A HIGH GAIN VIDEO AMPLIFIER

Thesis for the Degree of M. S.

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Robert Orlando Bolster

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THESIS

A HIGH GAIN VIDEO AMPLIFIER

by

Robert Orlando Bolster

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R.C.B.

Contents

Introduction

The purpose of this thesis is to present the theory, design, construction, and operation of a new type of video amplifier. In television equipment the performance of the video amplifier is always a question because of the tremendous band-width involved. For excellent reproduction of the original image, using standard transmission, this band-width spreads from about 30 cycles to about 4,000,000 cycles per second. Actually the band allotted in the radio frequency spectrum for the televised image is 4,500,000 cycles per second.

In order that the picture elements will form a normally shaded reproduction with a proper distinction between details, it is necessary that the gain and the time-delay of all amplifiers be as nearly constant as possible. The maintenance of a constant gain and time-delay across the 4 megacycle per second band is the foremost problem encountered in video amplifiers. The greatest evil in the problem of constant gain is the shunt capacitance of the video amplifier's plate circuit to ground. This shunt capacitance which limits the allowable plate load to a very low value is composed of the amplifier tube's output capacitance, the input capacitance of the following tube, and the capacitance to ground of the plate circuit elements and wiring.

In this thesis an attempt is made to eliminate this shunt capacitance through the use of a reactance tube displaying negative capacitance. A discussion of the application of negative capacitance to the video amplifier is presented in Part I.

The low frequency response of the video amplifier will not be considered in this thesis.

Part I

Theory of the Amplifier

As brought out in the Introduction, the foremost problems encountered in the video amplifier are the maintenance of a constant gain and of a constant time delay across the tremendous band-width. These two factors will be treated separately in the following work; the problem of constant gain being discussed first.

1. Problem of Constant Gain.

Concept of Negative Capacitance.

Since the shunt capacitance to ground of the video amplifier's radio frequency plate circuit limits the allowable plate load resistance and thus the gain, a solution to this problem would be effected if this shunt capacitance were eliminated or neutralized. To accomplish this elimination or neutralization, the phenomenon of negative capacitance, as defined below, is introduced.

With a sinusoidal voltage applied, a positive capacitance is characterized by a current vector which leads the voltage vector and a reactance which decreases with increasing frequency. A negative capacitance, as defined for this work, is characterized by a current

vector which lags the applied voltage vector and a reactance which also decreases with increasing frequency just as does a positive capacitance. Thus negative capacitance and inductance differ in the manner in which their reactances vary with frequency.

There is a fundamental difference, however, between negative capacitance and ordinary circuit elements; negative capacitance can appear only as a dynamic circuit element. This fact at once suggests the use of a reactance tube network to obtain the desired negative capacitance.

A reactance tube is merely a vacuum tube supplied with the correct alternating grid voltage to exhibit the desired relationship between plate voltage and plate current. Thus the tube may exhibit positive and negative resistance, inductance, or capacitance depending upon the arrangement of the grid circuit.

Example of the Possibilities of the Reactance Tube.

Consider the circuit of the elementary phase modulator of Figure 1.

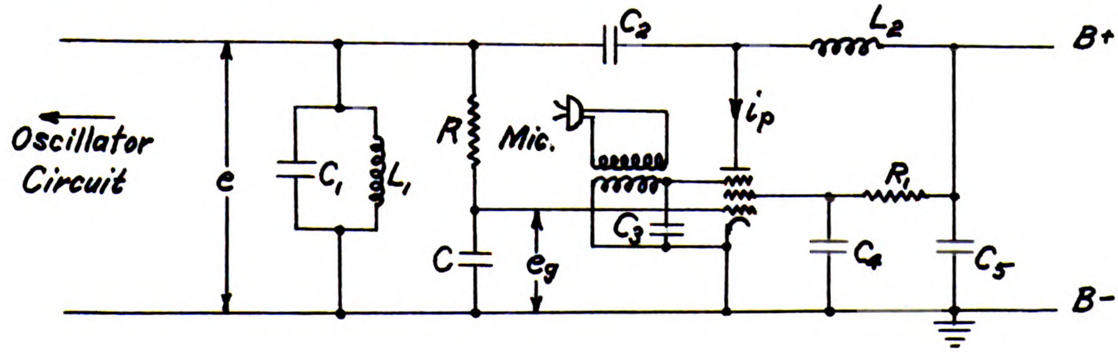


Figure 1

L_1C_1 is assumed to determine the oscillator frequency. C is a large condenser having a reactance which is in the operating range, very small in comparison with the resistance of R. Then the current through RC is in phase with e. i_p, in phase with e_g, thus lags e by 90°.. Now, if an increase in frequency occurs in the oscillator, e_g decreases; and i_p, being directly proportional to e_g, also decreases. These conditions indicate that the tube is acting as a dynamic positive inductance since the current vector lags the applied voltage vector by 90°, and the reactance increases (current decreases) with increasing frequency. Thus the frequency of oscillation actually depends not only upon L_1C_1, but also upon the reactance tube circuit.

If now an audio frequency signal is applied to the suppressor grid through the microphone and transformer, the amplitude of i_p varies in accordance with the audio signal voltage. As i_p varies, so varies the inductance in parallel with L_1; and, as the inductance varies, so varies the frequency. Thus the frequency of oscillation varies with the audio signal, resulting in a form of phase modulation.

Incidentally, it will be noticed that the variation of the amplitude of i_p with the audio signal is amplitude modulation which is necessary for the successful operation of this circuit.

Determination of Grid Circuit Arrangement.

The example presented above indicates the method of connection of the reactance tube in the circuit. This method is carried over directly into the video amplifier circuit shown in Figure 2. For the purpose of simplification, triode tubes are shown here, although pentodes are used in practice because of their high plate resistance and mutual conductance.

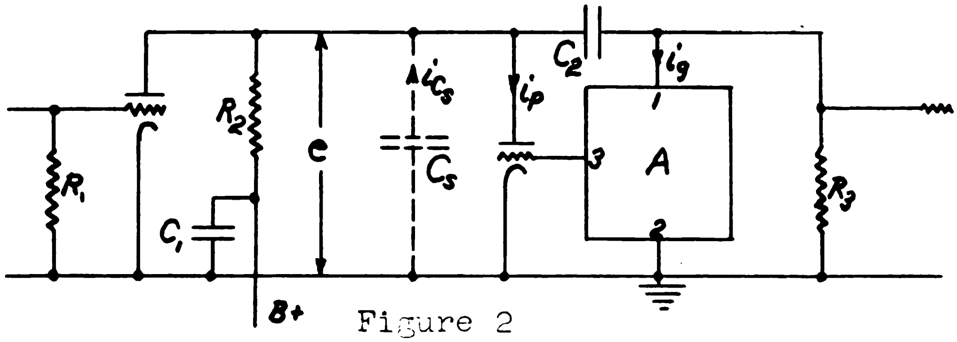


Figure 2

In the above figure C_s is the shunt capacitance of the high potential radio frequency circuit to ground. It is desired to neutralize C_s with the reactance tube shown to the right of C_s. In order that this may be accomplished, the vector diagram of Figure 3 must hold for all frequencies.

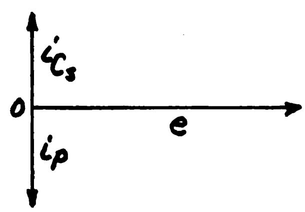


Figure 3

Remembering that i_{C_s} increases with increasing frequency (assuming e constant), the problem now is to find a three terminal network, A, which will give the vector relation above and at the same time insure that i_p and i_{C_s} increase with increasing frequency in the same manner.

The solution to this problem is found by means illustrated in Figure 4.

