FREQUENCY DISCRIMINATION BY INVERSE FEEDBACK

A THESIS

Submitted in partial fulfillment
of the requirements for the Degree of
Master of Science in Electrical Engineering

by

Donald M. Leslie

Georgia School of Technology
Atlanta, Georgia
1941

FREQUENCY DISCRIMINATION BY INVERSE FEEDBACK

Approved:

Date Approved by Chairman May 22 1941

ACKNOWLEDGMENT

I wish to express my sincerest thanks to Assistant Professor M. A. Honnell for his valuable aid and guidance in the prosecution of this work.

TABLE OF CONTENTS

DEFINITIONS OF SYMBOLS USED

A = Amplification of the amplifier without feedback.

A^1 = Amplification of the amplifier with feedback.

β = Attenuation of the feedback control resistance, or ratio of the voltage supplied to the feedback network to the total output voltage. A pure numeric, independent of frequency.

γ = Attenuation of the feedback network. A vector quantity.

ϕ = $\beta\gamma$, a vector quantity.

e_o = Total output voltage.

e_s = Signal input voltage.

R_g = Resistance across which e_s is developed.

R_c = Cathode biasing resistance.

R_t = Resistance terminating the feedback network.

R_f, R'_f = Feedback control resistances such that $\beta \cong \frac{R_f}{R_f + R'_f}$

R_b = Resistance through which the plate voltage is applied to the tube.

R_{sg} = Screen grid voltage supply resistance.

C_c = Cathode by-pass condenser.

C_b = Blocking condenser for the feedback control circuit.

C_{sg} = Screen grid by-pass condenser.

C = Coupling condenser between stages.

Z_o = Characteristic impedance of network.

Z_t = Transfer impedance of network.

f_o = Null frequency, cycles per second.

The term "regeneration" refers to the effect of feedback voltage upon amplifier characteristics when the vector sum of the input signal voltage and the feedback voltage present in the amplifier is greater than the input signal voltage.

The terms "degeneration", "inverse feedback", and "negative feedback" are synonomous in this thesis, and refer to the effect of feedback voltage upon amplifier characteristics when the vector sum of the input signal voltage and the feedback voltage present in the amplifier is less than the input signal voltage.

FREQUENCY DISCRIMINATION BY INVERSE FEEDBACK

INTRODUCTION

The use of negative or inverse feedback in amplifiers has become widespread since the appearance of H. S. Black's paper[1] on the subject in 1934. Much improvement in the operating characteristics of amplifiers may be realized by inverse feedback. Essentially the method of use consists in taking part of the output voltage of the amplifier and returning it to the input circuit in the proper phase relative to the impressed signal. When this is done the re-amplified feedback voltage will tend to cancel the distortion and the noise originating in the amplifier. This advantage is, of course, obtained at the expense of reduced voltage amplification but with the high gain possible with modern tubes this gives little cause for concern. Usually inverse feedback may be added to an existing amplifier or designed in the circuit of an amplifier by the addition of only one condenser and one or two resistances. The benefits may be summarized as follows[2]:

(1) Reduction of amplitude distortion.

(2) Reduction of noise.

(3) Reduction of frequency and phase distortion.

---

[1] Black, H.S., Elec. Eng., vol. 53, Jan. 1934, p. 114

[2] Reich, H.J., Theory and Applications of Electron Tubes, McGraw-Hill, 1939, p.220

(4) Increase of stability (reduction of variation of amplification with operating voltages and tube age).

(5) Reduction of variation of amplification with input voltage.

(6) Reduction of variation of amplification with load impedance.

(7) Increased damping of loud speaker transients and resonance.

In some cases it may be desirable to have a gain-frequency characteristic which is not flat but discriminates against certain frequencies. One way in which this may be accomplished is by the insertion in the feedback circuit of a network having a response characteristic the inverse of that desired for the amplifier. In particular the amplifier may be made highly selective by the use of a frequency-null bridge or network such as the parallel-T network. It is the purpose of this thesis to design and test several amplifiers which achieve frequency discrimination through the use of inverse feedback networks.

ANALYSIS OF INVERSE FEEDBACK

In figure 1, which represents a feedback amplifier in general form, a portion of the output voltage is returned through the feedback network to the input circuit. "A" represents the amplification without feedback, ϕ is the ratio of feedback voltage to the total output voltage. In general both A and ϕ are complex quantities. Neglecting the harmonic and noise components, the output voltage, expressed in terms of the signal voltage, will take the form of the series:

$$e_o = Ae_s\ (1+A\phi + A^2\phi^2 + - - -) \qquad (1)$$

which is convergent if $|A\phi|<1$. The overall amplification, with feedback, is

$$A^1 = \frac{e_o}{e_s} = \frac{Ae_s}{e_s}(1+A\phi+A^2\phi^2 + - - -) \qquad (2)$$

The quantity in parenthesis in the above equation is the power series development of

$$\frac{1}{1 - A\phi}$$

Equation (2) may therefore be written

$$A^1 = \frac{A}{1 - A\phi} \qquad (3)$$

Thus the original amplification is changed by the factor

$$\frac{1}{1 - A\phi}$$

The product $A\phi$ is called the feedback factor and is considered positive when the vector sum of e_s and ϕe_o is greater than e_s, negative when the sum is less than e_s.

The feedback is positive, resulting in increased amplification when $|1-A\phi| < 1$, negative, resulting in decreased amplification when $|1-A\phi| > 1$, and there is no change in gain when $|1-A\phi| = 1$. Thus the magnitude of $1-A\phi$ is the criterion for determining whether feedback is positive or negative, and not the feedback factor.

It may be shown that when the feedback is negative the change in amplification is always such as to result in a decrease in amplitude distortion and frequency distortion, and in some types of noise. Since it is beyond the purpose of this paper to analyze in detail the effect of feedback on noise and distortion these components were omitted from the diagram in Figure 1. A complete analysis from this standpoint may be found in the references listed in the bibliography.

As $A\phi$ is increased beyond the value of three or four the importance of 1 in the gain expression

$$A^1 = \frac{A}{1-A\phi}$$

becomes small in comparison with $A\phi$, particularly if the imaginary component of $A\phi$ is large in comparison with the real component. Thus if the feedback is increased or is applied over a higher gain the overall amplification approaches the limiting value:

$$A^1 = -\frac{1}{\phi} \tag{4}$$

and is independent of A. If $|A\phi| \gg 1$ the amplifier gain is inversely proportional to ϕ, which means that the frequency response characteristic of the amplifier with high feedback is the inverse of the response characteristic of the feedback network. Since $\phi = \beta\gamma$, and β does not introduce any phase shift, the frequency response of the amplifier should be flat when $\gamma = 1$. If γ has a null point the amplifier will have a frequency response peak, or if γ is a low-pass filter the amplifier will have a high-pass characteristic. Or, by proper design it is possible to correct the characteristic due to some section at a point beyond the amplifier by putting a network with identical characteristics in the feedback circuit.

If $|A\phi| \gg 1$ the amplification is independent of load impedance provided the load does not form a part of the feedback circuit. Also the amplification is independent of A and hence of supply voltage and tube factors, which greatly improves the stability of the amplifier.

Nyquist has shown[3] that a great deal may be learned about the operation of a feedback amplifier by plotting a polar diagram of $A\phi$ at all frequencies from zero to infinity. The magnitude of $A\phi$ is plotted to a linear scale against the total phase shift angle. If this curve encloses the point $1\underline{/0}$ the feedback is positive and in sufficient amount to cause the amplifier to oscillate at a particular frequency. If the curve does not enclose $1\underline{/0}$ but does intersect a circle of unit radius with center at $1\underline{/0}$ the feedback is positive at these frequencies but oscillation will not occur. If the curve does not intersect this

[3] Nyquist, H., Bell System Tech. Jour., vol. 11, p. 126

circle at any point the feedback is negative while tangency to the circle means there is no change in gain. The polar diagram for the triode feedback amplifier with pure resistance load is a circle with center at $\frac{A\phi}{2} \angle 180^\circ$ and radius of $\frac{A\phi}{2}$. In general the high values of relative phase shift occur at frequencies for which A is small, so that it is possible to prevent oscillation in one-, two-, or three-stage resistance coupled amplifiers.

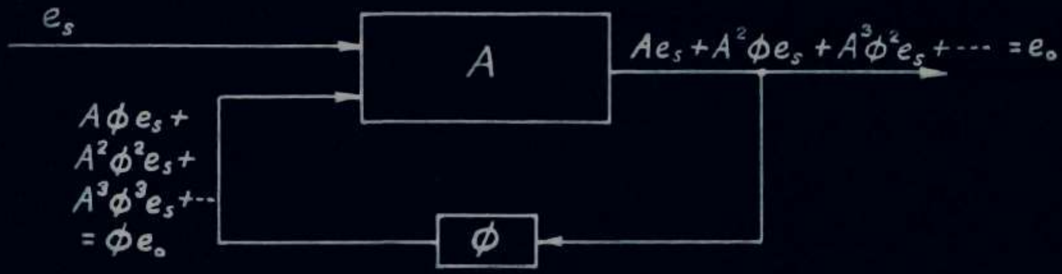


Fig. 1

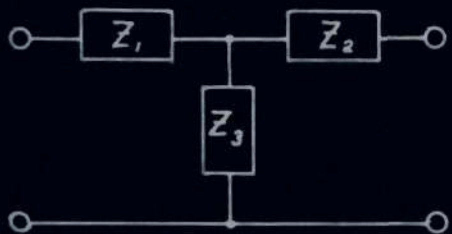


Fig. 2

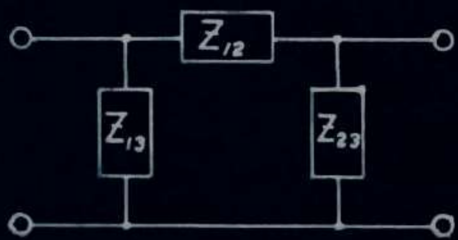


Fig. 3

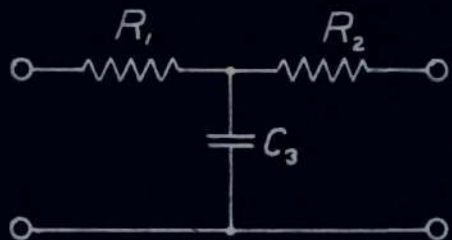


Fig. 4

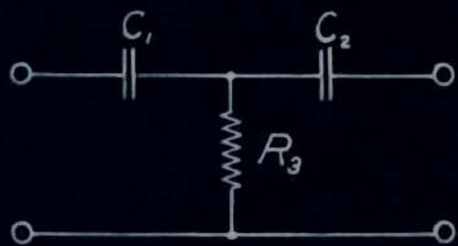


Fig. 5

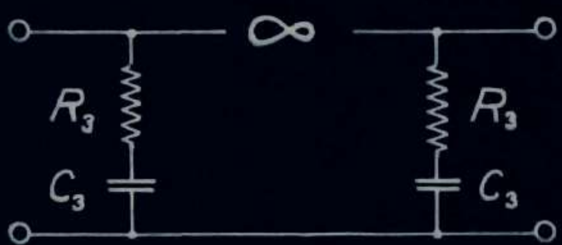


Fig. 6

DML
Ga Tech 41

THE PARALLEL-T NETWORK

In the section on feedback analysis it was shown that if $|A\phi| >> 1$ the frequency response curve of a feedback amplifier will be the inverse of the response characteristic of the network used in the feedback circuit. The types of networks that may be used are practically unlimited, the type used depending on the response characteristic desired. In Figures 9 and 10 are shown configurations used by Fritzinger[4] to obtain high- and low-pass response respectively in an amplifier. Additional sections may be added to increase the sharpness of cut-off.

