AN ABSTRACT OF THE THESIS OF

JOHNSTON, HERBERT RAYMOND ---for the MASTER'S in ELECTRICAL ENG.
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Title--- THE CALCULATION AND MEASUREMENT OF PHASE SHIFT IN ---
-------------- WIDE-BAND AMPLIFIERS --------------------

Abstract Approved: Redacted for privacy
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Phase shift is very important in television amplifiers. Circuits in the amplifier must be arranged so that the phase shift or delay in the amplifier is directly proportional to the frequency, or distortion of the television picture results.

In an uncompensated amplifier the phase shift is positive at low frequencies. Positive phase shift is due to the grid coupling circuits and to the cathode impedance circuits. This phase shift may be reduced or eliminated by a capacitance-resistance network in the plate circuit of the amplifier tube. Generalized design curves for phase shift and gain of low frequency compensation networks are given together with the mathematical developement of the generalized equation from which the curves were calculated.

The measurement of low-frequency phase shift was accomplished by the use of the cathode-ray tube as a null indicator and the use of a negative phase shifting bridge for the measurement of positive phase shift. The positive phase shifting bridge was used to measure negative phase shift at low frequencies. Several curves showing measured against calculated values of phase shift are included.

In an uncompensated amplifier the phase shift is negative at high frequencies. This negative phase shift is due to the stray shunt capacitances in the load circuit of the tube. It may be reduced by several methods. One method involves the use of a small bypass capacitor across the cathode biasing resistor. This method was fully developed in this paper. The generalized equation was derived and several sets of generalized curves are shown.

The measurements of high frequency phase shift were made by means of a cathode-ray tube and a camera. The elliptical trace was measured to determine the phase shift of the amplifier. Several curves showing the results of phase shift measurements are shown.

The use of square waves for the testing of the phase shift and gain characteristics of wide-band amplifiers was explained. The low frequency response of the amplifier was tested by means of a 30-cycle square wave. The distortion resulting from the deficiencies in high frequency response are given.

THE CALCULATION AND MEASUREMENT OF PHASE SHIFT IN WIDE BAND AMPLIFIERS

By

HERBERT RAYMOND JOHNSTON

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TABLE OF CONTENTS

TABLE OF CONTENTS (continued)

ILLUSTRATIONS

Foreword

The need for further investigation of the theoretical and practical problems of the measurement and calculation of phase shift in wide-band amplifiers was brought to light during the design of a special high-gain television amplifier.* A survey of the literature dealing with this subject revealed considerable material but there seemed to be a definite lack of practical information on ways and means of actual measurement of phase shift at both low and high frequencies.

Other deficiencies in methods of calculating and designing compensating circuits to minimize the deficiencies of ordinary capacitance-resistance coupled amplifiers were also revealed. Particularly, the lack of universal design curves for low-frequency compensation circuits was noted. Certain other problems in connection with the practical application of negative feedback seemed to be inadequately treated in the literature.

This thesis was undertaken to solve certain of these problems.

* This amplifier was designed and constructed by Mr. F. A. Everest of the department of Electrical Engineering at Oregon State College and the writer during the summer of 1938.

THE CALCULATION AND MEASUREMENT OF PHASE SHIFT IN WIDE-BAND AMPLIFIERS

By

HERBERT R. JOHNSTON

INTRODUCTION

In all amplifiers, the presence of reactive components in the circuit causes the output voltage to have an angular displacement with respect to the input voltage. This angular difference is commonly known as phase distortion or phase shift.

In audio amplifiers, it is not necessary to have an exact knowledge of the phase distortion. The ear does not detect the phase shift which does occur in any audio amplifier at each end of the frequency amplification range. In recent years it has become necessary to construct amplifiers, for various purposes, with a minimum of phase shift or delay (30). The most important of the applications of amplifiers with a minimum phase shift is in video amplifiers for use in television.

Very severe requirements are placed on amplifiers by present day television standards. Excessive phase shift at high and low frequencies can not be tolerated if a high definition picture is to be obtained. Distortion of the picture results when the amplifiers are not properly designed or adjusted for television signals.

In a compensated amplifier, it is possible to perfectly correct or compensate the circuit out to a certain frequency for either gain or phase shift but this ideal cannot be achieved for both at the same time. However, improving either phase shift or frequency response always tends to improve the other. Generally speaking, if the phase shift is made as perfect as possible, a peak in the frequency response curve will always be present. A compromise is usually made between gain and phase delay in practical amplifiers.

It can be shown (12)(29) that the criterion for the transmission of transients through an amplifier without distortion is that the time delay must be substantially independent of frequency. This may be expressed by the equation

$$d\theta/df = \text{constant}, \qquad (1)$$

where θ = phase delay in degrees, and

f = frequency in cycles per second.

It is found that a maximum variation in phase delay of one microsecond between the low and high frequencies is tolerable over the whole television system. (12) The actual time delay in seconds for any frequency is found from the equation

$$t(\text{seconds}) = \theta/2\pi f, \qquad (2)$$

where θ and f have the same meaning as in equation (1).

The following material will deal, first, with the calculation and measurement of gain and phase shift in amplifiers at low frequencies and then with the high-frequency phase shift and gain considerations.

Because of lack of time and limitations of available apparatus, a large number of measurements at high frequencies were not obtained. Enough work was done to show the limitations of methods which have been proposed (29) and to show the means of applying a straightforward but less accurate method detailed in reference (17).

LOW-FREQUENCY CONSIDERATIONS

DISCUSSION

Television requirements are such that video amplifiers must have very good low-frequency characteristics. The amplifier must pass frequencies from 30 cycles up with substantially no phase shift and constant gain. A deficiency in low-frequency response shows up in a television picture by its effect on the general background illumination. The background will vary in brightness from the top to the bottom and the contrast will be affected adversely.

The phase shift in the low-frequency end of the amplifier characteristic is more important than a perfectly flat response curve. The practice is to allow a small hump or peak at the lowest frequency to be amplified and make the phase shift zero out to this frequency. The requirements of a wide-band amplifiers are such that anything done to improve the low-frequency response must not affect the high-frequency response adversely. It is therefore necessary to keep the high-frequency requirements in mind even when dealing with the low-frequency end of the response curve.

The elements of a wide-band resistance-capacitance coupled amplifier which cause loss of gain and phase shift at low frequencies are the grid capacitor and the grid resistor coupling elements between stages and the cathode biasing resistors with their associated bypass capacitors. Any compensating arrangements must take into account both of these circuits.

The loss of gain and the phase shift at any frequency which occurs in the grid coupling circuit depends on the product of the coupling capacitance in farads times the coupling resistance in ohms. This product is known as the time constant of the circuit. The general expression in complex notation is readily obtained and is given by

$$\text{Response ratio} = \frac{1}{1 - j/wT_g}, \qquad (3)$$

where response ratio = the ratio between the output voltage at low frequency and the output voltage at mid-band frequency,

$w = 2\pi f$, and

$T_g = R_g C_g$ time constant of the grid coupling circuit.

Figures (1) and (2) are the result of plotting wT_g as an independent variable and loss in decibels and phase shift in degrees as dependent variables. From these general curves may be found the decibel loss and the phase shift for any frequency and for any particular value of T_g.

The maximum value of T_g that can be used in a television amplifier is limited by a number of practical considerations. If a large capacitor is used, the

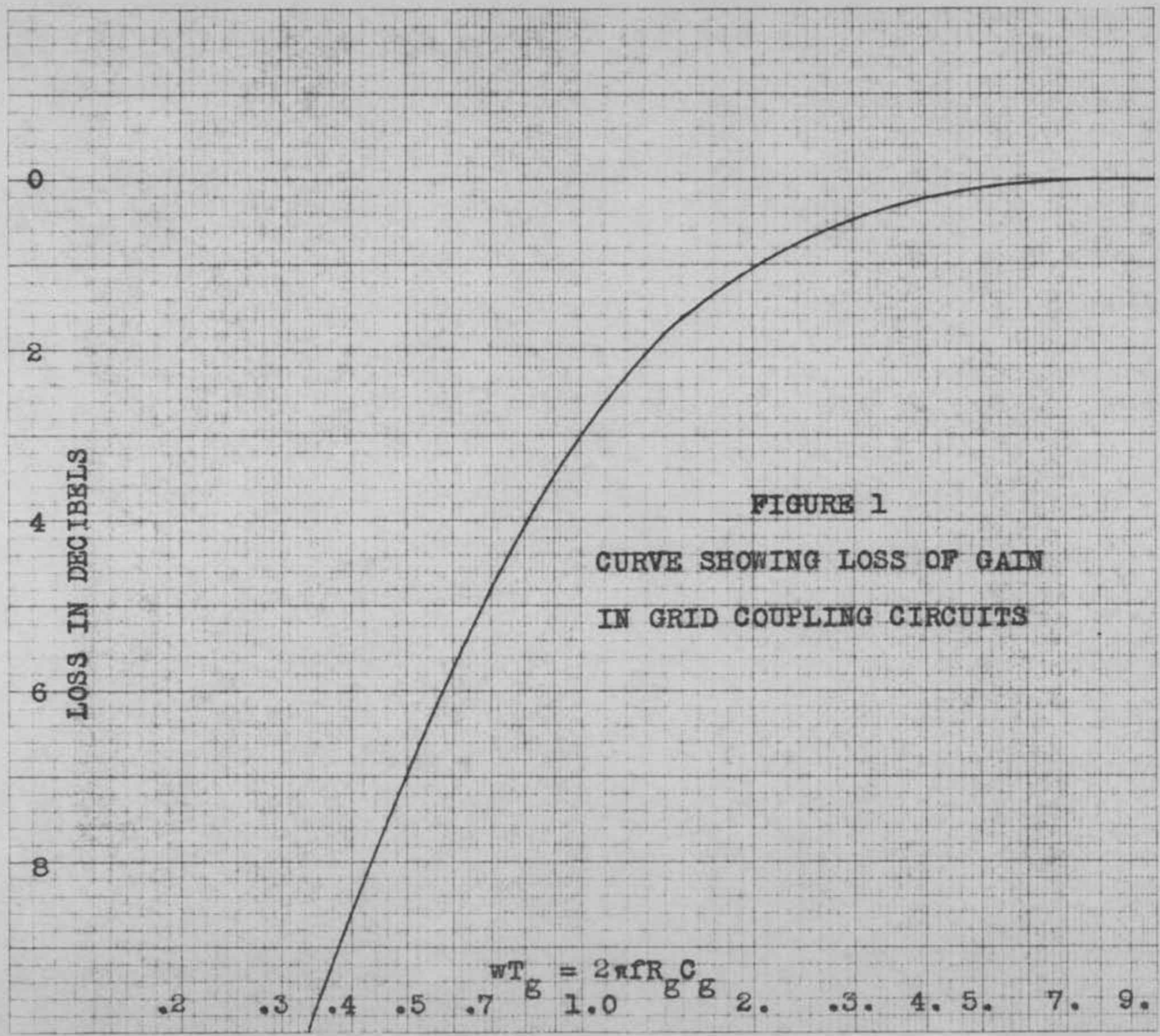


FIGURE 1

CURVE SHOWING LOSS OF GAIN

IN GRID COUPLING CIRCUITS

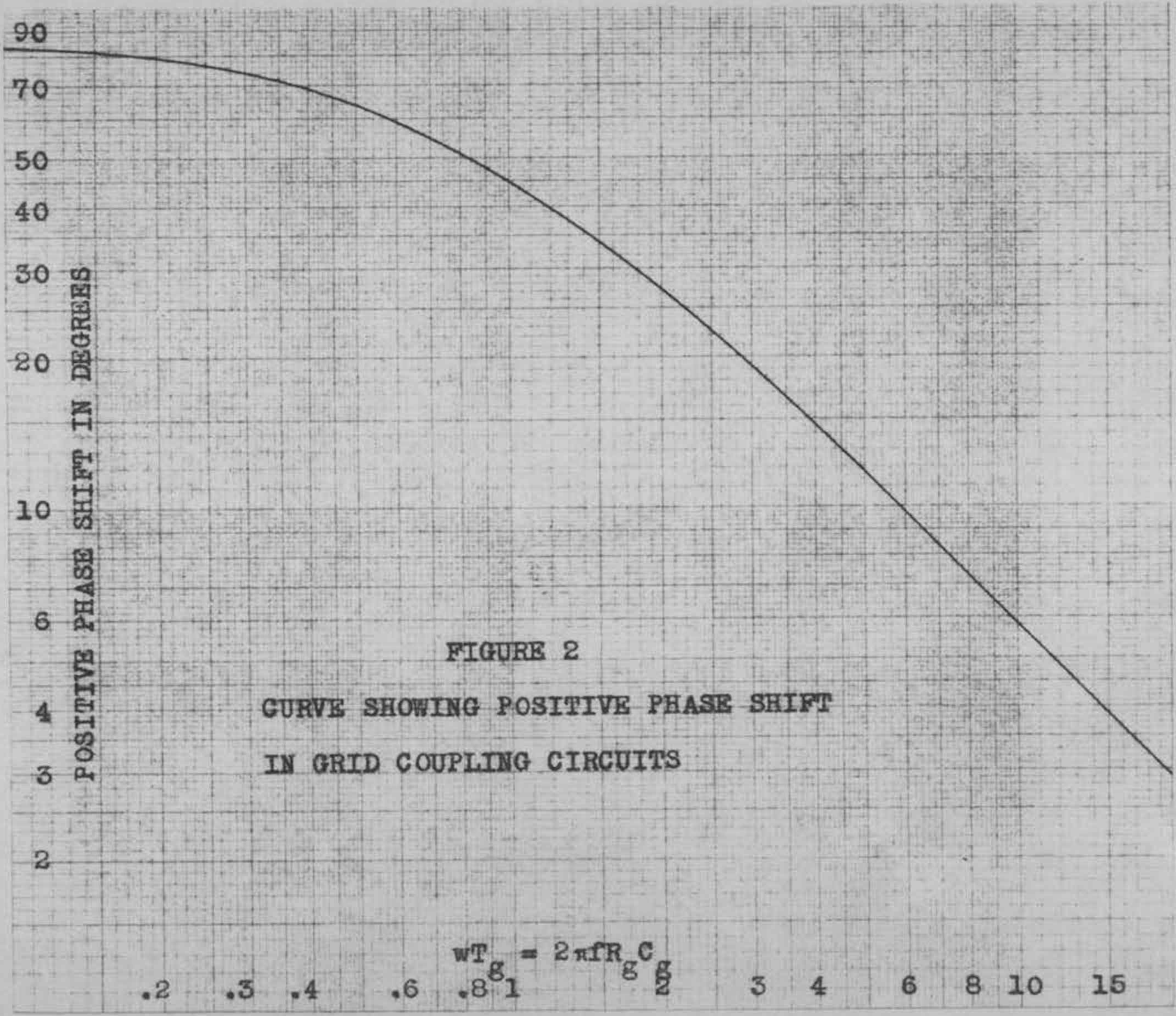


FIGURE 2

CURVE SHOWING POSITIVE PHASE SHIFT

IN GRID COUPLING CIRCUITS

mechanical size will be excessive and it will have an excessively high stray capacitance to ground which will lower the high-frequency response of the amplifier. The maximum value of the grid resistor is limited by the type of tube employed and it depends on a number of design factors of the tube. It is well to have the grid coupling resistance as large as possible, consistant with the tube input characteristics, and use a smaller condenser for the same value of the time constant T_g. The grid resistor is, in effect, in parallel with the plate load resistor of the previous stage.

Two other practical considerations limit the value of the grid circuit time constant T_g. These considerations are the possibility of sustained low-frequency oscillations commonly known as "motor-boating" (11) and the possibility of having the amplifier block (31) and become paralyzed under strong excitation or shock. A large time constant increases the susceptibility of the amplifier to both of these conditions. A good practical limit has been found to be of the order of $T_g = 0.01$.

The effect on frequency response and phase shift of the cathode resistor and bypass capacitor network depends on its time constant or the product of the cathode resistor in ohms times the capacitance of the bypass capacitor in farads. The general equation for the response ratio in complex notation is (15) as follows

$$\text{Response ratio} = \frac{1 + jwT_k}{(1 + M) + jwT_k}, \tag{4}$$

where response ratio = ratio between the output voltage at low frequencies and the output voltage at mid-band frequency,

$T_k = C_k R_k$ = time constant of cathode circuit, and

$M = G_m R_k$ = transconductance of tube times cathode resistance

This equation is derived for high transconductance tubes under the assumption that the plate resistance is very much larger than the plate load resistance of the stage and that the amplification factor of the tube is very much larger than one.

Figures (3) and (4) give the loss in decibels and the phase shift in degrees plotted against wT_k as the independent variable. The parameter M is used to indicate the product of the transconductance of the tube in ohms times the cathode biasing resistance in ohms.

The value of R_k depends on the tube and must be the correct amount to furnish the necessary grid bias to the tube in the stage. The capacitor C_k may be the largest available considering the mechanical size and the space

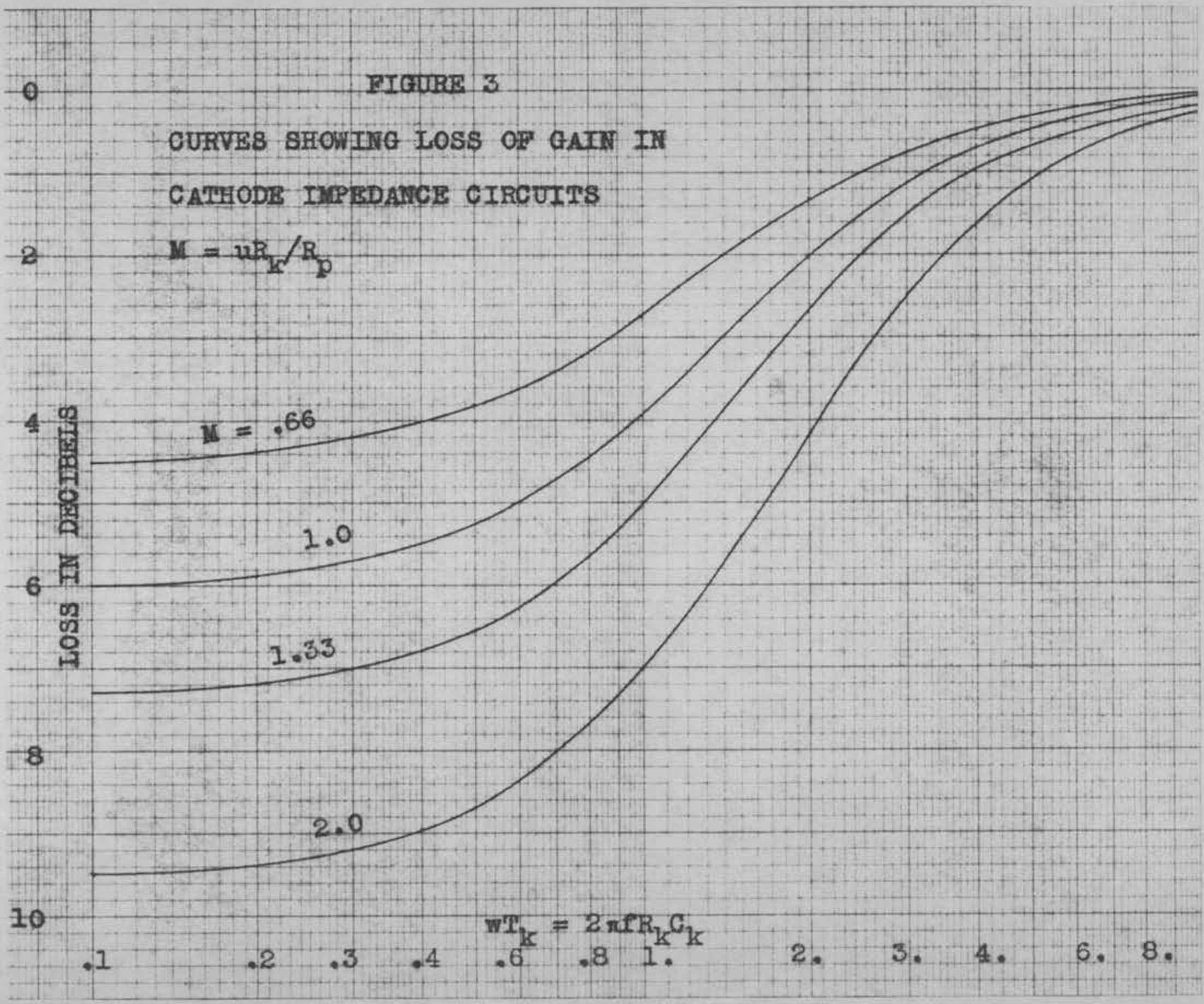


FIGURE 3

CURVES SHOWING LOSS OF GAIN IN CATHODE IMPEDANCE CIRCUITS

$M = uR_k/R_p$

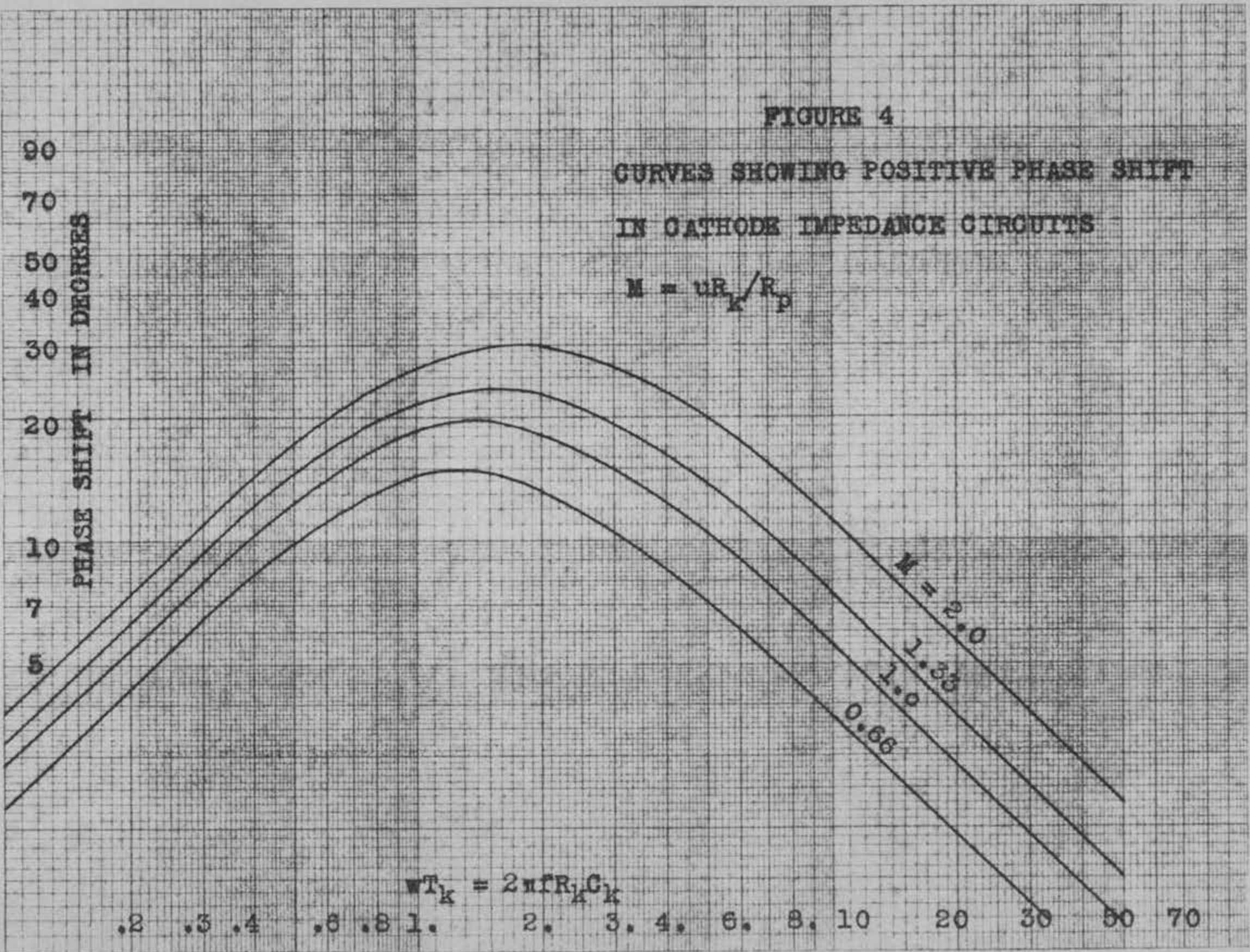


FIGURE 4

CURVES SHOWING POSITIVE PHASE SHIFT IN CATHODE IMPEDANCE CIRCUITS

$M = uR_k/R_p$

required. For high transconductance tubes such as the type 1851, 1852, and 1853 this condenser should be at least several hundred microfarads to minimize phase distortion.

