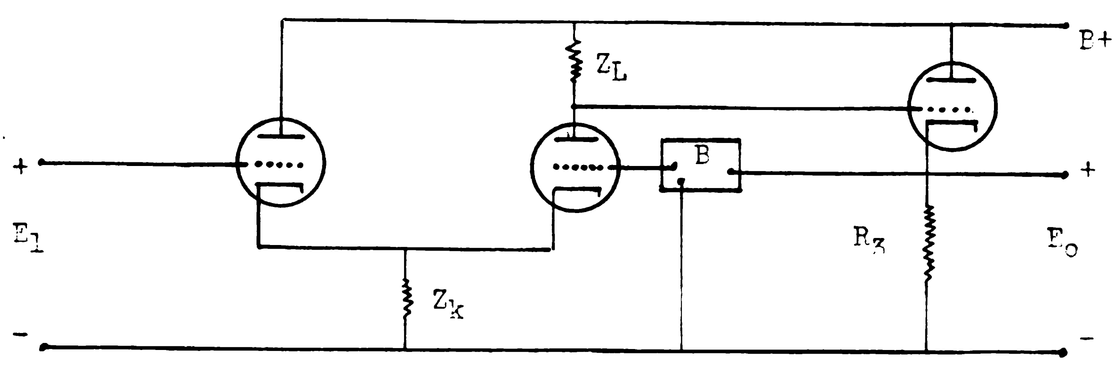
$$g_m' = \frac{g_{m_2}}{\dfrac{g_{m_2}}{g_{m_1}} + \dfrac{(1 + \frac{1}{u_1})}{(1 + \frac{1}{u_2})}\left[1 + \dfrac{1}{g_{m_1}(1 + \frac{1}{u_1})Z_k}\right]}$$

$$r_p' = \frac{r_{p_2}\left[\dfrac{g_{m_2}(1 + \frac{1}{u_1})}{g_{m_1}(1 + \frac{1}{u_2})} + 1 + \dfrac{1}{g_{m_1}(1 + \frac{1}{u_1})Z_k}\right]}{1 + \dfrac{1}{g_{m_1}(1 + \frac{1}{u_1})Z_k}}$$

The output impedance of this amplifier can be seen to be $\frac{r'Z_L}{r_p' + Z_L}$. This is ordinarily considerably greater than 300 ohms using conventional tubes. For this reason, the amplifier is direct coupled to a cathode follower. (Ref. No. 14) The output impedance of a cathode follower is approximately $\frac{1}{g_m}$ for the tube used and can be easily made to be less than 300 ohms.

The gain equation for the amplifier indicates that negative feedback could be utilized by making $E_2 = BE_o$. Strictly speaking, this is not negative feedback since the signal is not returned to the input of the amplifier. The advantages of applying negative feedback to this amplifier are a smaller output impedance, less distortion, and a more stable amplifier. (Ref. No. 15) Although in this case the feedback signal is not returned to the input, these advantages are retained to a high degree since the input stage is a cathode follower -- a very stable, low distortion circuit.

The amplifier with a cathode follower output and negative feedback network are shown in Fig. 11.



Amplifier with Cathode Follower Output Stage and Negative Feedback Network

Fig. 11

The gain of this circuit is $$\frac{E_o}{E_1} = \frac{g'_m r'_p Z_L A_{oe}}{(Z_L + r'_p)\left(1 + \frac{g''_m r'_p Z_L A_{oe} B}{Z_L + r'_p}\right)}$$

Where A_{cf} is the gain of the cathode follower.

Define $A = \dfrac{g'_m r'_p Z_L A_{cf}}{r'_p + Z_L}$

Then $\dfrac{E_o}{E_1} = \dfrac{A}{1 + \dfrac{g''_m}{g'_m} AB}$

This will be a real number if Z_L , Z_k , and B are real.

In general, the larger the quantity $BA\dfrac{g''}{g'}$, the greater will be the reduction of distortion and output impedance of the amplifier. This is accomplished by choosing tubes with a large transconductance and plate resistance, by choosing a large Z_L , and by making $Z_k \gg \dfrac{1}{g_{m_1}}$. This combination of conditions together with the conditions for a real gain invariably make it necessary to provide a grid bias supply. This complicates the power supply and introduces additional sources of noise and hum into the circuit. Greater band width for the amplifier is also partially assured by choosing tubes with a large transconductance. (Ref. No. 16)

It has been suggested that the plate resistance of a pentode be used for Z_k. (Ref. No. 6) This would insure that Z_k would be very much greater than $\dfrac{1}{g_{m_1}}$.

Fig. 12 gives a circuit that is a fair compromise of these various considerations. A conservative estimate of the tube parameters and stray capacity of the circuit gives the following values with B = 0:

$g'_m = g''_m = 3500$ u mho

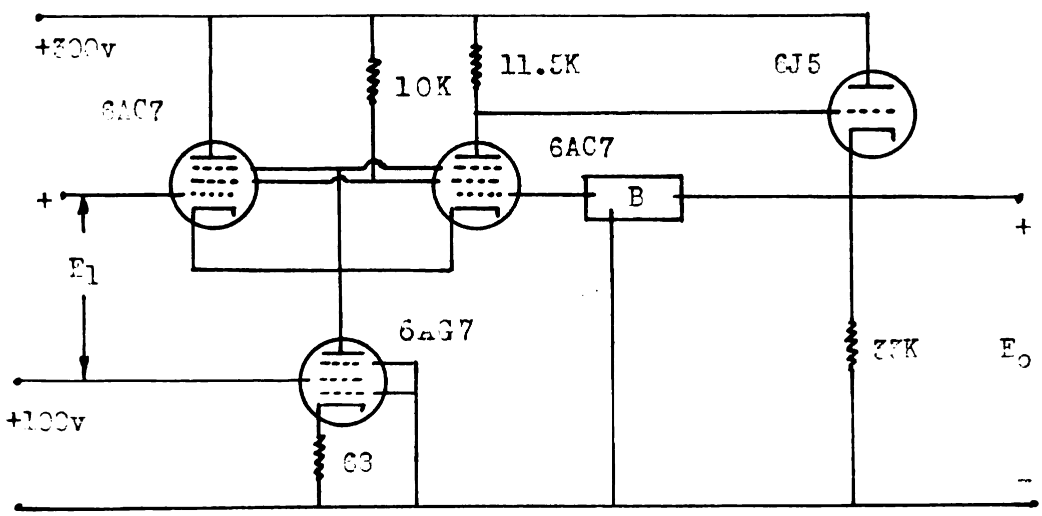
r_p' = 2 megohm

A = 35

Output Impedance = 400 ohms.