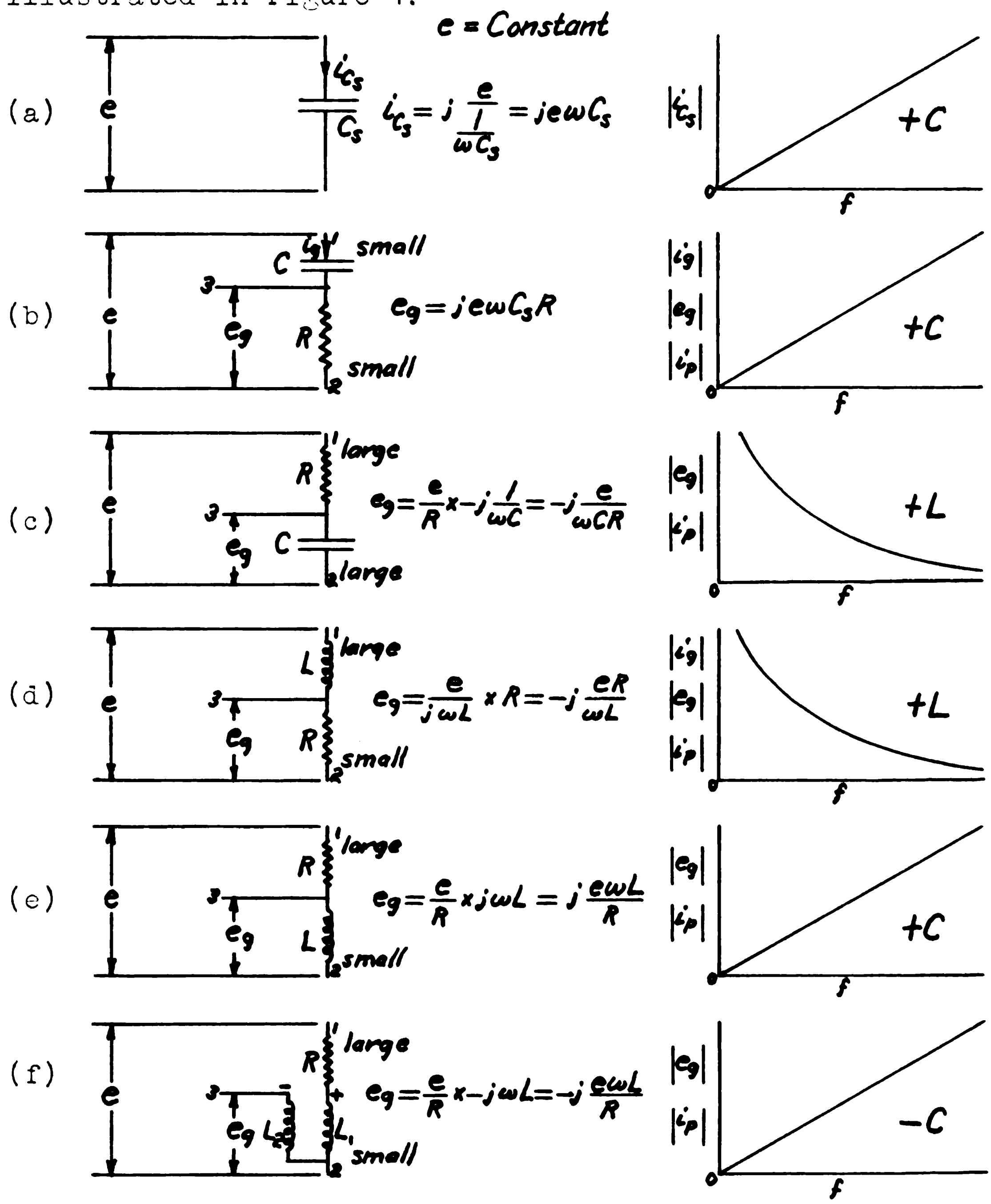


Figure 4

From Figure 4 it is apparent that the only possible form of the three terminal network, A, is that which consists of a large resistance in series with a radio frequency transformer so connected that the polarity of the secondary is reversed from that of the primary. Thus Figure 2 becomes Figure 5.

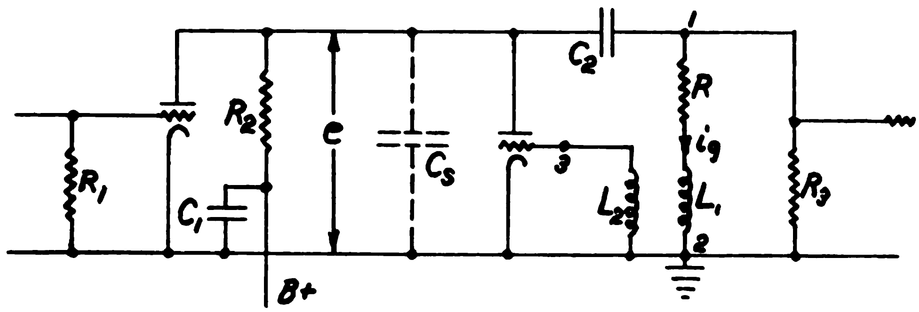


Figure 5

Examination of Grid Circuit Radio Frequency Transformer.

It is of interest to make a close inspection of the transformer of Figure 4 (f) and Figure 5. Since the apparent reactance of the transformer on the primary side must be small, the inductance of L_1 and L_2 must be small. For this reason the values of distributed capacitance across L_1 and L_2 become of importance because both primary and secondary must have resonant frequencies well above 4.5 megacycles in order that the apparent reactance will be inductive throughout the band-width. If the capacitance between windings is neglected here, the circuit under consideration is shown in Figure 6 (a).

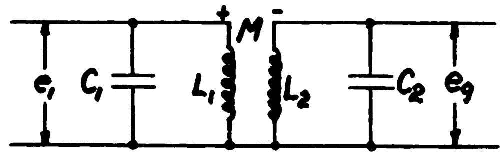


Figure 6 (a)

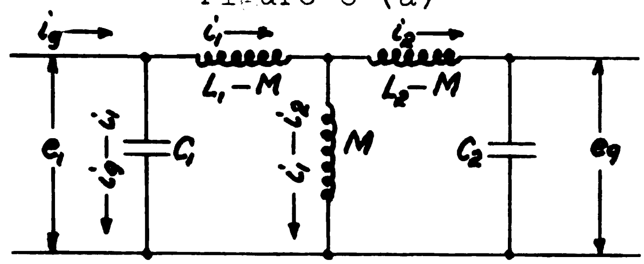


Figure 6 (b)

The equivalent circuit of Figure 6 (a) is shown in Figure 6 (b) with the assumed positive directions of currents as shown. Now the gain of the circuit can be calculated as follows:

(1) Gain $= \frac{e_g}{e_1}$,

but

$$e_g = -j\frac{i_2}{\omega C_2}$$

and

$$e_1 = -j\frac{i_g - i_1}{\omega C_1}$$

so that

(2) $$\text{Gain} = \frac{e_g}{e_1} = \frac{-j\frac{i_2}{\omega C_2}}{-j\frac{i_g - i_1}{\omega C_1}} = \frac{i_2 C_1}{(i_g - i_1)C_2}$$

Since the currents in parallel branches divide inversely as the respective impedances,

(3) $$i_2 = i_1 \frac{\frac{j\omega M\left[j\omega(L_2 - M) - j\frac{1}{\omega C_2}\right]}{j\omega M + j\omega L_2 - j\omega M - j\frac{1}{\omega C_2}}}{j\omega(L_2 - M) - j\frac{1}{\omega C_2}} = i_1\frac{\omega M}{\omega L_2 - \frac{1}{\omega C_2}}$$

and

$$i_g - i_1 = i_1 \frac{j\omega(L_1 - M) + \frac{j\omega M\left[j\omega(L_2 - M) - j\frac{1}{\omega C_2}\right]}{j\omega L_2 - j\frac{1}{\omega C_2}}}{-j\frac{1}{\omega C_1}}$$

(4) $$i_g - i_1 = i_1 \frac{\omega^2(M^2 - L_1L_2) + \frac{L_1}{C_2}}{\frac{1}{\omega C_1}\left(\omega L_2 - \frac{1}{\omega C_2}\right)}$$

Substituting equations (3) and (4) in (2)

(5) $$\text{Gain} = \frac{e_g}{e_1} = \frac{M}{C_2\omega^2(M^2 - L_1L_2) + L_1}$$

To find the actual value of e_g, e_1 must be evaluated. Thus

$$e_1 = i_g\left[\frac{\left(-j\frac{1}{\omega C_1}\left[j\omega(L_1 - M) + \frac{j\omega M\left[j\omega(L_2 - M) - j\frac{1}{\omega C_2}\right]}{j\omega L_2 - j\frac{1}{\omega C_2}}\right]\right)}{-j\frac{1}{\omega C_1} + j\omega(L_1 - M) + \frac{j\omega M\left[j\omega(L_2 - M) - j\frac{1}{\omega C_2}\right]}{j\omega L_2 - j\frac{1}{\omega C_2}}}\right]$$

$$= i_g\left[\frac{-j\frac{1}{\omega C_1}\left(\omega^2(M^2 - L_1L_2) + \frac{L_1}{C_2}\right)}{\frac{L_2}{C_1} + \frac{L_1}{C_2} + \omega^2(M^2 - L_1L_2) - \frac{1}{\omega^2 C_1 C_2}}\right]$$

(6) $$e_1 = -j\frac{i_g}{\omega C_1 C_2}\left[\frac{C_2\omega^2(M^2 - L_1L_2) + L_1}{\frac{L_2}{C_1} + \frac{L_1}{C_2} + \omega^2(M^2 - L_1L_2) - \frac{1}{\omega^2 C_1 C_2}}\right]$$

Therefore, from equations (1) and (6)

(7) $$e_g = -j\frac{i_g M}{\omega C_1 C_2}\left[\frac{1}{\frac{L_2}{C_1} + \frac{L_1}{C_2} + \omega^2(M^2 - L_1L_2) - \frac{1}{\omega^2 C_1 C_2}}\right]$$

If M, L_1, L_2, C_1, and C_2 are assigned representative values, the expressions for gain and e_g may be simplified to quite an extent. Let

$L_1 = L_2 = 80$ microhenries

$M = K\sqrt{L_1 L_2} = 0.2 \times 80 = 16$ microhenries

$C_1 = 5$ micromicrofarads

$C_2 = 10$ micromicrofarads (distributed capacitance of coil plus input capacitance of tube.)

Then the gain is

$$\text{gain} = \frac{16\times10^{-6}}{10\times10^{-12}\,\omega^2(256-6400)\times10^{-12} + 80\times10^{-6}} \cong \frac{16}{80} = 0.2$$

Thus, providing the anti-resonance frequencies of L_1C_1 and L_2C_2 are well above 4.5 megacycles, the gain may be represented by

$$(8)\quad \text{gain} = \frac{M}{L_1}$$

Now

$$e_g = -j\frac{i_g\,16\times10^{-6}}{\omega\times50\times10^{-24}}\left[\frac{1}{24\times10^{-6} - 6144\,\omega^2\times10^{-12} - \dfrac{0.02\times10^{-24}}{\omega^2}}\right]$$

It will be seen that the last term in the denominator of the bracketed factor is predominant. Thus it is possible to write

$$(9)\qquad e_g = -j\frac{i_g M}{\omega C_1 C_2}\times\frac{1}{-\dfrac{1}{\omega^2 C_1 C_2}} = j\omega i_g M$$

assuming the same conditions as given under equation (8).

Equation (9) indicates that e_g is shifted from i_g by a positive 90°. Actually this shift may be made the desired negative 90° simply by choosing the correct connection of the secondary. Also, equation (9) shows that e_g is directly proportional to frequency as has been shown necessary in Figure 4.

To date it has been tacitly assumed that the reactance tube's plate current is in phase with the grid voltage. This is strictly true only when the load is purely resistive. Since the load in this case is highly capacitive, the alternative is to have a plate resistance sufficiently high to overshadow the capacitive reactance. Thus the reactance tube must be a pentode.

The Value of the Negative Capacitance

Now the alternating current circuit of the video amplifier may be drawn as in Figure (7).

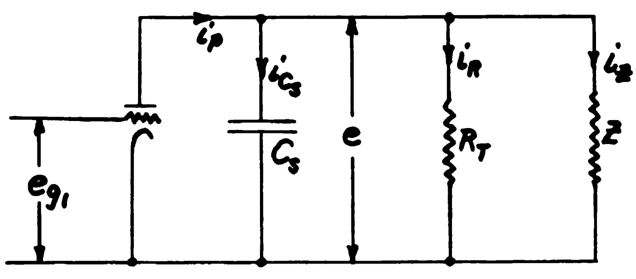


Figure 7.