LOW-FREQUENCY PHASE SHIFT AND GAIN COMPENSATION CIRCUITS.

Two methods of low-frequency phase shift and gain compensation make use of capacitance-resistance networks in the plate and in the grid circuit. (15) The most widely used and the most practical method is the use of the plate circuit compensation network. The latter is the only one treated in this paper. See figure (5).

The plate circuit low-frequency compensation network consists of an additional plate load resistor which is shunted by a relatively large capacitor. As the frequency is reduced, the reactance of the capacitor becomes larger and the plate load impedance increases and tends to raise the gain of the stage at low frequencies. The phase shift and gain characteristics of this network should be exactly opposite to the characteristics of the rest of the circuit for exact compensation.

The method of attack of the problem of the proper design of the low-frequency network seems to be largely that of the trial and error method. O. E. Keal gives in reference (15) a generalized equation and a method of calculation which, it appears, leaves much to be desired in the way of simplicity and of direct approach to the problem.

The desire for a simple approach to the problem of

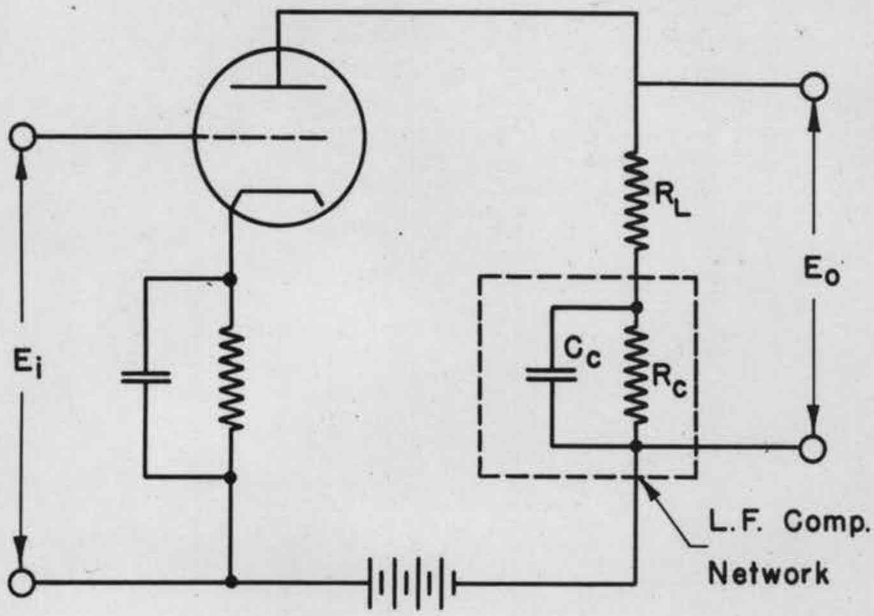


Figure 5

CIRCUIT DIAGRAM SHOWING PLATE CIRCUIT

LOW FREQUENCY COMPENSATION NETWORK

designing low-frequency compensation networks led the writer to derive a general parametric equation for the gain and phase shift characteristics of plate circuit compensation networks. This equation was derived for use in television amplifiers since the assumption is used that the magnitude of the plate load resistor is negligible in comparison with the dynamic plate resistance of the tube. This assumption is entirely valid for high transconductance pentodes that are used in wide-band amplifiers.

The equation for the response ratio of the low-frequency networks is derived in Appendix (1) and is as follows

$$\text{Response ratio} = \frac{(n + 1) + jwT_c}{1 + jwT_c}, \qquad (5)$$

where response ratio = ratio between the output voltage at low frequency and output voltage at mid-band frequency,

$T_c = C_cR_c$ = time constant of low-frequency compensation network, and

$n = R_c/R_L$ = ratio between resistance in network to plate load resistance.

Figures (6) and (7) give the results of plotting equation (5) using wT_c as an independent variable and decibels gain and phase shift in degrees as dependent variables. Various values of n are used which cover the

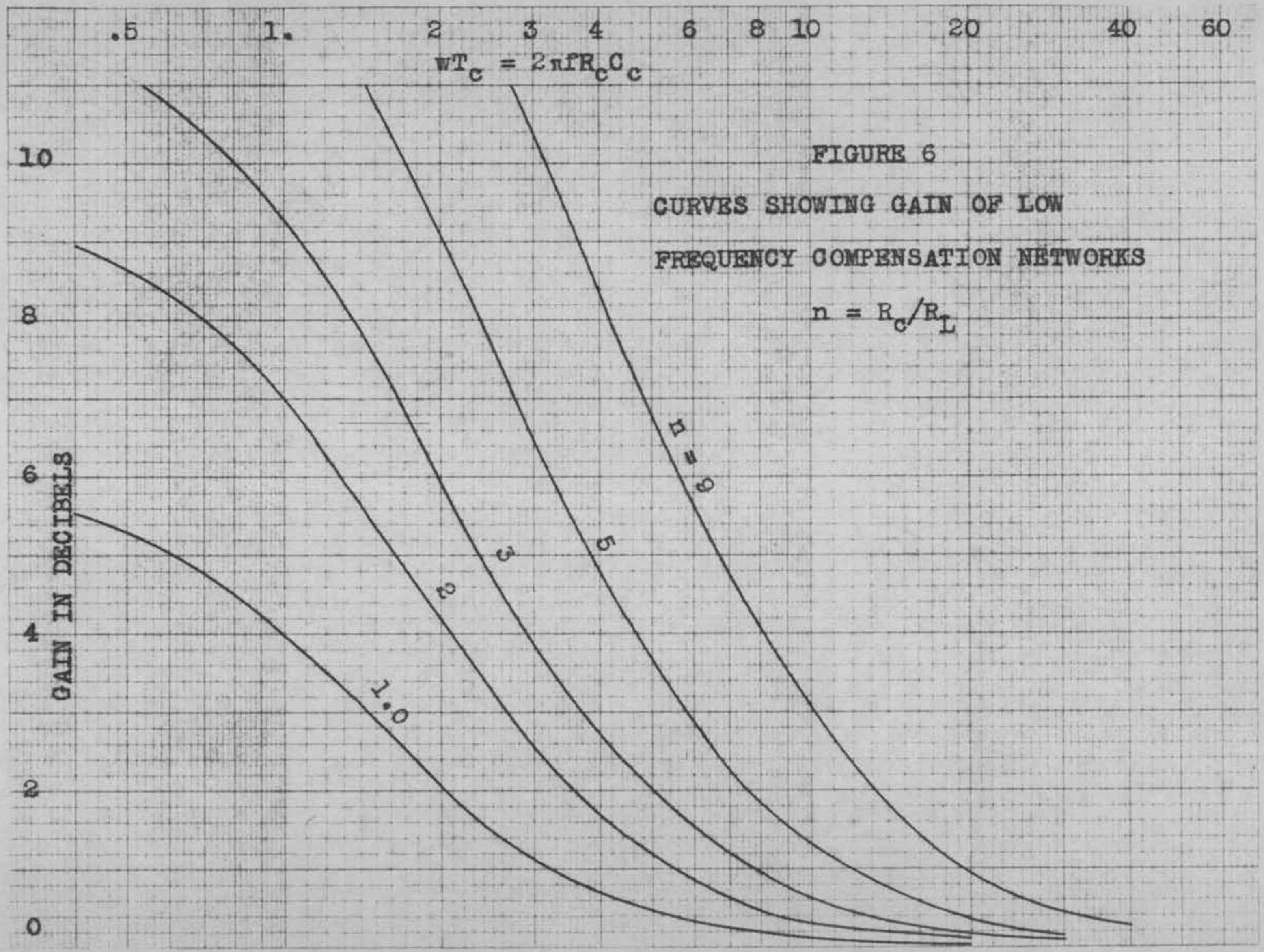


FIGURE 6

CURVES SHOWING GAIN OF LOW FREQUENCY COMPENSATION NETWORKS

$n = R_c/R_L$

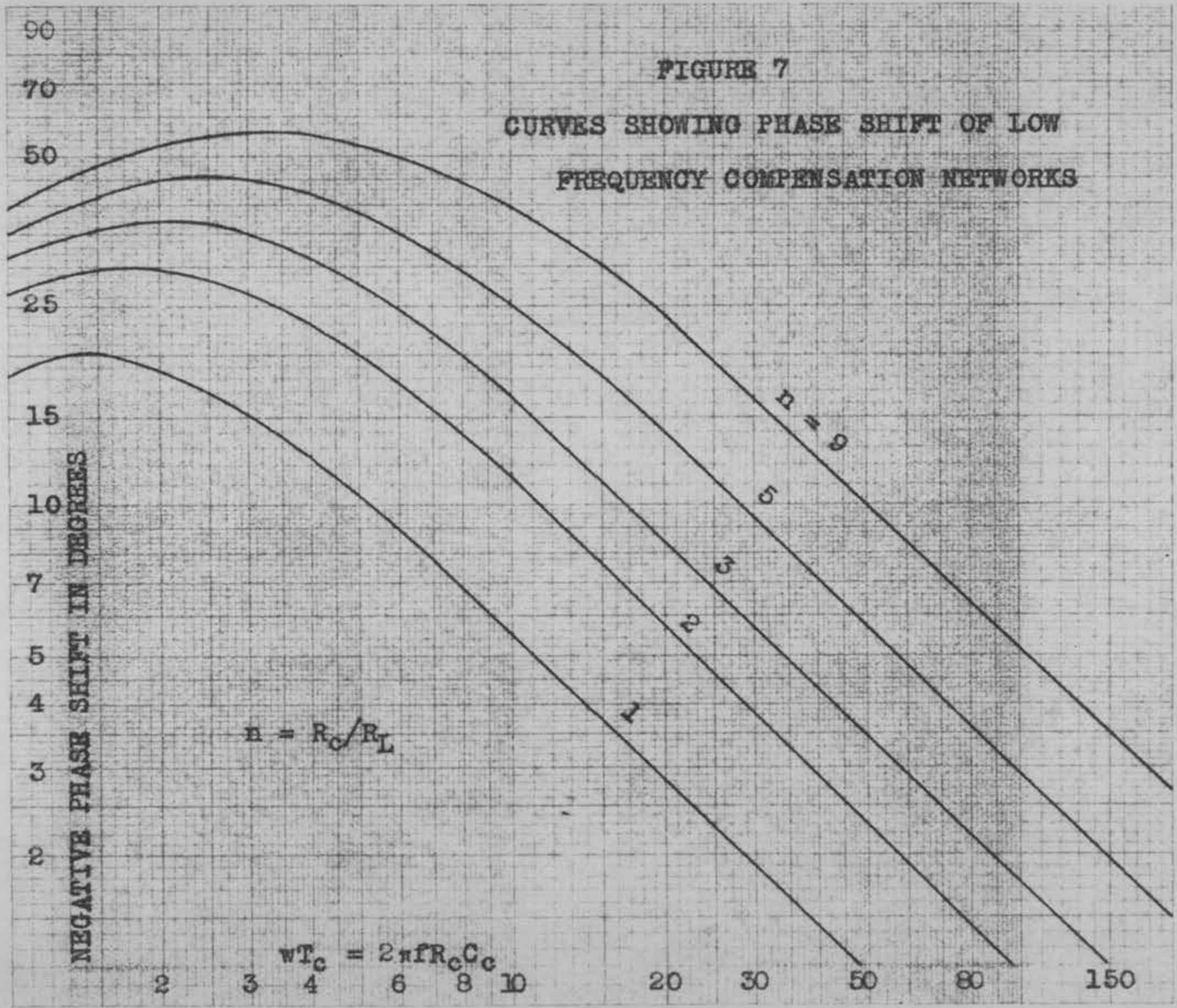


FIGURE 7

CURVES SHOWING PHASE SHIFT OF LOW FREQUENCY COMPENSATION NETWORKS useful range of design factors. For other values of n not given on these curves, it is possible to interpolate between the curves or calculate a new curve from the general equation.

These curves may be used to determine quickly the phase shift in degrees and the gain in decibels for any combination of plate load resistance and compensation network resistance and capacitance, within the limits imposed by the assumptions used in the derivation of the equation.

The procedure for the design of a low-frequency compensation network using the general curves is as follows. First it is necessary to assume the time constant for the grid coupling circuit and assume a reasonably large time constant for the cathode circuit. This latter should be of the order of 0.05 or better. Then by the use of the curves of figures (1), (2), (3), and (4), the phase and gain characteristics of the entire stage is plotted on the same scale transluscent coordinate paper as is used for figures (6) and (7). This may be done easily since the loss in decibels of the two circuits may be added directly at any one frequency. The phase shift in degrees may also be added directly at any one frequency. The curve showing the loss of the assumed stage should be plotted with the ordinates reversed so that the curve may be superimposed upon the gain curves of fugure (6). In these curves, the independent variable should be frequency in cycles per second.

If the best possible phase-shift characteristics are desired, the curve for the amplifier which shows the variation of phase shift with frequency should be superimposed upon figure (7). The curve is shifted from left to right or from right to left until one of the parametric curves coincides with the phase shift curve of the amplifier. The parameter n that should be used is the one which shows no more than 10 degrees phase shift at the lowest frequency of importance to be amplified. If the phase shift is made zero up to the lowest frequency of importance, the final gain curve will show a considerable peak or hump and the gain will be sustained to very low frequencies. The value of n should be as small as consistant with the desired phase-shift characteristics.

With the curve showing the phase characteristics of the amplifier superimposed upon figure (7), the value of wT_c on figure (7) should be noted under any convenient frequency say f cycles per second. The value of T_c is then found from the relation

$$T_c = wT_c/2\pi f. \qquad (6)$$

The value of R_L is fixed by high frequency considerations and thus R_c is found by the relation given below,

$$R_c = nR_L. \qquad (7)$$

The capacitance is now found from the relation

$$C_c = T_c/R_c. \qquad (8)$$

Finally, the complete gain and phase shift characteristics of the compensated stage are plotted against frequency. If the gain is too high at low frequencies, a smaller value of the time constant in the grid circuit T_g must be assumed and the work repeated until a satisfactory response curve is obtained.

It should be emphasized again that a compromise between phase and gain characteristics must be reached in the design of the low-frequency network. It is impossible to have a perfectly flat gain characteristic with a perfect phase shift characteristic.

It is believed that the use of these curves greatly simplifies this design procedure over other methods.

THE MEASUREMENT OF LOW-FREQUENCY PHASE SHIFT IN AMPLIFIERS

METHODS

Under certain conditions it is difficult or impossible to calculate accurately the phase shift of a multistage amplifier. In order that this important characteristic may be determined, it is necessary to devise some method of measuring the phase shift with a fair degree of accuracy.

There are a number of methods which can be used for the measurement of phase shift (40). One of these methods is described in reference (17). In this reference, the author makes use of a cathode-ray tube for the purpose of measuring the Lissajous figures produced by the phase shift in the amplifier. This method was used by the present writer for the measurement of phase shift at high frequencies and it will be explained in detail under that heading.

The method of low-frequency phase shift measurement used in this investigation is one which makes use of the cathode-ray tube as a null indicator. The figure on the fluorescent screen of the tube is used to indicate the point of zero phase shift between the voltages applied to the horizontal and vertical deflection plates of the cathode-ray tube.

It can be shown that when two sinusoidal voltages of the same frequency are applied to the horizontal and vertical deflection plates of a cathode-ray tube, (27) a Lissajous figure will be formed whose exact shape will depend on the phase difference between the two voltages. The figure will be a circle if the magnitudes of the two voltages are the same and they are 90 degrees out of phase. When the figure is an ellipse, the phase shift is less than 90 degrees but more than zero degrees. When the figure is a straight line, the two voltages are in phase or 180 degrees out of phase. (27)

Figure (8) shows, in outline, the general method used for the measurement of phase shift at low frequencies. If the phase shift of the amplifier is positive, enough negative phase shift is introduced into the circuit by means of the phase-shifting bridge to reduce the total phase shift in the circuit between the two sets of deflection plates to zero. This condition of zero phase shift is indicated by a straight line on the florescent screen of the cathode-ray tube. This line will be inclined at an angle to the horizontal, depending on the relative magnitudes of the two voltages.

This method provides an easy and fairly accurate means of phase shift measurement. The accuracy depends a great deal on the skill of the operator in finding the exact null point and also on the focus of the beam on the

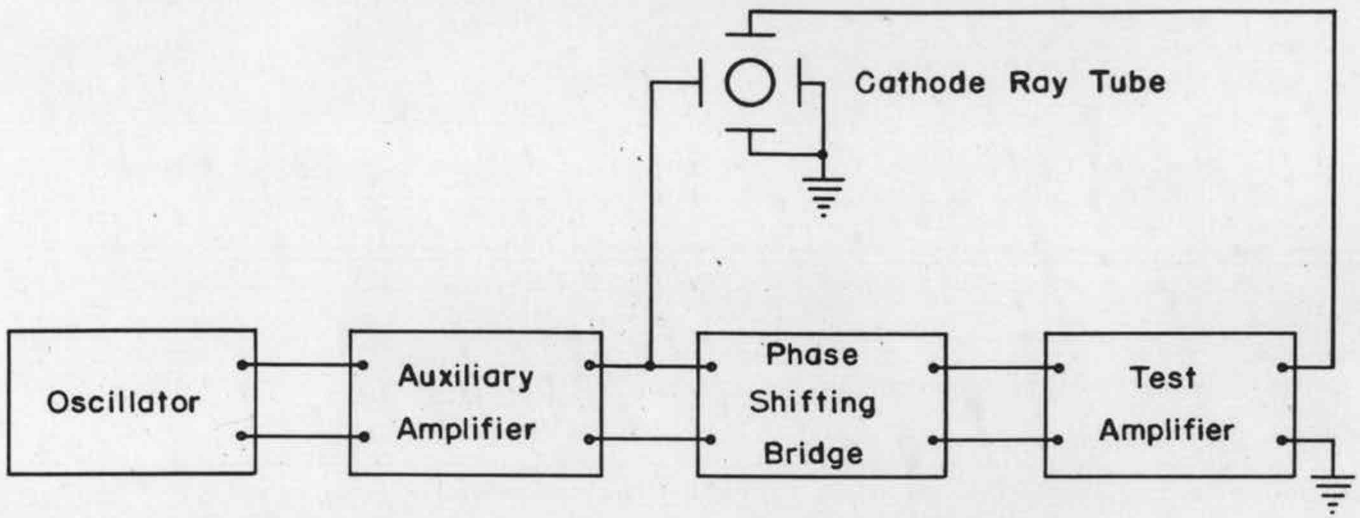


Figure 8.

BLOCK DIAGRAM FOR GENERAL
METHOD OF MEASURING PHASE
SHIFT AT LOW FREQUENCY

fluorescent screen. Figure (9) shows a series of photographs of the trace formed on the end of the cathode-ray tube. The top trace (a) shows an unbalanced condition and indicates a definite phase shift; trace (b) shows a partly balanced bridge and trace (c) is a straight line and indicates zero phase shift with the bridge perfectly balanced.

In the uncompensated or properly compensated amplifier, there should not be any negative phase shift at low frequencies. In order that positive phase shift may be measured, it is necessary to use a bridge which has a negative phase shift. Thus when the negative shift of the bridge is equal and opposite to the positive phase shift of the amplifier, the resultant phase shift is zero. The trace on the cathode-ray tube is then a straight line.

Figure (10) illustrates the circuit diagram of the negative phase-shifting bridge used in this investigation. This bridge also serves as a voltage divider to attenuate the output of the auxiliary amplifier which must be of considerable magnitude since it is directly connected to the horizontal deflection plates of the cathode-ray tube. The optimum conditions are obtained when the two voltages impressed on the two sets of deflection plates of the oscilloscope are equal in magnitude. Thus it is advisable to have the attenuation of the phase-shifting bridge

(a)

(b)

(c)

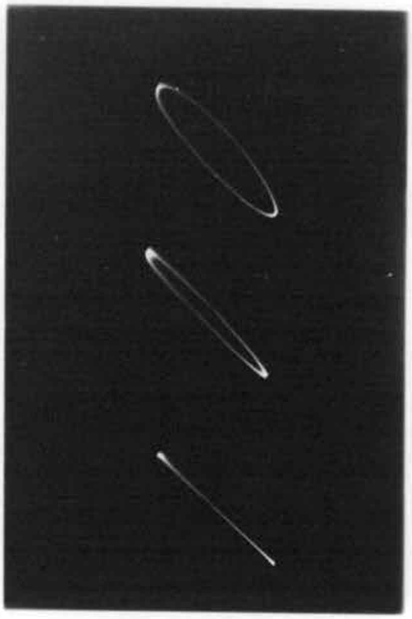


FIGURE 9

(a) CATHODE RAY TUBE TRACE SHOWING A CERTAIN PHASE SHIFT BRIDGE UNBALANCED

(b) TRACE SHOWING BRIDGE PARTLY BALANCED

(c) PERFECTLY BALANCED BRIDGE SHOWING ZERO PHASE SHIFT

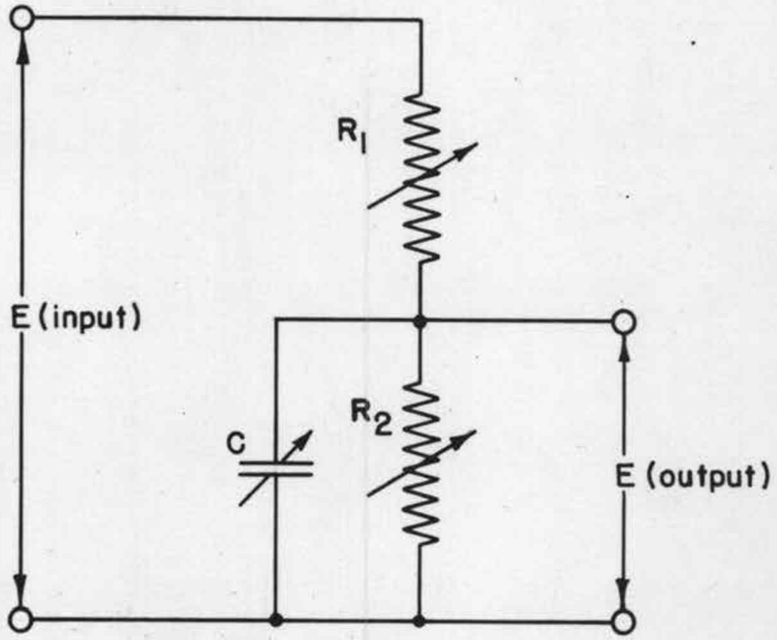


Figure 10

PARALLEL TYPE NEGATIVE PHASE SHIFTING BRIDGE AND ATTENUATOR equal to the amplification factor of the amplifier to be tested. This may easily be done by varying the two resistances R_1 and R_2 and adjusting the phase by means of the capacitor C. See figure (10).

The equation showing the phase shift of the negative phase-shifting bridge is derived in Appendix (3) and is

$$\theta = (-) \arctan wCR_1R_2/(R_1 + R_2), \qquad (9)$$

where $w = 2\pi f$.

If the amplifier to be tested has an amplification factor such that R_1 must be much greater than R_2, then the above equation may be written as follows

$$\theta = (-) \arctan wCR_2. \qquad (10)$$

In case equation (10) is applicable to the problem in hand, the capacitance C must be of considerable magnitude. The resistances R_1 and R_2 must be standard resistances whose values are accurately known and they must have negligible inductance.