Input Impedance - open circuit.

Bandwidth - 0-500kcs.



Amplifier Circuit

Fig. 12

With B = .33

Gain = 3

Input Impedance - open circuit

Output Impedance - 40 ohms

Bandwidth 0-5mcs.

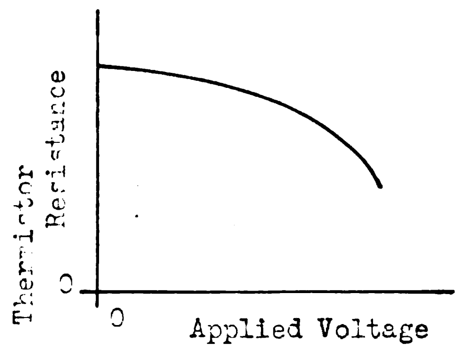
This value of bandwidth is not satisfactory for producing oscillations at 1 mcs, but it is just about all that can be obtained with this circuit.

Amplitude Limit Control

The amplitude of oscillation is determined by the Barkhausen conditions of oscillation. As long as these conditions are satisfied and the amplifier operates in a linear fashion, the amplitude will continue to increase. Obviously, the amplitude cannot continue to increase indefinitely. In many oscillators, amplitude control is accomplished by allowing the oscillations to build up until the amplifier is operating in a non-linear fashion. This generates distortion components which are fed back and amplified again. If the frequency determining network has a high Q, the loop gain for these distortion components is low and a reasonably pure sine wave results. If the Q of the frequency determining network is low, the loop gain may be nearly the same for all frequencies, and the output will suffer considerable distortion.

A more satisfactory method of amplitude control is to operate the amplifier in a linear range and control the amplitude by some external non-linear element. A number of circuits have been tested and all of them give excellent results. (Ref. No. 17, 18, 19, 20, 21, 22 and 11) A method suggested by Becker, Green, and Pearson seemed well adapted for this problem. (Ref. No. 23) This method uses a thermistor for the non-linear element.

The thermistor is a thermally sensitive resistor with a large, negative, temperature coefficient. (Ref. No. 24) As its temperature increases the resistance decreases. The heating of the thermistor can be accomplished by applying a voltage to its terminals. The effect is to produce a non-linear resistance. Fig. 13 gives a typical characteristic of thermistors.



Typical Thermistor Characteristics

Fig. 13

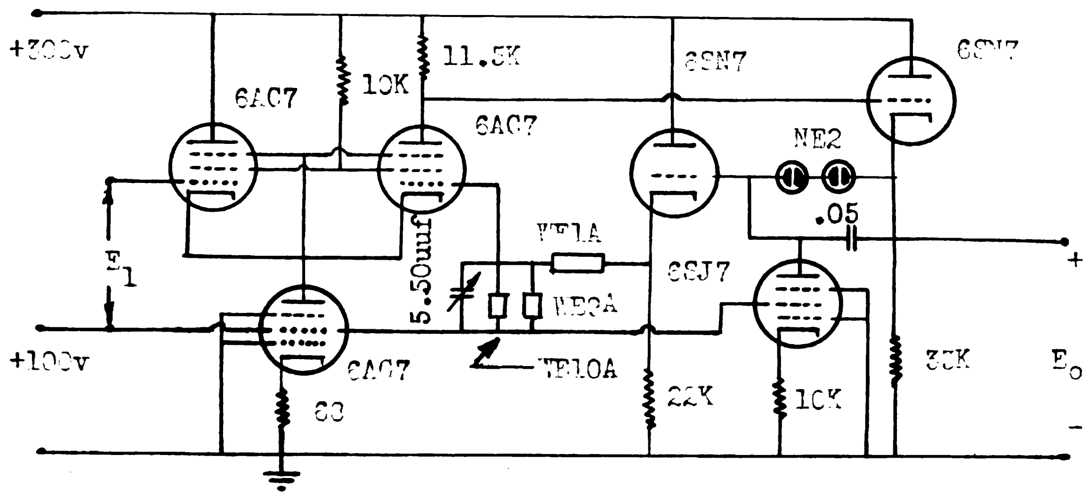
This curve is obtained by applying D-C voltages of various values, allowing the thermistor to come to a thermal equilibrium and measuring the resistance. It would be observed equally well using periodic voltages providing the period of the voltage were short compared with the thermal time constant of the thermistor. This condition is necessary to insure that the thermistor temperature remains constant throughout the voltage cycle. Under this condition the thermistor would appear like an ordinary linear resistor at any given value of applied voltage. If the voltage were increased, and if the thermistor were allowed to come to thermal equilibrium, the thermistor would again appear like an ordinary resistor of smaller value than that observed previously. The application of this property to the present problem can best be understood by considering the gain equation of the amplifier.

The gain of amplifiers with a large amount of negative feedback is given approximately by $\frac{1}{B}$. Where B is the feedback factor. (Ref. No. 15) The gain is nearly independent of tube parameters and circuit constants other than the feedback network. The logical place to use the thermistor then, is in the negative feedback network.

The requirements of this network are that B shall be a constant for cycalic variations of voltage of short period and fixed amplitude, but B shall increase with an increase in amplitude.