In Figure (7) Z represents the reactance tube plate circuit. R_T includes R_2, R_3, and R of Figure 5. C_s is the shunt capacitance.

Since the reactance tube is a pentode, it may, in general, be regarded as a constant current generator. Whence

$$i_z = g_m e_g .$$

Substituting the value of e grom equation (9)

(10) $$i_z = -j\omega\, g_m i_g M = \frac{-j\omega\, g_m e M}{R} .$$

Now

$$i_p = i_R + i_{C_s} + i_z$$

or

(11) $$i_p = \frac{e}{R_T} + \frac{e}{-j\frac{1}{\omega C_s}} + \frac{e}{j\frac{R}{\omega g_m M}} .$$

It is desirable that the last two terms of equation (11) should be equal and opposite so that i_p will be in phase with e. If this condition holds, then the apparent negative capacitance must be

(12) $$-C = -\frac{g_m M}{R} .$$

From equation (12) it may be seen that the value of negative capacitance is dependent directly upon the mutual conductance of the reactance tube and the mutual inductance of the grid transformer. The value of R is more or less fixed in the neighborhood of 25,000 ohms as R must always be much greater than the equivalent reactance of the radio frequency transformer primary.

Effect of the Distributed Capacitance between Windings of Grid Circuit Transformer

It is now of interest to investigate the effect of the distributed capacitance between the windings of the reactance tube grid transformer. Since the lower ends of the two windings are tied together, the maximum voltage between coils will be at the free ends; so that the distributed capacitance may be considered as being lumped across the free ends of the transformer. This is shown in Figure (8) (a) and (b)

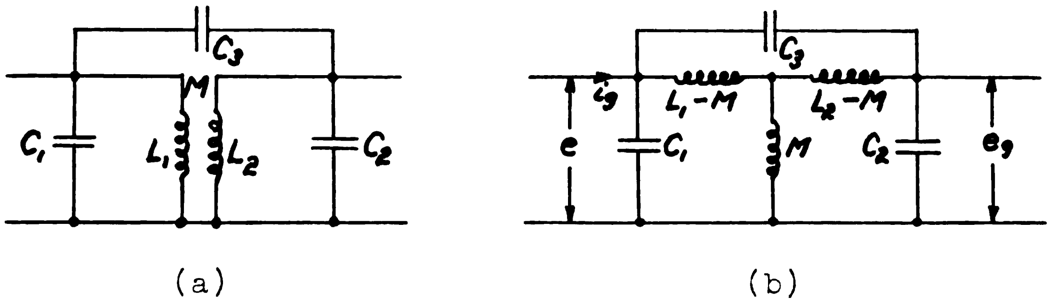


Figure 8

The circuit of Figure (8) b is not easily solved for e_g ; therefore use is made of the equivalent π section of the transformer. The conversion from T to π may be made according to equation (13) with reference to Figure 9.

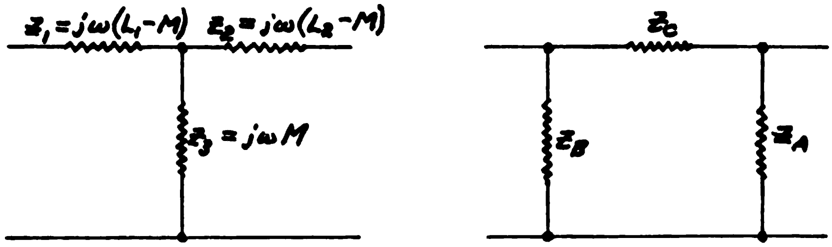


Figure 9

(a) $$Z_A=\frac{Z_1Z_2+Z_2Z_3+Z_3Z_1}{Z_1}$$

(13) (b) $$Z_B=\frac{Z_1Z_2+Z_2Z_3+Z_3Z_1}{Z_2}$$

(c) $$Z_C=\frac{Z_1Z_2+Z_2Z_3+Z_3Z_1}{Z_3}$$.

Thus

$$(14)\quad \begin{aligned} &(a)\ Z_A = j\omega \frac{L_1 L_2 - M^2}{L_1 - M} \\ &(b)\ Z_B = j\omega \frac{L_1 L_2 - M^2}{L_2 - M} \\ &(c)\ Z_C = j\omega \frac{L_1 L_2 - M^2}{M} \ . \end{aligned}$$

Now the equivalent circuit of Figure 8 (a) may be drawn as shown in Figure 10.

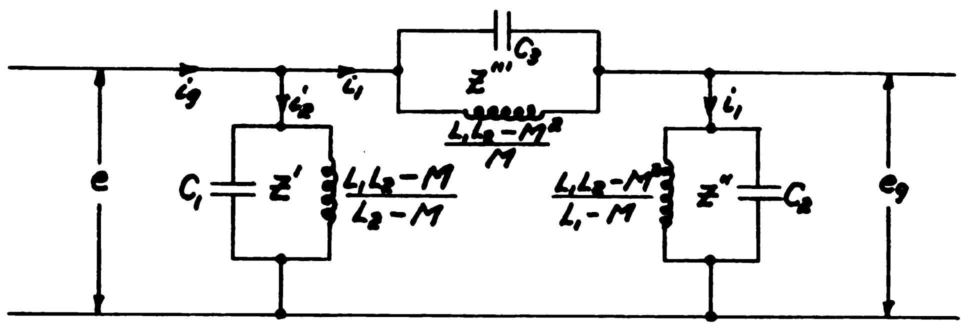


Figure 10

Let

$$(15)\qquad A = L_1 L_2 - M^2 \ .$$

Now the values of the parallel impedances, Z', Z'', and Z''', shown in Figure 10 may be found as follows:

$$Z' = \frac{\left(-j\frac{1}{\omega C_1}\right)\left(j\omega\frac{A}{L_2 - M}\right)}{-j\frac{1}{\omega C_1} + j\omega\frac{A}{L_2 - M}}$$

or

$$(16)\quad \begin{aligned} &(a)\ Z' = \frac{j\omega A}{L_2 - M - \omega^2 C_1 A} \\ &(b)\ Z'' = \frac{j\omega A}{L_1 - M - \omega^2 C_2 A} \\ &(c)\ Z''' = \frac{j\omega A}{+M - \omega^2 C_3 A} \ . \end{aligned}$$

The gain of the transformer circuit may now be evaluated.

$$(17)\ \text{Gain} = \frac{e_g}{e} = \frac{i_1 Z''}{i_2 Z'}$$

$$i_2 = i_1 \frac{z'' + z'''}{z'} = i_1 \frac{\left[L_1 - \omega^2 A \; C_2 + C_3\right]\left[L_2 - M - \omega^2 C_1 A\right]}{\left[L_1 - M - \omega^2 C_2 A\right]\left[M - \omega^2 C_3 A \quad\right]} .$$

Substituting i_2 in equation (17)

$$\text{Gain} = \frac{M - \omega^2 C_3 A}{L_1 - \omega^2 A(C_2 + C_3)}$$

(18) $$= \frac{M - \omega^2 C_3 (L_1 L_2 - M^2)}{L_1 - \omega^2 (C_2 + C_3)(L_1 L_2 - M^2)} .$$

From equation (17)

$e_g = e$ gain.

But

$$e = i_g \frac{z'(z'' + z''')}{z' + z'' + z'''}$$

or

$$e = i_g \frac{j\omega \frac{A}{L_2 - M - \omega^2 C_1 A}\left[\frac{1}{L_1 - M - \omega^2 C_2 A} + \frac{1}{M - \omega^2 C_3 A}\right]}{\frac{1}{L_2 - M - \omega^2 C_1 A} + \frac{1}{L_1 - M - \omega^2 C_2 A} + \frac{1}{M - \omega^2 C_3 A}} .$$

Simplifying this expression for e, there results

$$e = i_g \frac{j\omega \left[L_1 - \omega^2 (C_2 + C_3) A\right]}{1 - L_1 \omega^2 (C_1 + C_3) - L_2 \omega^2 (C_2 + C_3) + \omega^4 A(C_1 C_2 + C_2 C_3 + C_3 C_1) + 2M\omega^2 C_3} .$$

Multiplying by the gain as given in equation (18),

(19) $$e_g = \frac{j\omega M i_g}{1 - L_1 \omega^2 (C_1 + C_3) - L_2 \omega^2 (C_2 + C_3) + \omega^4 (L_1 L_2 - M^2)(C_1 C_2 + C_2 C_3 + C_3 C_1) + 2M\omega^2 C_3} .$$

If $C_3 = 0$, equations (18) and (19) become

(20) $$\text{Gain} = \frac{M}{L_1 - \omega^2 C_2 (L_1 L_2 - M^2)}$$

and $e_g = \dfrac{j\omega M i_g'}{1-\omega^2 L_1 C_1 - \omega^2 L_2 C_2 + \omega^4 C_1 C_2 (L_1 L_2 - M^2)}$.

Dividing numerator and denominator by $-\omega^2 C_1 C_2$

(21) $e_g = -j\dfrac{i_g' M}{\omega C_1 C_2}\left[\dfrac{1}{\frac{L_1}{C_2} + \frac{L_2}{C_1} - \omega^2(L_1 L_2 - M^2) - \frac{1}{\omega^2 C_1 C_2}}\right]$.

Equations (20) and (21) are identical with equations (5) and (7).

Now, as before, representative values of M, L_1, L_2, C_1, C_2, and C_g are assumed as follows:

$L_1 = L_2 = 80$ microhenries

$M = K\sqrt{L_1 L_2} = 0.2 \times 80 = 16$ microhenries

$C_1 = 5$ micromicrofarads

$C_2 = 10$ micromicrofarads

$C_g = 5$ micromicrofarads

Substituting the above values in equation (18),

$$\text{Gain} = \frac{16\times10^{-6} - \omega^2\times10^{-12}(6400-256)\times10^{-12}}{80\times10^{-6} - \omega^2\times20\times10^{-12}(6400-256)\times10^{-12}}$$

$$\cong \frac{16}{80} = 0.2 \ .$$

Thus, as before, the gain may well be represented as

(22) $\text{Gain} = \dfrac{M}{L_1}$.