Figure (11) is a photograph of the parallel phase-shifting bridge used in these investigations. Use was made of a 1.0 microfarad standard mica decade capacitor in parallel with a bank of paper capicitors with a switch bank for varying the capacitance between 1.0 microfarad and 32 microfarads. This bank of capacitors was first

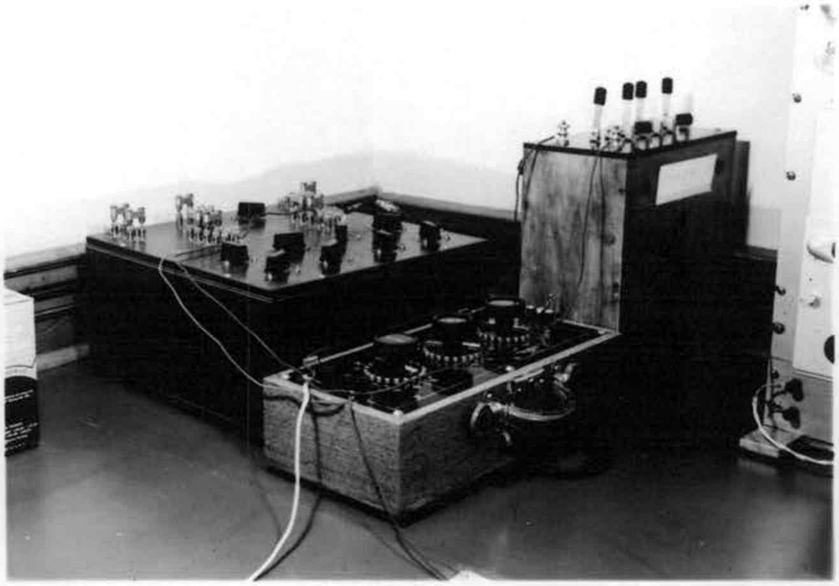


FIGURE 11

PHOTOGRAPH SHOWING PARALLEL TYPE
NEGATIVE PHASE SHIFTING BRIDGE

accurately calibrated by means of a sensitive capacitance bridge. The resistances were built into a universal bridge as shown in the photograph.

If it is desirable to measure negative phase shift in an amplifier, it is necessary to use a positive phase-shifting bridge (29). A series type bridge is shown in figure (12). If it is desirable to make R a fixed resistance, as it was in this investigation, the relative positions of the phase-shifting bridge and the amplifier may be interchanged as in figure (13). The input to the test amplifier must be through a resistance voltage divider as shown in figure (13).

It may easily be shown that the phase shift in degrees of the bridge shown in figure (12) is given by the equation

$$\theta = (+) \arctan 1/wRC \, , \qquad (11)$$

where $w = 2\pi f$, R is in ohms, and C in farads.

Certain precautions must be taken in the measurement of phase shift if accuracy is to be obtained. In the first place, the input resistance of the test amplifier must be much higher than the resistance R_2 (see figure (10)) of the negative phase-shifting bridge. The wave form of the low-frequency oscillator must be free from harmonics and

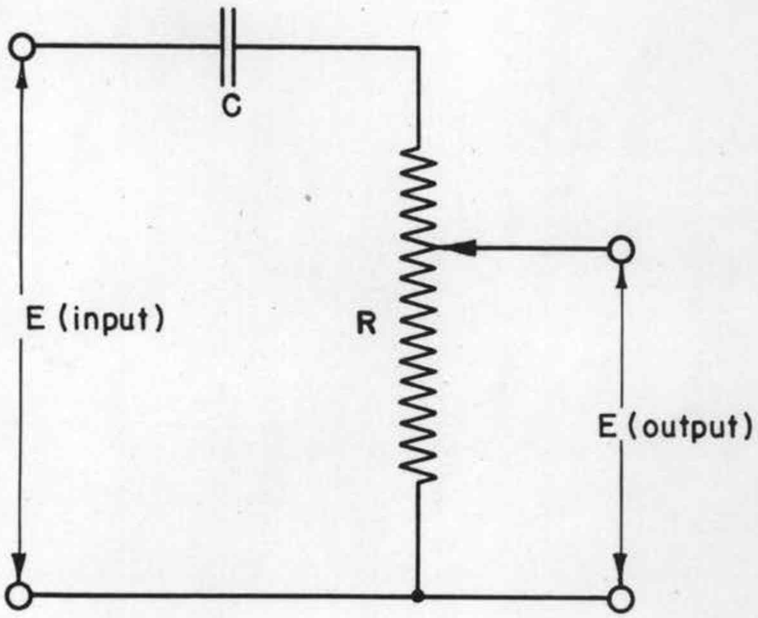


Figure 12.

SERIES TYPE POSITIVE PHASE SHIFTING BRIDGE

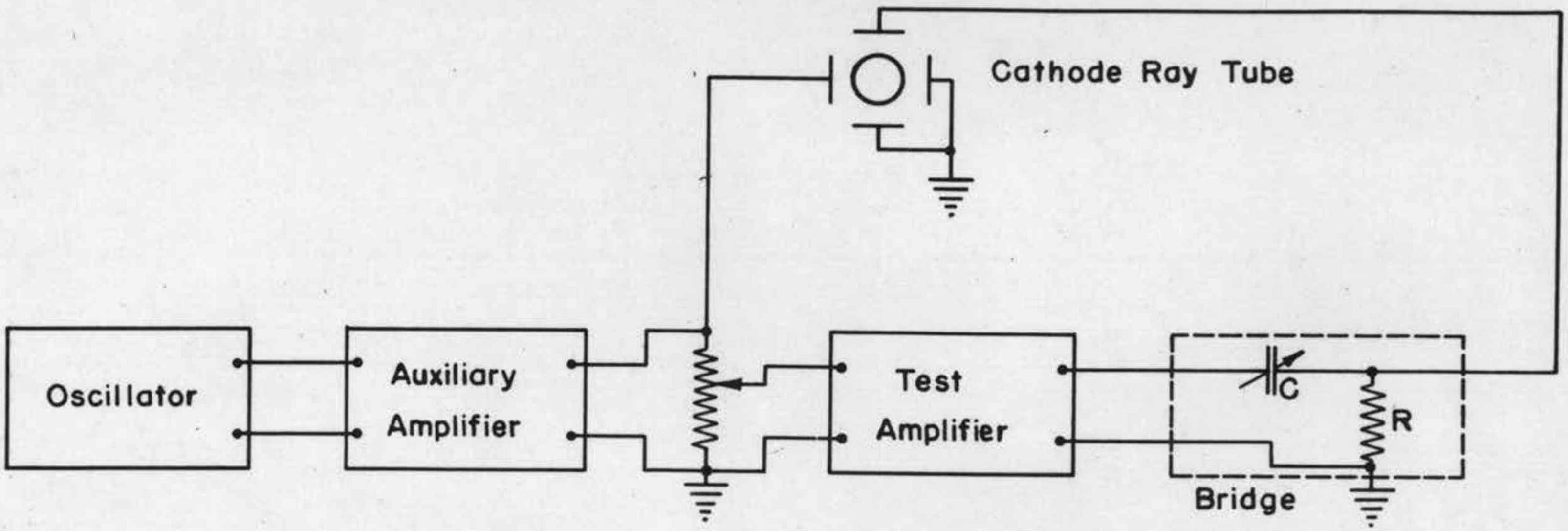


Figure 13.

CIRCUIT DIAGRAM FOR THE MEASUREMENT
OF NEGATIVE PHASE SHIFT AT
LOW FREQUENCIES

the auxiliary amplifier must be free from harmonic distortion. The test amplifier itself must be operated at a point where its harmonic output is a minimum.

This might necessitate the use of another amplifier between the test amplifier and the vertical deflection plates of the cathode-ray tube. In this event, it is necessary to calibrate this second amplifier with the test amplifier removed from the circuit. The actual value for the phase shift of the test amplifier is then obtained by subtracting the phase shift of the second auxiliary amplifier from the measured values with the test amplifier in the circuit.

This precaution against harmonics is necessary if a straight line is to be obtained on the oscilloscope screen when the phase-shifting bridge is balanced. If harmonics are present, it is impossible to balance the bridge accurately. A typical figure obtained on the screen if a second harmonic is present is a small figure eight instead of a straight line. In case the presence of harmonics is unavoidable, it is necessary to cause the middle of the figure to close and ignore the ends of the line.

Figure (14) shows a close view of the test amplifier used in the low-frequency measurements and also a general view of the laboratory apparatus. The three-inch RCA Model TMV 122-B cathode-ray oscilloscope shown in the

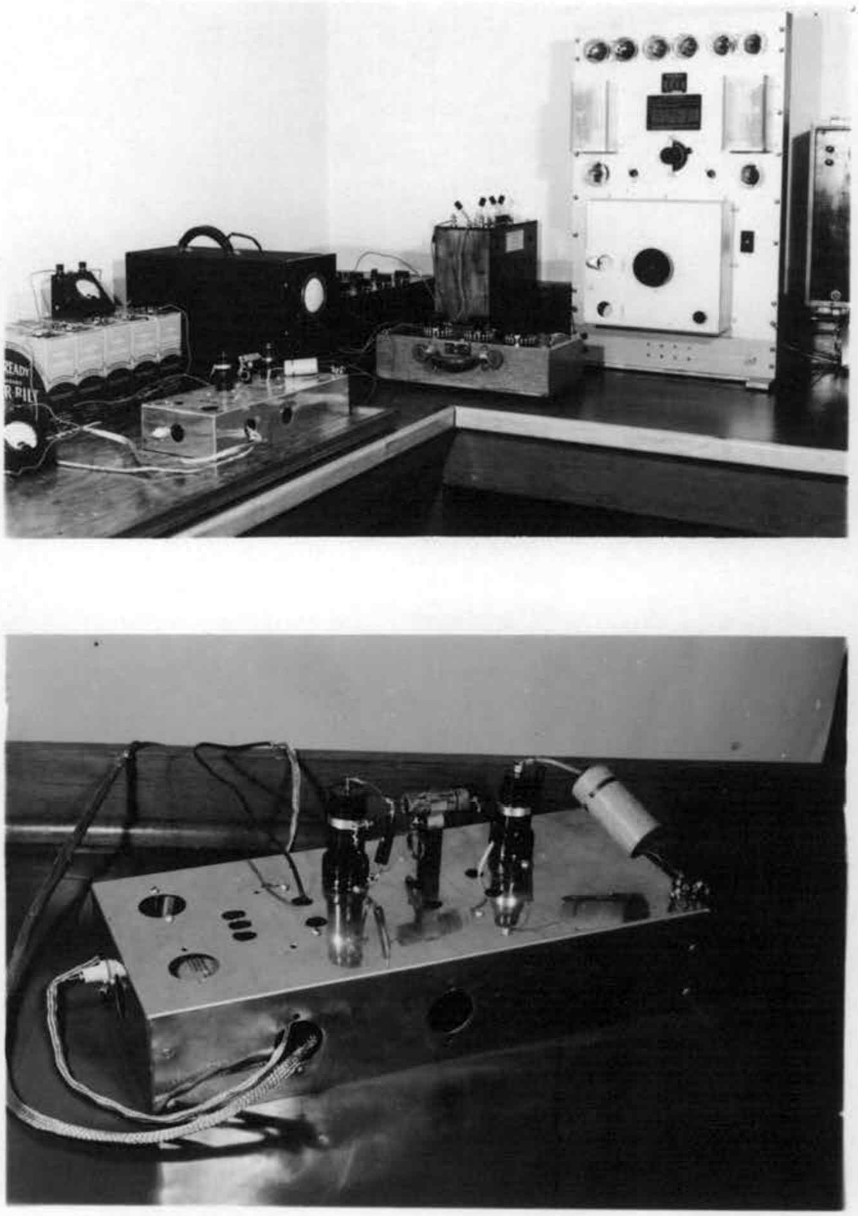


FIGURE 14

PHOTOGRAPH OF LABORATORY APPARATUS AND TEST AMPLIFIER USED IN THE LOW FREQUENCY PHASE SHIFT MEASUREMENTS photograph is slightly modified to allow direct access to the ungrounded vertical and horizontal deflection plates.

Figure (15) shows a complete circuit diagram for the test amplifier shown in figure (14). The first tube is a type 6J7, a metal shell pentode, which it was found necessary to use when taking any measurements with one of the new high transconductance pentodes. The type 1851, which is the second tube, is a recent addition to the vacuum-tube family and it is made especially for use in television amplifiers. It has a transconductance of 9000 micromhos which is over seven times that of the ordinary pentode receiving tube in common use in radio sets today. Thus, the type 1851 gives seven times the gain of an ordinary pentode tube using the same plate load resistance. This enables the plate load resistance to be reduced which results in the extension of the high-frequency gain characteristics of the amplifier. The 1851 grid circuit must work out of a circuit of a few thousand ohms impedance or instability and oscillations will result due to its high sensitivity. In this amplifier, the impedance in the plate circuit of first tube is 3000 ohms which provides the necessary equivalent loading effect on the grid of the type 1851 tube.

It was found that it was necessary to provide an additional ground connection to the metal shell of the two

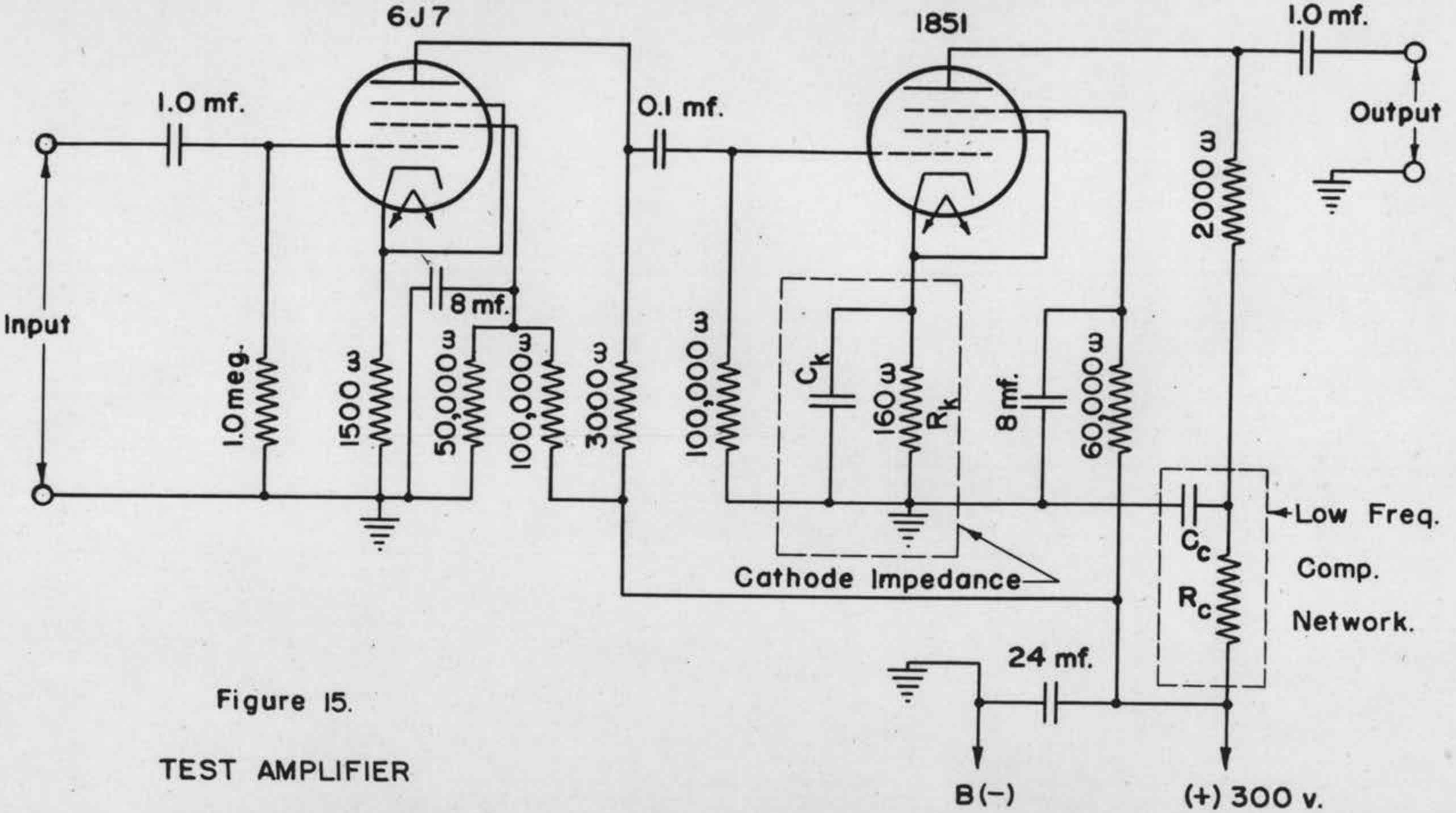


Figure 15.

TEST AMPLIFIER

CIRCUIT DIAGRAM

tubes used in the amplifier. The paint was sandpapered away from a narrow strip around the metal shells and bands of metal fastened about these unpainted strips. These bands were then connected to the nearest point of the chassis.

Provisions were made under the chassis for quickly changing values of the low-frequency compensation network in the plate circuit of the 1851. A lead from the cathode of the type 1851 tube was provided so that various values of cathode bypass capacitor could be easily connected into the circuit.

The low-frequency oscillator used in this work was a Western Electric Model 13A beat-frequency oscillator. This oscillator is directly calibrated down to 20 cycles per second and up to 10,000 cycles per second. It proved very satisfactory for this work due to its low harmonic content and ample output.

THE MEASUREMENT OF LOW-FREQUENCY PHASE SHIFT IN AMPLIFIERS

RESULTS AND CONCLUSIONS

A considerable number of measurements at low frequencies were made of the phase shift and gain characteristics of the test amplifier under different conditions. Some of the results of the measurements are illustrated by figures (16) to (21) inclusive. Figures (16) and (17) show the measured (solid line) and the calculated (dotted line) phase shift and gain characteristics of a low-frequency network whose constants are shown on the figures. It may be seen that the measured and the calculated values check very well and any difference may easily be accounted for by the errors in measurement that are inherent to the method.

These curves, shown in figures (16) to (21), are difference curves which were obtained as follows. The test amplifier and the output amplifier were calibrated both for phase shift and for gain at the low frequencies. Then the low-frequency network, for example, was inserted into the test amplifier and the circuit again measured. The difference between the two measurements was taken to be the amount of phase shift and gain introduced into the circuit by the inserted network. The calculated values, shown as dotted lines, were taken from the corresponding curves which were calculated from the general parametric equation (5) on page 16.

Of particular interest are the curves showing the characteristics of the low-frequency compensation networks. The close check between the calculated and the measured values serves to check both the accuracy and the consistancy of the method used to measure the phase shift and also the accuracy of the general design curves for the low-frequency compensation networks.