A resonant circuit or a frequency bridge may be used to provide a null in the feedback path and hence a peak in the amplifier characteristic. Figure 7 shows the Wien bridge which is well suited for this purpose. At balance:

$$\frac{R_1}{R_2} = \frac{\dfrac{R_4 \left(-j\dfrac{1}{wc_4}\right)}{R_4 \quad -j\dfrac{1}{wc_4}}}{R_3 \quad -j\dfrac{1}{wc_3}}$$

$$R_1R_3R_4 - j\frac{R_1R_4}{wc_3} - j\frac{R_1R_3}{wc_4} - \frac{R_1}{w^2c_3c_4} = -j\frac{R_2R_4}{wc_4}$$

In order for this equation to hold the sum of the reals and the sum of the imaginaries must equal zero.

---

[4]Fritzinger, G.H., Proc. I. R. E., vol. 26, Jan. 1938, p. 207

Equating reals:

$$R_1R_3R_4 = \frac{R_1}{w^2c_3c_4}$$

$$w^2 = \frac{1}{R_3R_4c_3c_4} \qquad (5)$$

Equating imaginaries:

$$\frac{R_1R_4}{c_3} + \frac{R_1R_3}{c_4} = \frac{R_2R_4}{c_4}$$

$$\frac{R_1R_4}{c_3} = \frac{R_2R_4 - R_1R_3}{c_4}$$

$$\frac{c_4}{c_3} = \frac{R_2R_4 - R_1R_3}{R_1R_4} = \frac{R_2}{R_1} - \frac{R_3}{R_4} \qquad (6)$$

In many cases it is not convenient to use a bridge because of the necessity of isolating the input terminals. Furthermore the additional frequency distortion and phase shift introduced by the isolating medium may be objectional in many cases. For these reasons the equivalent parallel-T (Figure 8) or bridged-T 3-terminal networks are more desirable. The parallel-T will be considered in some detail herewith.

To show the equivalent circuit of the parallel-T network at the null frequency each component T is first reduced to an equivalent pi section and then the two sections are paralleled. Using the T to pi transformation equations (notations in Figures 2 and 3):

$$Z_{12} = \frac{Z_1Z_2 + Z_2Z_3 + Z_3Z_1}{Z_3}$$

$$Z_{23} = \frac{Z_1Z_2 + Z_2Z_3 + Z_3Z_1}{Z_1}$$

$$Z_{13} = \frac{Z_1Z_2 + Z_2Z_3 + Z_3Z_1}{Z_2}$$

For the T section shown in Figure 4

$$Z_{12} = \frac{R_1R_2 - jR_2X_3 - jR_1X_3}{-jX_3}$$

Rationalizing

$$Z_{12} = \frac{jR_1R_2 + R_2X_3 + R_1X_3}{X_3}$$

$$= j\,\frac{R_1R_2}{X_3} + R_2 + R_1 = (R_1 + R_2) + j\,\frac{R_1R_2}{X_3}$$

$$Z_{23} = \frac{R_1R_2 - jR_2X_3 - jR_1X_3}{R_1} = R_2 - j\left(\frac{R_2X_3}{R_1} + X_3\right)$$

$$Z_{31} = \frac{R_1R_2 - jR_2X_3 - jR_1X_3}{R_2} = R_1 - j\left(\frac{R_1X_3}{R_2} + X_3\right)$$

For the T section shown in Figure 5

$$Z_{12} = \frac{-X_1X_2 - jX_2R_3 - jX_1R_3}{R_3} = -\frac{X_1X_2}{R_3} - j\,(X_1 + X_2)$$

$$Z_{23} = \frac{-X_1X_2 - jX_2R_3 - jX_1R_3}{-jX_1}$$

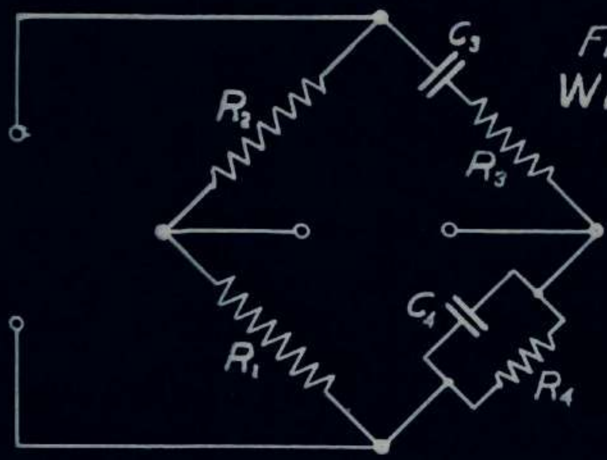


Fig. 7
Wien Bridge

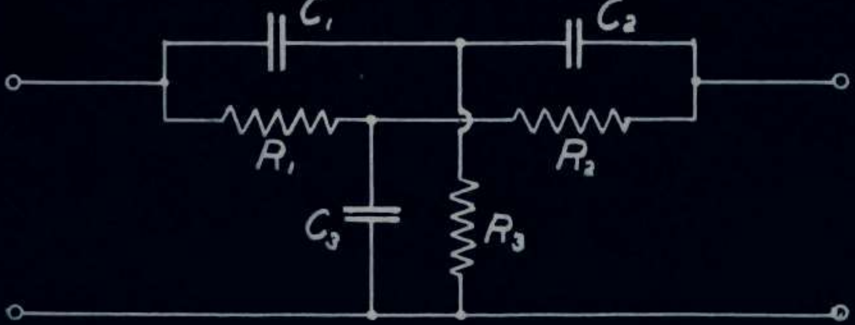


Fig. 8
Parallel T Network

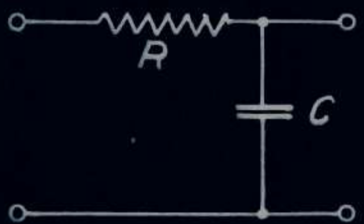


Fig. 9
Low-pass Section

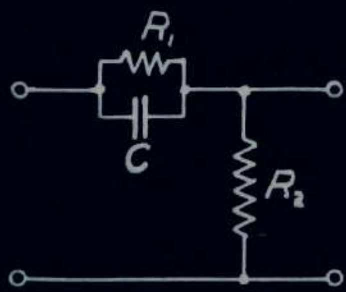


Fig. 10
High-pass Section

DML
Ga Tech-41

Rationalizing

$$Z_{23} = \frac{-jX_1X_2 + X_2R_3 + X_1R_3}{X_1} = \left(\frac{R_3X_2}{X_1} + R_3\right) - jX_2$$

$$Z_{31} = \frac{-X_1X_2 - jX_2R_3 - jX_1R_3}{-jX_2}$$

Rationalizing

$$Z_{31} = \frac{-jX_1X_2 + X_2R_3 + X_1R_3}{X_2} = \left(\frac{R_3X_1}{X_2} + R_3\right) - jX_1$$

For design convenience let[5] $R_1 = R_2 = 2R_3 = R$

$X_1 = X_2 = 2X_3 = X$

Then the equivalent pi for figure 4 becomes:

$$Z_{12} = 2R + j\frac{2R^2}{X}$$

$$Z_{23} = Z_{31} = R - jX$$

And for figure 5:

$$Z_{12} = -\frac{2X^2}{R} - j2X$$

$$Z_{23} = Z_{31} = R - jX$$

---

[5]This assumption will be justified on page 16

Paralleling the two pi's the series element is:

$$Z'_{12} = \frac{\left(2R + j\,\frac{2R^2}{X}\right)\left(-\frac{2X^2}{R} - j2X\right)}{2R - \frac{2X^2}{R} + j\left(\frac{2R^2}{X} - 2X\right)}$$

$$= \frac{-4X^2 - j8RX + 4R^2}{\frac{2}{R}(R^2 - X^2) + j\frac{2}{X}(R^2 - X^2)} = \frac{R^2 - X^2 - j2RX}{(R^2 - X^2)\left(\frac{1}{2R} + j\frac{1}{2X}\right)}$$

Rationalizing and simplifying:

$$Z'_{12} = \frac{-\frac{R}{2} - j\frac{X}{2} - \frac{X^2}{2R} - j\frac{R^2}{2X}}{\frac{R^2}{4X^2} - \frac{X^2}{4R^2}}$$

$$= \frac{R + \frac{X^2}{R}}{\frac{X^2}{2R^2} - \frac{R^2}{2X^2}} + j\,\frac{X + \frac{R^2}{X}}{\frac{X^2}{2R^2} - \frac{R^2}{2X^2}} \tag{7}$$

The shunt elements of the equivalent pi, equal if balanced, are:

$$Z'_{23} = Z'_{13} = \frac{(R - jX)\,(R - jX)}{2(R - jX)} = \frac{R - jX}{2}$$

In terms of the original resistance and reactance:

$$Z'_{23} = Z'_{13} = R_3 - jX_3 \tag{8}$$

At one particular frequency, the balance frequency, $R = X$. The deonminator of Z'_{12} becomes zero and Z'_{12} becomes infinite. Then the equivalent pi circuit is that shown in Figure 6.