Under the same condition

$e_g = j\omega i_g' \times 16\times10^{-6}$.

Therefore,

(23) $e_g = j\omega i_g' M$

where the sign of e_g may be either positive or negative according to the connection of the transformer secondary.

The effect of the distributed capacitance between windings of the transformer is thus negligible if the value of the capacitance is of the same order of magnitude as the distributed capacitance of one winding alone.

2. Problem of Constant Time-Delay.

A constant time delay for varying frequency means that the phase shift must be directly proportional to frequency. The phase shift of an amplifier is inherently 180° so that the phase shift which must be proportional to frequency is given by equation (24).

(24) $\phi_a = \phi - 180°$

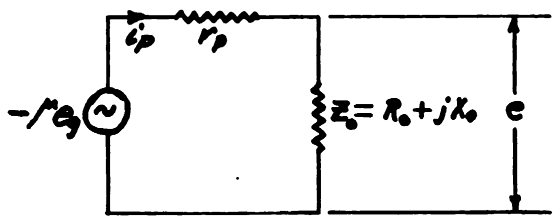


Figure 11

Using the equivalent circuit of Figure 11,

$$i_p = \frac{-\mu e_g}{r_p + Z_o}$$

and

$$e = i_p Z_o = \frac{-\mu e_g Z_o}{r_p + Z_o}$$

so that

(25) $$\text{gain} = \frac{e}{-e_g} = \frac{\mu Z_o}{r_p + Z_o} = \frac{\mu(R_o + jX_o)}{r_p + R_o + jX_o}$$

Now the negative sign attached to e_g takes care of the inherent 180° shift, so that the added phase shift, ϕ_a , is given by the phase angle of the gain. The tangent of ϕ_a may be found by separating the gain into its real and imaginary parts as below:

$$\text{gain} = \frac{\mu(R_o + jX_o)}{r_p + R_o + jX_o} = \frac{\mu\left(R_o r_p + R_o^2 + X_o^2\right) + j\mu(X_o r_p + R_o X_o - R_o X_o)}{(r_p + R_o)^2 + X_o^2}.$$

Therefore

(26) $$\phi_a = \tan^{-1}\frac{X_o r_p}{R_o^2 + R_o r_p + X_o^2} .$$

Since the amplifier tube used is a pentode with a plate resistance much greater than the plate load,

(27) $$\phi_a = \tan^{-1}\frac{X_o}{R_o} .$$

If the phase shift is restricted to small angles,

(28) $$\phi_a \cong \frac{X_o}{R_o} .$$

Now, by inspecting the equivalent load shown in Figure 12, the values of series reactance and resistance, R_o and X_o , may be determined.

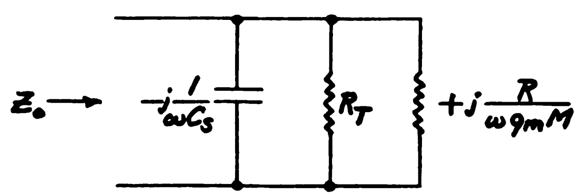


Figure 12

thus

$$Z_o = \frac{1}{\dfrac{1}{-j\frac{1}{\omega C_s}} + \dfrac{1}{R_T} + \dfrac{1}{\frac{jR}{\omega g_m M}}}$$

$$= \frac{R_T R^2 - j\omega R_T^2 R(C_s R - g_m M)}{R^2 + (\omega C_s R_T R - \omega g_m M R_T)^2} .$$

Substituting in equation (28),

$$(29) \quad \phi_a = \frac{-\omega R_T (C_0 R - g_m M)}{R} \ .$$

Therefore the phase shift of the amplifier is always proportional to the frequency, or the time-delay is always constant.

Part II

Design, Construction, and Operation of the Amplifier

In this section are described the preliminary design, the construction and operation of the first circuit, and the construction and operation of the second circuit.

1. Preliminary Design.

In the design of the amplifier, a number of distinct problems were faced. These problems and their solutions will be presented in as logical an order as possible.

The circuit of the amplifier as first proposed is illustrated in Figure 13.

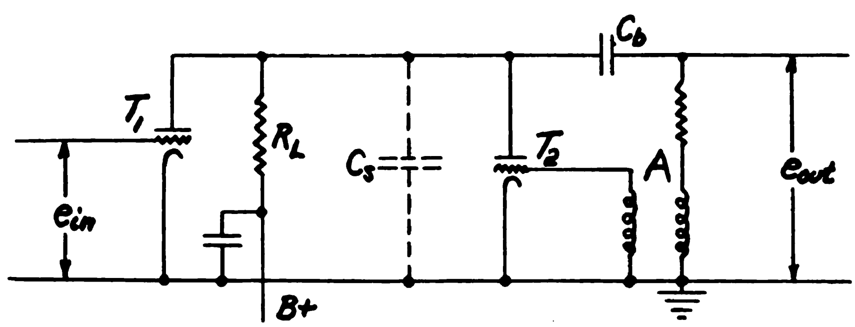


Figure 13

In this Figure

T_1 is the amplifier or gain tube, a pentode.

T_2 is the reactance tube, a pentode.

C_s is the shunt capacitance to ground.

C_b is the blocking condenser.

R_L is the resistor through which direct current is supplied.

A is the coupling network of the reactance tube grid.

e_{in} is the input voltage.

e_{out} is the output voltage.

The 6AC7/1852 television pentode as T_1 seemed the best tube available. It is commonly used in video amplification circuits because of its high plate resistance and extremely high mutual conductance. The output capacitance of this tube, a factor contributing to C_S, is but 5 micromicrofarads. The normal operating conditions of this tube as a class A amplifier are given below.

Heater Voltage	6.3 volts
Plate Voltage	300 volts
Suppressor Voltage	0 volts
Screen Voltage	150 volts
Cathode Bias Resistor	160 ohms minimum
Plate Resistance (Approx.)	0.75 megohms
Transconductance	9000 micromhos
Plate Current	10 Milliamperes
Screen Current	2.5 Milliamperes

Next the tube to be used as T_2 was determined. Since the grid voltage to this tube must necessarily be low because of the connecting network, the tube must have a high mutual conductance. Now a consideration of the magnitude of the current taken by C_s at 4.5 megacycles per second will indicate the necessary current capacity of the reactance tube. Let

$C_s = 40$ micromicrofarads

$e_{out}=15$ volts .

Then

$$i_{C_s}= \frac{e_{out}}{\frac{1}{\omega C_s}} = e_{out}\omega C_s =15 \times 2\pi \times 4.5 \times 10^{6} \times 40 \times 10^{-12}$$

$$=17 \text{ milliamperes.}$$

Multiplying by $\sqrt{2}$

(30) $i_{C_s max} = 24$ milliamperes.

Because of the requirements of high alternating plate current, and high mutual conductance, and high plate resistance (discussed in Part I); the tube chosen as T_2 was the 6AG7, a television power pentode. The output capacitance of this tube is 12 micromicrofarads. A bad feature of this tube is its high input capacitance of 12 micromicrofarads. This value, however, is typical of the television pentodes. The normal operating conditions of the 6AG7 as a class A amplifier are given below.

Heater voltage	6.3 volts
Plate voltage	300 volts
Suppressor voltage	0 volts
Screen voltage	300 volts
Grid voltage	-10.5 volts
Plate current	25 milliamperes
Screen current	6.5 milliamperes

Plate resistance	0.1 megohm
Transconductance	7700 micromhos

Next encountered was the problem of plate feed for the 6AC7 and 6AG7. The combined direct plate current of the two tubes was 35 milliamperes. It was desired to make the load resistance as high as possible in order that most of the alternating plate current would be shunted through the reactance tube's grid circuit. Because of the requirement that the current through this grid circuit be in phase with the voltage across it, the value of the resistor here was fixed in the neighborhood of 25,000 ohms. Thus the plate feed resistor had to be at least 40,000 ohms. A 40,000 ohms resistor carrying 35 milliamperes would have had a direct current voltage across it of

(31) $E_{DC} = IR = 0.035 \times 40,000 = 1400$ volts.

Thus the power supply would have had to supply 35 milliamperes at 1700 volts direct current. This was a practical impossibility. Incidentally, the power dissipated by the above resistor would have been

$$P = I^2R = EI = 1400 \times 0.035 = 49 \text{ watts}$$

Although no tests were run, it was feared that a resistor capable of dissipating 49 watts would have had an anti-resonant frequency below 4.5 megacycles per second due to distributed inductance and capacitance. Obviously this type of parallel feed was ruled out.

At this point it was recalled that pentodes have a very high alternating current plate resistance. Thus if a pentode could be selected which was capable of passing 35 milliamperes, the plate feed problem was solved. The 6F6G power pentode was selected. The output capacitance of this tube is less than the equivalent 6F6 metal tube, and its other characteristics meet the requirements as is demonstrated by its normal operating conditions as a class A_1 amplifier.

Heater voltage	6.3 volts
Plate voltage	250 volts
Screen voltage	250 volts
Cathode resistor	410 volts
Plate current	34 milliamperes
Screen current	6.5 milliamperes
Plate resistance	80,000 ohms

Since the available power supply had a 500 volt output, it was decided to operate the 6AC7/1852 and 6AG7 at a plate voltage of 250 volts. At this voltage, the operating conditions as class A amplifiers are as follows:

	6AC7/1852	6AG7
Heater voltage	6.3	6.3 volts
Plate voltage	250	250 volts
Screen voltage	150	140 volts
Grid voltage	-1.67	-3.0 volts

	6AC7/1852	6AG7	
Plate Current	8.0	35.0	milliamperes
Screen current	1.72	8.7	milliamperes
Plate resistance	0.7	0.0667	megohms
Transconductance	9000	10,000	micromhos

The values of plate resistance given in the above table were found experimentally and the values of transconductance were determined from data given in the RCA Receiving Tube Manual, 1940.

Next undertaken was the design of perhaps the most important single part of the amplifier, that of the reactance tube's grid transformer. As shown in Part I, if the anti-resonant frequency is above 4.5 megacycles per second, the output voltage of the transformer is directly proportional to the mutual inductance, M. Thus it was desired to measure the anit-resonant frequency of primary and secondary, and the mutual inductance.