Fig. 14 shows the feedback and amplitude control network and its connection into the amplifier circuit. The feedback factor, B, is determined by the voltage divider network composed of the thermistors and the trimmer condenser. The function of the WE9A and WE10A thermistors is to compensate the network for changes in ambient temperature. These thermistors all have about the same temperature coefficient. However, the WE1A thermistor shows a much more rapid change of resistance with low voltage than do the other two. From the curves of Fig. 15 it can be seen that for voltages up to 30 volts across the network, the parallel combination resistance of the WE9A and the WE10A thermistors does not change, while the resistance of the WE1A thermistor will change appreciably. Since the resistance-voltage characteristic of a thermistor is a function of ambient temperature, this combination is virtually independent of temperature.

The function of the two neon lamps (NE2) is to direct couple the two halves of the 6SN7 and maintain a fixed potential difference between their cathodes. (Ref. No. 25). The plate resistance of the 6SJ7 completes this coupling scheme. The purpose of this method of



Circuit of Amplifier and Amplitude
Control Network

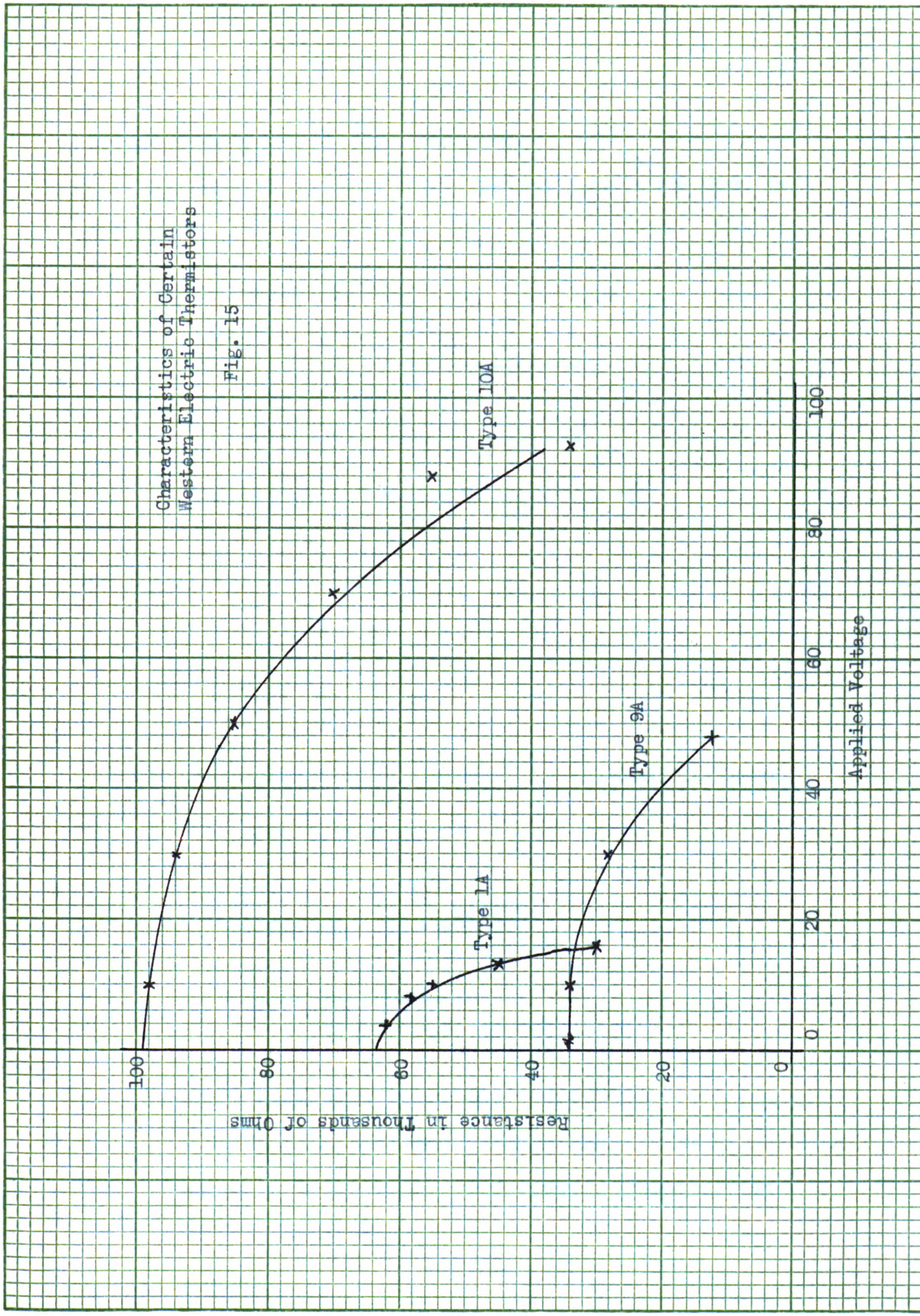
Fig. 14

coupling is to balance the direct current component of voltage out of the thermistor network and hence increase the sensitivity of the amplitude control.

The trimmer condenser compensates the amplifier for phase shift at high frequencies. Its operation is as follows. Due to the stray capacity of the circuit, the gain of the amplifier becomes a complex number at high frequencies. It is given by the equation

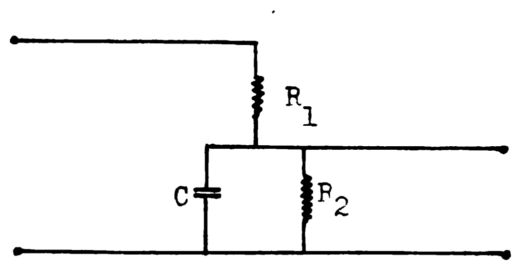
$$\frac{E_o}{E_1} = \frac{A}{1 + j\frac{w}{w''} + B\,\frac{A g_m''}{g_m}} \qquad \text{(Ref. No. 5)}$$

where A and B are defined as before, and w'' is the upper half-power frequency of the amplifier. If B is a real number, the gain is complex; and, as pointed out earlier, the phase shift becomes excessive at high frequencies. At 500 kcs this phase shift becomes great enough to cause an error of 15% in the frequency calibration. A


Characteristics of Certain
Western Electric Thermistors
Fig. 15
Type 10A
Type 1A
Type 9A
Applied Voltage
0
20
40
60
80
100
Resistance in Thousands of Ohms
100
80
60
40
20
0

complex B could improve this considerably.