The characteristic impedance at the null frequency is equal to

$$Z_o = R_3 - jX_3 \tag{9}$$

since $Z_o = \sqrt{Z_{oc}\,Z_{sc}}$ and $Z_{oc} = Z_{sc} = R_3 - jX_3$

Tuttle[6] has shown that at balance each of the component T networks plays its part independently of the other, and the null condition is simply that corresponding to equal and opposite transmission through the two components. Thus as in the case of a balanced bridge the impedance of the generator or of the common output circuit has no effect on the balance condition. No voltage is developed across the output circuit because the currents from the two component sections are equal and opposite and hence balance out. Hence in calculating the transfer impedance the output may be assumed to be shorted. From Figure 2, with the output terminals shorted, currents assumed in a right-hand rotation, the transfer impedance is

$$Z_{t12} = \frac{e_s}{i_o} = Z_1 + Z_2 + \frac{Z_1\,Z_2}{Z_3} \tag{10}$$

For the parallel combination of the two T networks the null condition is

$$i_o + i'_o = \frac{e_s}{Z_{t12}} + \frac{e_s}{Z'_{t12}} \tag{11}$$

---

[6] Tuttle, W.N., Proc. I.R.E., vol 28, Jan. 1940, p.23

giving

$$Z_{t12} + Z'_{t12} = Z_1 + Z_2 + \frac{Z_1 Z_2}{Z_3} + Z'_1 + Z'_2 + \frac{Z'_1 Z'_2}{Z'_3} = 0 \quad (12)$$

For two components this equation states that the sum of the two transfer impedances must be zero. For three or more components this is not true, but in the general case a balance requires that the sum of the transfer admittances must be zero, as may be determined from equation 11 above.

Tuttle has shown that a network of the type shown in Figure 5 has an equivalent transfer impedance circuit formed by a condenser in series with a negative resistance

$$\left(- \frac{1}{R_3 C_1^2 W^2} - j \frac{2}{W C_1} \right)$$

while the network of Figure 4 has a transfer impedance equal to an inductance in series with a positive resistance $(2R_1 + jR^2{}_1 WC_3)$. Equating real and reactive components:

$$\frac{2}{C_1 W} = R^2{}_1 C_3 W$$

$$\frac{1}{R_3 C_1{}^2 W^2} = 2R_1$$

Dividing the first by the second:

$$\frac{C_3}{2C_1} = \frac{2R_3}{R_1} = Q^2 \quad (13)$$

where Q is the ratio of reactance to resistance of the transfer impedance, and is the same for both T's at balance.

Q makes a convenient design parameter and, when unity,

$$R_1 = R_2 = 2R_3$$

$$C_1 = C_2 = 1/2\ C_3$$

By ganging 3 resistances the null frequency may be easily varied. The three resistances will be equal if Q is chosen to be $\sqrt{2}$

$$R_1 = R_2 = R_3$$

$$C_1 = C_2 = 1/4\ C_3$$

Or, what is more convenient at high frequencies, the three condensers may be ganged and made equal if $Q = \frac{1}{\sqrt{2}}$

Then:

$$R_1 = R_2 = 4R_3$$

$$C_1 = C_2 = C_3$$

It may be shown also that a null may be obtained when the two series branches are not equal or that the null frequency may be varied by control of a single element. The effect of unbalance is to increase the attenuation of the network especially of frequencies on the low side of the null point.

The parallel-T networks used in these tests were constructed in shielded compartments in small plug-in cans for convenience in changing.

To solve for the balance frequency the condition for which the denominator of equation (7) becomes zero is used:

$$R = X$$

$$R = \frac{1}{2\pi f_o C}$$

$$f_o = \frac{1}{2\pi RC}$$

or, in terms of the original network values:

$$f_o = \frac{1}{2\pi R_1 C_1} \tag{14}$$

Or, in the more general case the sum of the transfer impedances of the two T components is set equal to zero:

$$R_1 + R_2 + \frac{R_1\ R_2}{-j\frac{1}{WC_3}} - j\frac{1}{WC_1} - j\frac{1}{WC_2} - \frac{1}{W^2 C_1 C_2 R_3} = 0$$

If $Q = 1$ and the network is balanced,

$$R_1 = R_2 = 2R_3 = R$$

$$C_1 = C_2 = 1/2C_3 = C$$

Then:

$$2R + j\frac{R^2}{\frac{1}{2WC}} \quad - j\frac{2}{WC} \quad - \frac{2}{W^2 RC^2} = 0$$

Equating reals: $$R = \frac{1}{W^2 RC^2}$$

$$W = \frac{1}{R\ C}$$

or,

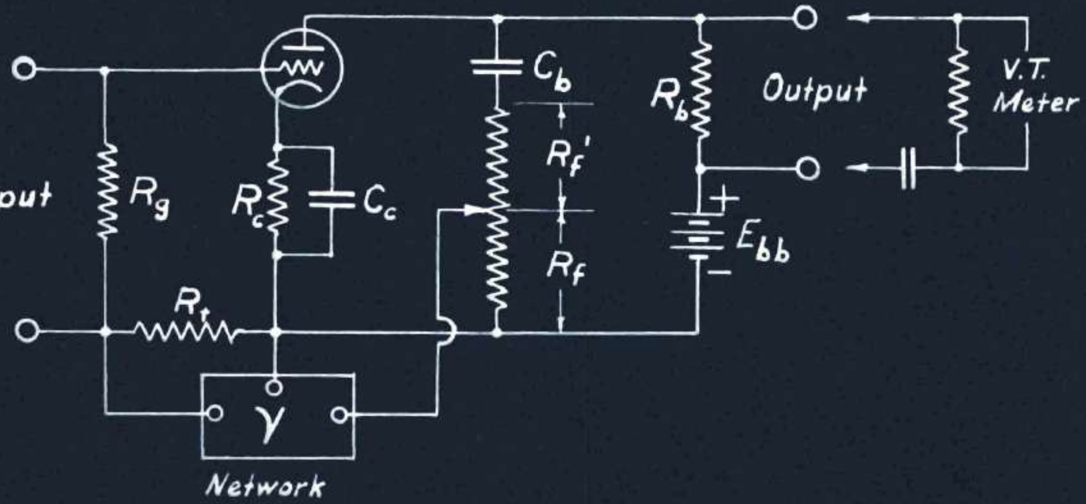
$$f_o = \frac{1}{2\pi R_1 C_1}$$

DESIGN OF SELECTIVE TRIODE AMPLIFIER

The tube employed in the single-stage amplifier (Figure 11) is a 6J5, a triode with comparatively high mutual conductance and amplification factor. It is important that the tube selected be capable of working into a low-impedance load if a wide variety of networks is to be used in the feedback path, since the feedback network should have a characteristic impedance well above the normal load impedance of the tube. Since it is desirable, in most types of networks, to keep the internal dissipation down to as low a value as possible it is the usual practice to make the characteristic impedance of networks rather low. For general test runs, therefore, the use of a regular pentode as the sole amplifying tube is ruled out because of its high plate resistance, unless some impedance transforming device is used. This fact together with the desirability of testing amplifiers of both low and high gain led to the selection of the 6J5 for the first tests. In general the values of circuit elements used in the triode amplifier are not critical, and ordinary commercial tolerances in the values of resistances and condensers suffice.

To simplify the test runs a battery plate supply, with substantially constant voltage output, was used rather than a rectifier-filter supply. An a.c. supply would have to be designed for close voltage regulation or a variable control used to reset the voltage at each reading. In use with a null network, however, any supply that meets the normal voltage and filter requirements should be satisfactory since the negative feedback present except at the null frequency minimizes the

Fig. 11
Selective Triode Amplifier



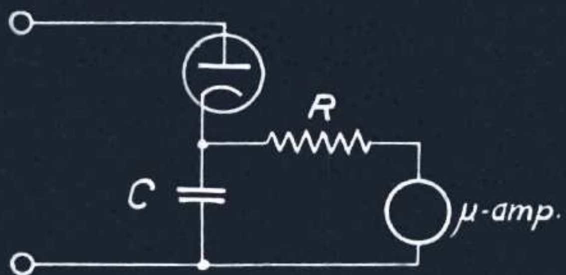


Fig. 12
Diode Voltmeter

DML
Ga Tech 41

effect of supply line fluctuations. Used as a band-pass, high-pass, or low-pass amplifier any precaution dictated by the required stability over the transmitted band should be observed in choosing, or designing, a supply.

Since neither maximum gain nor minimum distortion is sought in the particular amplifier tested, a fairly low value of plate-supply resistance was used, permitting a lower plate-supply voltage to be used. Excluding the output circuit the load impedance is made up of the elements R_b , Xc_b , R_f , R_f^1, and the input impedance of the network (see Figure 11). It may be necessary in calculating some network impedances to consider also the reflected impedance of the network termination. C_b should be made large enough to have a negligible reactance compared with the load impedance at the lowest frequency to be considered. In most cases, however, unless the network input impedance is very low the load impedance will be approximately equal to the feedback resistance $R_f + R_f^1$ in parallel with R_b. It should be noted that in a permanent set-up where no control over Beta is desired R_f or R_f plus R_f^1 might be omitted entirely leaving only R_b in parallel with the input impedance of the feedback network as the load resistance. The plate resistance of the 6J5 is approximately 7000 ohms. In this particular case R_b was made 20,000 ohms, R_f plus R_{f^1} 20,000 ohms, and C_b one microfarad. The input impedance of the 1100 cycle network most used in these tests was approximately 21,000 ohms. Therefore, when Beta is zero $R_L \cong 10,000$ ohms. When Beta is unity $R_L \cong 6,700$ ohms. If another stage had followed the triode or if a low impedance output circuit had been used it would be desirable to make R_b considerably greater. In the tests made an output circuit having more than ten times the value of R_b was used and so this impedance was not considered in these calculations. It may be noted that the Beta control could as well have been placed at the terminating end of the network, and in some cases this might be more desirable.

The blocking condenser C_b should be a high-quality paper or mica condenser with low d.c. leakage current. Since at least a portion of the leakage current will flow through the feedback network, which is in series with the d. c. grid bias, a considerable change in operating point may be effected by it, especially if the feedback is aplied to a pentode stage. The bias voltage introduced by this leakage current will be positive with respect to ground causing an increase in average plate current, possibly introducing non-linear distortion and overloading the tube. Both mica and paper dykanol-filled condensers were tried and found to be satisfactory.

The cathode bias resistance of 1000 ohms was selected so as to give seven volts negative bias at seven milliamperes plate current. In order to prevent stray constant-current feedback across R_c the by-pass condenser C_c employed had a value of 26 microfarads. The reactance of C_c should be very small compared with R_c at the lowest frequency to be amplified.