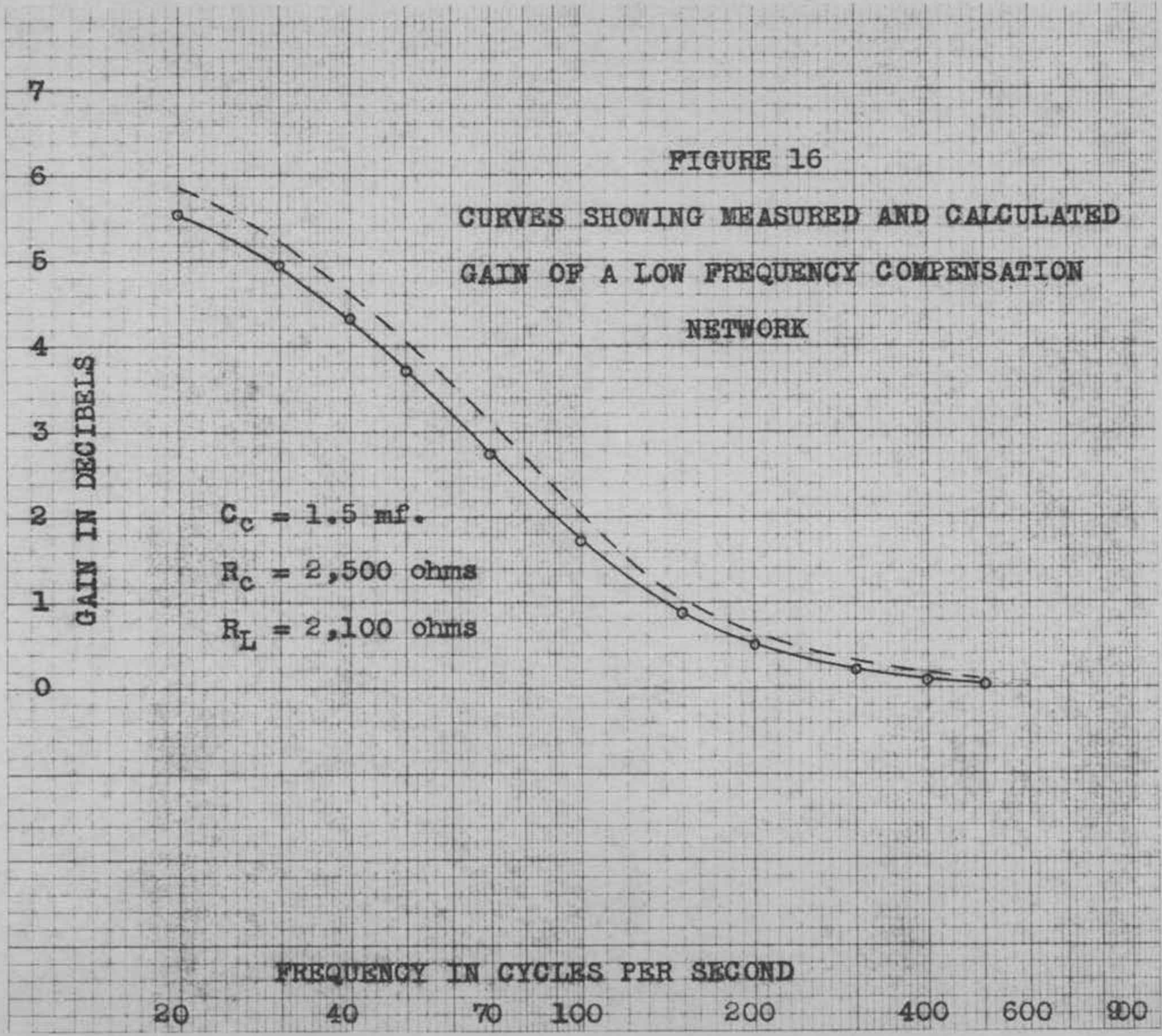


FIGURE 16

CURVES SHOWING MEASURED AND CALCULATED GAIN OF A LOW FREQUENCY COMPENSATION NETWORK

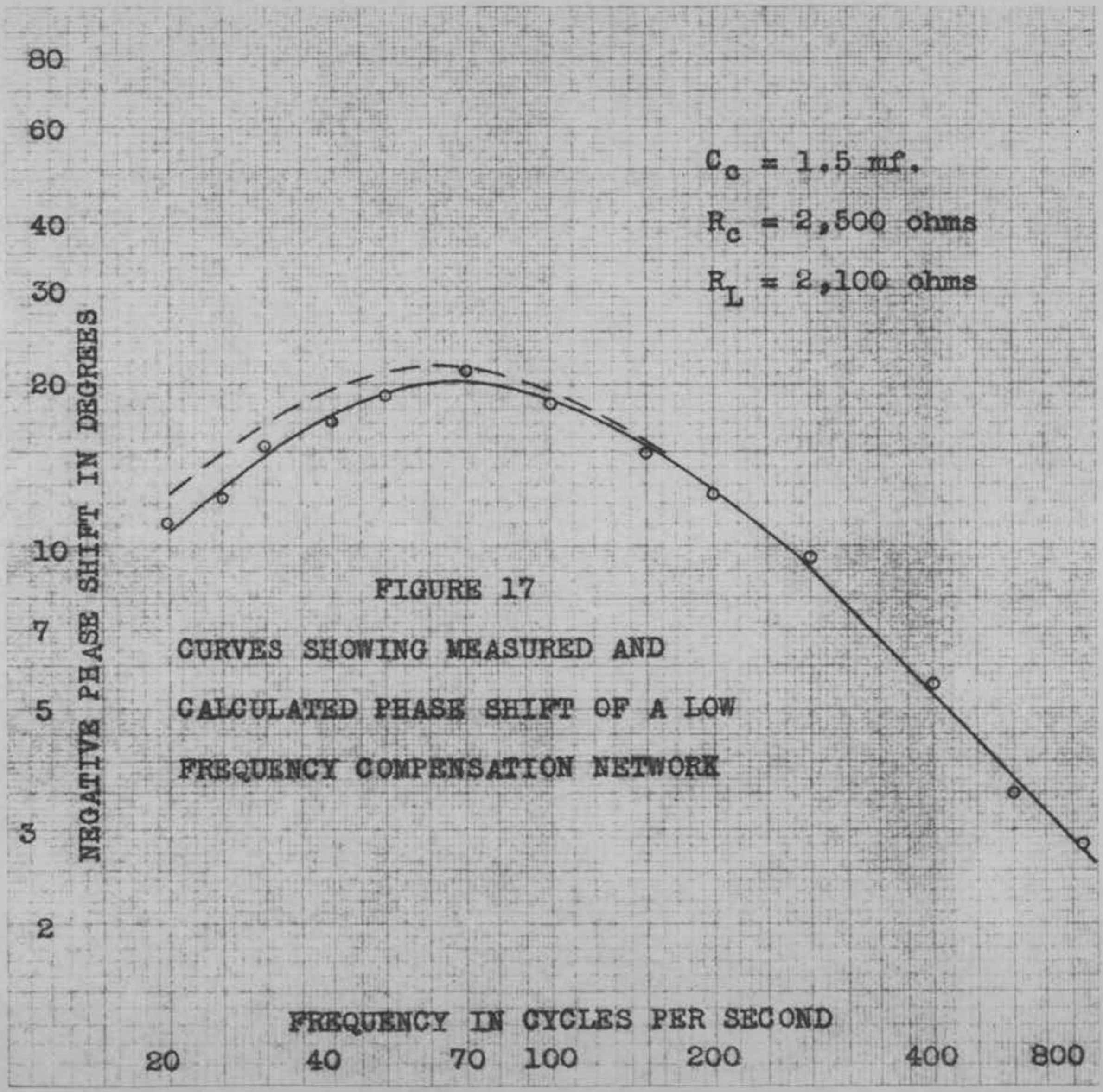


FIGURE 17

CURVES SHOWING MEASURED AND CALCULATED PHASE SHIFT OF A LOW FREQUENCY COMPENSATION NETWORK

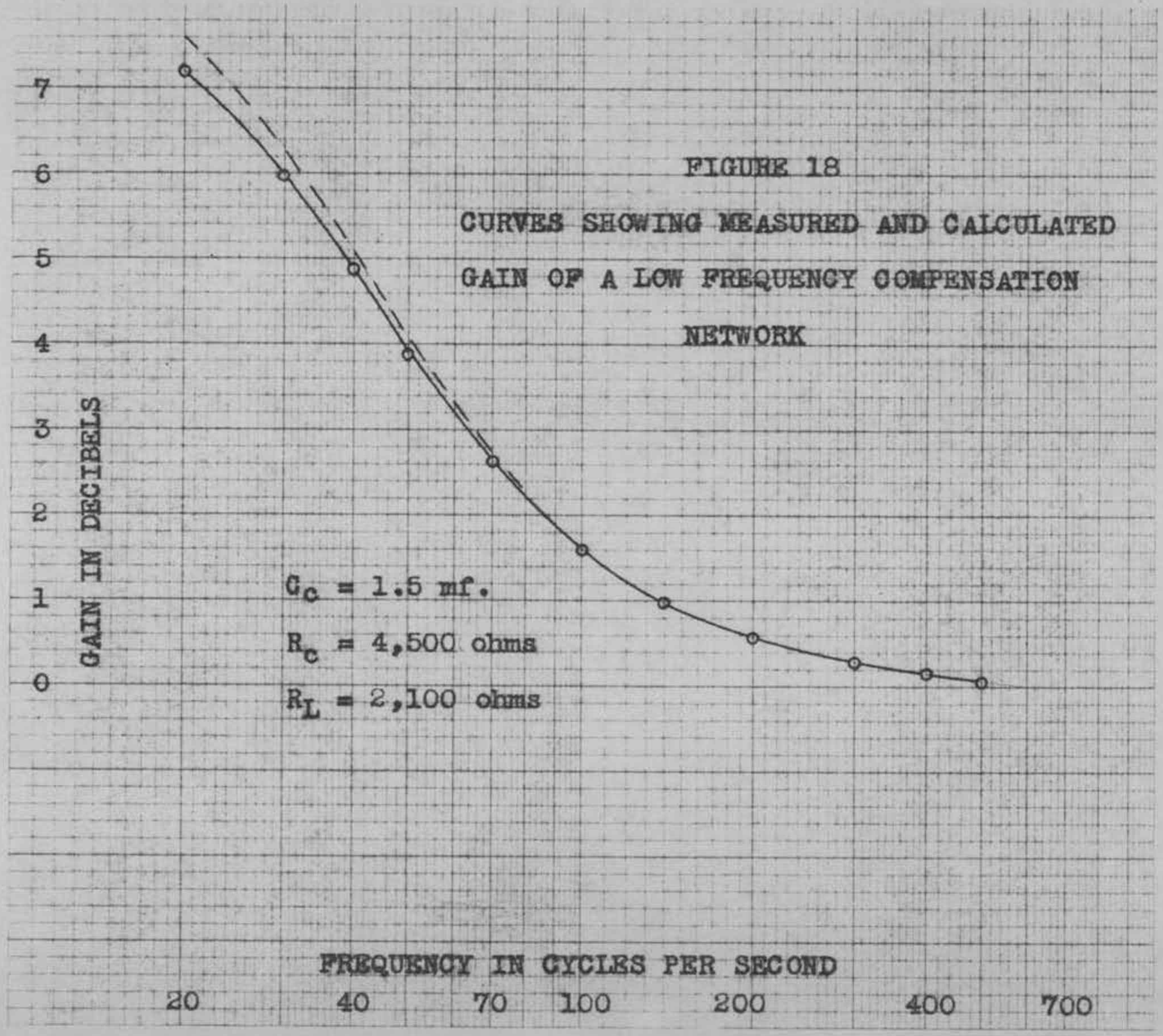


FIGURE 18

CURVES SHOWING MEASURED AND CALCULATED GAIN OF A LOW FREQUENCY COMPENSATION NETWORK

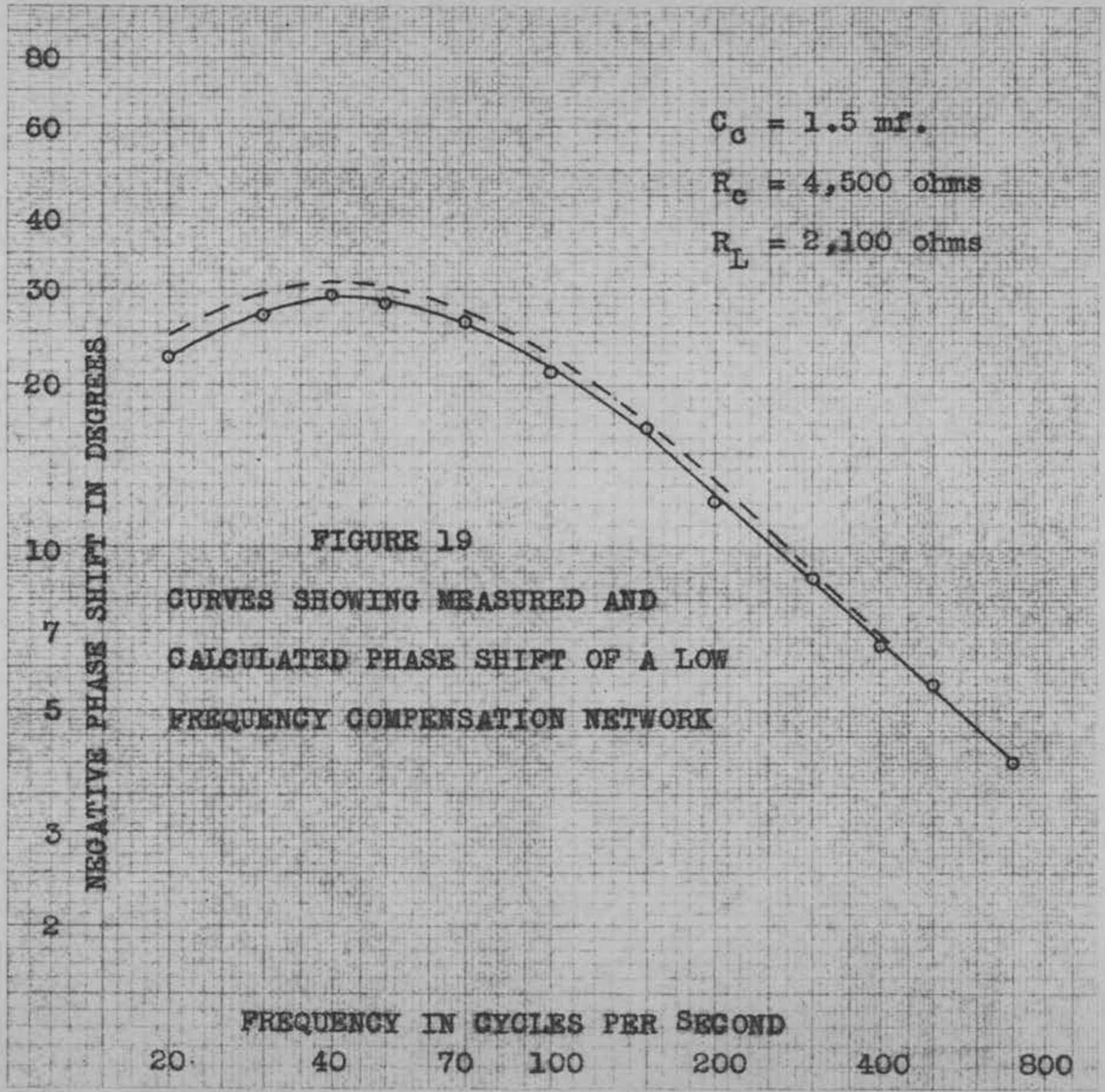


FIGURE 19

CURVES SHOWING MEASURED AND CALCULATED PHASE SHIFT OF A LOW FREQUENCY COMPENSATION NETWORK

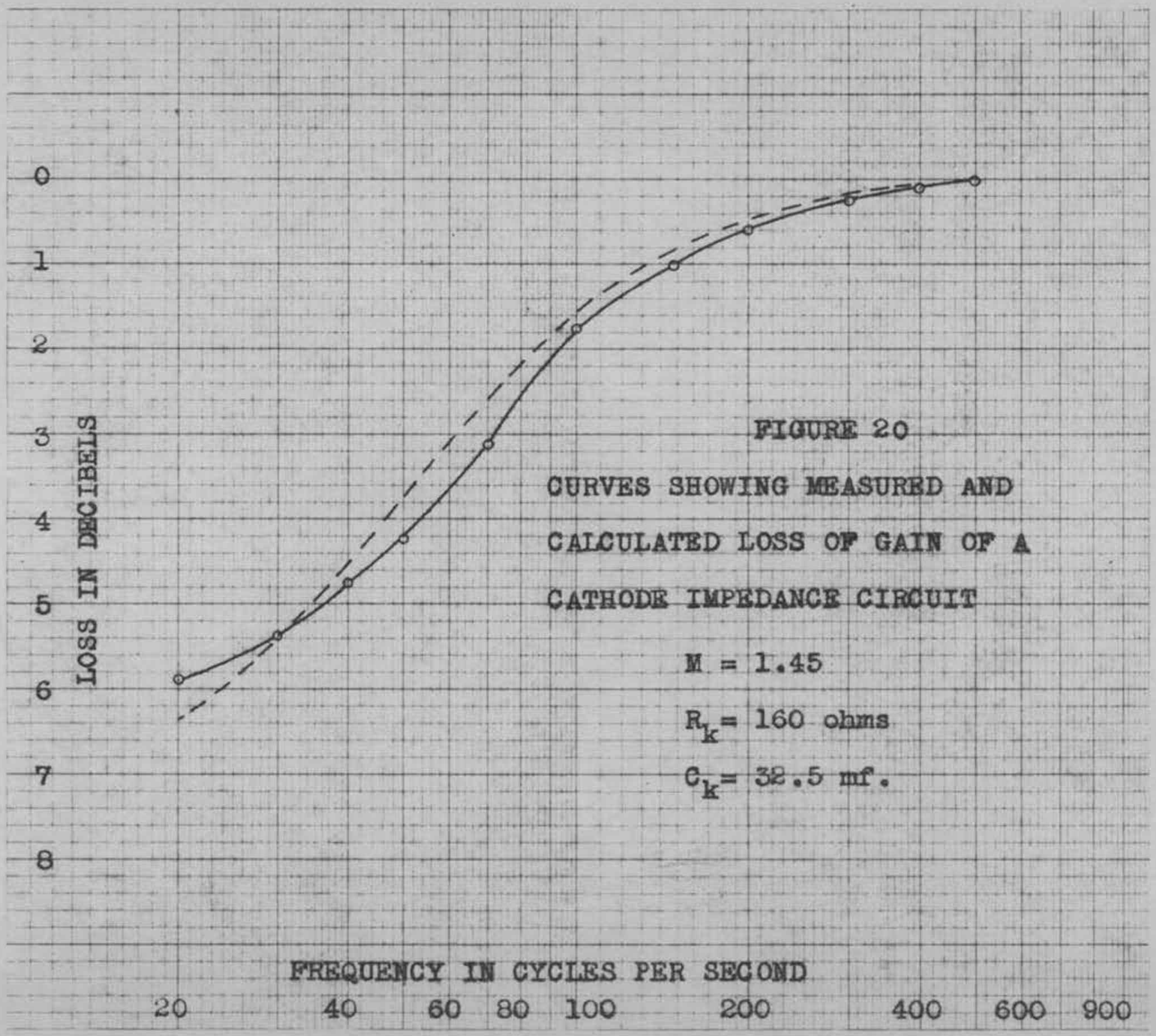


FIGURE 20

CURVES SHOWING MEASURED AND CALCULATED LOSS OF GAIN OF A CATHODE IMPEDANCE CIRCUIT

$M = 1.45$

$R_k = 160$ ohms

$C_k = 32.5$ mf.

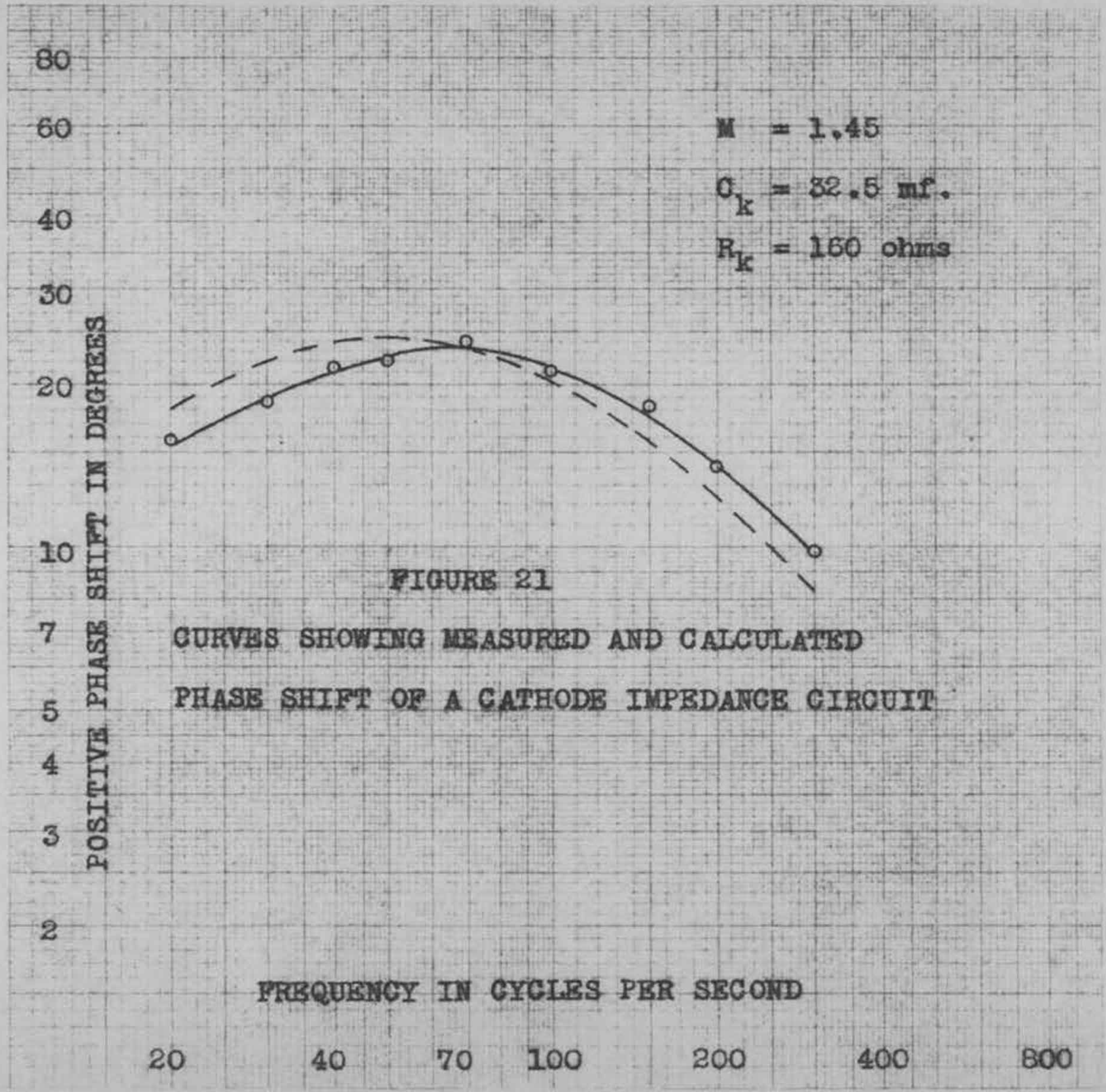


FIGURE 21

CURVES SHOWING MEASURED AND CALCULATED PHASE SHIFT OF A CATHODE IMPEDANCE CIRCUIT

NEGATIVE FEEDBACK AMPLIFIERS AT LOW FREQUENCIES

The use of negative feedback for the reduction of noise, hum and harmonic distortion is well known. (4) (36). However its use in television amplifiers has been avoided due to a lack of exact knowledge of the behavior of feedback amplifiers at the extreme ends of the frequency band.

In reference (37), F. E. Terman points out that peaks in the response curve occur at both high and low frequencies in negative feedback amplifiers. He assumes in his discussion of the problem that the feedback circuit is a resistance network and consequently has no phase shift. This assumed condition is mainly of academic interest in the case of resistance-coupled amplifiers which require a blocking condenser in nearly all negative feedback arrangements. The present writer investigated this problem of frequency response and phase shift in a two-stage amplifier using a capacitance and resistance network in the feedback circuit. The following discussion is confined to the low-frequency end of the response curves.

In order that the mathematical investigation might be simplified as much as possible, it is assumed that the only source of loss of gain and phase shift in the amplifier proper is in the grid-coupling circuit between the two stages. The negative-feedback circuit feeds the output voltage from the plate of the last stage through a capacitance and a resistance voltage divider made up in part of the cathode resistor in the cathode circuit of the first tube. (see figure (22)). This is the ordinary circuit employed for two-stage negative feedback resistance coupled amplifiers.

Using the familiar feedback equation

$$\text{Gain} = \frac{A}{1 - A\beta}, \tag{12}$$

where A = complex amplification factor of the amplifier without feedback, and

β = complex feedback ratio (negative for negative feedback).

The writer developed a general parametric equation for the gain and phase shift of this negative feedback amplifier. Two parameters are involved, one being the feedback factor $A_o\beta_o$ at the middle of the frequency band and the other the ration T_β/T_g or the ratio between the time constant of the feedback network and the time constant of the grid-coupling circuit. This equation is derived in detail in Appendix (2) and is

$$\text{Response ratio} = \frac{wT_\beta(wT_\beta - j1)}{(wT_\beta)^2 - p/(1+d) - jwT_\beta(1+p)/(1+d)}. \tag{13}$$

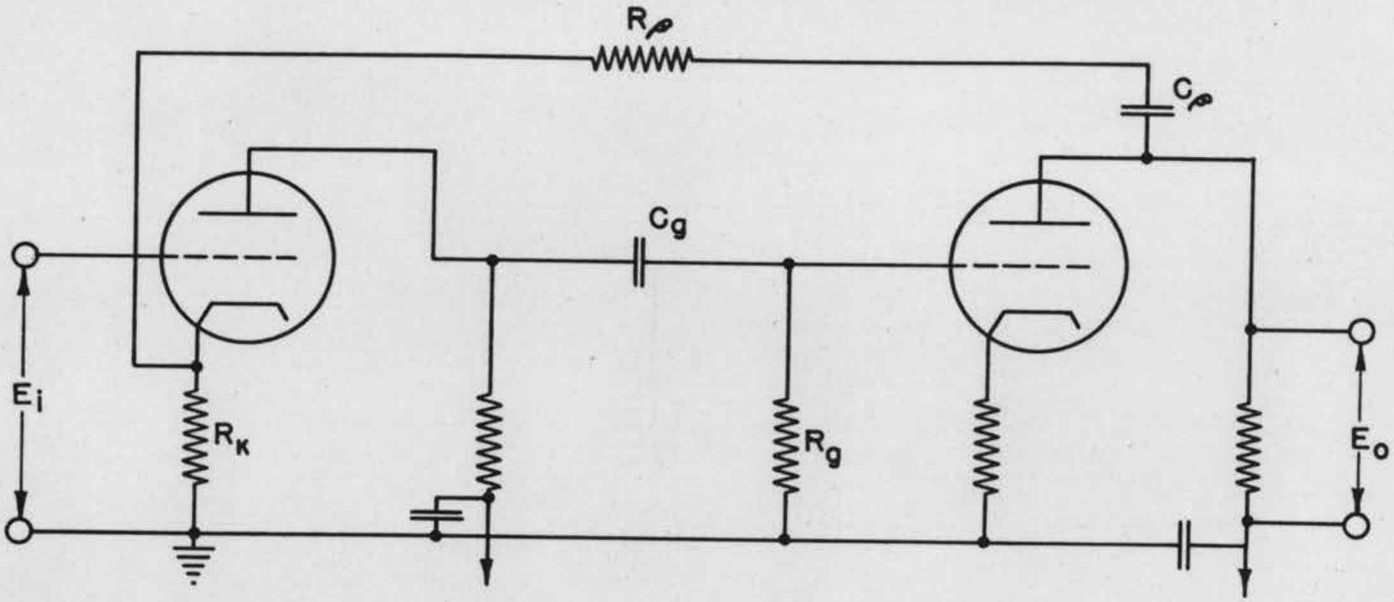


Figure 22

CIRCUIT DIAGRAM FOR MATHEMATICAL INVESTIGATION
OF NEGATIVE FEEDBACK AMPLIFIERS

where

response ratio = ratio between the output voltage at low frequency and the output voltage at mid-band frequency,

$w = 2\pi f$,

f = frequency in cycles per second,

$T_\beta = C_\beta(R_\beta + R_k)$ time constant of the negative feedback network,

j = operator square root of (-1),

$p = T_\beta / T_g$ = ratio of the time constant of the feedback circuit and the time constant of the grid circuit,

$d = A_o\beta_o$ feedback factor at mid-band frequency,

A_o = amplification factor of amplifier without feedback at mid-band frequency, and

β_o = feedback ratio at mid-band frequency.