The essential elements of the feedback network are shown in Fig. 13.



Elements of the Negative Feedback Network

Fig. 16

B for this network is given by

$$B = \frac{R_2}{R_1 + R_2}\left[\frac{1 - j\dfrac{wR_1R_2C}{R_1 + R_2}}{1 + \left(\dfrac{wR_1R_2C}{R_1 + R_2}\right)^2}\right]$$

This is inserted into the gain equation for the amplifier below.

$\dfrac{E_o}{E_1} = \dfrac{A}{D}$ where D is given as

$$D = 1 + \frac{g_m'' A R_2}{g_m'(R_1 + R_2)\left[1 + \left(\dfrac{wCR_1R_2}{R_1 + R_2}\right)^2\right]}$$

$$+ \; j\left\{\frac{w}{w''} - \frac{w\, g_m'' A R_1 R_2 C}{g_m'(R_1 + R_2)^2\left[1 + \left(\dfrac{wCR_1R_2}{R_1 + R_2}\right)^2\right]}\right\}$$

From this equation it can be seen that at any given frequency the phase shift may be made zero, positive, or negative by adjusting C. It can not be made zero for all frequencies, however. For values of $\frac{\omega_n}{\omega} \leqslant .1$, the phase shift can be made practically zero.

The Power Supply:

The power supply requirements are as follows:

41 milliamps at 300 volts

8 milliamps at 100 volts

The 100 volt supply should have an internal impedance of less than 20 ohms. In addition, both of these voltages should be free from hum and ripple and short time fluctuations with line voltage. These latter requirements dictate the use of a voltage regulated power supply. (Ref. No. 28)

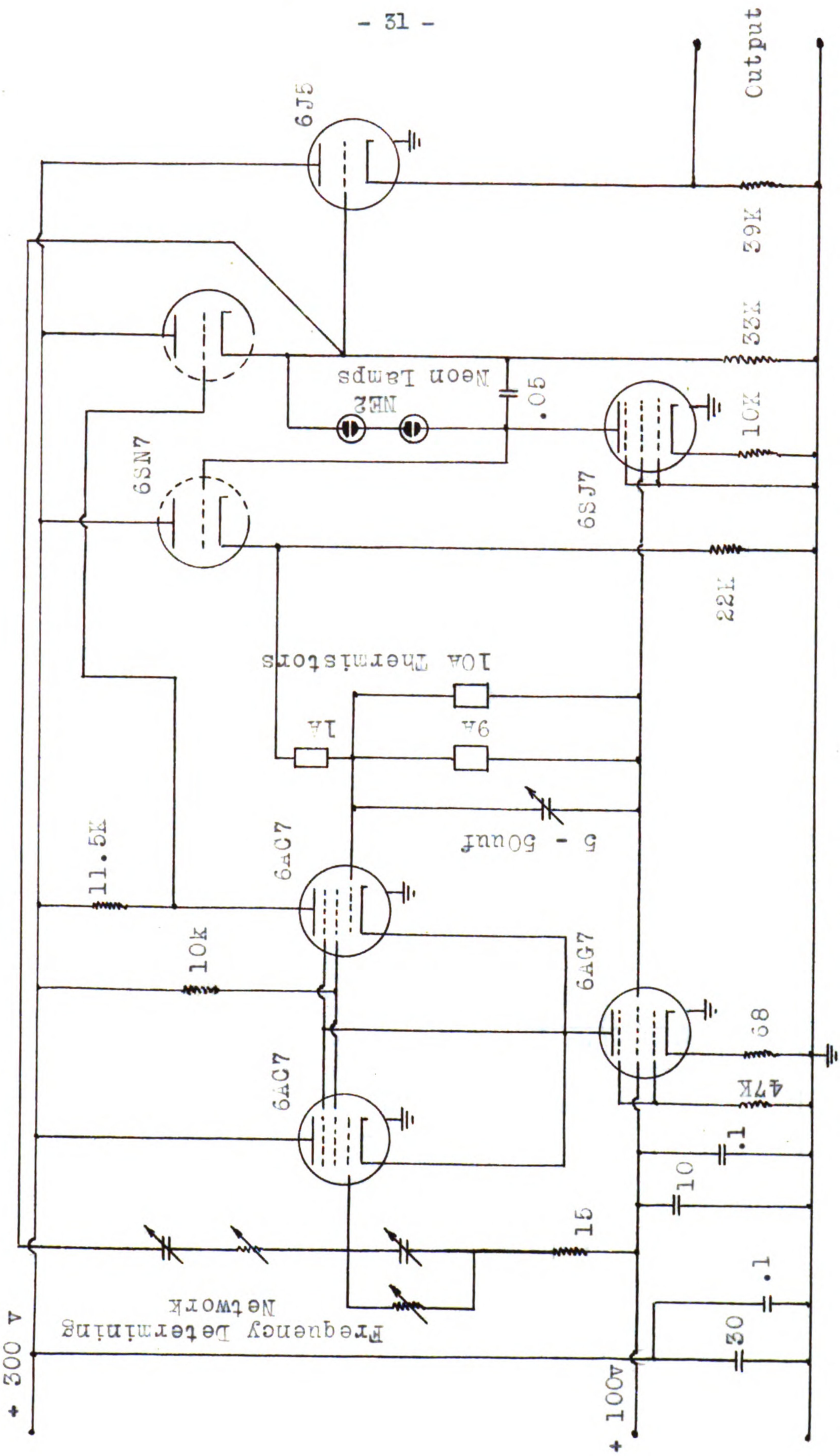
The circuit is given with the complete circuit diagram. It is quite conventional with the exception of the reference voltage tubes. Small neon lamps were used for this application rather than the more conventional "VR" tubes. The current required for these lamps is about .8 milliamps compared to 20 milliamps for the "VR" tubes. Their operation is very satisfactory.