The resistance R_t terminates the feedback network and, in general, should be about equal in magnitude to the characteristic impedance of the network. In the case of the parallel-T network it has been shown that the generator or terminating resistances do not affect the null frequency, although the shape of the attenuation curve may be altered. The lower the resistance the less the voltage developed across it and the less effective the feedback becomes. For certain other types of networks the cut-off frequencies are not independent of the generator or load resistances and it is desirable to terminate them in their characteristic impedances so that the simple design equations will hold.

The input resistance R_g provides an ever-present path for the d.c. bias voltage. It should be made sufficiently high so as to introduce negligible attenuation of the input signal. On running tests on the amplifier using a high-output oscillator such as the Western Electric 13-A, which was used in this case, care must be exercised to keep the peak input signal voltage to the amplifier well below the value of d.c. bias on the tube.

In Figure 11 is shown the circuit of the single-stage amplifier tested. If R_t were shunted out and the feedback network with its control resistance R_f plus R'_f and blocking condenser C_b were taken out, an amplifier stage of conventional design, with some slight modifications, would remain. As has been shown in the section on feedback theory the voltage fed back to the input circuit must have a component in inverse phase with the input signal voltage in order to produce degeneration. Since the normal phase shift in a single tube is 180° it is necessary only to take voltage directly from the plate circuit and conduct it back to the grid circuit to get degeneration. Because direct current from the plate-supply battery E_{bb} would flow through R_g or R_t a direct connection cannot be made and some type of blocking device such as C_b must be used. This introduces additional phase shift but if C_b is made large this shift is not excessive. If the feedback voltage is to be reduced or varied a voltage divider must be used. This is the purpose of R_f and R'_f. Then Beta, the fraction of the output voltage fed to the grid, or the controlling network, is approximately:

$$\beta \cong \frac{R_f}{R_f + R_f'}$$

If straight feedback without a controlling network were used the input terminal would be connected directly to the common junction point of R_f and R'_f. There are a number of modifications of this form of feedback. In the amplifier constructed R_f and R'_f consisted of a two-pole selector switch used with fixed resistances. This arrangement permits the resetting of Beta to precise values.

If a bridge or four-terminal network (Figure 7) is to be used in the feedback circuit at least one isolating transformer will be necessary. One transformer would permit the necessary isolation and balance to ground but it also introduces approximately 180° phase shift and hence the phase must be inverted or reversed again by some method. The use of an equivalent three-terminal network, (Figure 8), greatly simplifies the design of the feedback path since there is a common input and output connection. Used over a single stage, transformers are unnecessary. On the other hand the bridge has the advantage of requiring only two controls, R_3 and R_4, while the three terminal equivalent network requires three variable elements.

Thus in the circuit of figure 11 the output voltage across R_f is subjected to the frequency discrimination of the feedback network. The voltage passed by the network appears across R_t in series with the signal voltage, and, if the phase shift within the network is less than 90°, this voltage is out of phase with the signal voltage by more than 90°. If Beta is high this gives the overall amplifier a characteristic approximately the inverse of the feedback network characteristic. For convenience in testing amplifier response with different networks the network connections were brought to a tube socket and the networks were constructed in shielded aluminum cans (see photograph) mounted on plugs.

A simple arrangement of a peak-reading diode voltmeter as shown in Figure 12 was used to measure the voltage on both input and output circuits. A meter of this type was selected in preference to a thermocouple or copper oxide meter because of its flat frequency characteristic and its low damping. The output condenser and resistor (Figure 11), were used to provide an a.c. supply entirely free of d.c. for the meter. The condenser has a value of several microfarads and the resistance is of the order of one meg-ohm. Reasonably accurate results could be obtained with copper-oxide rectifier voltmeters. Two Weston 10,000 ohms-per-volt meters were tried and found to give results very similar to the vacuum tube meters. The tube used is a 6H6 with the two diode sections connected in parallel. R is 500,000 ohms, C is one microfarad, the meter used was a Sensitive Research Polyranger, 0-100 microamperes. Since identical voltmeters were used on the input and on the output of the amplifier it was not necessary to calibrate them to measure voltage amplification. Instead a voltage divider was used on the input circuit permitting the adjustment of the two meters to the same voltage reading so that the amplification was equal to the voltage divider ratio.

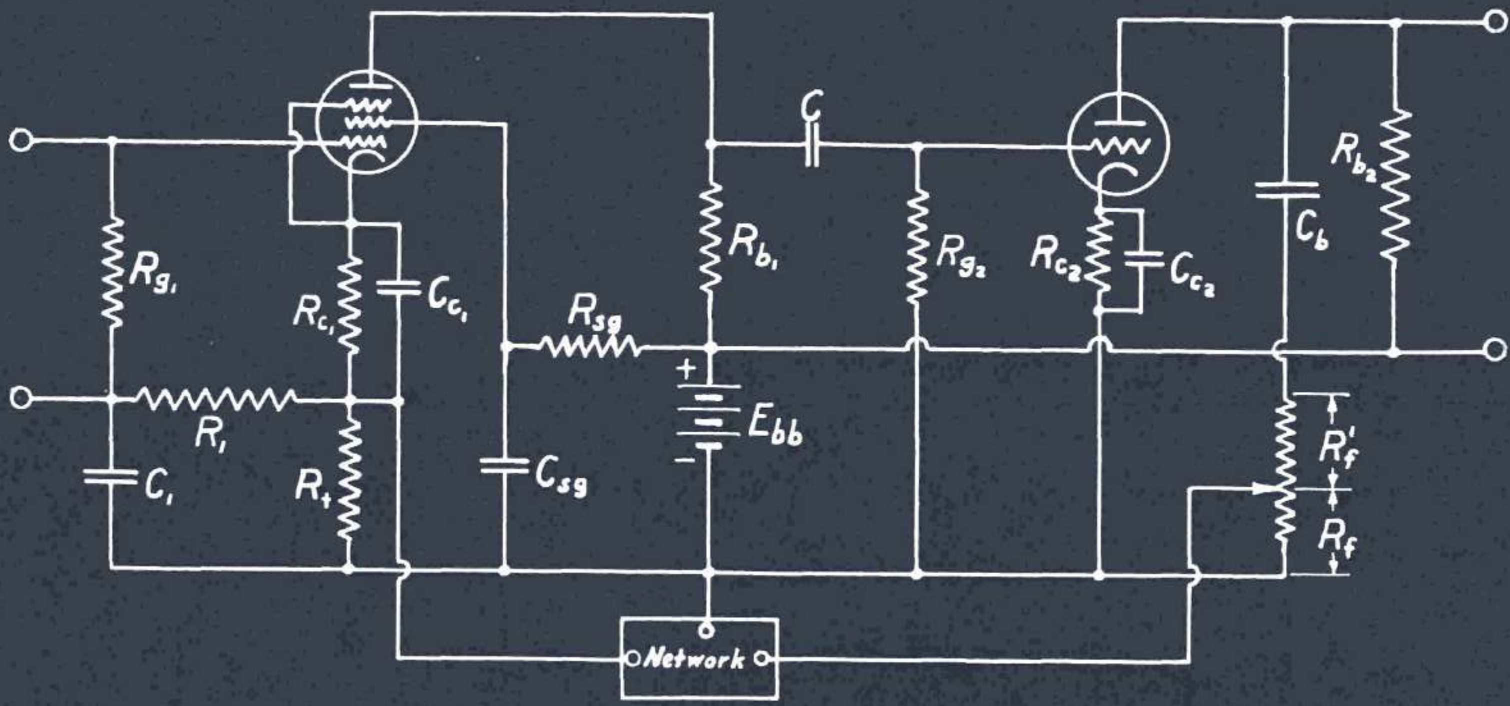
The voltage developed across the condenser in Figure 12 is a linear function of the peak volts applied and is nearly independent of frequency, provided C is large and R is high.

DESIGN OF TWO-STAGE SELECTIVE AMPLIFIER

In order to see the effect of increased gain on selectivity a two-stage amplifier was constructed. For the first stage a 6J7 pentode was selected because of its high amplification factor. The 6J5 triode was again used in the final stage because of its low plate resistance. If a pentode were used for this stage also care would have to be observed to prevent overloading of the tube by the comparatively low impedance of some networks. The beam power tube, however, is well suited to use in the stage loaded by the network.

Over two stages the normal phase shift in the signal voltage approximates 360°. Therefore, the feedback voltage cannot be introduced in exactly the same fashion as it was over a single stage unless a transformer or phase inverting network is used. Perhaps the simplest manner is to terminate the feedback circuit in a resistance in series with the signal voltage in such a manner that the phase of the feedback voltage with respect to the input signal voltage is reversed. As may be seen in Figure 13 this resistance must be placed in the cathode circuit. The plate current also will flow through the terminating resistance which, if provision is not made to prevent it, will bias the grid highly negative. To secure the proper bias a high resistance, R_1 (Figure 13), connects the grid circuit of the tube to a point in the cathode circuit such that the drop across R_c, is applied to the grid. Then a low reactance condenser C_1 is used to provide a path to ground for the a.c. voltage. The reactance of C_1 should be negligible, compared to the resistance of R_1, at the lowest frequency to be amplified so that the

Fig. 13



Two-Stage Selective Amplifier

DML
Ga Tech 41

effect of R_1 on the signal voltage will be negligible.

The terminating resistance R_t depends on the particular network used and should be made as low as is consistent with good feedback since the IR drop in R_1 caused by the plate current reduces the plate voltage and limits the current obtainable from the tube.

The only other unusual consideration in the design of the amplifier is in the selection of the coupling condenser C. This condenser should be made as large as possible to prevent undue additional phase shift at low frequencies. If made too large, however, the R_{g_2} C time constant may become too large and motorboating might take place, or the d.c. leakage current from E_{bb} flowing through R_{g_2} might upset the bias on the following stage. For these reasons C should be of good quality and should have a fairly high capacitance. A value of 0.1 microfarad was used in the amplifier tested and gave satisfactory results.