The circuit used to measure the anti-resonant frequencies of the coils is shown in Figure 14.

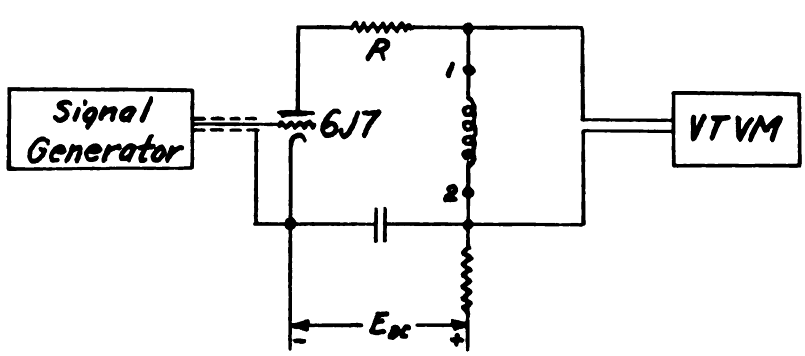


Figure 14

In Figure 14 the coil to be tested was connected across the terminals 1-2, and a signal which varied in frequency from 100 kilocycles per second to 31 megacycles per second was impressed upon the grid of the 6J7. The lowest frequency at which a peak was recorded on the vacuum tube voltmeter indicated the anti-resonant frequency of the coil in question. The resistor R (25,000 ohms) was included to prevent the output capacitance of 6J7 from playing a part in determining the anti-resonant frequency. Thus only the input capacitance of the vacuum tube voltmeter influenced the anti-resonant frequency.

This input capacitance was measured by determining the change in frequency of an oscillator when the vacuum tube voltmeter was first connected across the frequency determining circuit and then removed. Since the capacitance across the above circuit with the VTVM removed must be known in this method, a General Radio Precision Type Variable Condenser was used. The precision condenser was set at a value large enough that the distributed capacitance of the coil and the electrode capacitance of the oscillator might be neglected. Then

$$(32)\quad C_x = C_0\left[\left(\frac{f_0}{f_x}\right)^2 - 1\right]$$

where

C_x is the unknown capacitance

C_0 is the capacitance of the precision condenser.

f_x is the oscillator frequency with C_x in the circuit

f_0 is the oscillator frequency with C_x removed.

The resonant frequencies were determined using a radio receiver with a beat frequency oscillator. By this method

$C_x = 12.0$ micromicrofarads.

This value of C_x was checked by using the substitution method at 1000 cycles per second. Here a null was found on a General Radio bridge using only the precision condenser mentioned above. Next the vacuum tube voltmeter was connected in parallel with the precision condenser and the precision condenser adjusted again to the same null. The change in the precision condenser setting gave directly

$C_x = 11.8$ micromicrofarads.

Since this value was necessarily measured with the circuit cold and with the grid resistor disconnected, it is presumably slightly low.

The value of the mutual inductance was determined by measuring the primary, secondary, and total inductance with a 1000cycle per second bridge. At 1000 cycles per second the effect of the distributed capacitance was negligible. The total inductance was measured series aiding and series opposing as a check. From the values of total inductance

(33) $M = \pm\left[\dfrac{L_T - (L_p + L_s)}{2}\right]$

The positive sign was used when the coils were connected series aiding and the negative sign when the coils were connected series opposing.

In equation (33)

M is the mutual inductance

L_T is the total inductance

L_P is the primary inductance

L_S is the secondary inductance.

In Figure 15 are tabulated the values determined from the use of equation (33) and the circuit of Figure 14. The resonant frequencies given are those obtained with the vacuum tube voltmeter in the circuit. The input capacitance of the vacuum tube voltmeter, 12 micromicro-farads, has been subtracted from the total capacitance to give the distributed capacitances listed. For the secondary windings the actual resonant frequencies were nearly the same, when the coils were in the circuit, as those given in Figure 15 because the input capacitance of the 6AG7 tube is very nearly the same as that of the vacuum tube voltmeter.

Coil Number	Winding	Number of Turns Pri.	Number of Turns Sec.	Pri. r_1	Pri. r_2	Pri. l	Sep.	Sec. r_1	Sec. r_2	Sec. l
1	o.u.			1/4	3/4	7/16				
2	o.u.			7/32	5/16	1/4				
80A	c.s.	21	5	3/4	3/4	7/16		3/4	3/4	3/32
3	o.u.			1/4	3/8	5/16				
4	c.u.	100		3/8	7/16	5/16				
5	o.u.	100		3/8	7/16	3/8				
6	v.o.u.	100		3/8	13/32	5/8				
6*	v.o.u.	100		3/8	13/32	5/8				
7	c.s.	38		3/8	3/8	3/8				
8	o.u.	100	100	3/8	7/16	11/32	1/4	3/8	7/16	11/32
8*	o.u.	100	100	3/8	7/16	11/32	1/4	3/8	7/16	11/32
9	o.u.	50	50	3/8	13/32	1/4	1/4	3/8	13/32	1/4
9*	o.u.	50	50	3/8	13/32	1/4	1/4	3/8	13/32	1/4
10	v.o.u.	40	40	3/8	13/32	5/16	3/16	3/8	13/32	5/16
10*	v.o.u.	40	40	3/8	13/32	5/16	3/16	3/8	13/32	5/16
12	o.u.	40	40	3/8	7/16	3/16	3/16	11/32	13/32	3/16
12*	o.u.	40	40	3/8	7/16	3/16	3/16	11/32	13/32	3/16
14	c.s.	100		3/8	3/8	3/4				
14*	c.s.	100		3/8	3/8	3/4				
15	c.s.	75		3/8	3/8	5/8				
15*	c.s.	75		3/8	3/8	5/8				
16	c.s.	50		3/8	3/8	7/16				
17	o.u.	75		3/8	1/2	5/16				
18	o.u.	35		3/8	7/16	1/4				
19	o.u.	90	150	3/8	1/2	5/16	1/16	3/8	9/16	5/16
20	o.u.	80	100	5/16	7/16	1/4	1/16	3/8	1/2	1/4

Figure 15 (concluded on next page)

In Figure 15

* means powdered iron core

o.u. means open universal

c.s. means close solenoid

c.u. means close universal

v.o.u. means very open universal

Pri. means primary

Sec. means secondary

r_1 is inner radius

r_2 is outer radius

l is length of winding

Sep. means separation between windings.

Coil No.	Resonant Frequency, Megacycles		Inductance, Microhenries		Mutual Inductance, Microhenries	Distributed Capacitance, Micromicro-farads	
	Pri.	Sec.	Pri.	Sec.		Pri.	Sec.
1	0.19		17,100			28.1	
2	3.94		95			5.2	
80A	6.75	13.4	31	10		6.0	2.1
3	3.76		100			6.2	
4	3.29		125			6.7	
5	3.76		100			5.9	
6	4.28		88			3.7	
6*	3.63		112			5.2	
7	6.25		39			4.8	
8	3.23	3.23	135	135	21	6.0	6.0
8*	2.46	2.33	230	255	37	6.2	6.3
9	4.48	4.50	83	83	14	3.2	3.1
9*	4.47	4.36	69	72	19	6.4	6.5
10	5.07	5.07	72	72	29	1.7	1.7
10*	4.0	4.19	64	60	23	12.7	11.9
12	4.62	4.67	87	84	37	1.7	1.8
12*	3.86	3.75	65	68	17	14.2	14.5
14	3.75		82			12.0	
14*	2.4		125			23.5	
15	4.5		70			5.9	
15*	2.8		95			25.9	
16	5.6		57			2.1	
17	3.8		98			5.4	
18	5.1		70			1.9	
19	3.78	2.85	110	177	42	4.1	5.6
20	3.97	3.37	100	130	40	4.1	5.2

Figure 15 (concluded)

A very striking thing will be seen in the tabulations of primary and secondary inductances. The insertion of a powdered iron core increased the inductance of coils having 75 turns or better. This is the normal result. For coils having 50 turns or less, however, the inductance decreased with the insertion of the powdered iron core. The core, being a semi-conductor, increased the distributed capacitance sufficiently that the resonant frequency of the coil decreased. There is no apparent explanation of this phenomenon.

Coils number 12 and 20 were wound on concentric paper tubes which were tightly fitted so that the value of mutual inductance might be varied simply by sliding the inner tube up and down in the outer tube.

Coil No. 10 proved to be the only coil which had an anti-resonant frequency high enough that equations (8), (9), and (12) applied across the complete band.

2. Construction and Operation of the First Circuit.

Based upon the preliminary design of the preceding section, the circuit of Figure 16 was constructed upon a metal chassis.

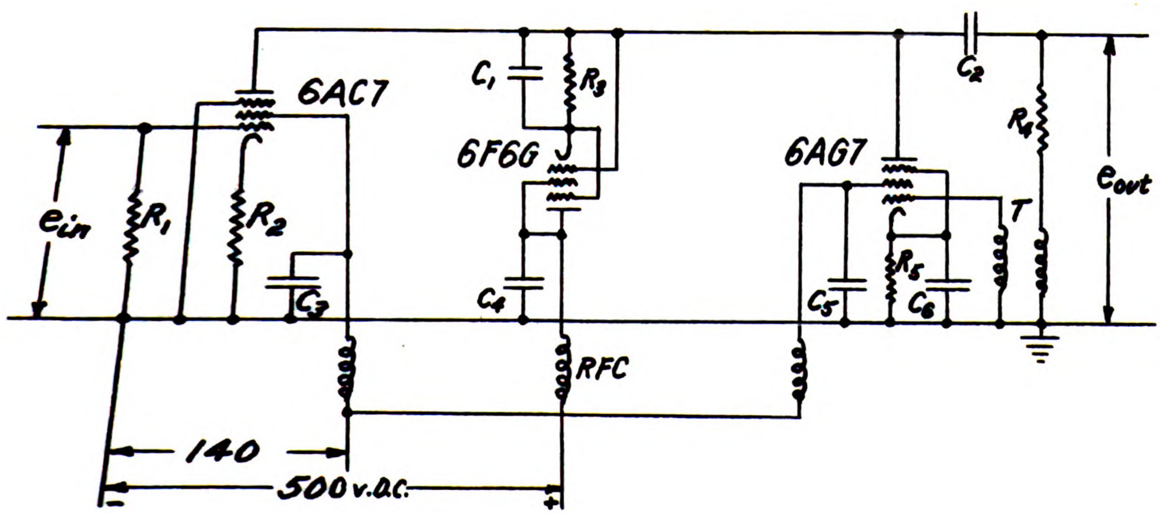


Figure 16

In Figure 16

R_1 = 1 megohm

R_2 = 160 ohms

R_3 = 460 ohms

R_4 = 25,000 ohms

R_5 = 50 ohms

C_1 = .002 mica + .01 + 25 electrlytic (microfarads)

$C_2 = C_6$ = .002 mica + .01 paper + 1 paper (microfarads)

$C_3 = C_4 = C_5$ = .01 paper

RFC = Radio Frequency Choke - 8 millihenries

T as determined from Figure 15.