Figures (23) to (28) inclusive, show the results of plotting the above equation (13) with wT_β as the independent variable and varying the parameter $A_o\beta_o$ from 2 to 10 and the parameter $p = T_\beta/T_g$ from 0.5 to 5.0 for each value of $A_o\beta_o$. These curves show the range of the variations of the gain and phase shift of the usual negative feedback amplifier.

It will be noted, when the curves are examined, that there exists a definite peak at the low frequencies and this peak occurs where the rate of change of phase shift is relatively large. The peaks are higher as the feedback factor $A_o\beta_o$ is increased in magnitude for the same value of the parameter p. For any fixed value of $A_o\beta_o$, the peaks in the low-frequency end of the band become smaller and smaller as the value of p increases. There is seen to be an optimum value of p for each value of the feedback factor $A_o\beta_o$ which makes the phase shift substantially zero out to a certain frequency and this optimum value of p also gives only a very small peak in the region of large rate of change of phase shift.

By the use of these curves, it is possible to so proportion the value of p for any feedback factor, that the phase shift is zero out to a certain frequency and have only a very small peak in the low-frequency response curve below the useful range of the amplifier.

In the design of a feedback amplifier for use in television, the time constant T_β is limited by the physical size of the capacitor C_β which is used to insulate the plate voltage of the second tube from ground. This capacitor must not be excessive in size or its stray capacitance to ground will have an adverse effect on the high-frequency response of the amplifier. The resistance R_β is limited by the amount of feedback desired. The cathode resistance R_k is limited by the tube which is used in the first stage of the amplifier since this resistor provides bias for the grid of the tube.

Thus the only way to increase the value of p is to decrease the time constant T_g of the grid-coupling circuit between the stages. This tends to improve the action of the amplifier under high amplitude shock and tends to prevent blocking or paralysis of the amplifier. Blocking is generally caused by excessive time constants in the grid-coupling circuits (31).

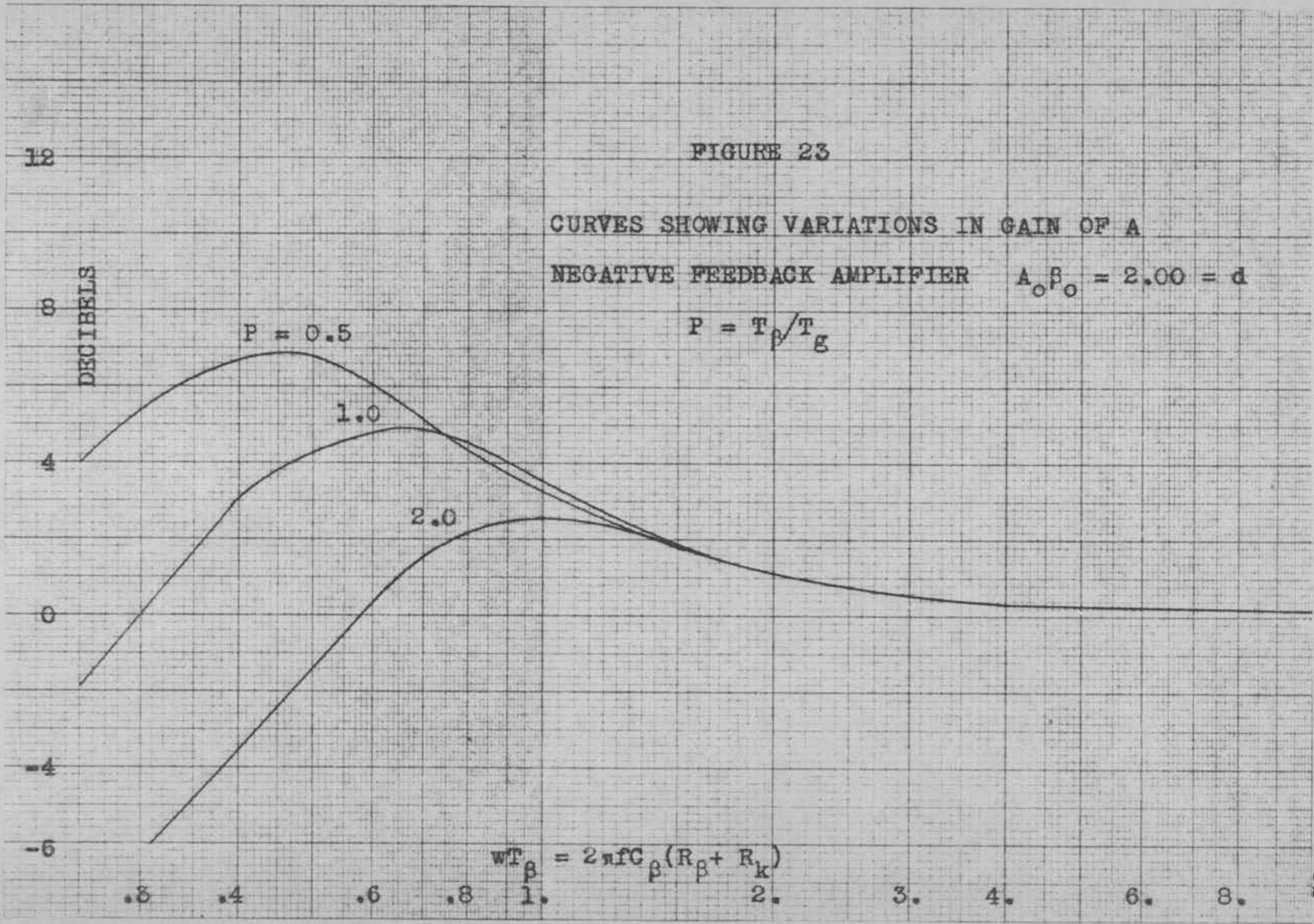


FIGURE 23

CURVES SHOWING VARIATIONS IN GAIN OF A NEGATIVE FEEDBACK AMPLIFIER $A_o\beta_o = 2.00 = d$

$P = T_\beta/T_g$

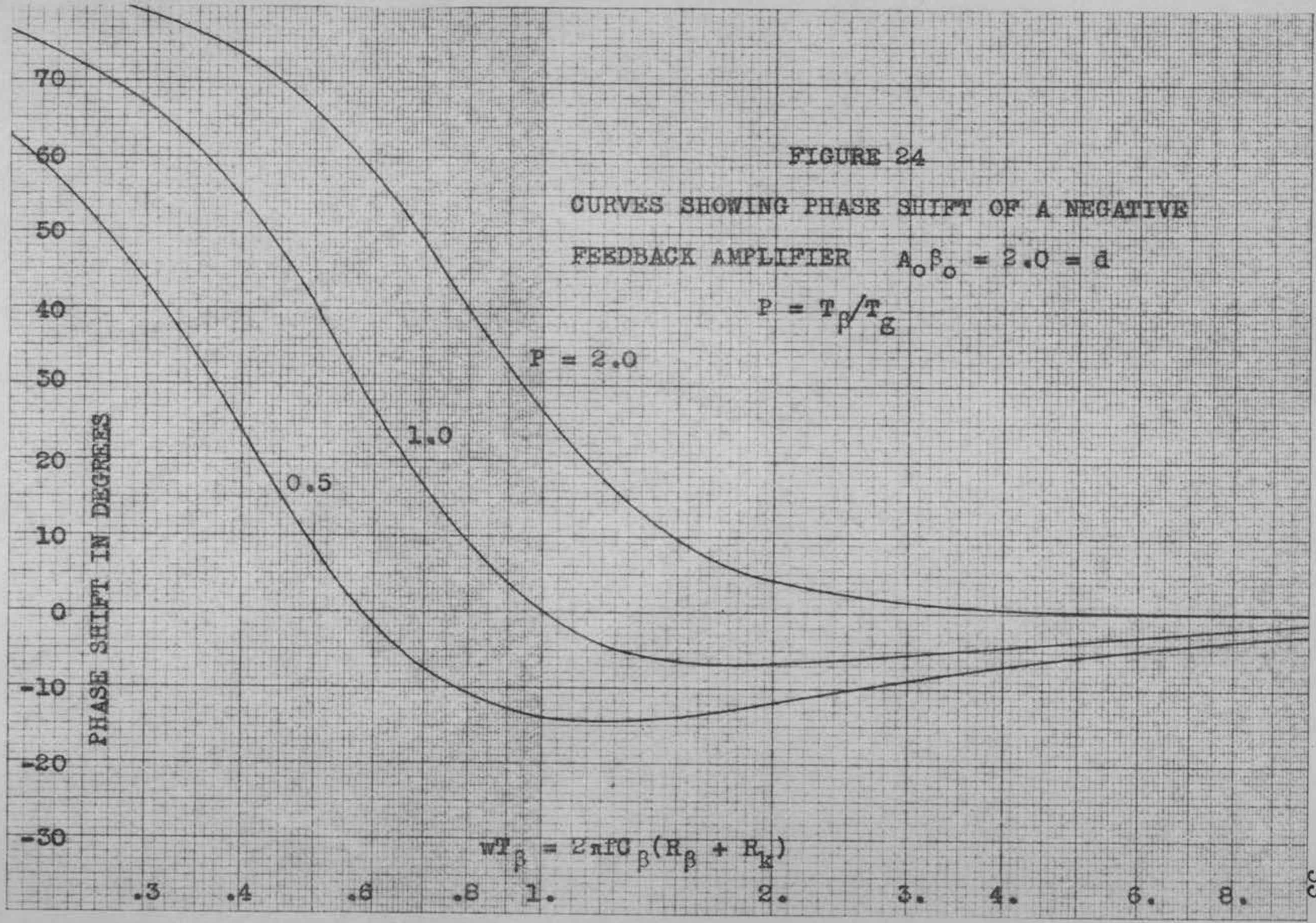


FIGURE 24

CURVES SHOWING PHASE SHIFT OF A NEGATIVE FEEDBACK AMPLIFIER $A_o\beta_o = 2.0 = d$

$P = T_\beta/T_g$

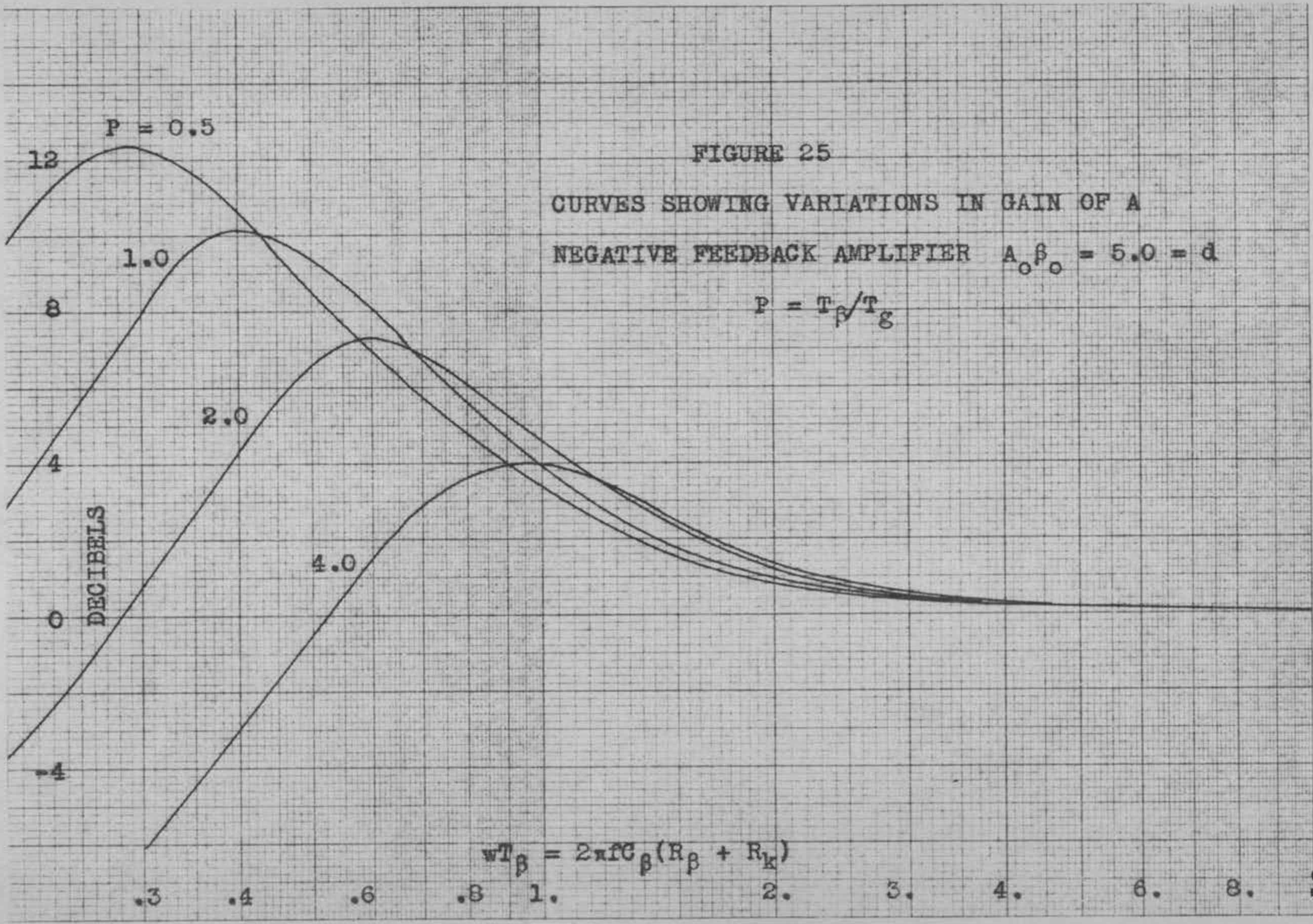


FIGURE 25

CURVES SHOWING VARIATIONS IN GAIN OF A NEGATIVE FEEDBACK AMPLIFIER $A_o\beta_o = 5.0 = d$

$P = T_\beta/T_g$

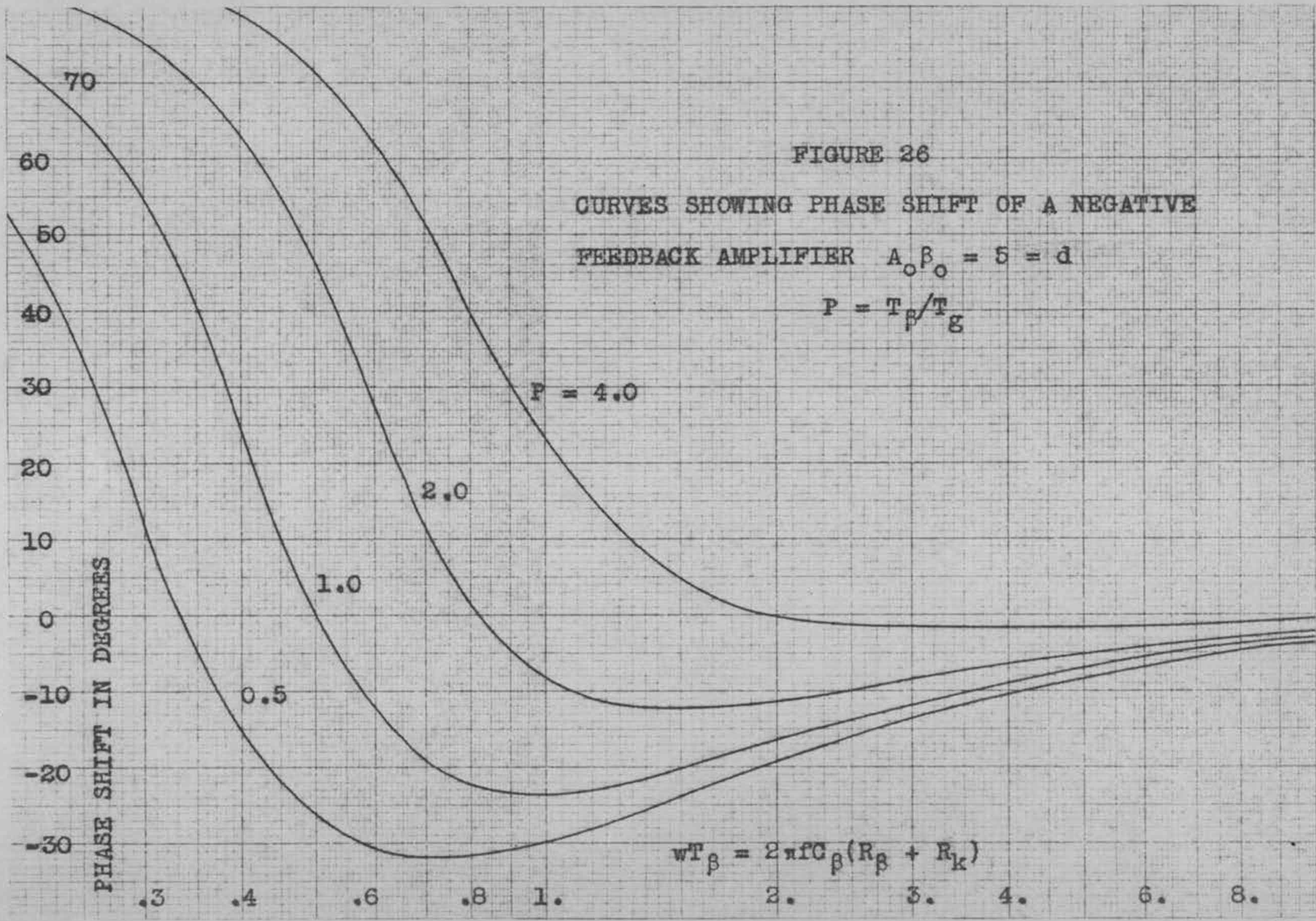


FIGURE 26

CURVES SHOWING PHASE SHIFT OF A NEGATIVE FEEDBACK AMPLIFIER $A_o \beta_o = 5 = d$

$P = T_\beta / T_g$

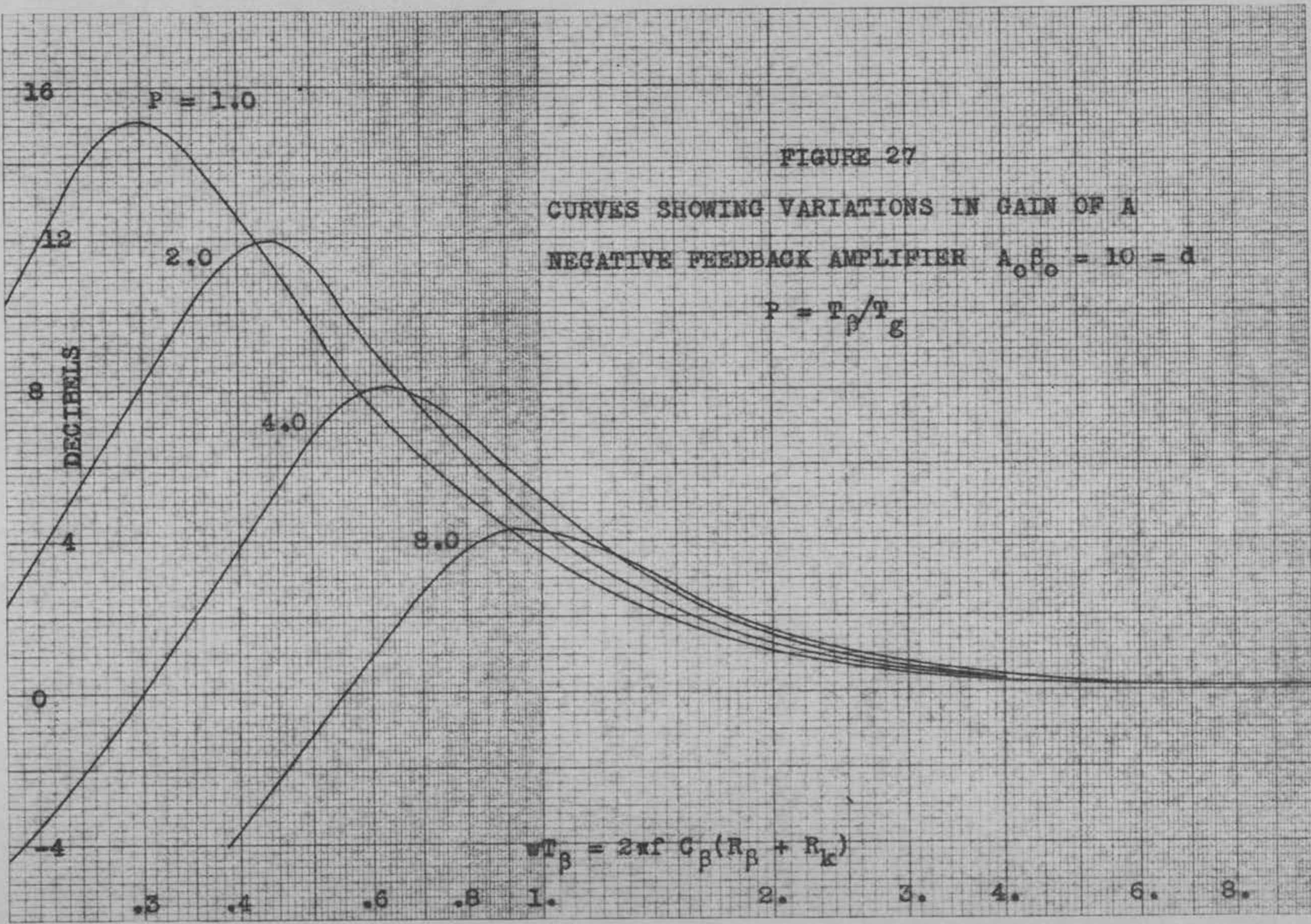


FIGURE 27

CURVES SHOWING VARIATIONS IN GAIN OF A NEGATIVE FEEDBACK AMPLIFIER $A_o\beta_o = 10 = d$

$P = T_\beta/T_g$

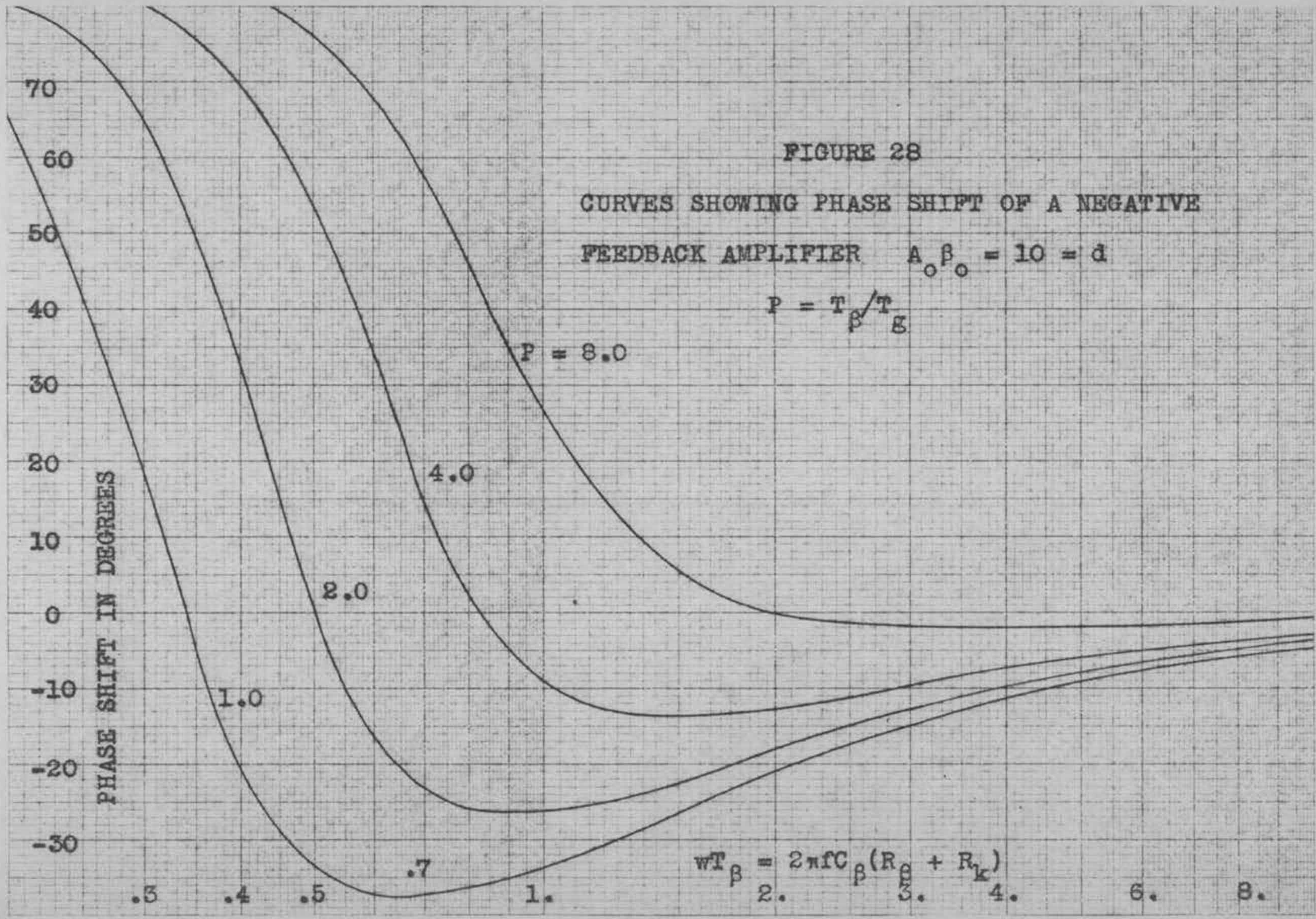


FIGURE 28

CURVES SHOWING PHASE SHIFT OF A NEGATIVE FEEDBACK AMPLIFIER $A_o\beta_o = 10 = d$

$P = T_{\beta}/T_g$

HIGH-FREQUENCY CONSIDERATIONS

DISCUSSION

The recently released Radio Manufactures Association (RMA) television standards require the frequency band of a video amplifier to be 2.5 megacycles in width. This means that the television video amplifier stages are required to amplify all frequencies up to 2.5 megacycles per second with substantially constant gain and constant delay. (12). This relatively large frequency band width is necessary to resolve effectively high-definition television pictures.