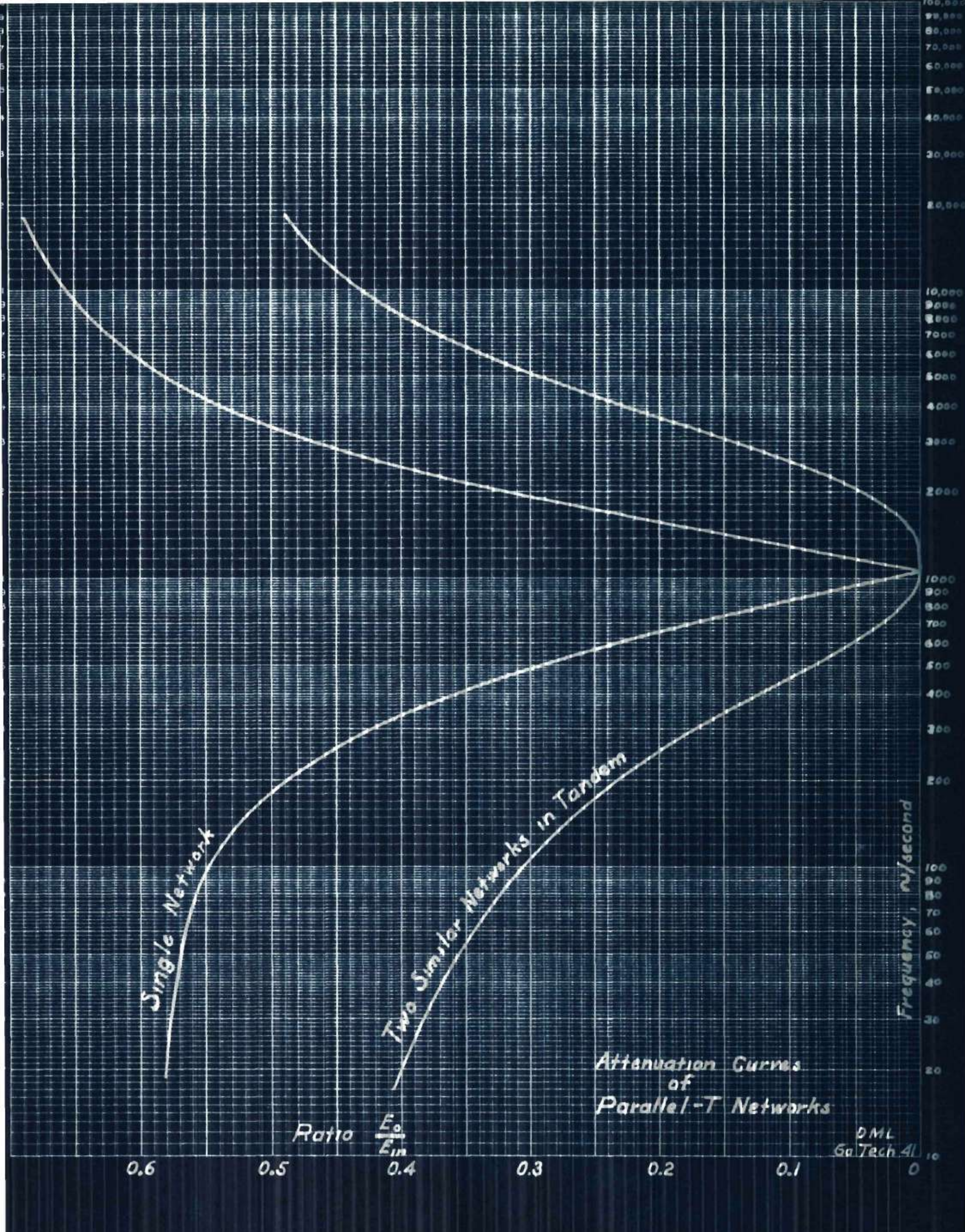
In medium or high gain amplifiers care must be taken to minimize stray feedback, especially positive feedback. The coupling between wires, tube elements, a common ground path, all may cause some feedback, which prevents the actual response from conforming with the calculated response. The higher the gain, the more sensitive an amplifier is to small amounts of feedback. Over two or more stages, special care should be observed to prevent oscillation from taking place. A small fixed amount of negative feedback may be of great help in stabilizing the operation of the amplifier.

EXPERIMENTAL VERIFICATION OF THEORY

Before any tests were made on the amplifiers, frequency runs were made on the networks to determine their attenuation characteristics. For these runs a Western Electric 13-A beat-frequency oscillator with continuously variable frequency control from 20 to 10,000 cycles per second was used. The input to the networks was held constant at 10 volts R.M.S. by means of a Weston 10,000 ohms-per-volt rectifier type meter. The output voltage was measured with a similar meter. In all the attenuation or response curves shown the terminating resistance was so high as to have a negligible loading effect on the networks tested. On page 30 are shown the response curves for a single parallel-T network with $R_1 = R_2 = 30{,}000$ ohms[7]. The terminating meter has a resistance of 100,000 ohms. It was found that the use of a terminating resistance several times the magnitude of the characteristic impedance gave essentially the same results as open circuit termination. If R_t is equal in magnitude to the characteristic impedance of the network, the attenuation is somewhat greater at frequencies away from the null frequency, especially below the null frequency. At half the null frequency it is about twice the value of the open circuit value while at twice the null frequency it is only slightly greater. Around the null point the curve is practically unchanged. A further reduction in R_t therefore decreases the effectiveness of the network, especially on the low frequency side of the null point. For most effective operation of the network, R_t should

---

[7] Number (1) in Appendix.



Attenuation Curves
of
Parallel-T Networks
Single Network
Two Similar Networks in Tandem
Ratio $\frac{E_o}{E_{in}}$
0.6
0.5
0.4
0.3
0.2
0.1
0
Frequency, ~/second
10,000
1000
100
10
DML
Ga Tech 41

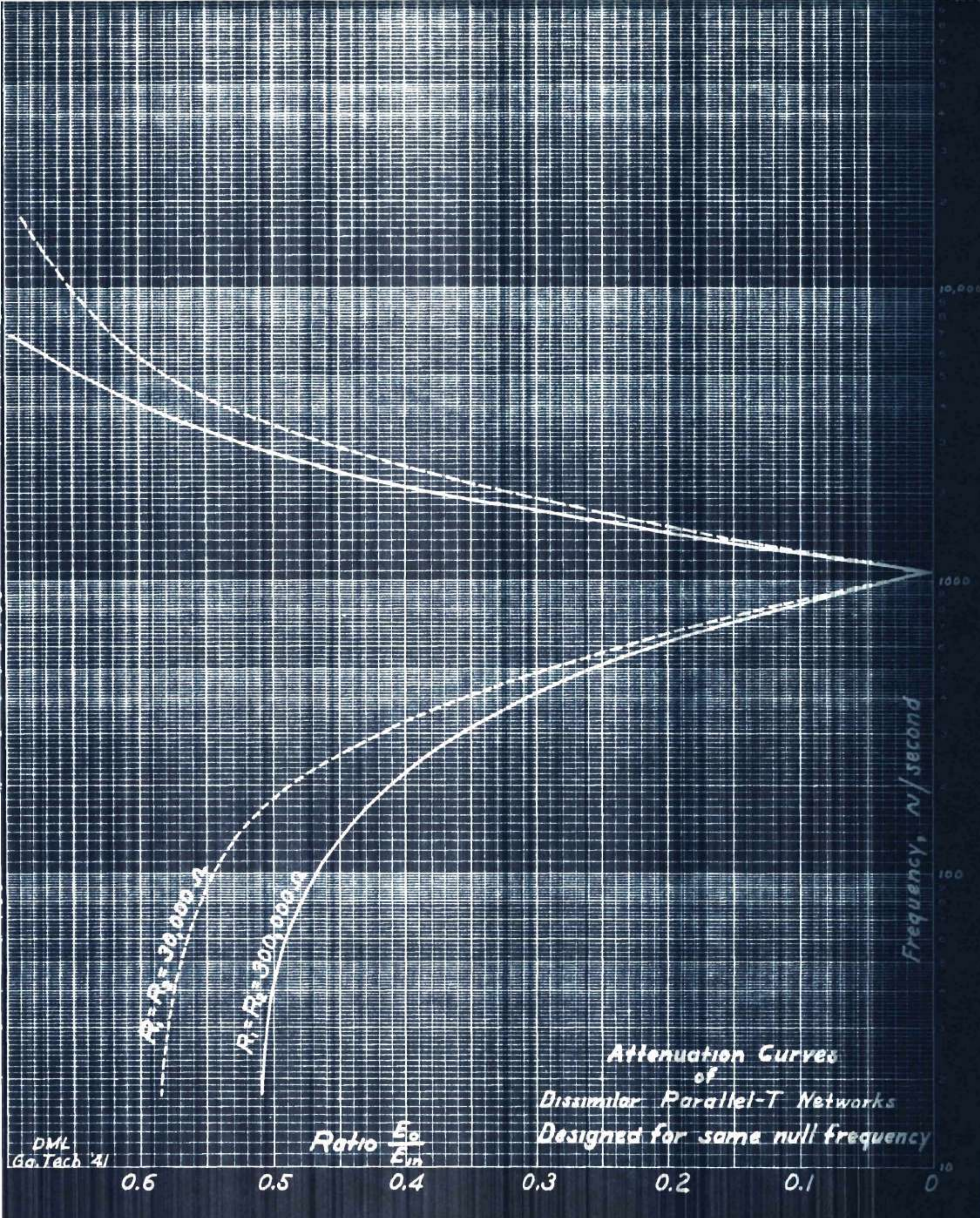
be as high as possible or the network should be designed so that Z_0 is only one-third or one-fourth the value of R_t demanded by the circuit.

The linearity of these curves near the null frequency, especially on the high-frequency side shows that the attenuation is logarithmic in this range. The effect of unbalancing the series elements of the T components is to increase the attenuation, especially at low frequencies, while a definite null is still obtained although at a different frequency. Thus any element can be used as a null frequency control although the effectiveness is reduced and the simple design equations no longer hold. It has been shown that the criterion for a null is that the sum of the transfer impedances of the two T components be zero.

The second curve on page 30 shows the effect of using two identical networks in tandem. As would be expected the response is approximately the square of that of either network alone.

In order to compare networks designed for the same frequency but with widely different circuit constants two 1100 cycle networks were constructed, one with $R_1 = R_2 = 30{,}000$ ohms and the other with $R_1 = R_2 = 300{,}000$ ohms. The curves for these two networks are shown on page 32. It will be noticed that provided R_t is several times the magnitude of the characteristic impedance the attenuation is substantially independent of the characteristic impedance of the network. For practical considerations the low-impedance network may be more useful.