Circuit Diagram and Construction Details:

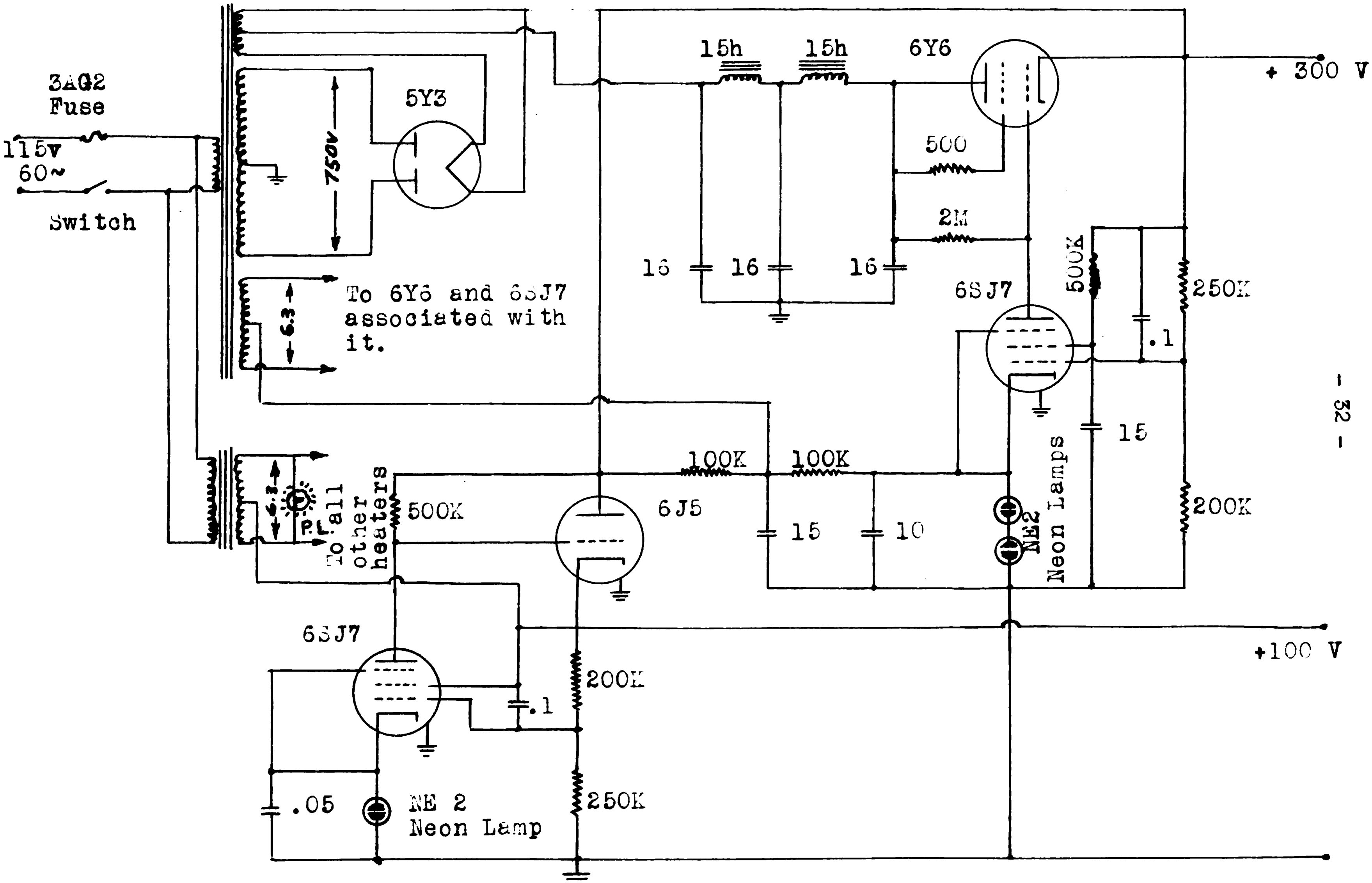
The complete circuit diagram is given on the following pages.

The component parts were mounted on a 17- by 8- by 3-in. steel chassis. Every attempt was made to keep stray capacity to a minimum, and the power supply was completely shielded from the remainder of the circuit in order to reduce hum in the output signal, The remaining construction was conventional.

The most time-consuming operation during the construction period



Oscillator Circuit Diagram



Power Supply Circuit Diagram

was the selection of resistors for the frequency determining network. This network consisted of 60 precision resistors with a tolerance of ± 1%. There is no commercial source for all the values so it was decided to sort them from standard 10% composition resistors.

The method in which this was carried out was as follows: The 3133 ohm resistor pair of R_a of the frequency determining network was sorted by using an impedance bridge. This was then soldered into the selector switch in the oscillator circuit. A convenient pair of condensers were also soldered into the oscillator, and the oscillator made to oscillate. The remaining resistor pairs were selected by choosing resistors slightly higher than the required value and then shurting them with other resistors until the correct Lissajous figure (Ref. No. 26, 27) was produced on an oscilloscope. It was necessary to use an auxiliary oscillator and make frequent checks against the original resistor pair.

Several pictures of the completed oscillator are shown following a discussion of the test results.