Insufficient gain at the higher video frequencies shows up in the television picture as a loss of fine detail. For example, the pattern of the weave of a man's coat would be lost if the video amplifiers were deficient in gain at high frequencies. The presence of negative phase shift or excessive time delay of the high frequencies results in the so-called ghosts or secondary images on the television screen. The effect is caused by the delayed arrival of the high-frequency components of the picture. This is because the details of the picture are misplaced in relation to the other lower-frequency components and thus a faint secondary image is produced. Since time delay and deficiency in frequency response go together, this double image is not bright except under exaggerated conditions designed to show the effect.

The cause of loss of gain and negative phase shift or excessive time delay at high frequencies in resistance-capacitance coupled amplifiers is the effective capacitance to ground in parallel with the plate load resistor. This effective capacitance consists of the output capacitance of the tube employed in the stage, the input capacitance of the tube in the following stage and the stray capacitance to ground of the coupling capacitor and the wiring.

The relative gain and the phase shift of an uncompensated amplifier depend on the time constant of the load which is the product of the plate load resistance in ohms and the stray capacitance in farads. The following equation for the relative response may easily be derived on the assumption that the dynamic plate resistance of the tube is very much larger than the plate load resistance.

$$\text{Response ratio} = \frac{1}{1 + jwT_L}, \tag{14}$$

where response ratio = ratio between the gain at high frequency and the gain at mid-band frequency,

$w = 2\pi f$, and

$T_L = C_L R_L$ = time constant of plate load.

Figures (29) and (30) are the result of plotting equation (14) with wT_L as an independent variable and loss of gain in decibels and phase shift in degrees as dependent variables. From these general curves may be found the loss of gain and the phase shift of any amplifier for any frequency and for any value of T_L the time constant. It may be seen that any amplifier must have a value of wT_L of less than 1.00 if it is to amplify effectively at this frequency ($w=2\pi f$). At this value, the gain is down 3 decibels and the phase shift is 45 degrees. With larger values of wT_L , the gain falls off and the phase shift is very bad.

Since the loss of gain and the increase of delay or negative phase shift are proportional to this time constant T_L, it is desirable to reduce the value of T_L to as small a value as practical. As previously stated, the time constant is equal to the product of the plate load resistance times the effective stray capacitance in parallel with the load resistor. Thus either one or both of these quantities may be decreased to reduce the time constant but not beyond certain practical limits. The output capacitance of the tube of the stage and the input capacitance of the tube in the following stage are more or less fixed quantities. The stray capacitance to ground may be reduced somewhat by extreme care in wiring and the use of

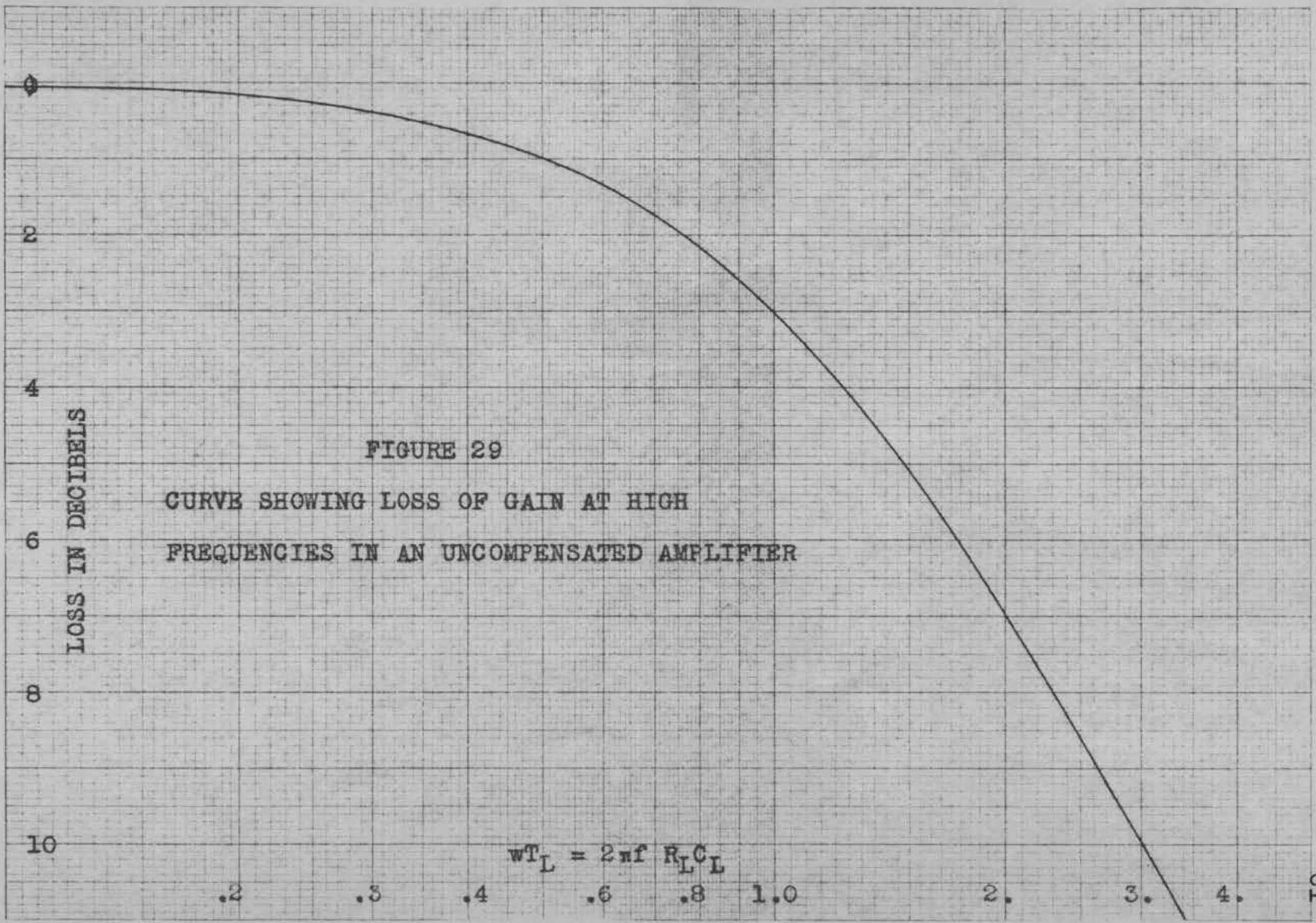


FIGURE 29

CURVE SHOWING LOSS OF GAIN AT HIGH FREQUENCIES IN AN UNCOMPENSATED AMPLIFIER

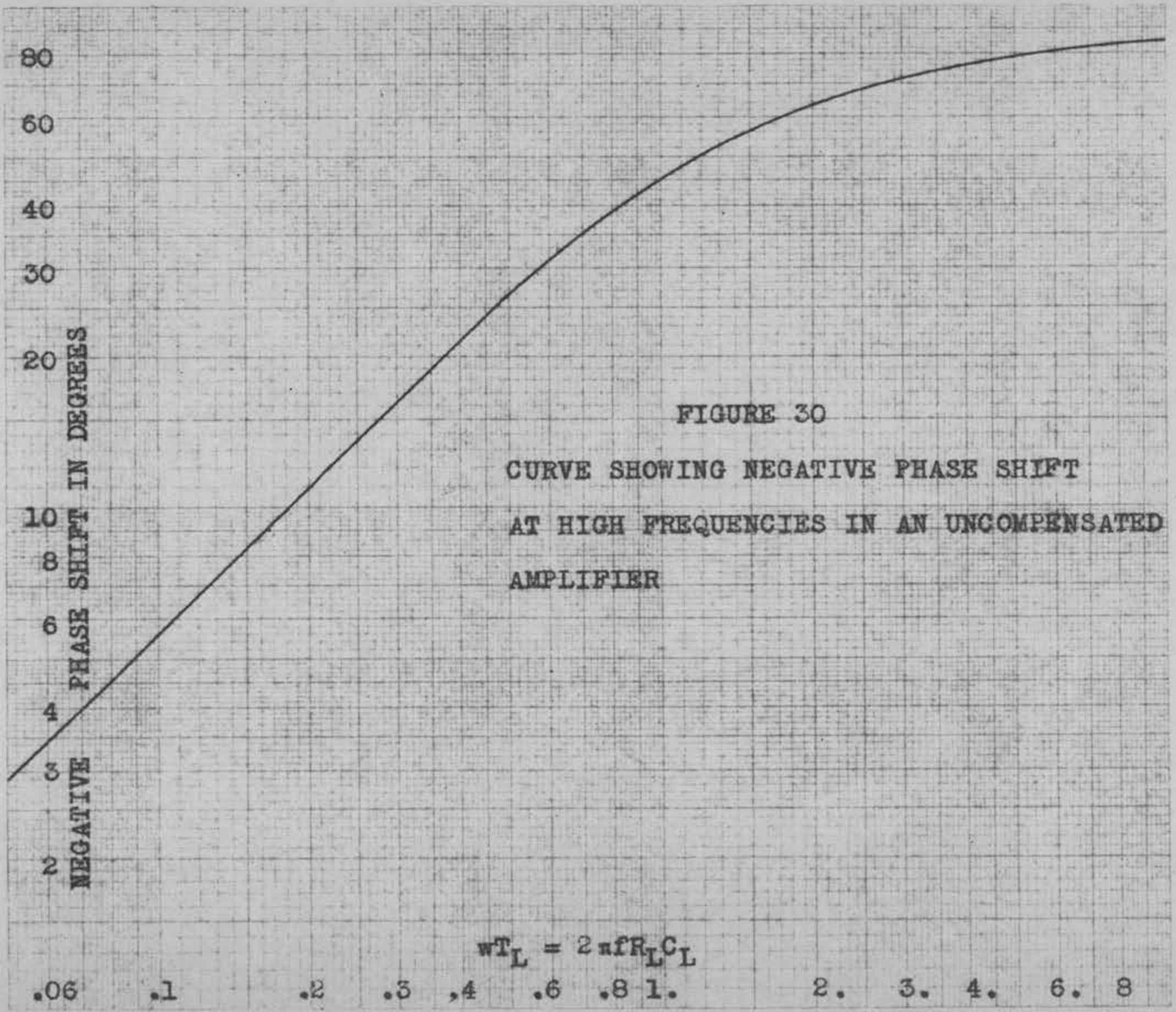


FIGURE 30

CURVE SHOWING NEGATIVE PHASE SHIFT AT HIGH FREQUENCIES IN AN UNCOMPENSATED AMPLIFIER very short leads and by the use of a physically small plate to grid coupling capacitor. After this is done there remains only the reduction of the value of the plate load resistance.

There exists, however, a very definite limit to this method of the improvement of high-frequency response. The gain of any amplifier stage using pentode tubes with a plate load resistance which is relatively small compared to the dynamic plate resistance of the tube is given by the equation

$$\text{Gain} = G_m R_L, \tag{15}$$

where the gain is that of the middle of the frequency band, G_m is the transconductance of the tube and R_L is the plate load resistance. As R_L is reduced, the mid-band frequency amplification becomes smaller. There must be a sort of "economic balance" between the high-frequency characteristics and the mid-band amplification of the stage.

The above considerations show that other means of improving the high-frequency response of an amplifier are necessary for satisfactory operation.

HIGH FREQUENCY GAIN AND PHASE SHIFT COMPENSATION CIRCUITS

There are a number of different methods (or circuits) for the improvement of high-frequency response and the reduction of phase delay. The most widely known method makes use of a small inductance which is inserted in series with the plate load resistor. This method is treated in detail in references (10), (12), and (28) and will not be given much space in this paper.

Another method consists of bypassing a portion of the plate load resistor with a small capacitor and by the proper proportioning of the constants, the high-frequency gain can be made to exceed the mid-band frequency amplification at the expense of the mid-band amplification. See figure (31) and reference (26) page 190.

A third method for high-frequency compensation consists of the use of only a small bypass capacitor across the cathode self-biasing resistor. This method improves the high-frequency response at the expense of the mid-band frequency amplification factor. At low frequencies, the impedance of the cathode bypass capacitor is practically infinite and the stage operates as a negative feedback amplifier. This tends to reduce harmonic distortion and noise as in any negative feedback arrangement but a certain amount of gain is sacrificed. However the

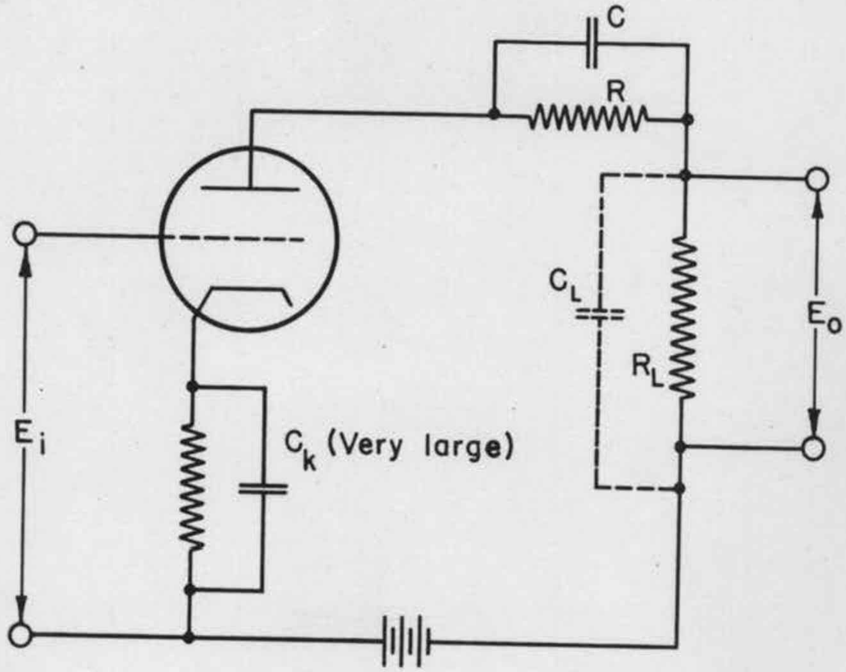


Figure 31.

USE OF RESISTANCE SHUNTED BY CAPACITANCE IN SERIES WITH THE PLATE LOAD RESISTOR TO IMPROVE HIGH-FREQUENCY RESPONSE introduction of the type 1851 and 1852 high transconductance pentodes makes the use of cathode impedance high-frequency compensation practical.

It has been realized for some time that the high-frequency response of an amplifier could be improved by the use of a small capacitor across the cathode self-biasing resistor but the exact effect on frequency response and phase shift has never been properly calculated (1), as far as the writer has been able to discover. The desire for a knowledge of the exact numerical relations between the various factors involved, led the writer to develop a general parametric equation for the relative response of an amplifier stage making use of this method of high-frequency compensation.

The general equation is derived in Appendix (4) and is shown below.

$$\text{Response ratio} = \frac{(1 + M)(P + jwT_L)}{(1 + jwT_L)(P + PM + jwT_L)}, \quad (16)$$

where response ratio = ratio between the output voltage at high frequencies and the output voltage at low frequencies,

$M = uR_k/R_p = G_m R_k$, and

$P = T_L/T_K = C_L R_L/C_k R_k$ = ratio of time constants of the plate load circuit and the cathode impedance circuit.

(See appendix (4) or the list of symbols for a complete list of all symbols used in the equation). The equation was derived on the assumption that the plate load resistance is negligible in comparison with the dynamic plate resistance of the tube.

Figures (32) to (37) inclusive show the result of plotting equation (16) with wT_L as an independent variable with loss of gain in decibels and phase shift in degrees for various values of the parameters M and P. It may be seen that for any particular value of M, there exists an optimum value of P which causes the stage to have the best phase shift characteristics. There is another value of P which gives the best gain characteristics. These values of P are not the same and a compromise value between the two would be the best possible solution. The value of P which gives the best phase characteristics may be seen to give a small peak of 2 or 3 decibels in the response curve.

The improvement in response and phase shift at high frequencies is easily seen by comparison with the dotted curves on figures (36) and (37) which shows the uncompensated stage characteristics. This method of high-frequency compensation compares favorably with the use of a small inductance in series with the plate load resistor. It has the advantage that the phenomenon of resonance at some

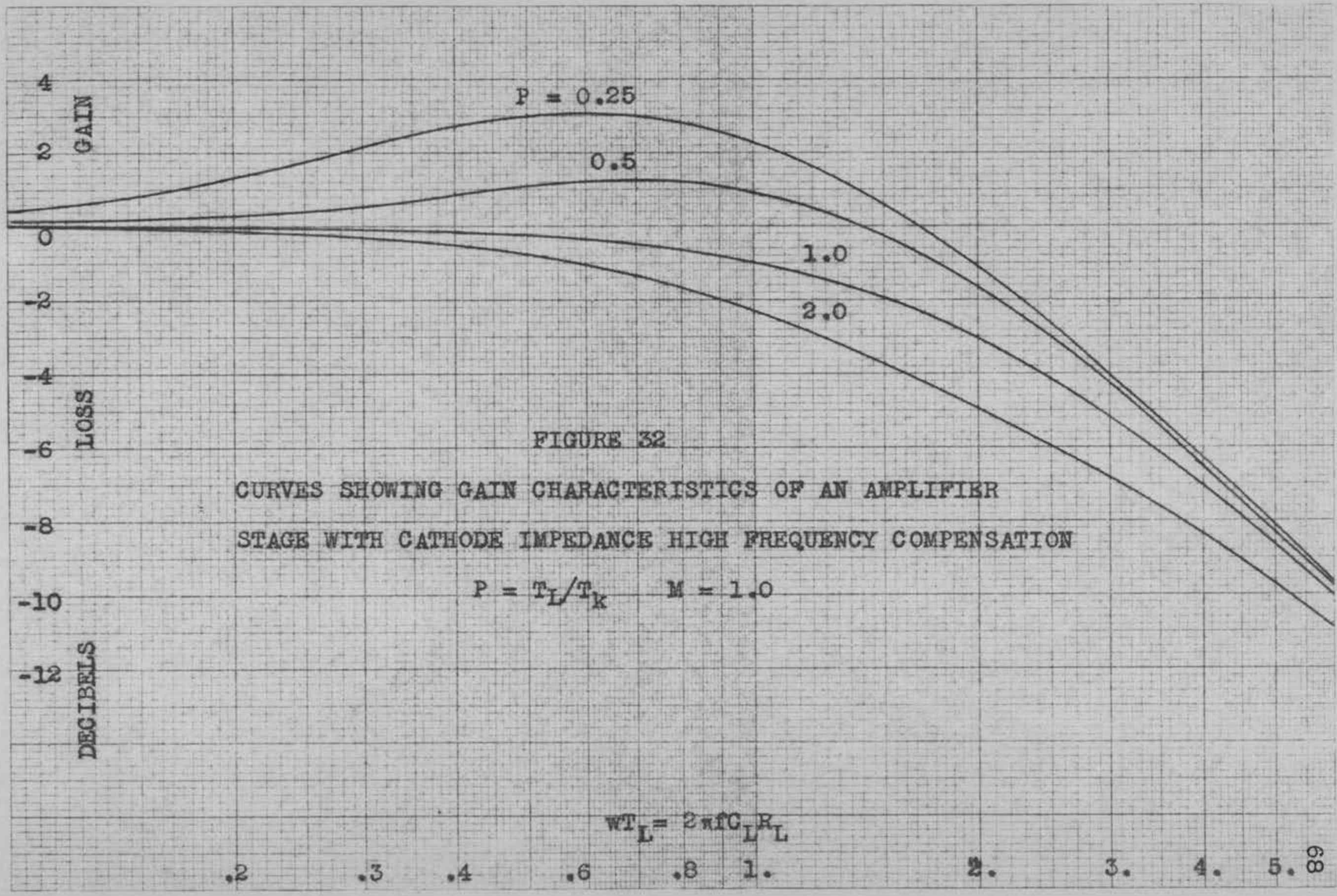


FIGURE 32

CURVES SHOWING GAIN CHARACTERISTICS OF AN AMPLIFIER STAGE WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION

$P = T_L/T_k \quad M = 1.0$

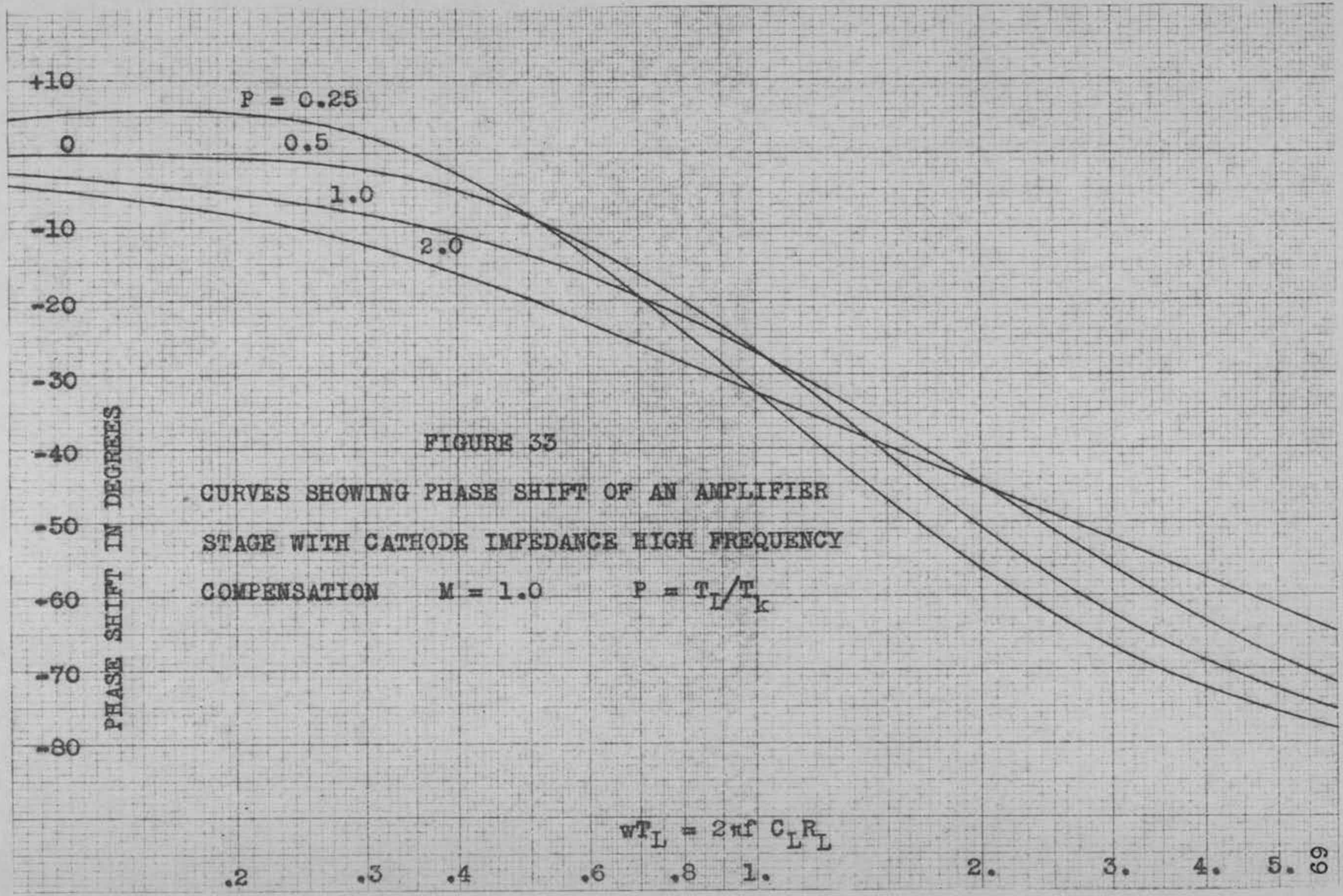


FIGURE 33

CURVES SHOWING PHASE SHIFT OF AN AMPLIFIER STAGE WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION $M = 1.0$ $P = T_L/T_k$