In all the network attenuation curves it may be seen that at one point the attenuation is infinite and a null is obtained. This will hold true at one frequency for any design of the parallel-T network of the type shown in Figure 8. If the signal contains appreciable harmonic



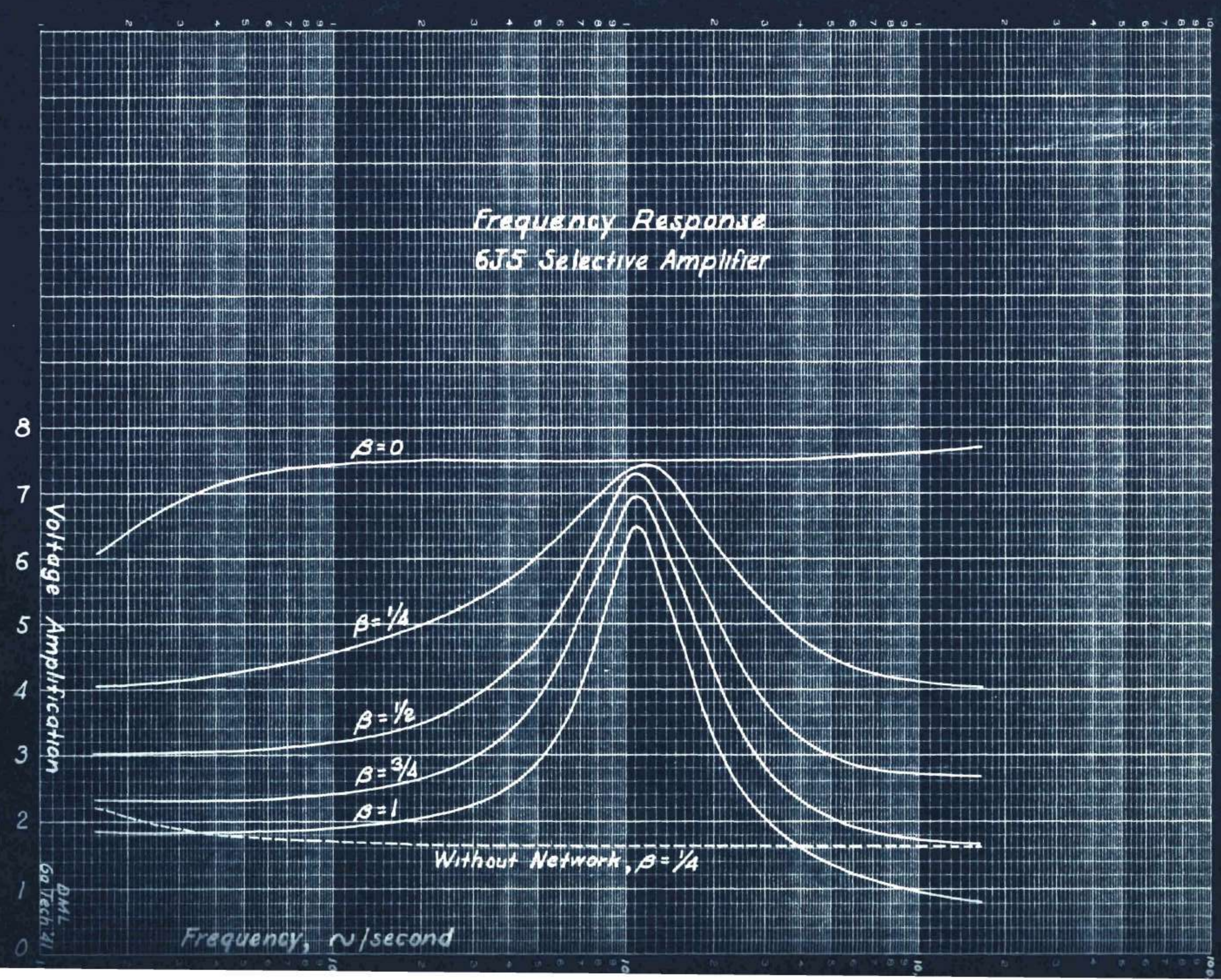
Attenuation Curves
of
Dissimilar Parallel-T Networks
Designed for same null frequency
$R_1 = R_3 = 30{,}000\ \Omega$
$R_1 = R_3 = 300{,}000\ \Omega$
Frequency, N/second
10,000
1000
100
10
Ratio $\frac{E_o}{E_{in}}$
0.6
0.5
0.4
0.3
0.2
0.1
0
DML
Ga. Tech '41

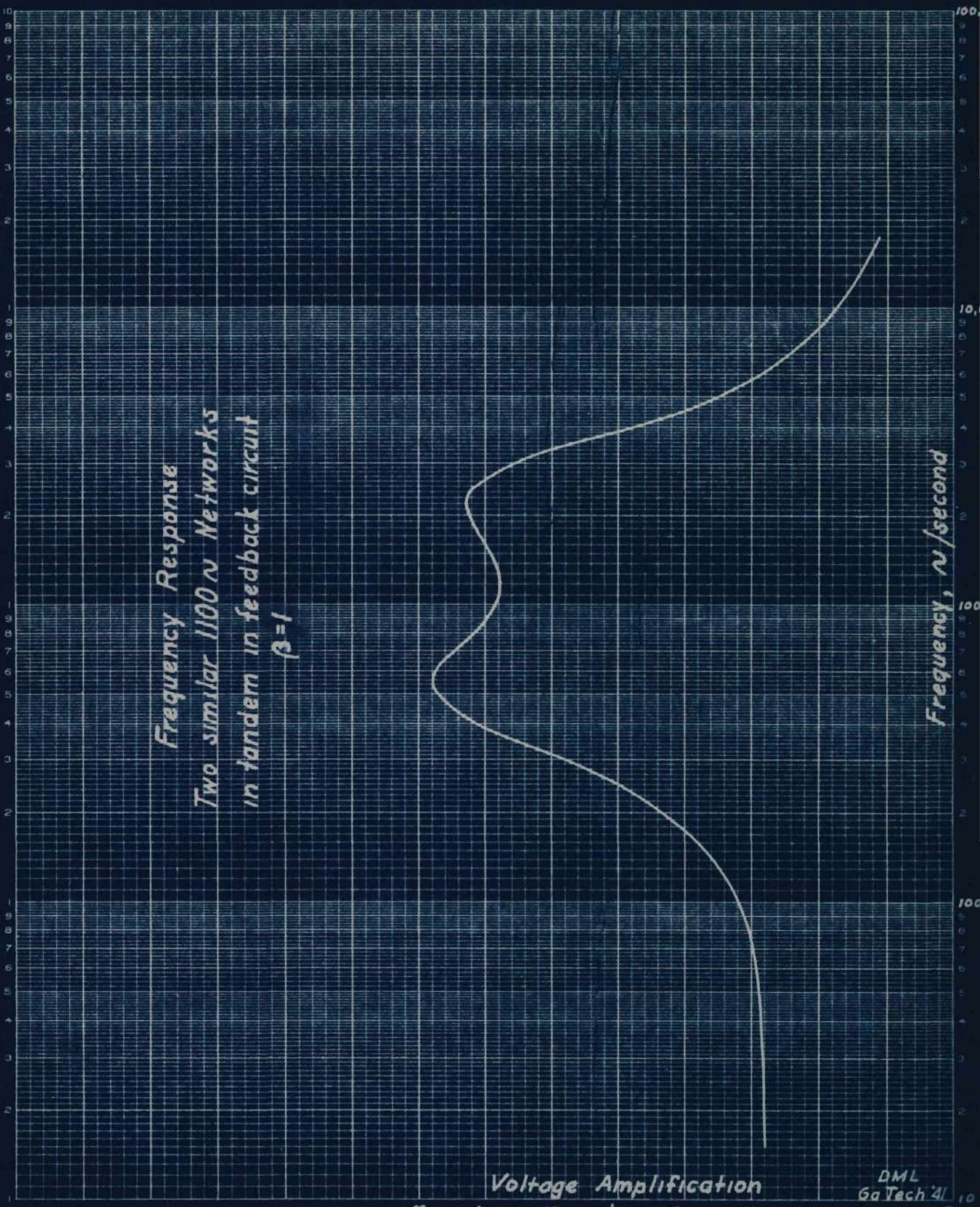
content it is difficult to obtain a good null, especially by acoustic detection, since the harmonics are passed through with little attenuation. When the network is used in the feedback circuit of an amplifier, however, the harmonics contained in a wave will pass through while the fundamental is attenuated. This provides degenerative feedback which will greatly reduce the amplification of harmonics, provided the phase shift in the network is not greater than 90°.

After making tests on the networks alone frequency runs were made on the single-stage amplifier with various values of Beta and with different network combinations. The response of the amplifier alone is shown on page 34 in the curve Beta equal to zero. The frequency response is flat over the normal range of audio frequencies from 100 to 10,000 cycles. Now the 1100 cycle parallel-T network is plugged in the circuit and Beta set at 1/4, 1/2, 3/4 and 1 successively. The effect of increasing the feedback voltage as shown on page 34 is to "pull in the sides" of the response curve, making it sharper and slightly reducing the peak amplification. The curves are very similar in appearance to resonance curves with variable Q.

The effect of linear degenerative feedback network over all frequencies is shown by the dotted curve. For this test a plug with jumper between input and output terminals is used instead of a network. With a small amount of feedback the amplification is greatly decreased. Since the original response curve was nearly flat, the effect of feedback on frequency distortion cannot be very well illustrated in this amplifier.