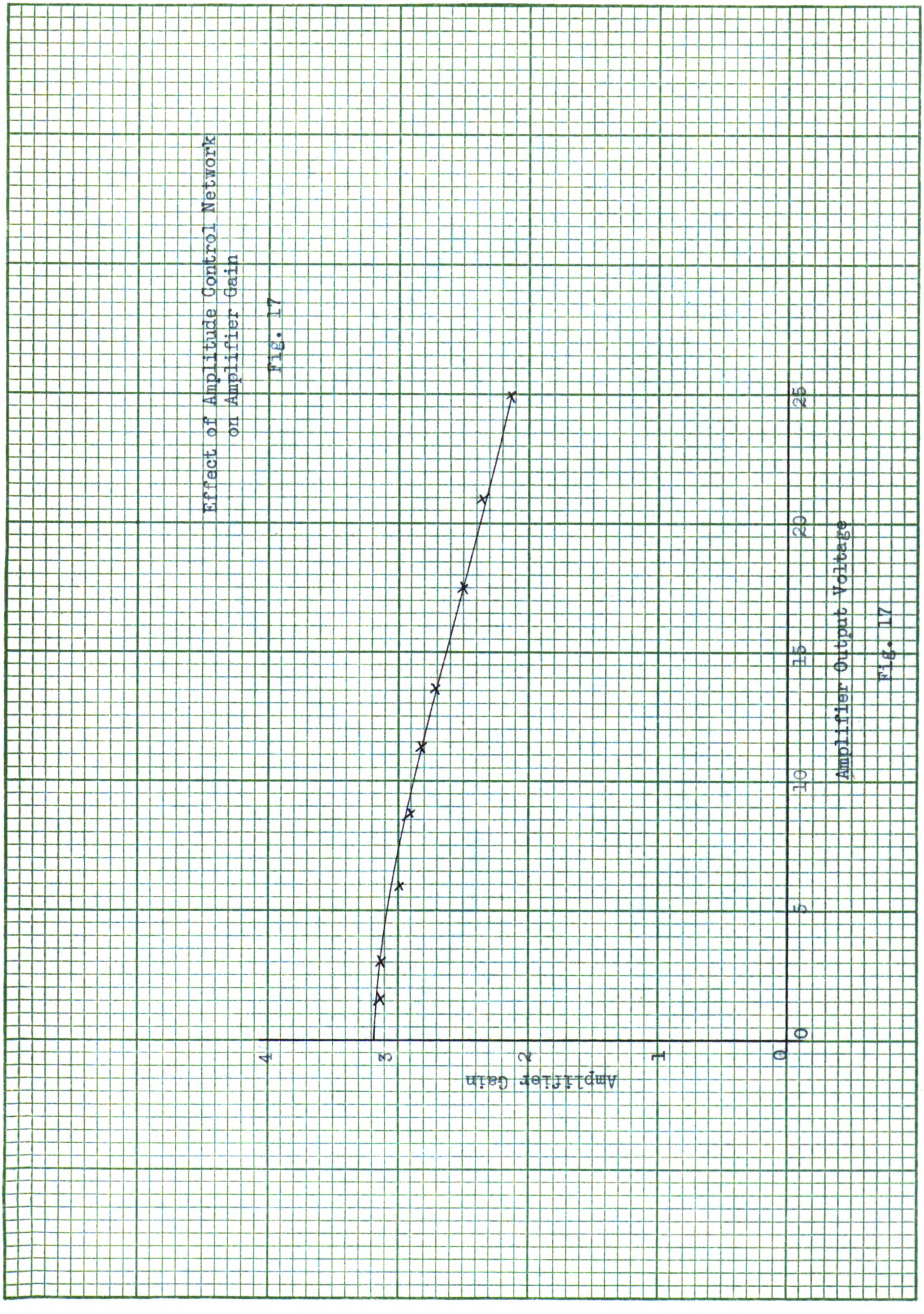
Test Results:

In general, this oscillator performs very well. The amplitude of oscillation is 7 volts ± .5 volts over its complete frequency range.

Between 20 cps and 20 kcs the distortion is less than 0.4%. No test equipment was available for making distoritiion measurements outside of this frequency band. No distortion could be observed on an oscilloscope at any frequency above 0.5 cps. No suitable thermistor was available with a time constant longer than that of the WF1A. Since the time constant of this thermistor is about 30 seconds, the amplifier might be expected to introduce more distortion at frequencies below 0.5 cps. A series of oscillograms is inclosed to illustrate the wave shape at various frequencies.

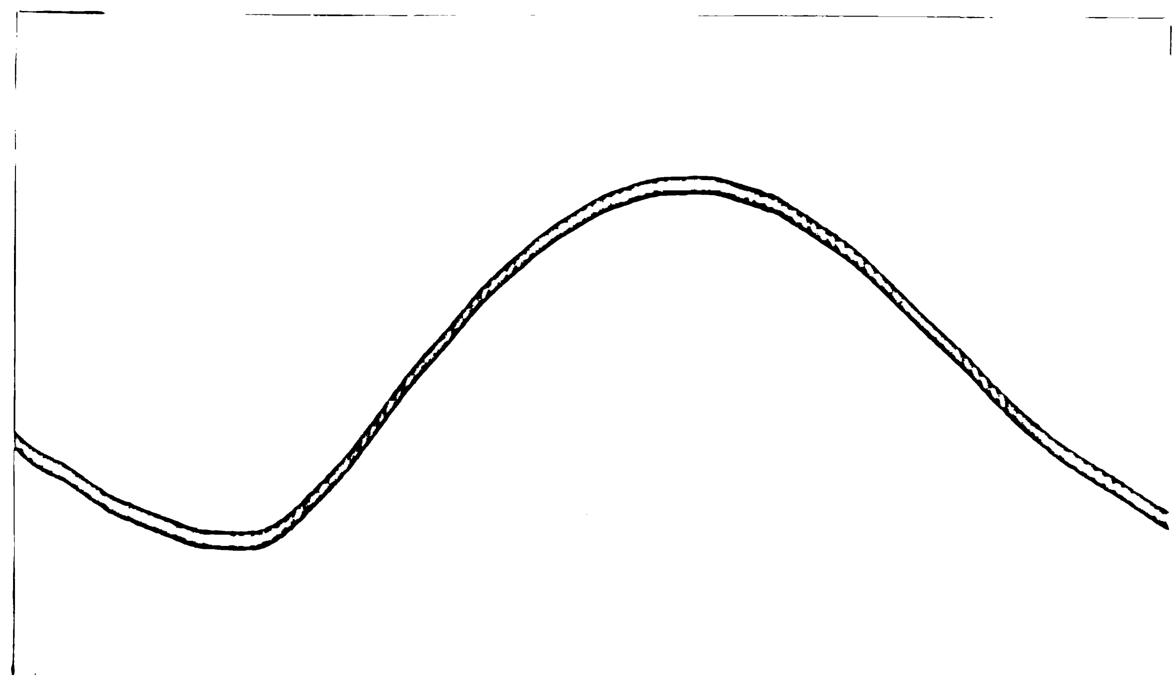
The frequency range of this oscillator is 0.01 cps to 500 kcs. It was hoped to make the frequency range 0.01 cps to 1 mcs; however, it was impossible to compensate the amplifier for phase shift above 600 kcs. Without the phase shift compensation the frequency range was 0.01 cps to 100 kcs.