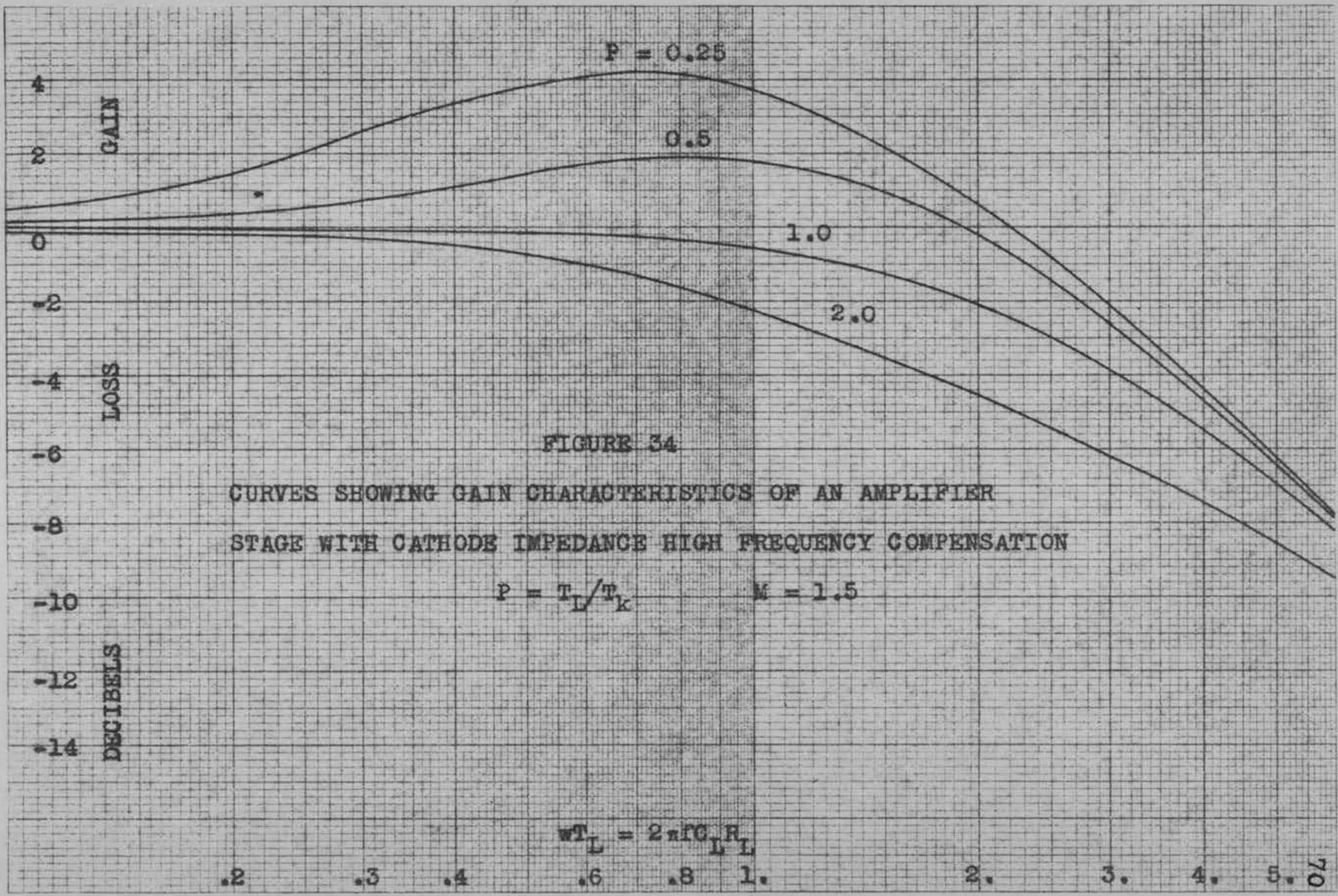


FIGURE 34

CURVES SHOWING GAIN CHARACTERISTICS OF AN AMPLIFIER STAGE WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION

$P = T_L/T_k$ $M = 1.5$

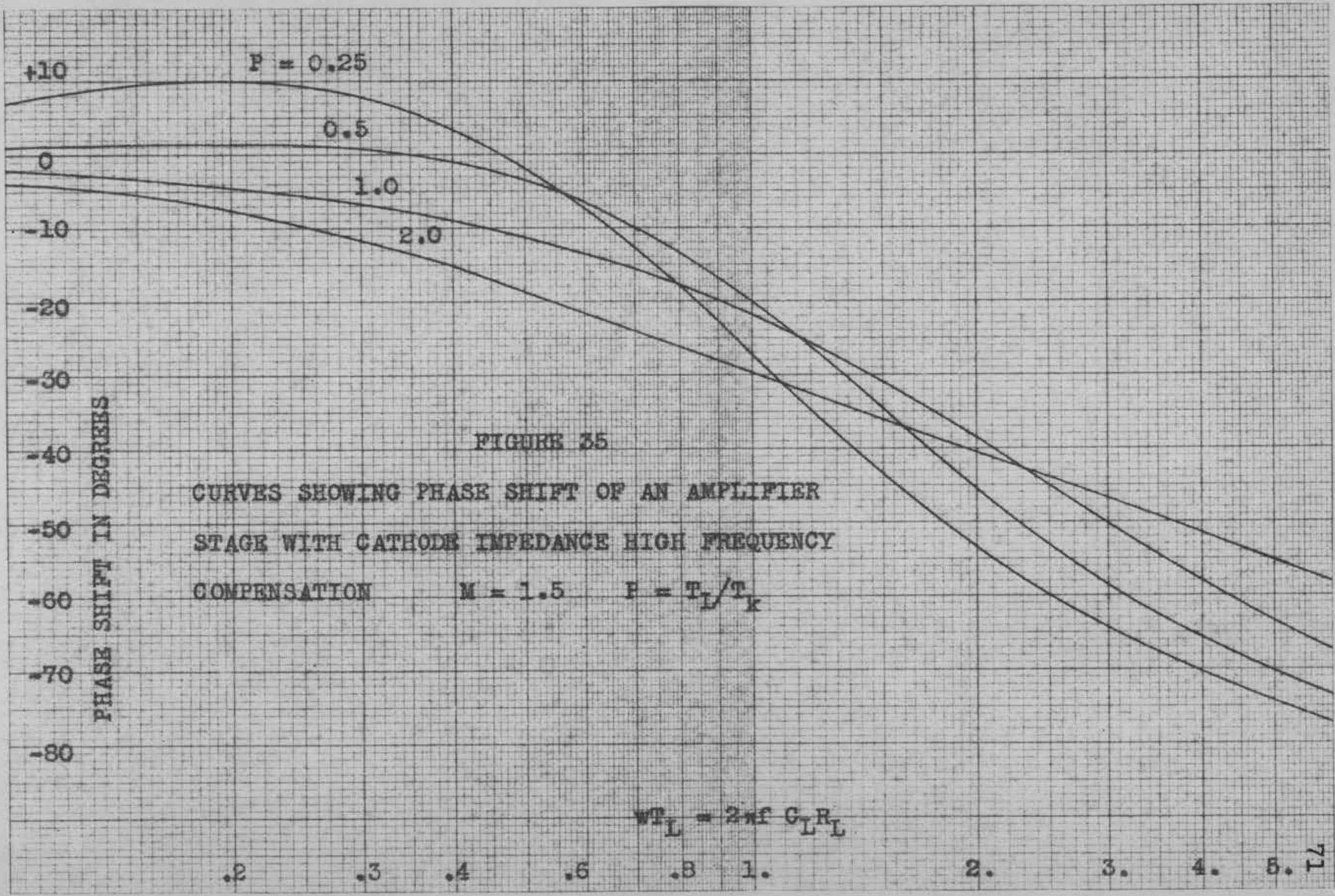


FIGURE 35

CURVES SHOWING PHASE SHIFT OF AN AMPLIFIER STAGE WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION $M = 1.5$ $P = T_L/T_k$

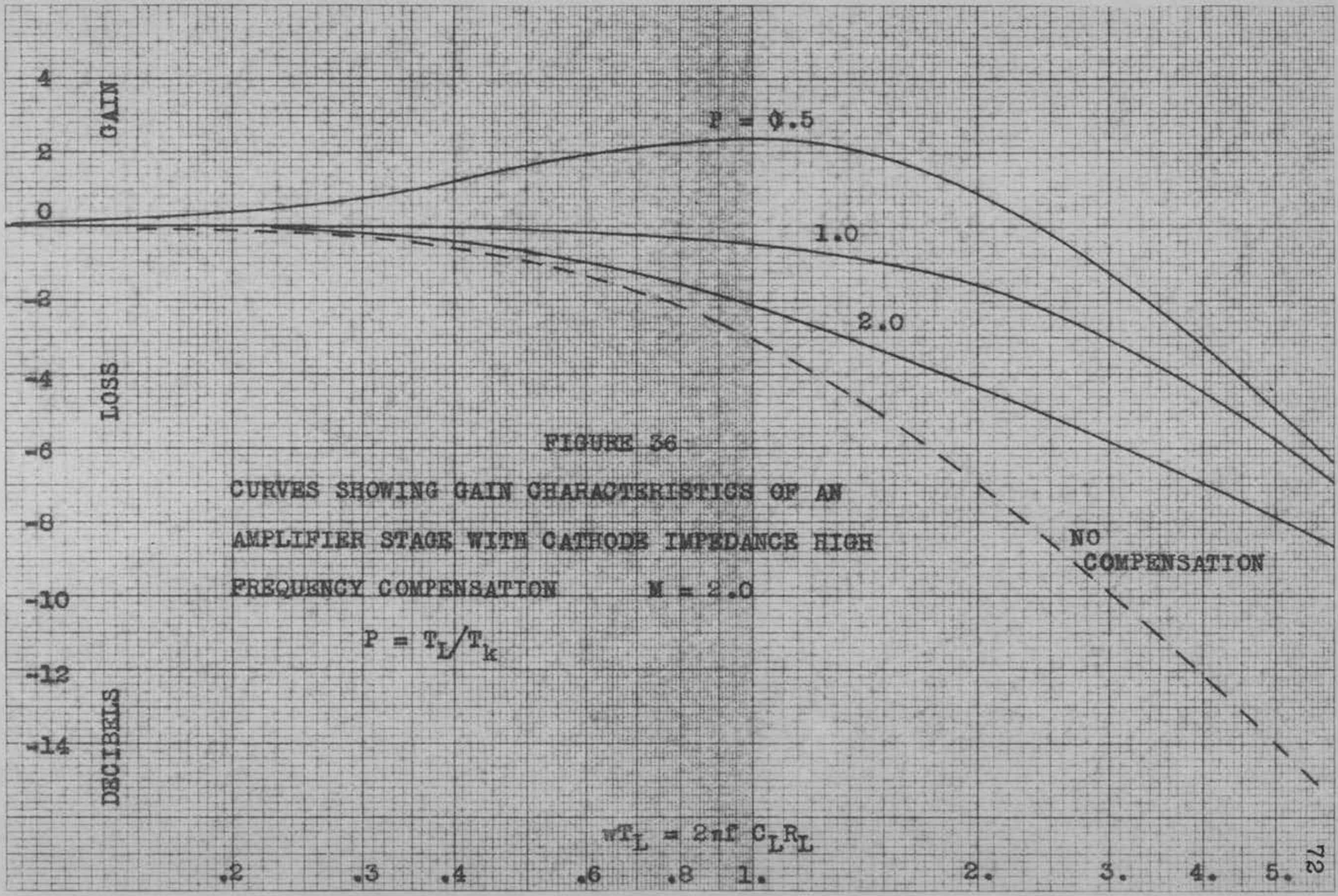


FIGURE 36

CURVES SHOWING GAIN CHARACTERISTICS OF AN AMPLIFIER STAGE WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION $M = 2.0$

$P = T_L/T_k$

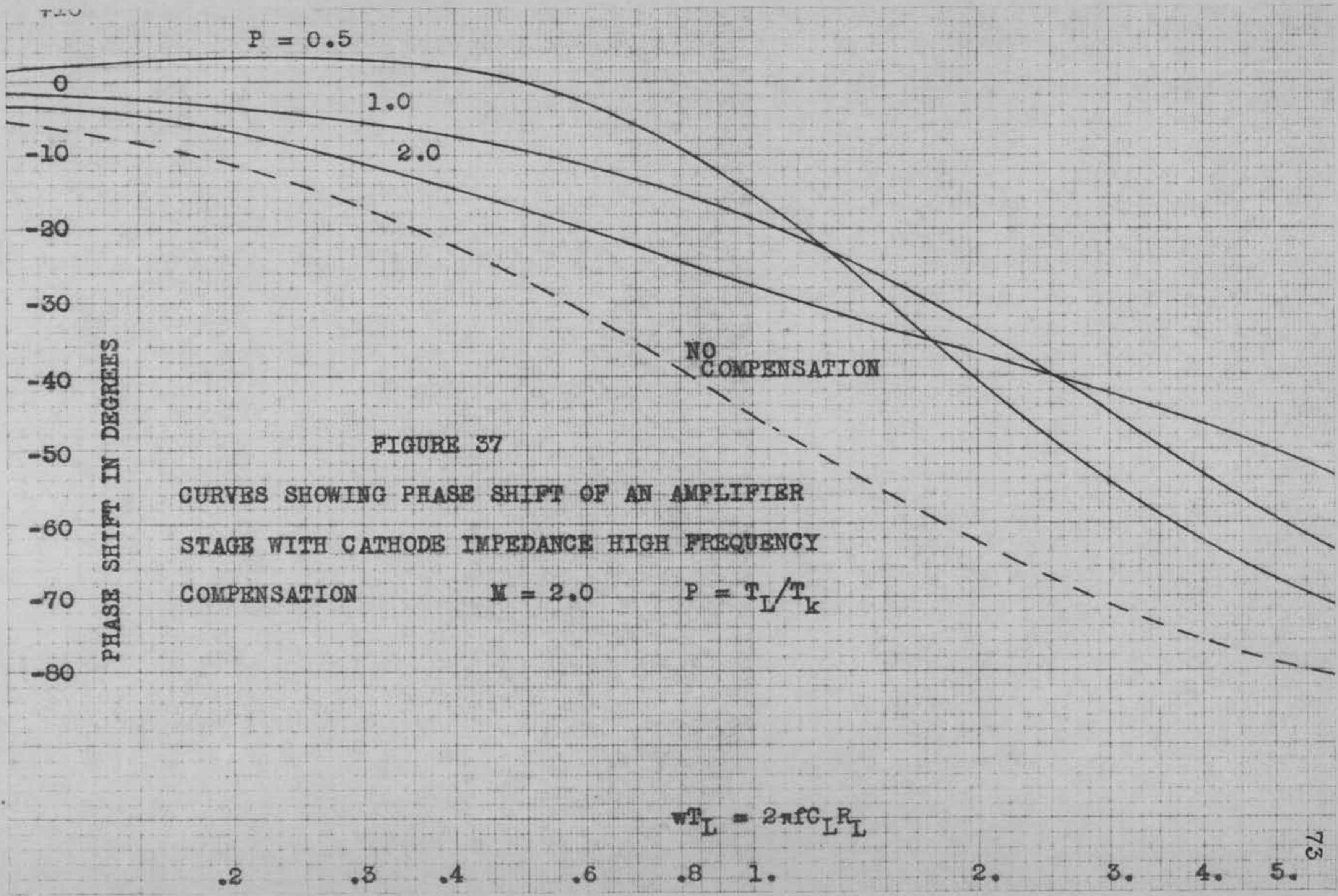


FIGURE 37

CURVES SHOWING PHASE SHIFT OF AN AMPLIFIER STAGE WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION $M = 2.0$ $P = T_L/T_k$

particular frequency is impossible. Troubles due to insufficient damping of the L-C-R circuit of the plate load impedance can not exist if there is no inductance because such a circuit cannot oscillate.

THE MEASUREMENT OF HIGH-FREQUENCY PHASE SHIFT IN AMPLIFIERS

METHODS

It is difficult to calculate the phase shift of a multistage amplifier, especially at high frequencies. It is therefore necessary to devise a method or methods for the measurement of negative phase shift at high video frequencies. The problem is somewhat more difficult than the measurement of negative phase shift at low frequencies for a number of reasons as will now be explained.

Figure (13) shows the circuit diagram used for the measurement of negative phase shift at low frequencies. This is essentially the circuit proposed in reference (29) for the measurement of negative phase shift at high frequencies. At first glance, it may seem that this circuit should be satisfactory for the measurement of high-frequency phase shift by only changing the values of the series type positive phase-shifting bridge for operation at high frequencies but this is not the case.

Calculation from the equation that shows the phase shift of the positive phase-shifting bridge (figure 12)

$$\theta = (+) \arctan 1/wRC, \qquad (17)$$

shows that for a value of f in the order of megacycles, the time constant RC must have a very small value for values of θ lying between 5 and 70 degrees which is the practical range of this variable. The resistance R in the positive phase-shifting bridge, should be very large in comparison with the plate load resistance of the output tube of the amplifier. This is to avoid changing the effective load on this tube. If R_L, the plate load resistance, is of the order of 2,000 ohms, the value of R in the bridge should be at least ten times this value or 20,000 ohms. Substitution in the phase shift equation for the positive phase-shifting bridge gives a value of C which is physically unrealizable with ordinary apparatus because it is so small.

The placing of this bridge in the amplifier circuit changes the effective capacitance to ground of the last stage of the amplifier and thus changes the phase shift of this stage. These considerations were realized when measurements were attempted using different values of R comparable with the plate load resistance of the last stage of the test amplifier. The values obtained were very erratic and varied with different values of R. This method was then dropped in favor of another.

The method which was finally adopted for use in this investigation is the one which is described in reference (17). It consists of the application of the input and the output voltages of the amplifier to be tested to the horizontal and vertical deflection plates of an oscilloscope. The resulting ellipse is then photographed together with superimposed horizontal and vertical deflection lines to establish a reference axis. The phase angle between the two voltages is then found from the relations given in figure (38). In figure (38) is shown a drawing of an ellipse together with a reference axis. The sine of the angle between the two voltages is found by dividing the length CD, the distance between the two intercepts on the Y axis by the length AB which is the total vertical height of the rectangle enclosing the ellipse.

The complete diagram of the circuit which was used to measure high-frequency phase shift is given in figure (39). The voltage output of the auxiliary amplifier had to be of sufficient amplitude to give an adequate deflection on the cathode-ray tube. The input to the test amplifier had to be attenuated by a voltage divider as shown. It was necessary to use a resistance-capacitance voltage divider to attenuate this voltage. It may be shown that if the product C_1R_1 is equal to the product C_2R_2 (see figure 39), then the attenuation of the voltage divider will be constant at all frequencies (21) and the phase shift will be zero. Then the phase of the voltage applied to the horizontal deflection plates of the oscilloscope and the voltage applied to the grid of the type 802 is the same. Under

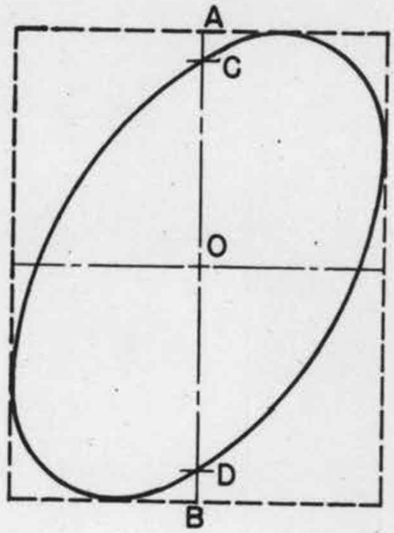


Figure 38.

SHOWING METHOD OF FINDING

PHASE SHIFT FROM OSCILLOGRAM

$$\text{sine } \theta = \frac{CD}{AB}$$

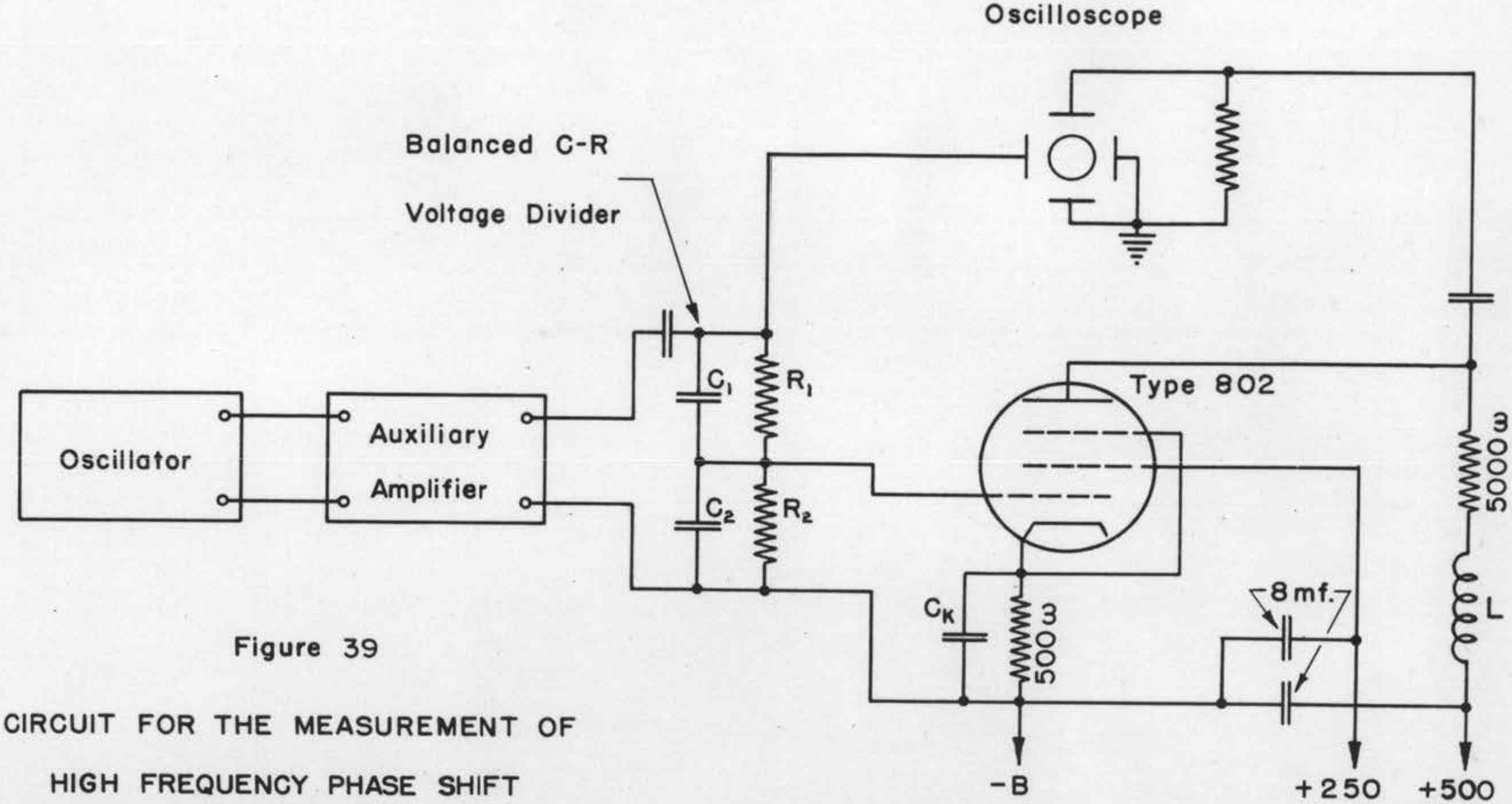


Figure 39

CIRCUIT FOR THE MEASUREMENT OF

HIGH FREQUENCY PHASE SHIFT

these conditions, the only phase shift indicated by the ellipse on the oscilloscope screen is the phase shift in the test amplifier.