The heater circuits are not shown in Figure 16. Since the cathode of the 6F6G was 250 direct current volts above ground potential, it was necessary to use a separate 6.3 volt transformer with the secondary at the same direct potential as the 6F6G cathode. Here another problem was encountered in the form of the cathode-heater capacitance. The impedance of the transformer secondary to ground was very small so that the cathode-heater capacitance, which was thought to be large, practically shortcircuited the 6AC7/1852 alternating plate current to ground. The approximate value of this capacitance was calculated from the formula for the capacitance for two infinite concentric cylinders.

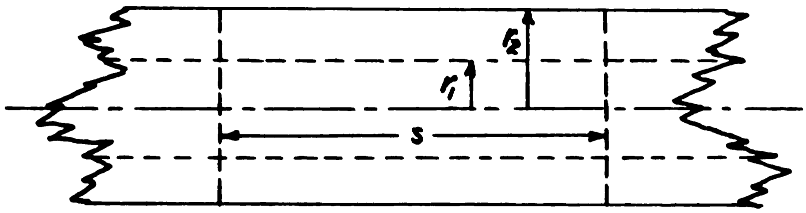


Figure 17

(34) $$C=\frac{ks}{2\log\frac{r_2}{r_1}}\times\frac{10}{9}=\frac{1.11ks}{2\log\frac{r_2}{r_1}}$$

Here

C = capacitance in micromicrofarads

r_1 = 0.025 centimeters

r_2 = 0.1 centimeters

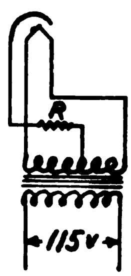
s = 3.0 centimeters

k = dielectric constant.

Hence

(35) $$C=\frac{1.11\times 3k}{2\log\frac{0.1}{0.025}} = 1.2\ k \text{ micromicrofarads.}$$

Equation (35) gave such a low value of the heater-cathode capacitance that it was measured by the substitution method and found to be 9 micromicrofarads. Thus this problem was satisfactorily solved by the 6F6G cathode connection of Figure 18.



$R = 250{,}000$ ohms.

Figure 18

The gain of the circuit of Figure 16 was of the order of 4 at 100 kilocycles per second. Under normal operating conditions

$$(36)\quad \text{Gain} = \frac{i_p Z_L}{e_g} = \frac{g_m e_g Z_L}{e_g} = g_m Z_L = 0.009 \times Z_L .$$

$$\text{or}\quad Z_L = \frac{\text{Gain}}{g_m} = \frac{4.0}{0.009} = 444 \text{ ohms.}$$

As calculated from the circuit constants, Z_L should have been of the order of 8000 ohms. Thus it was evident that operating conditions were not normal and that steps must be taken to raise the gain. All of the following work was done with the 6AG7 grid grounded.

First the removal of the 6F6G cathode by-pass condensers raised the gain from 4.5 to 10.5 at 100 kilocycles per second. Next the 6AC7/1852 cathode resistor was changed from 160 ohms to 180 ohms changing the gain from 10.5 to 15.0. The screen of the 6F6G was next by-passed to the cathode with a 0.01 microfarad paper condenser, and a 10 millihenry radio frequency choke was inserted in the B + lead to the screen. The measured gain at 100 kilocycles was now 36.8.

Now a measurement of the shunt capacitance, C_s , at 100 cycles per second by the substitution method gave 123 micromicrofarads with the 6F6G screen by-pass condenser disconnected, and 17,000 micromicrofarads with the by-pass condenser connected. These measurements were taken with the vacuum tube voltmeter connected to the output.

Finally a 25,000 ohm resistor was substituted for the 6F6G screen choke and, with a 1700 ohm resistor in the 6F6G plate lead, the gain was measured at 51.8.

Since the 6AC7/1852 cathode resistor had not been passed to this point, a certain amount of negative feedback had been present. When this cathode resistor was by-passed by a 1.0 microfarad paper condenser, the gain rose to 71.

The load circuit of the 6AC7/1852 was of the type shown in Figure 19.

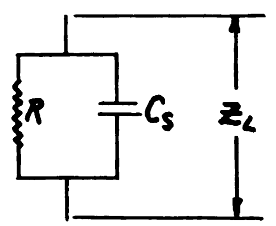


Figure 19

Here

R = load resistance

C_s = shunt capacitance ≅ 123 micromicrofarads.

Now

$$Z_L = \frac{-j\frac{R}{\omega C_s}}{R - j\frac{1}{\omega C_s}} = \frac{R - j\omega C_s R^2}{1 + \omega^2 C_s^2 R^2}$$

and the absolute value of Z_L is

(37) $$|Z_L| = \frac{R}{\sqrt{1 + \omega^2 C_s^2 R^2}}.$$

Rearranging

$$(38) \qquad R = \frac{|Z_L|}{\sqrt{1 - \omega^2 C_s^2 |Z_L|^2}} .$$

Making use of equation (36)

$$|Z_L| = \frac{Gain}{g_m} = \frac{71}{0.009} = 7{,}890 \text{ ohms} .$$

Substituting this value of $|Z_L|$, C_s=123 micromicrofarads, and $\omega = 2\pi \times 10^5$ radians per second in equation (38) gives

$$(39) \qquad R = 9{,}950 \text{ ohms.}$$

The load resistance was made up of the 6F6G screen resistor, 6F6G plate resistance, 6AG7 grid resistor, and 6AG7 plate resistance all acting in parallel and was approximately equal to 9,920 ohms. Thus the amplifier was operating normally at this point.

Now, with the 6AG7 grid grounded so that no negative capacitance was acting, a set of readings of gain versus frequency was taken and recorded in Figure 20. Next the 6AG7 grid was connected to the secondary of radio frequency transformer No. 12 so that the reactance tube was functioning as a negative capacitance. A second set of readings was now taken and also recorded in Figure 20. These data were then plotted in Figure 21. Since the low frequency response is not being considered here, it was not thought adviseable to use a logarithmic frequency scale.

Frequency in Megacycles.	Coil #12 out.	Coil #12 in.
0.1	70	72
0.5	45	45
1.0	31	32
1.5	25	27
2.0	23	25
2.5	19	23
3.0	15	20
3.5	14.3	17
4.0	13.8	17
4.5	12.5	16

Figure 20

It will be seen from Figure 21 that some negative capacitance was acting. When the data on coil No. 12 was substituted in equation (7) of Part I, it was found that the value of negative capacitance at 4.5 mega cycles per second was theoretically 407 micromicrofarads if i_g of equation (7) was considered to be equal to $\frac{e}{R}$ as given in Figure 5. Actually i_g was not equal to $\frac{e}{R}$ but was much smaller due to the reflected impedance of the transformer secondary which was approaching antiresonance. Thus the acting negative capacitance must be determined from a rearrangement of equation (37).