The effect of the amplitude control circuit is shown in Fig. 17.



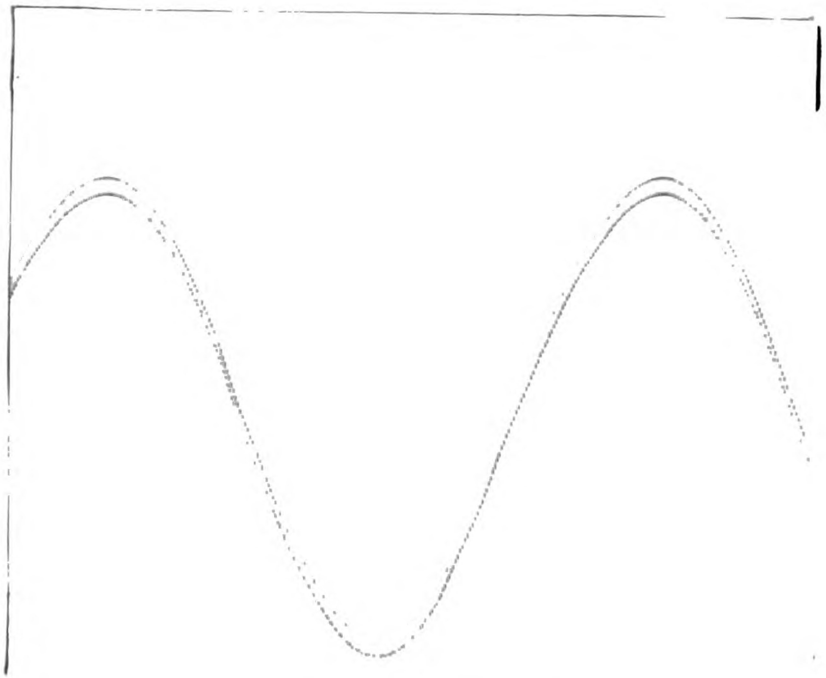
Effect of Amplitude Control Network
on Amplifier Gain

Fig. 17

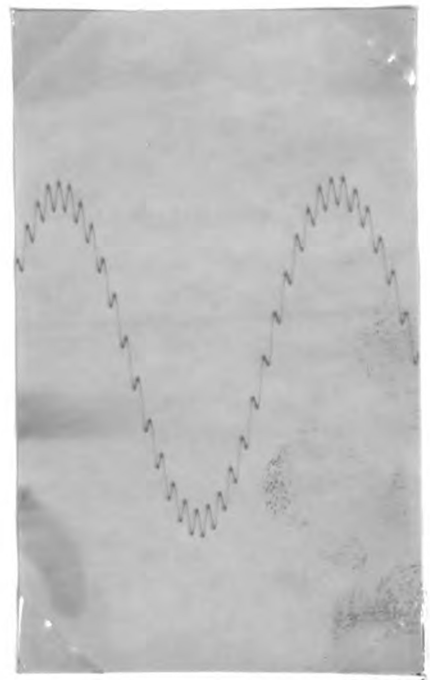


Oscillator Waveform at .05 cps

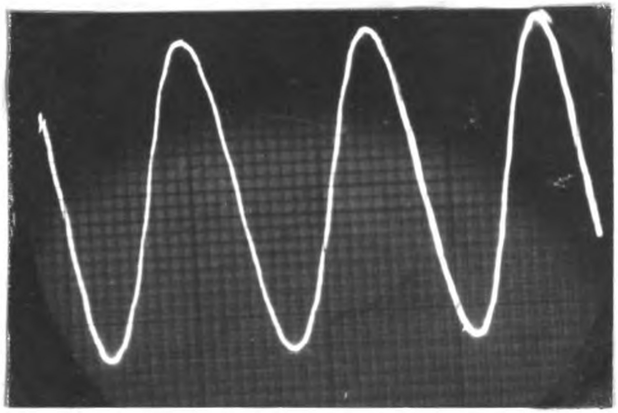
Note: These oscillograms were taken on a Hathaway Model 14 Oscillograph. An amplifier was used between the oscillator and the oscillograph to obtain the necessary driving power. The amplifier had considerable hum in its output and this distorted the waveform somewhat. This can best be seen on the 5 cps oscillogram.