On page 35 is shown the band-pass effect obtained by using two


Frequency Response
6J5 Selective Amplifier
β=0
β=1/4
β=1/2
β=3/4
β=1
Without Network, β=1/4
Voltage Amplification
Frequency, ~/second
8
7
6
5
4
3
2
1
0
DML
Go Tech '41

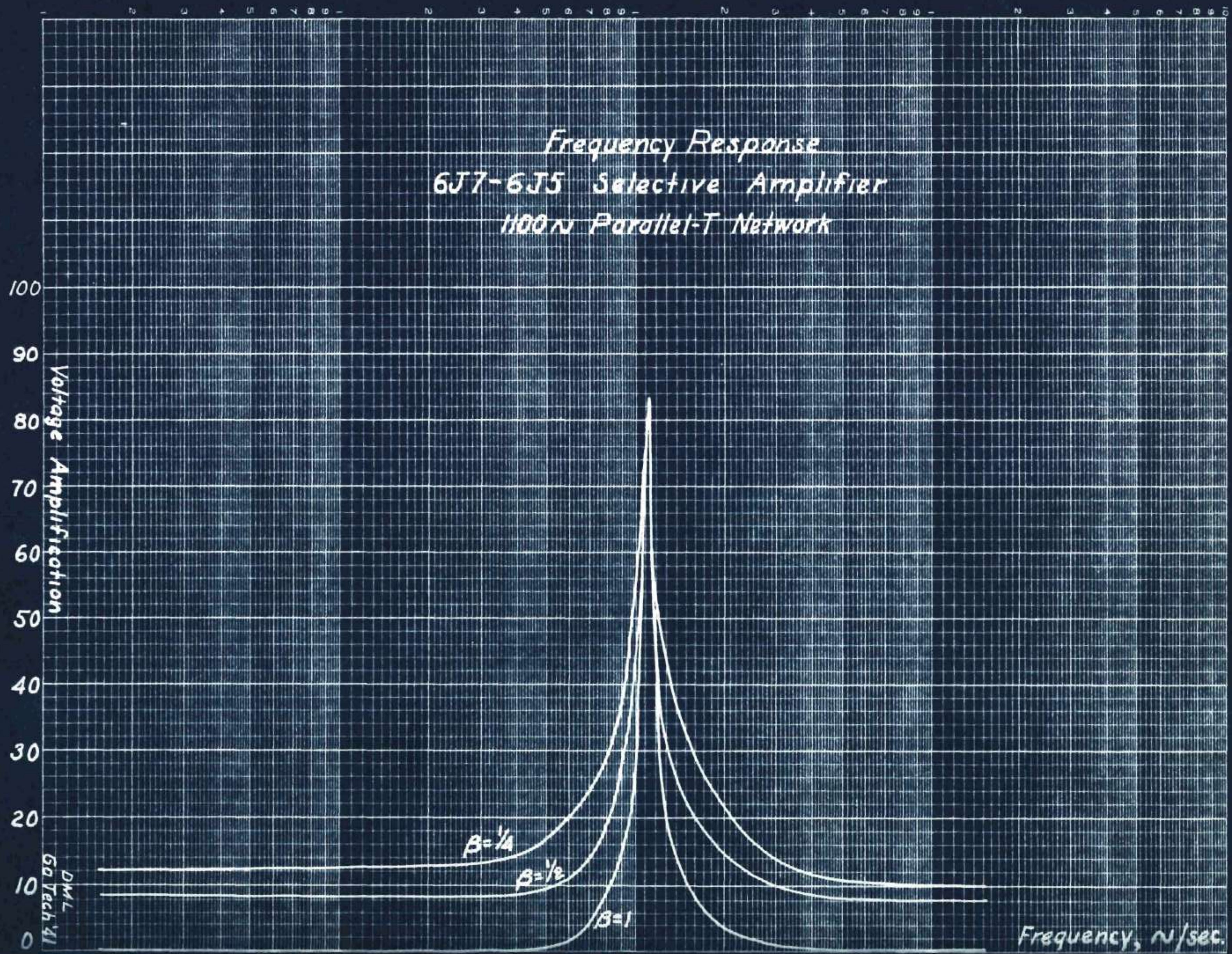


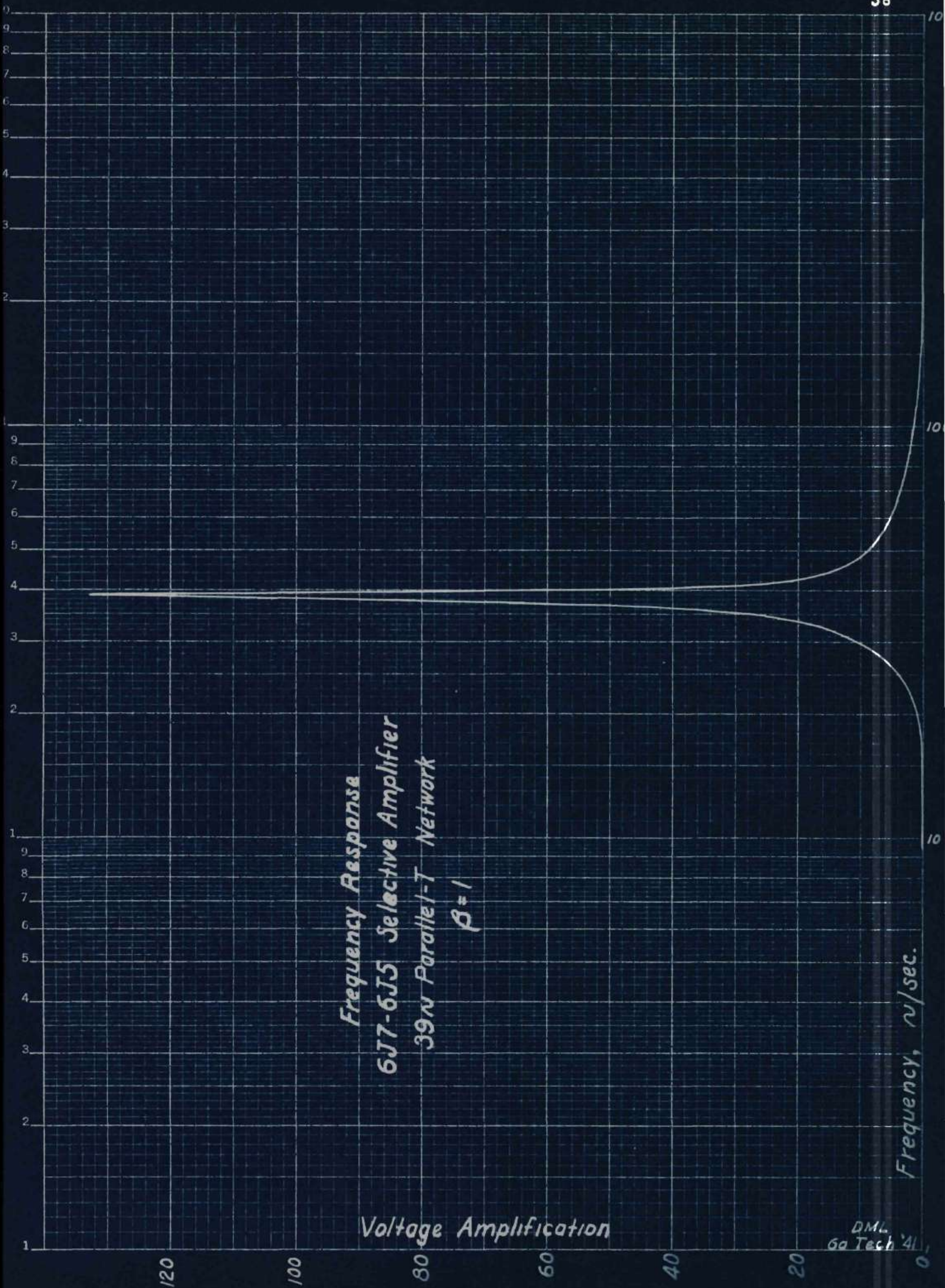
Frequency Response
Two similar 1100 ~ Networks
in tandem in feedback circuit
β=1
Frequency, ~/second
Voltage Amplification
8
7
6
5
4
3
2
1
0
DML
Ga Tech '41

identical parallel-T networks in tandem in the feedback circuit. The response curve of the two networks alone is shown on page 30.

To find the effect of increased gain on the response curve, and to try a second method of applying negative feedback a two stage amplifier (Figure 13) was used. The overall gain of this amplifier is about ten times the gain of the single triode. It is important that the signal voltage be held low here to prevent overloading of the second stage, so as not to introduce distortion.

The curves on page 37 and 38 show graphically the effect of increased gain in peaking the response curves. If $A\beta$ is high, the overall selectivity is accentuated, becoming more pronounced than in the network itself. A dedrease in Beta has greatest effect on frequencies well away from the peak frequency. Instead of raising the peak response with a decrease in Beta it was slightly reduced. The explanation for this apparent contradiction might be that some positive feedback is also present due to capacity coupling between network elements. Therefore a decrease in Beta also decreases the regeneration, slightly reducing the response. The curves with Beta equal to one show a remarkably high selectivity and could be approached only by resonant circuits having very high Q, a quality difficult to obtain at low frequencies. The lower frequency limit to which the parallel-T network could be constructed with small receiver parts is well below one cycle per second.


Frequency Response
6J7-6J5 Selective Amplifier
1100~ Parallel-T Network
Voltage Amplification
100
90
80
70
60
50
40
30
20
10
0
β=1/4
β=1/2
β=1
Frequency, ~/sec.
DML
Ga Tech '41



Frequency Response
6J7-6J5 Selective Amplifier
39~ Parallel-T Network
β=1
Voltage Amplification
Frequency, ~/sec.
DML
Ga Tech '41
120
100
80
60
40
20
0

CONCLUSIONS

The use of a network in the feedback circuit of an amplifier offers a convenient method to obtain frequency discrimination of almost any desired form in the amplifier. The discrimination is essentially gained by a reduction in amplification over the unwanted range. If the product of gain and network response is much greater than unity, then the overall frequency response of the amplifier will be approximately equal in form to the inverted response characteristic of the feedback network.

The sharpness of the cut-off frequency is a direct function of gain. If a null network is employed in the feedback circuit, the selectivity of the amplifier response curve can be increased by raising the gain of that portion of the amplifier over which feedback is applied. A parallel-T network used in a feedback amplifier yields a response curve similar to that which might be obtained from a high-Q resonant circuit. A great advantage over the resonant circuit is the fact that ordinary small-size condensers and resistances may be used in the parallel-T network to get a null at extremely low frequencies of the order of one cycle per second. An equivalent resonant circuit would involve bulky and expensive coils and condensers and the selectivity would ordinarily be of a low quality.

The use of two identical parallel-T networks in tandem gives a band-pass effect for the feedback amplifier, substantially passing all frequencies from one-half to twice the null frequency.

Since the network used to get frequency discrimination is in the feedback circuit, and not in the line, the loss of power in the network is negligible. Thus the network may be conveniently designed with little attention given to dissipation in the elements. The parallel-T network is unaffected by the amount of resistance in its elements so long as its termination is several times greater than its characteristic impedance. Because of these reasons the system is applicable to power amplifiers also, especially those employing beam power tubes, which are capable of giving full output and high efficiency on low driving voltage.

By introducing a small, constant amount of positive feedback in the amplifier, in addition to the negative feedback with parallel-T network, an oscillator with resistance frequency control, no coils, and an unusually good wave form may be obtained[8].