(40) $$C_s = \frac{\sqrt{R^2 - |Z_d|^2}}{\omega |Z_d| R} .$$

Now, if

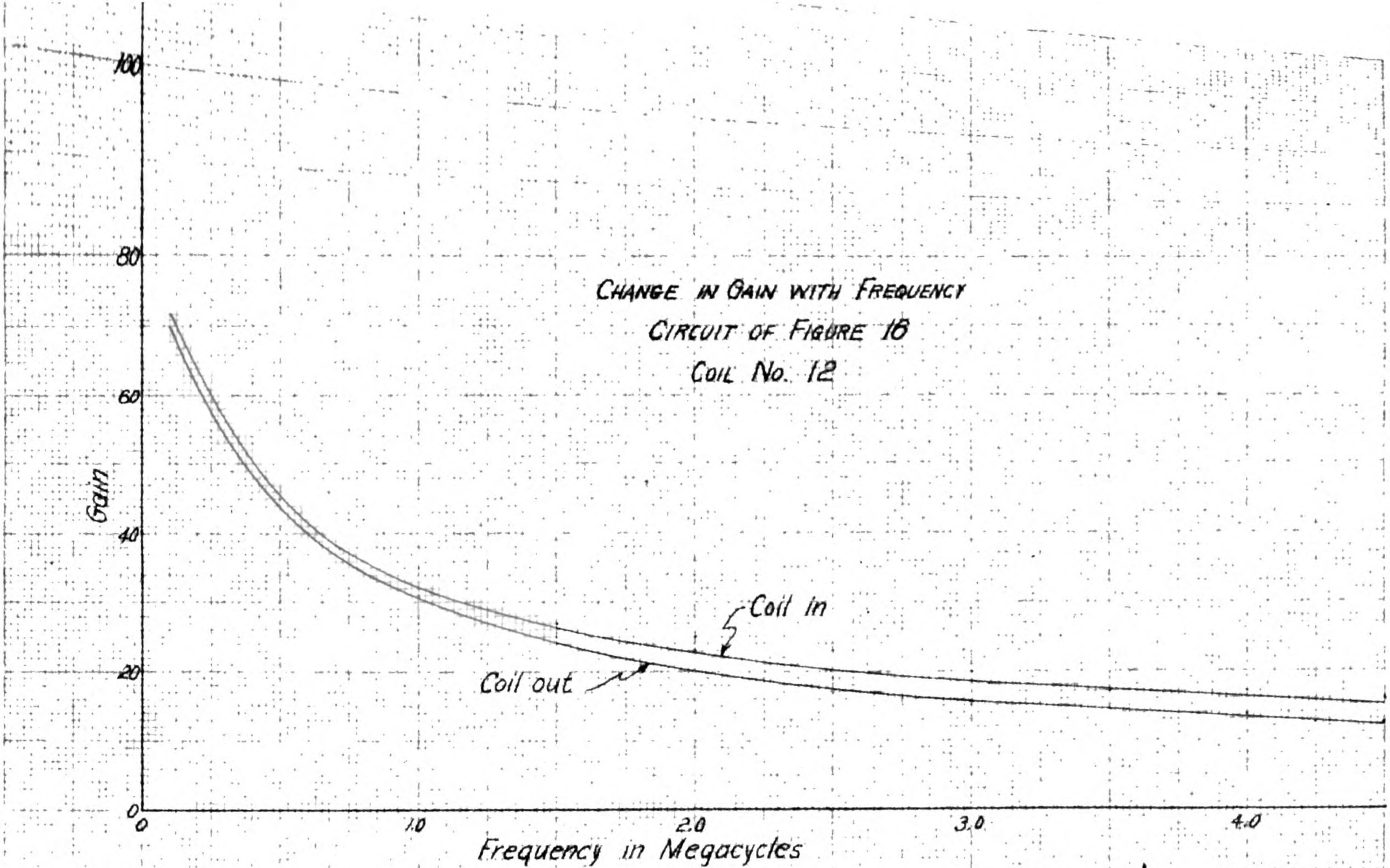

CHANGE IN GAIN WITH FREQUENCY
CIRCUIT OF FIGURE 16
COIL No. 12
Gain
100
80
60
40
20
0
0
1.0
2.0
3.0
4.0
Coil in
Coil out
Frequency in Megacycles

FIGURE 21

C_1=shunt capacitance with coil out = 123 micro-microfarads

C_2=shunt capacitance with coil in

Z_1=shunt impedance with coil out at 4.5 mega-cycles per second

Z_2=shunt impedance with coil in at 4.5 mega-cycles per second

R = 9,950 ohms,

then

$$\frac{C_2}{C_1} = \sqrt{\frac{R^2 - Z_2^2}{R^2 - Z_1^2}}\ \frac{Z_1}{Z_2} \quad . \tag{41}$$

Now from equation (36)

$$Z_1 = \frac{Gain}{g_m} = \frac{12.5}{0.01} = 1250 \text{ ohms}$$

$$Z_2 = \frac{16}{0.01} = 1600 \text{ ohms.}$$

and

$$C_2 = 123\sqrt{\frac{(9.95)^2 - (1.6)^2}{(9.95)^2 - (1.25)^2}}\ \frac{1.25}{1.6} = 95.6 \text{ micromicrofarads}$$

or $C_1 - C_2 = 123 - 95.6 = 27.4$ micromicrofarads.

Thus the negative capacitance acting in this case was in the neighborhood of 25 micromicrofarads. Since equation (12) did not apply in this case where the anti-resonant frequency of the coil was so near 4.5 mega-cycles, the value of the negative capacitance varied according to equation (7) with frequency.

Coil No. 12 had the highest mutual inductance of any of the coils which anti-resonated above 4.5

megacycles per second, so that 27.4 micromicrofarads was the highest value of negative capacitance to be expected from this circuit.

Evidently the only way to increase the gain of this circuit was to decrease the shunt capacitance of the circuit. A new circuit and construction layout was thus adopted as described in the next section.

3. Construction and Operation of the Second Circuit.

The main reason for constructing this second circuit was to decrease the shunt capacitance to ground. This meant that the whole circuit had to be compacted, and in particular the high potential radio frequency leads had to be made as short as possible. With this in mind the circuit of Figure 22 was constructed on a metal chassis.

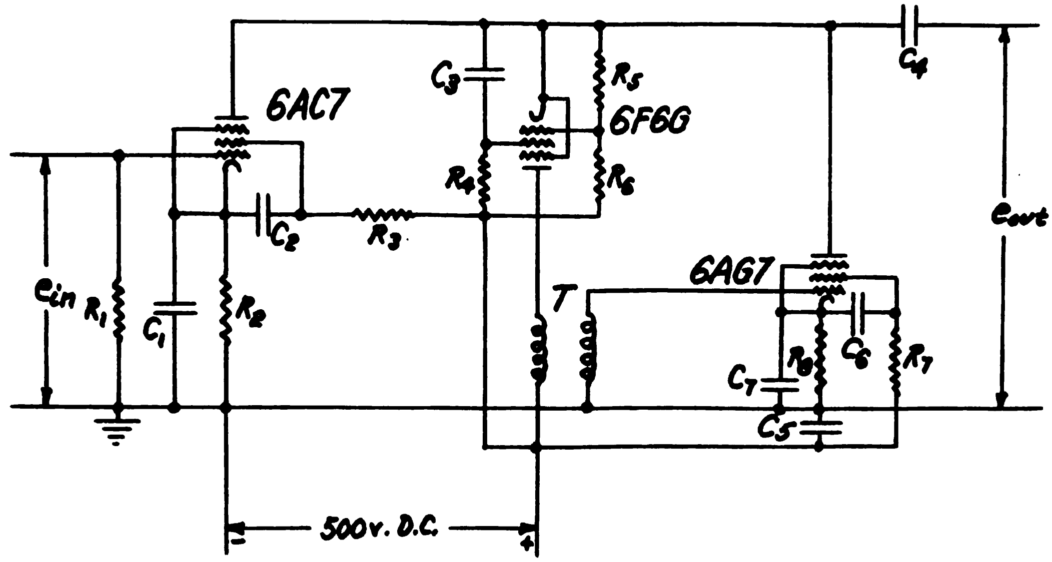


Figure 22

In Figure 22,

R_1=1 megohm

R_2=127 ohms

R_3=203,500 ohms

R_4=50,000 ohms

R_5=100,000 ohms

R_6=65,000 ohms

R_7=42,400 ohms

R_8=48.2 ohms

T as determined from Figure 15.

C_1=C_7=1.0 microfarad, paper

C_2=C_3=C_4=C_5=C_6=0.1 microfarad, paper.

It will be noted that the separate 6AG7 grid resistor used in the circuit of Figure 16 was replaced in the circuit of Figure 22 by the alternating current plate resistance of the 6F6G. In order to do this a new set of operating conditions had to be developed such that the plate resistance of the 6F6G would be approximately 25,000 ohms, while the tube still passed the 43 milliamperes direct plate current of the 6AC7/1852 and 6AG7. Also it was necessary to keep the 6F6G screen resistor as large as possible because the condenser, C_3 , of Figure 22 maintained the screen at the same radio frequency potential as the 6AC7/1852 plate. Experimentally the following operating conditions were found to satisfy the requirements stated above:

Heater voltage	6.3 volts
Plate voltage	250 volts
Grid voltage (control)	9 volts
Screen resistor	50,000 ohms
Plate current	35.5 milliamperes
Grid current	3.6 milliamperes
Screen current	3.9 milliamperes
Plate resistance	23,000 ohms.

To obtain the positive 9 volts on the 6F6G control grid, the voltage divider composed of R_5 and R_6 in Figure 22 was devised. As a special precaution against shunt capacitance, the output lead from C_4 was brought out through a hole in a mica sheet centered in a one inch hole in the metal chassis.

The shunt capacitance of the video amplifier now measured by the substitution method was 42.7 micromicrofarads. Added to this, of course, was the input capacitance of the vacuum tube voltmeter used to measure the output. Thus the total shunt capacitance was 54.7 micromicrofarads as compared to 123 micromicrofarads for the circuit of Figure 16.

No adjustment of the circuit constants was necessary because the direct current voltages were correct. With the 6AG7 grid grounded the gain of the amplifier at 100 kilocycles per second was 109.

When the 6AG7 grid was connected to coil number 12, the circuit went into violent oscillation with an output voltage greater than 16 volts. On a receiver using a beat frequency oscillator, the oscillation was heard as a broad signal at 2.3 megacycles. The low frequency involved and the broadness of the signal indicated that the primary of the radio frequency transformer number 12 with its associated circuit, the 6F6G plate resistance and the shunt capacitance of the amplifier, was anti-resonant.

Various methods were employed to stop the oscillation all of which so altered the circuit parameters that the equations of Part I no longer applied. The same trouble was encountered with coils number 8, 9, 19, and 20.

No oscillation occurred when the 6AG7 grid was connected to coil number 10. Readings of gain versus frequency, with the 6AG7 first connected to ground and then to coil number 10, were recorded in Figure 23.

Frequency (megacycles)	Coil No. 10 out	Coil No 10 in
0.1	105.5	107.5
0.5	45.5	58.5
1.0	24.5	32.5
2.0	13.2	17.0

Figure 23 (concluded on next page)

Frequency (megacycles)	Coil No. 10 out	Coil No. 10 in
3.0	8.5	11.2
4.0	6.4	8.45
4.5	5.6	7.52

Figure 23 (concluded)

The data of Figure 23 is plotted in Figure 24.

From the data of Figure 23 it will be seen that some negative capacitance was acting. Using equations (38) and (36)

$$R = \frac{\frac{Gain}{g_m}}{\sqrt{1-\left(\omega C_3 \frac{Gain}{g_m}\right)^2}} .$$

At 0.1 megacycle per second with $C_3 = 54.7$ micromicrofarads

$R = 12{,}710$ ohms .

Let

$C_1 =$ shunt capacitance with coil out $= 54.7$ micromicrofarads

$C_2 =$ shunt capacitance with coil in

$Z_1 =$ shunt impedance with coil out at 4.5 megacycles per second

$Z_2 =$ shunt impedance with coil in at 4.5 megacycles per second.