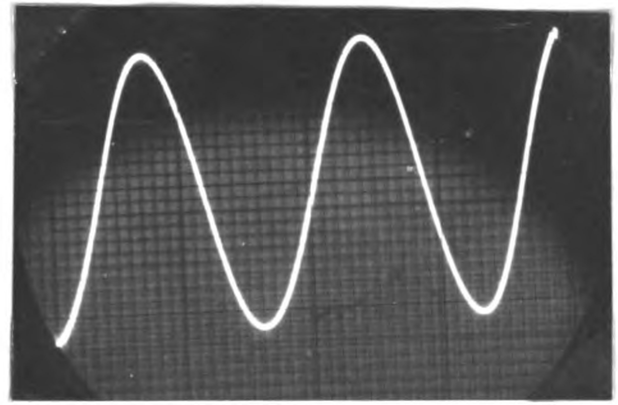
Oscillator Waveform at .5 cps



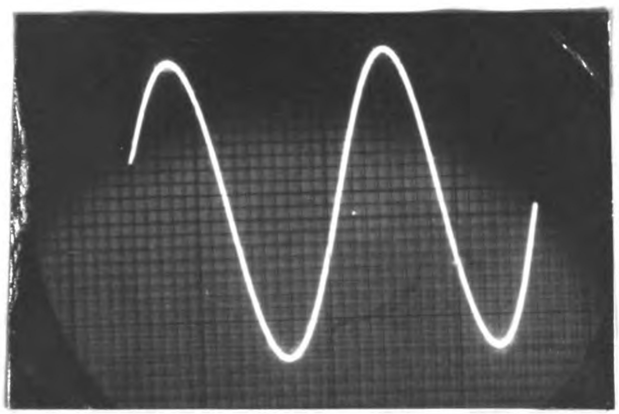
Oscillator Waveform
at 5 cps



Oscillator Waveform
at 5 cps

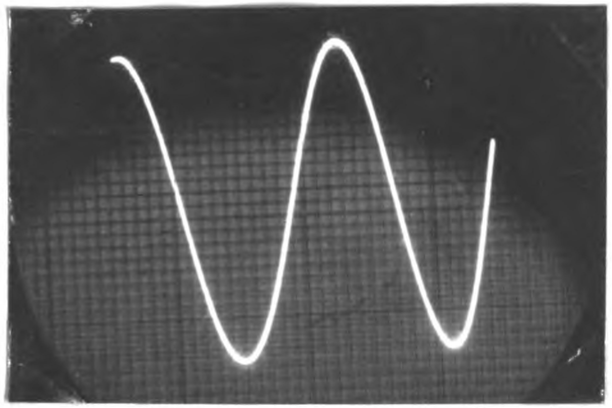


Oscillator Waveform at
50 cps

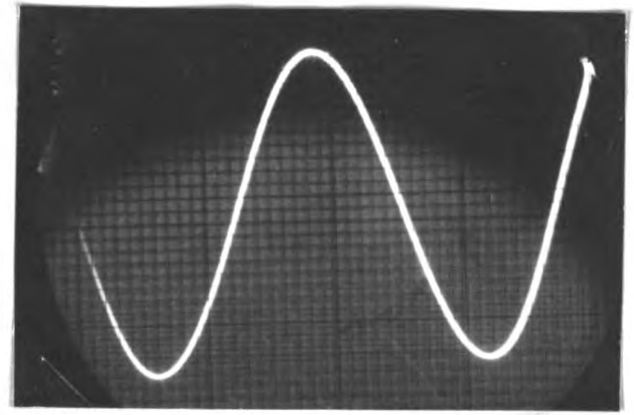


Oscillator Waveform
at 500 cps

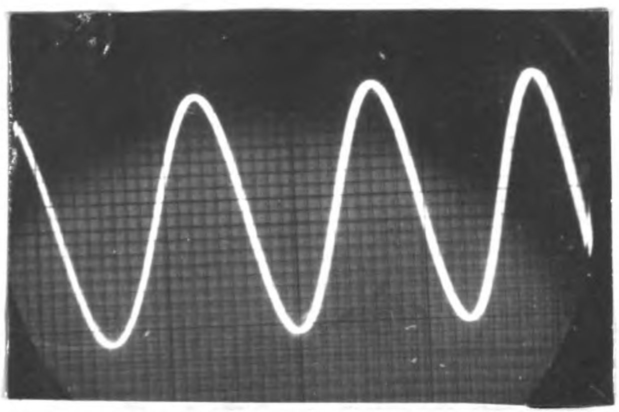
Note: These pictures were taken on Panchromatic film with a time exposure of 1 second. The shutter opening was f 4.5 and the ojbect distance was 30 inches. A DuMont Model 250 oscilloscope was used without the high voltage power supply. The room was nearly dark.



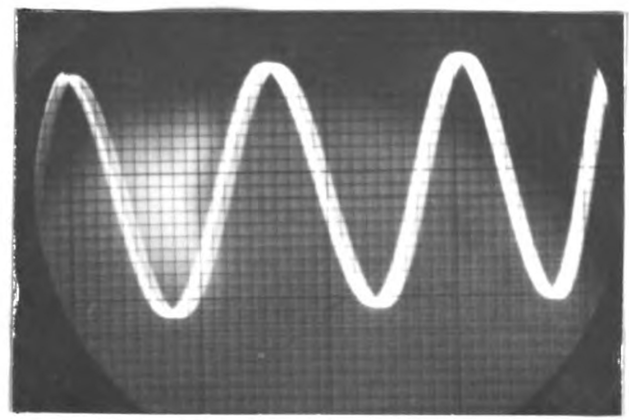
Oscillator Waveform
at 5 kcs



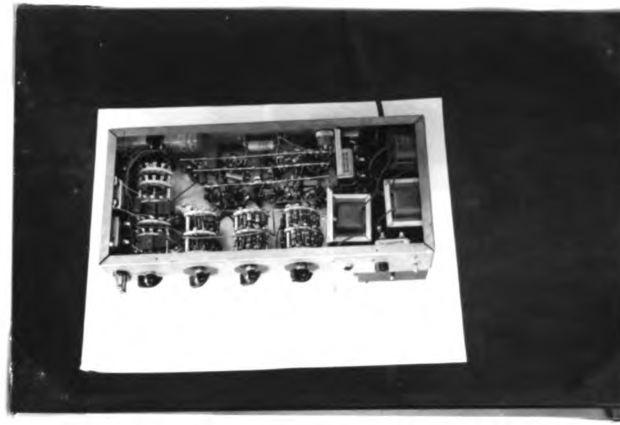
Oscillator Waveform
at 50 kcs

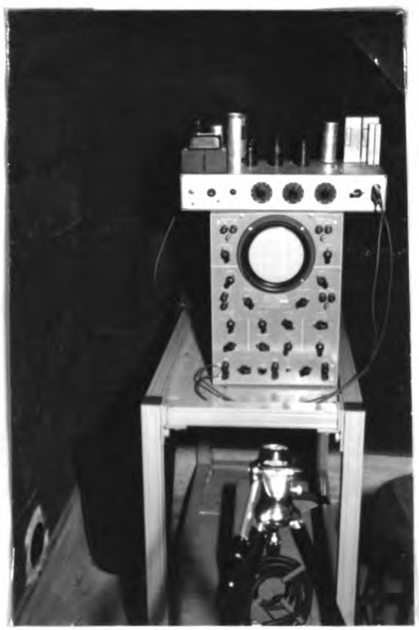


Oscillator Waveform
at 100 kcs



Oscillator Waveform
at 500 kcs





Views showing construction details and method of photographing oscilloscope.

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