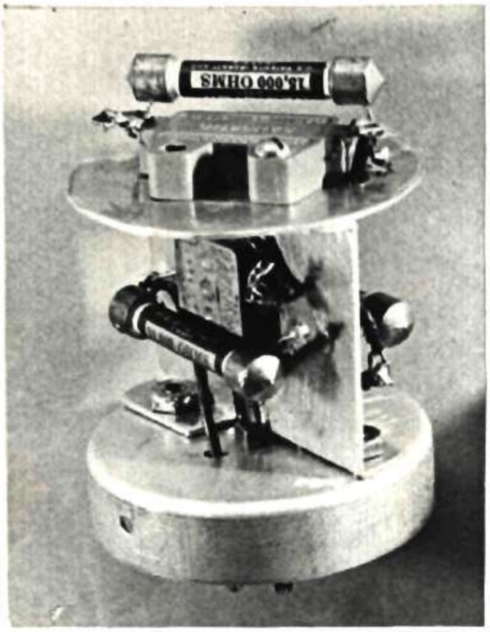
The selective amplifier may be used as a signal analyser at audio frequencies. The parallel-T network is not limited to audio frequencies but works equally well at radio frequencies. It is believed that circuits of the type described in this thesis will be increasingly used and should be considered for possible advantages whenever new measurement or selectivity problems are encountered.

---

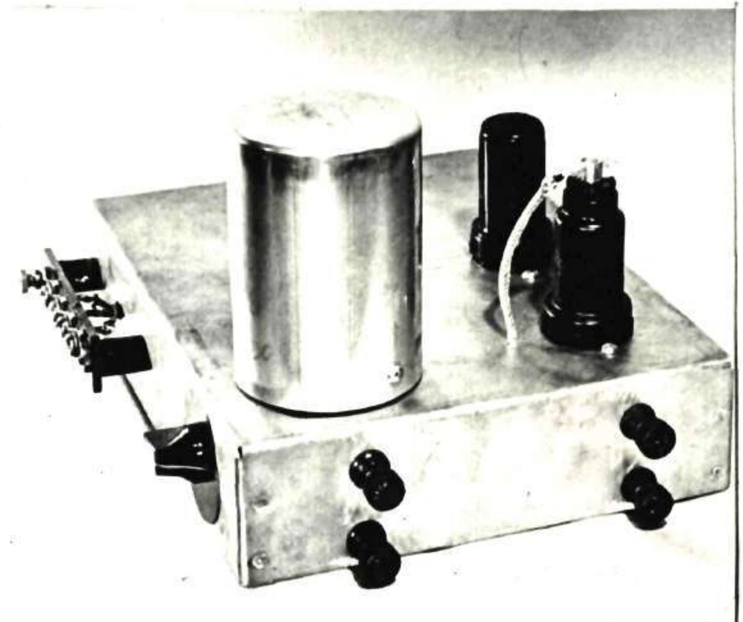
[8]Caywood, R.W., QST, Jan. 1941, p. 22

Scott, H.H., Proc. I.R.E., vol. 26, Jan. 1938, p. 226

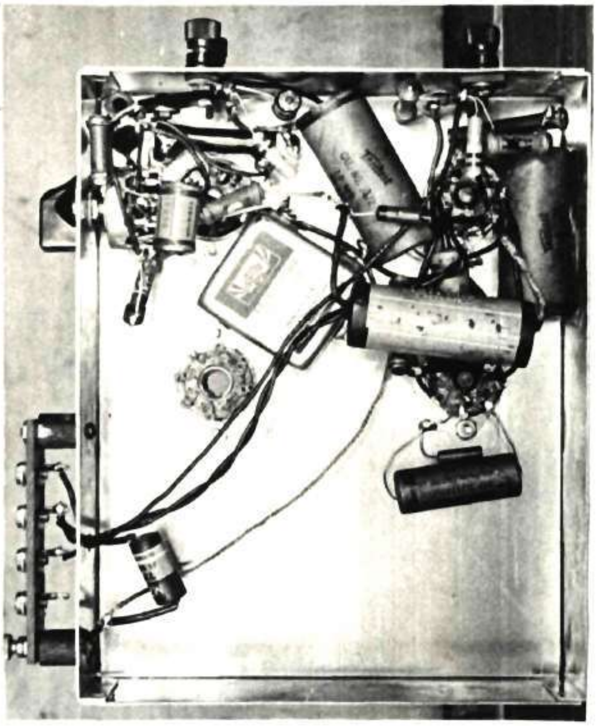
Terman, F.E., Buss, R.R., Hewlett, W.R., and Cahill, F.C., Proc. I.R.E., vol. 27, Oct. 1939, p. 649



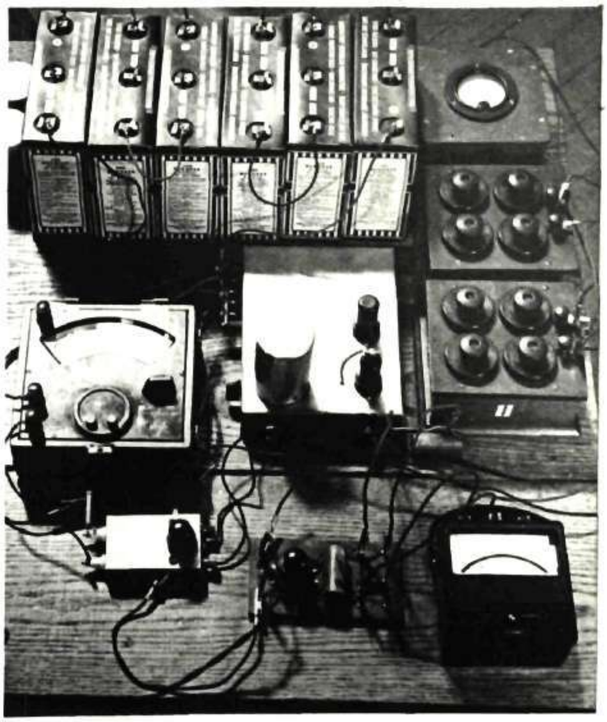
Parallel-T Network
with shield cover removed



Two Stage
Selective Amplifier



Under Chassis View
of two stage Amplifier



Parts Layout
in test run

BIBLIOGRAPHY

FEEDBACK THEORY

Black, H. S., Stabilized Feedback Amplifiers, *Elec. Eng.*, vol. 53, Jan. 1934, p. 114

Day, J. R., and Russel, J. B., Practical Feedback Amplifiers, *Electronics*, vol. 10, Apr. 1937, p. 16

Everitt, W. T., *Communication Engineering*, McGraw-Hill, 1937, p. 463

Nyquist, H, Regeneration Theory, *Bell System Tech. Journal*, vol. 11, p. 126

Reich, H. J., *Theory and Applications of Electron Tubes* McGraw-Hill, 1939, p. 220

Rumble, A. R., Audio Feedback, *Comm. and Bdcast Eng.* vol. 4, Apr., 1937, p. 14

Terman, F. E., Feedback Amplifier Design, *Electronics*, vol. 10, Jan. 1937, p. 12

NETWORK ANALYSIS

Everitt, W. T., *Communication Engineering*, McGraw-Hill, 1937, pp. 1 - 58

Guillemin, E. A., *Communication Networks*, vol. I, " The Classical Theory of Lumped Constant Networks", John Wiley and Sons, 1931

Tuttle, W. N., Bridged-T and Parallel-T Null Circuits for Measurements at Radio Frequencies, *Proc. I.R.E.* vol. 28, Jan. 1940, p. 23

Woodward, R. W., Hetrofil - An Aid to Selectivity, *QST*, Sept. 1939, p. 11

SELECTIVE FEEDBACK APPLICATIONS

Caywood, R. W., An Amateur Application of the Wien Bridge, *QST*, Jan. 1941, p. 22

Dawley, R. L., A Wide Range Audio Oscillator, *Radio*, June 1940, p. 17

______ The Improved Wide Range Audio Oscillator, *Radio*, Dec. 1940, p. 16

Fritzinger, G. H., Frequency Discrimination by Inverse Feedback, *Proc. I.R.E.*, vol. 26, Jan. 1938, p. 226

Scott, H. H., A New Type of Selective Circuit and Some Applications, Proc. *I.R.E.*, *vol.* 26, Jan. 1938, p. 226

Terman, F. E., Buss, R. R., Hewlett, W. R., and Cahill, F. C., Some Applications of Negative Feedback, *Proc. I.R.E.*, vol. 27, Oct. 1939, p. 649

APPENDIX

Table of values used in the triode selective amplifier shown in figure 5 :

R_g = 100,000 Ω	R_b = 20,000 Ω
R_c = 1000 Ω	C_c = 26 μf
R_t = 20,000 Ω (net.#1)	C_b = 1 μf
$R_f + R_f'$ = 20,000	E_{bb} = 270 volts

Triode = 6J5, μ = 20, G_m = 3000 μmhos

In the output circuit used with the diode meter :

R = 1 megΩ	C = 1 μf

In the diode meter, figure 6 :

R = 500,000Ω	C = 1 μf

μ_amp = 100 μ_ampere meter

Diode = 6H6

Table of valued used in the two-stage selective amplifier shown in figure 7 :

R_{g_1} = 100,000 Ω	R_{g_2} = 250,000 Ω
R_{c_1} = 1100 Ω	R_{c_2} = 1000 Ω
R_t = 20,000Ω(net.#1)	$R_f + R_f'$ = 20,000 Ω
R_{sg} = 1.1 megΩ	R_{b_2} = 20,000 Ω
R_{b_1} = 250,000 Ω	C_{c_1} = 10 μf
R_1 = 250,000 Ω	C_b = 1 μf
C_{c_1} = 25 μf	E_{bb} = 315 volts
C_{sg} = 1 μf	C_1 = 1 μf
C_c = 0.1 μf	Triode = 6J5

Pentode = 6J7, G_m = 1200 μmhos, R_p = One megΩ

Table of values used in the parallel-T networks. Notations from figure 2 :

Three networks having a null frequency of approximately 1100 cycles/sec. were used :

(1) and (2)

$R_1 = R_2 =$ 20,000 Ω	$R_3 =$ 15,000 Ω
$C_1 = C_2 =$ 0.005 µf	$C_3 =$ 0.01 µf

(3)

$R_1 = R_2 =$ 200,000 Ω	$R_3 =$ 150,000 Ω
$C_1 = C_2 =$ 0.0005 µf	$C_3 =$ 0.001 µf

In the network having a null frequency of approximately 39 cycles/sec. the following values were used :

(4)

$R_1 = R_2 =$ 40,000 Ω	$R_3 =$ 20,000 Ω
$C_1 = C_2 =$ 0.1 µf	$C_3 =$ 0.2 µf