From equation (36)

$$Z_1 = \frac{5.6}{0.009} = 622 \text{ ohms}$$

$$Z_2 = \frac{7.52}{0.009} = 835 \text{ ohms} .$$

Now utilizing equation (41)

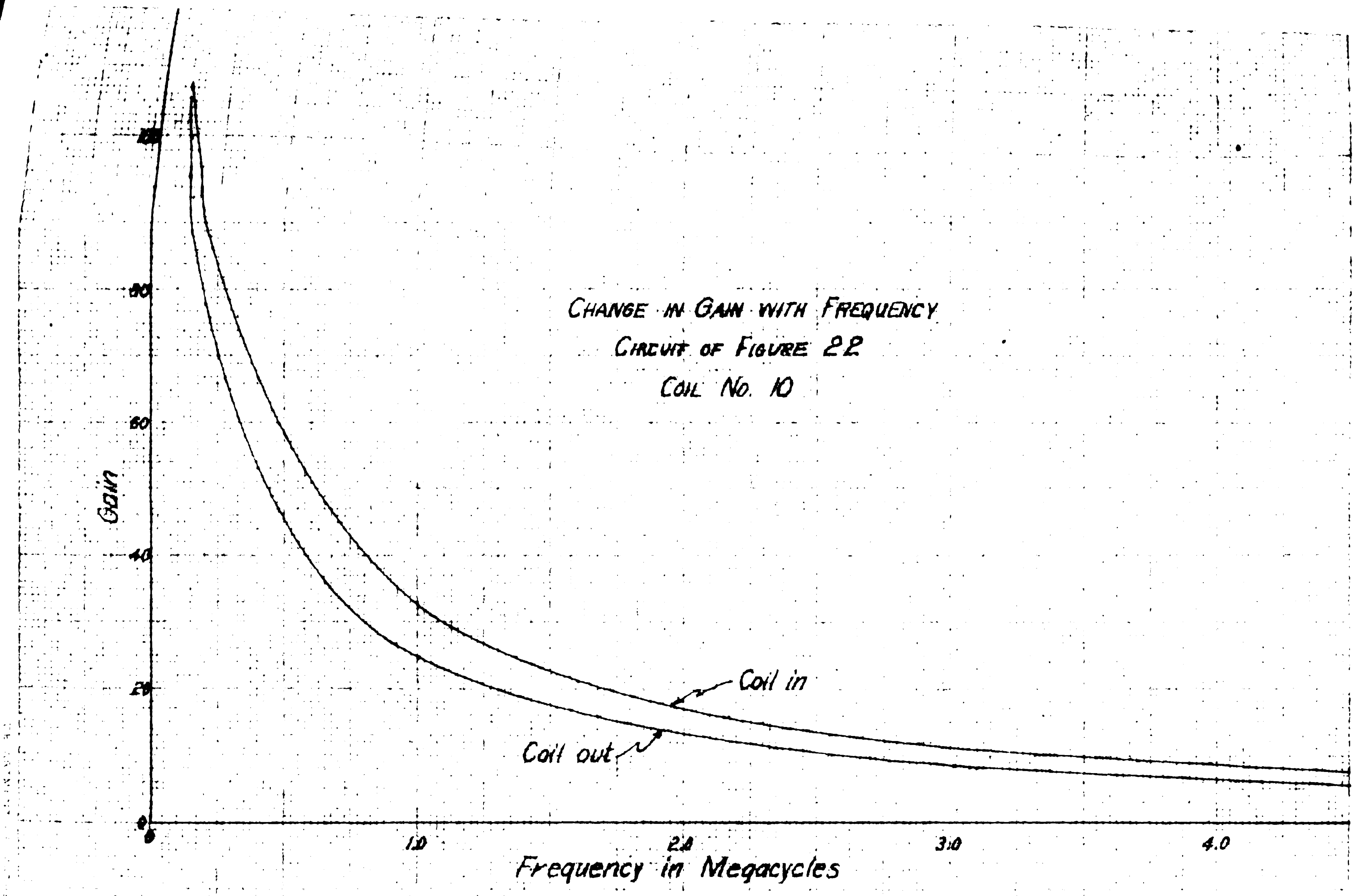

CHANGE IN GAIN WITH FREQUENCY
CIRCUIT OF FIGURE 22
COIL No. 10
Gain
0
20
40
60
80
100
1.0
2.0
3.0
4.0
Frequency in Megacycles
Coil in
Coil out

FIGURE 24

$$C_2 = C_1\sqrt{\frac{R^2 - Z_2^2}{R^2 - Z_1^2}}\ \frac{Z_1}{Z_2} = 54.7\sqrt{\frac{(12.71)^2 - (.835)^2}{(12.71)^2 - (.622)^2}}\ \frac{622}{835} = 40.8$$

micromicrofarads.

or

(42) $C_2 - C_1 = -54.7 + 40.8 = -13.9$ micromicrofarads.

Equation (42) gives the value of the negative capacitance.

Another way to find the negative capacitance is to make use of equation (12) of Part I. Thus

(43) $-C = -\frac{9mM}{R} = -\frac{0.01 \times 29 \times 10^{-6}}{23{,}000} = -12.6$ Micromicrofarads.

These two values of negative capacitance, found by two different methods, check closely.

Figures 25 and 26 are two views of the apparatus used in the experimental work. In Figure 25 the unit at the left is the signal generator, the unit at the right the vacuum tube voltmeter, the amplifier is in the center foreground, and the power supply is in the background. The large glass tube is the 6F6G, the small metal tube at the right is the 6AC7/1852, and the metal tube at the rear is the 6AG7. To the left of the 6AG7 is the coil shield can.

Figure 26 shows a close-up of the coil assembly with the shield can removed. The coil in position here is No. 20.

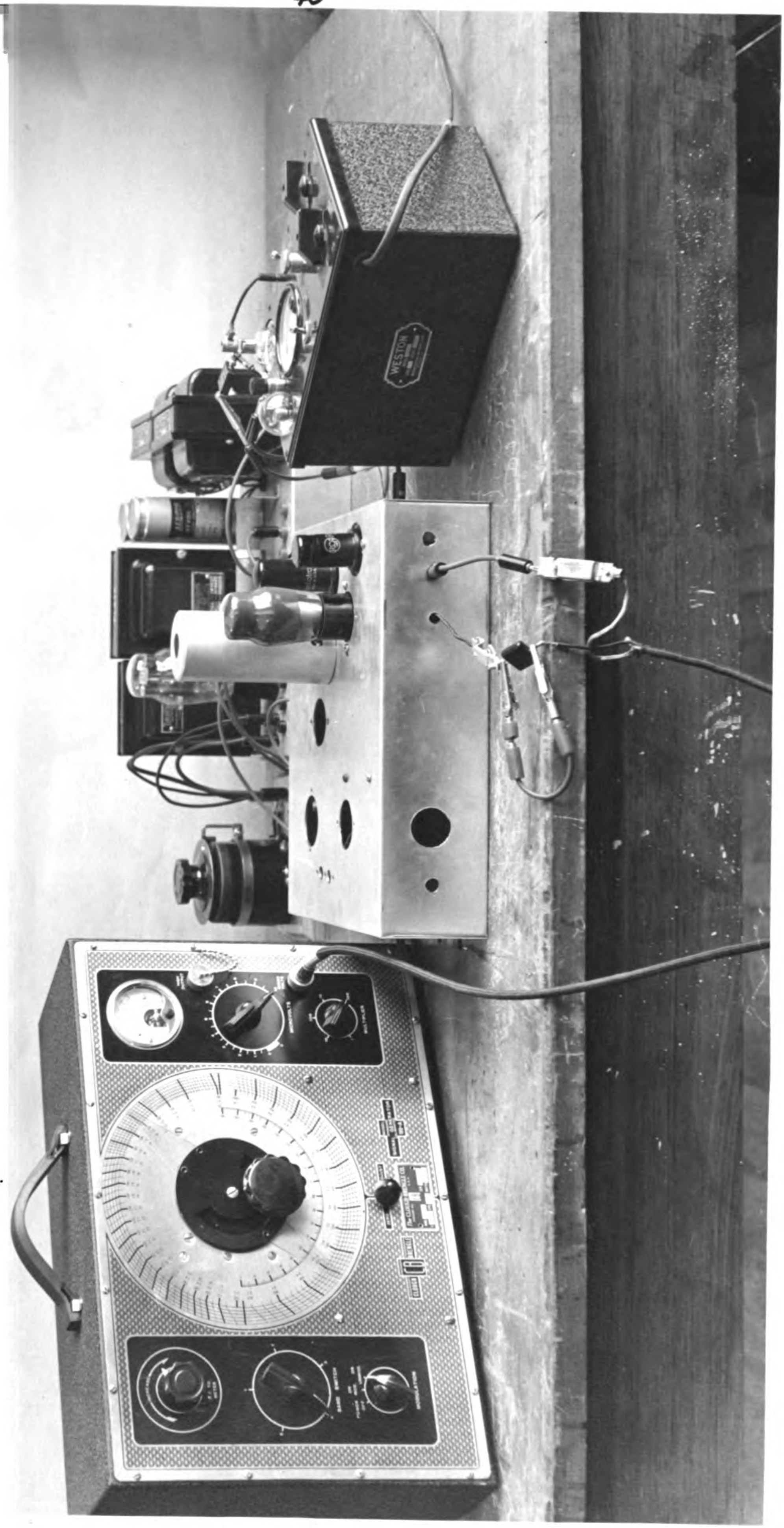


Figure 25

Figure 26

Conclusion

It was found impossible to completely neutralize the positive shunt capacitance of the video amplifier. The concept of negative capacitance was introduced, however, and this negative capacitance was actually used to neutralize a part of the shunt capacitance. The actual value of negative capacitance is given by equation (12) of Part I. This equation shows that, to increase the negative capacitance, the mutual inductance must be increased. To increase the mutual inductance means to increase the primary and secondary inductances which means that both primary and secondary will anti-resonate far below 4.5 megacycles per second.

Thus the main obstruction to the successful operation of the amplifier was the design of coils having a high mutual inductance and still anti-resonating above 4.5 megacycles per second. If the mutual inductance was increased too far, however, oscillation broke out. The 6AG7 tube proved to be a very good oscillator.

If a lower voltage input to the amplifier had been used, a 6AC7/1852 pentode with its lower input capacitance could have been substituted for the 6AG7.

This would have allowed the use of a coil having a greater secondary inductance and thus a greater mutual inductance.

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