The high-frequency oscillator used in this work was a type 605-B Standard Signal Generator made by General Radio Company. This oscillator provides a source of high frequencies from 10 kilocycles to 50 megacycles. The output is calibrated from 1.0 microvolt to 0.1 volt with a constant 1.0 volt output available. See figure (40) which shows a general view of the laboratory apparatus used in the high-frequency phase shift measurements. Gain measurements were made with this calibrated source and the output measured with a model 1252 Triplett vacuum-tube voltmeter which has the tube mounted on the end of the probe for minimum input capacitance.

The oscilloscope used in the high-frequency phase shift measurements was constructed in the laboratory especially for this type of work. It used a cathode-ray tube with a viewing screen 5 inches in diameter. The screen was of the short persistance type and gave a blue fluorescent light especially suited for photography. The tube (RCA-907) was fitted with separate connections to each of the four deflection plates. When operated with 2,000 volts on the accelerating anode, it was possible to secure a very sharp and brilliant trace which photographed very

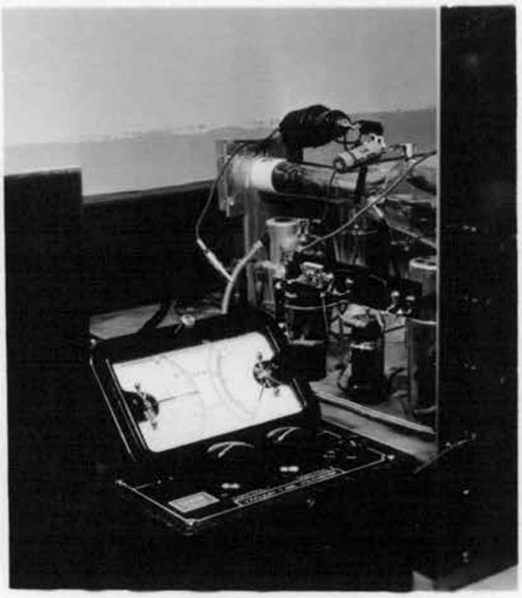


FIGURE 40

APPARATUS USED IN THE MEASUREMENT OF HIGH FREQUENCY PHASE SHIFT WITH CLOSE VIEW well. Using an F/4.5 camera with super-pancro-press cut film, it was found that excellent exposures were obtained at 1/25 of a second.

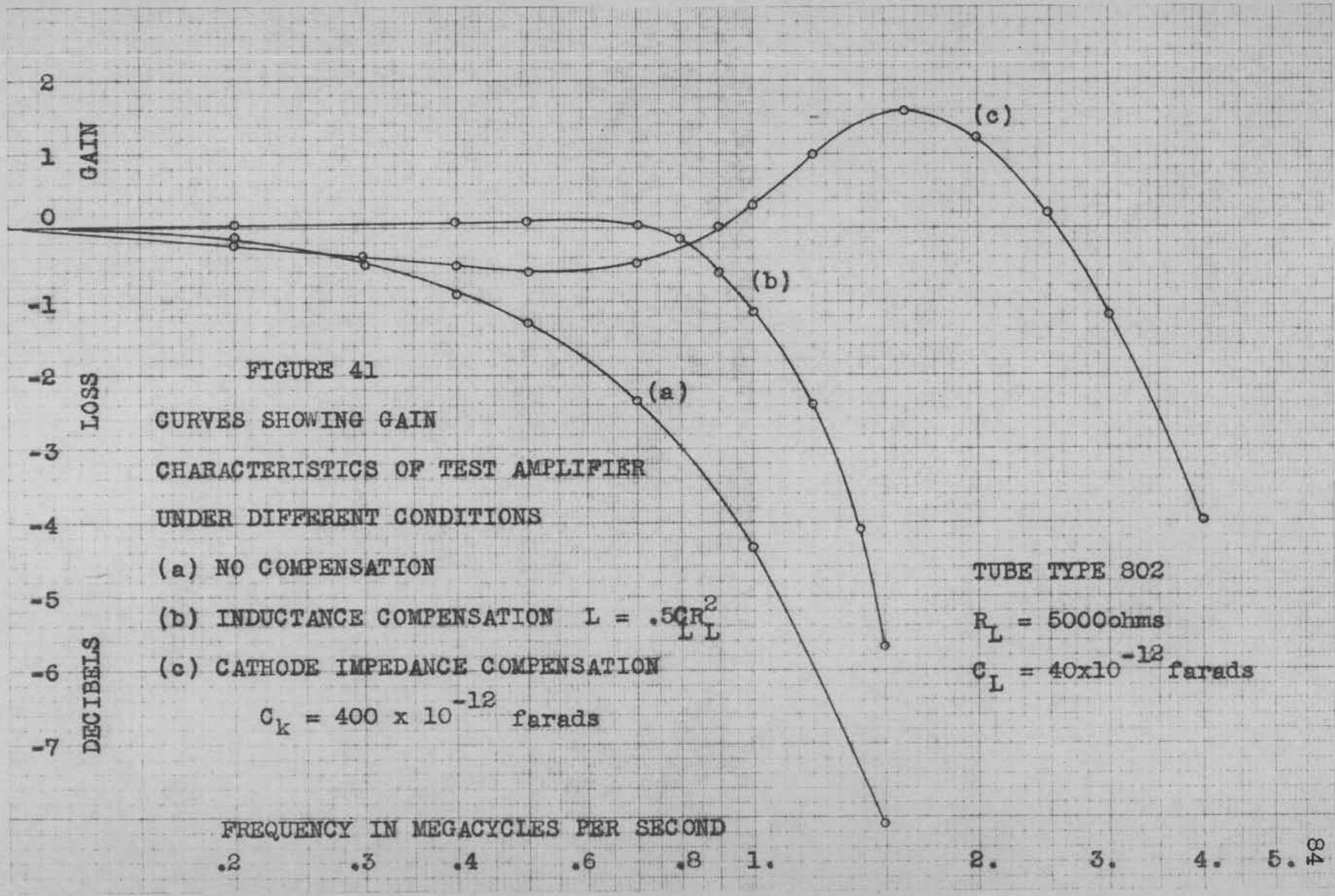
The experimental amplifier itself was built into the oscilloscope used in the measurements. It consisted of a type 802 power pentode with a plate load resistance of 5,000 ohms. The bias was provided by an unbypassed cathode resistor of 500 ohms resistance. The plate supply was in the neighborhood of 500 volts with 250 volts applied to the screen grid of the tube. This type tube has a nominal transconductance of about 2,000 micromhos.

THE MEASUREMENT OF HIGH-FREQUENCY
PHASE SHIFT IN AMPLIFIERS

RESULTS AND CONCLUSIONS

Three sets of data were obtained for three different conditions of the test amplifier. The first condition was with no high-frequency compensation and an unbypassed cathode resistor. The results are illustrated in figures (41) and (42) curves (a). From the gain curve in figure (41) the value of the plate load shunting capacitance was calculated. This capacitance was calculated from the fact that the gain is down 3 decibels at the frequency where the plate load resistance is equal in magnitude to the reactance of the shunt capacitance. The frequency where the gain was down 3 decibels was found to be 800 kilocycles in this case. The stray capacitance was calculated to be 40 micromicrofarads. Figure (43) shows the oscillogram from which the phase shift was calculated for the uncompensated amplifier.

The oscillograms of figures (43) and (44) were made at three different frequencies which were determined by the gain curve of the uncompensated amplifier. Frequency (1) of 430 kilocycles was the point where the uncompensated amplifier was down 1.0 decibel in gain; frequency (2) was at the point where the amplifier was down 3 decibels (800 kilocycles) and frequency (3) of 1,300 kilocycles was at


2
1
0
-1
-2
-3
-4
-5
-6
-7
GAIN
LOSS
DECIBELS
(a)
(b)
(c)
FIGURE 41
CURVES SHOWING GAIN
CHARACTERISTICS OF TEST AMPLIFIER
UNDER DIFFERENT CONDITIONS
(a) NO COMPENSATION
(b) INDUCTANCE COMPENSATION $L = .5C_LR_L^2$
(c) CATHODE IMPEDANCE COMPENSATION
$C_k = 400 \times 10^{-12}$ farads
TUBE TYPE 802
R_L = 5000ohms
$C_L = 40\times10^{-12}$ farads
FREQUENCY IN MEGACYCLES PER SECOND
.2
.3
.4
.6
.8
1.
2.
3.
4.
5.

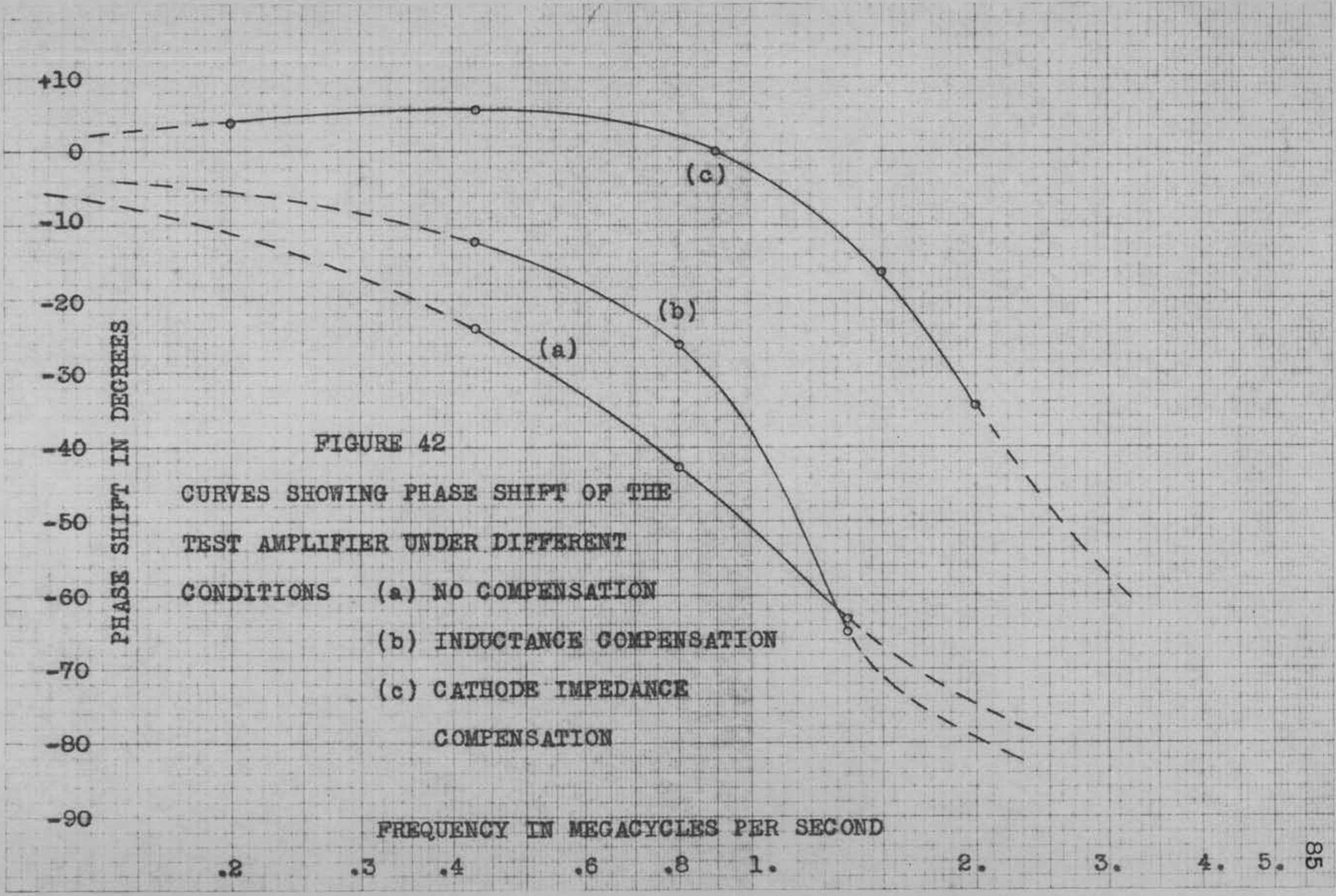


FIGURE 42

CURVES SHOWING PHASE SHIFT OF THE TEST AMPLIFIER UNDER DIFFERENT CONDITIONS (a) NO COMPENSATION (b) INDUCTANCE COMPENSATION (c) CATHODE IMPEDANCE COMPENSATION

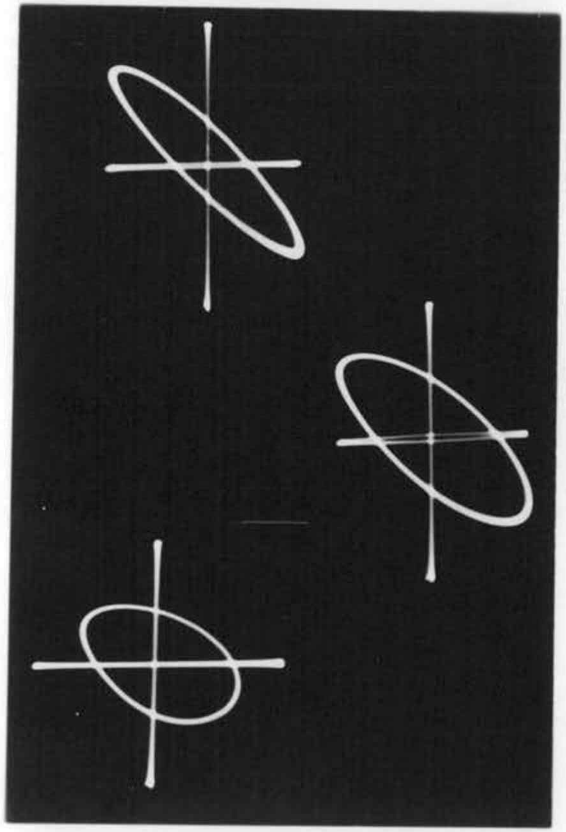


FIGURE 43

PHASE SHIFT OF UNCOMPENSATED AMPLIFIER

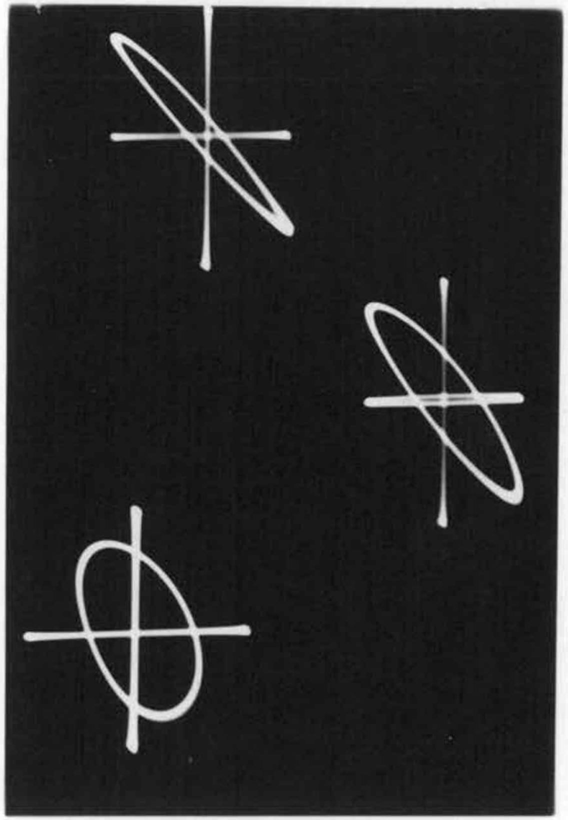


FIGURE 44

PHASE SHIFT OF INDUCTANCE COMPENSATED AMPLIFIER the point where the amplifier was down 7 decibels.

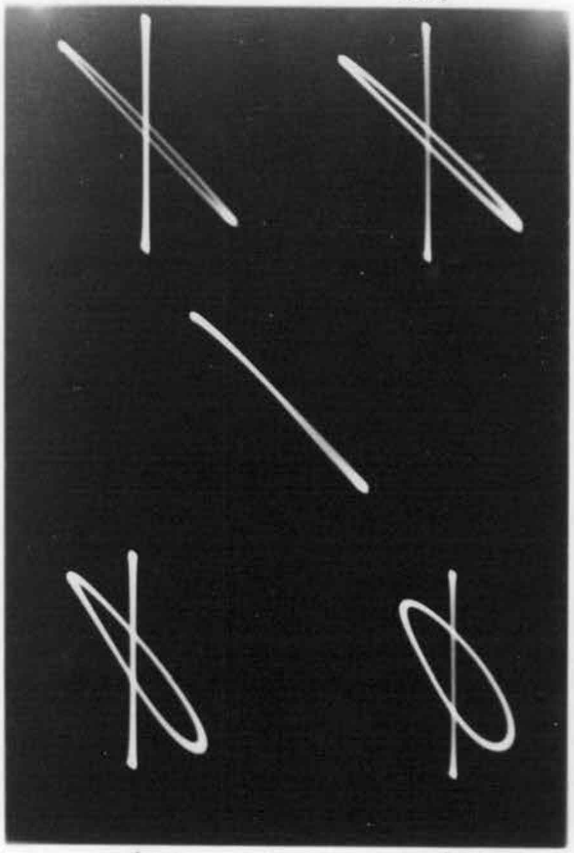
Test condition number two was the use of inductance compensation utilizing a small inductance in series with the plate load resistor. The value of the inductance was chosen so that the gain would be approximately flat to 800 kilocycles (reference (28) and reference (26) page 192). This value is given by the equation $L = 0.5\, C_L R_L^2$ where C_L and R_L are the plate load constants in farads and ohms and L is the required inductance in henries. The results of this test are illustrated in figures (41) and (42) curves (b). Figure (44) shows the oscillogram from which the phase shift curve was calculated.

The third condition was designed to show the effect of bypassing the cathode bias resistor with a small capacitor. A small mica capicitor of 400 micromicrofarads was connected in parallel with the cathode biasing resistor of the tube. The results are shown in fugures (41) and (42) curves (c). The oscillogram from which the phase shift was calculated is shown in figure (45). In this case, five different ellipses are shown on the oscillogram. These represent phase shift measurements taken at five different frequencies which are listed at the bottom of figure (45).

It will be noted that the phase shift is positive

(1) (2)

(3)



(4) (5)

FIGURE 45

SHOWING PHASE SHIFT OF AMPLIFIER WITH CATHODE IMPEDANCE HIGH FREQUENCY COMPENSATION

(1) 0.2 megacycles (3) 0.9 megacycles

(2) 0.43 megacycles (4) 1.5 megacycles

(5) 2.0 megacycles

from about 100 kilocycles to 900 kilocycles where the curve crosses the zero axis and the phase shift becomes negative. This is not a desirable condition since the positive phase shift means that the requirement of constant delay is not attained in the amplifier.

In all high-frequency measurements, the results become more and more uncertain as the frequency is increased. Thus the measurement of gain in any high-frequency amplifier is subject to various sources of error. One of the most important sources of error is the effect of the input capacitance of the device used to measure the gain characteristics. In this investigation, use was made of a vacuum tube voltmeter whose input capacitance was of the order of 10 to 12 micromicrofarads. Although this is relatively small, it has a large effect at frequencies of one or two megacycles, especially when in shunt with a circuit whose stray capacitance is only about 35 to 40 micromicrofarads.

The ideal vacuum-tube voltmeter for use in this type of measurement would be one using a type RCA-954 acorn tube which has an input capacitance of only 3 micromicrofarads. The tube should be mounted on the end of the test probe so that the grid lead from the top of the tube could be connected directly to the point of measurement.

TESTING OF THE TRANSMISSION CHARACTERISTICS OF AMPLIFIERS BY MEANS OF SQUARE WAVES

The use of square waves for the testing of amplifiers is comparitively new. It provides a means of quickly estimating the type and amount of frequency and phase distortion in an amplifier. (35)(39).

The generation of square waves of sufficiently steep sides and flat top is not particularly difficult and the square wave generator used in this investigation is treated in Appendix (5). It may be shown (25) that the mathematical expression for a perfectly symmetrical flat top wave of the type used to test amplifiers is given by the expression

$$e = \frac{4E}{\pi}(\sin wt + 1/3 \sin 3wt + 1/5 \sin 5wt + 1/7 \sin 7wt \quad -----),$$

where e = voltage at any instant,

E = peak amplitude of square wave,

$w = 2\pi f$,

t = time in seconds

The square wave is seen to be composed of the fundamental frequency together with all of the odd harmonics. The amplitude of the harmonics are inversely proportional to their harmonic frequency.

The relationship between the transient response of an amplifier and its response to a square wave is very close. The transient response of an amplifier may be defined as the response resulting from a sudden application of a voltage which is thereafter maintained constant. The transient response of an amplifier is limited by two effects; the ability of the amplifier to transmit first, the high frequencies and second, the low frequencies. (6). The response of the amplifier to the application of the voltage is determined initially by its high-frequency characteristics and later by its low-frequency characteristics.

A square wave is a succession of suddenly applied voltage impulses and thus it serves to test the transient characteristics of an amplifier. All amplifiers have a certain band of frequencies which they transmit with constant gain and with a phase shift that is proportional to frequency (constant delay). If a square wave whose fundamental frequency and harmonics up to the twenty-first fall within this band, the wave shape will be preserved through the amplifier and, for all practical purposes, the output voltage will be a perfect square wave.

As the fundamental frequency is reduced, the low-frequency transmission characteristics become the determining factor and the voltage output is no longer a square wave. As the frequency of the square wave becomes higher, the high-frequency response of the amplifier determines the wave shape of the output. Thus it is necessary to test an amplifier with at least two different frequencies of square waves; one to test the low-frequency response and another to test the high-frequency end of the band.

The low-frequency characteristics of the amplifier may be determined by observing the average shape of the top of the output wave when a square wave is applied to the input. An uncompensated amplifier has a loss of gain and a phase delay that is less than normal (positive phase shift) at low frequencies. The top of the output wave when a low-frequency square wave is impressed on the input has a definite slope from left to right if the delay is below normal and a slope with a curve in it if there is attenuation. See figure (46).

As the frequency of the square wave is increased, the high-frequency transmission characteristics come into effect. If the region of impaired high and low-frequency response are close together, the analysis of the output wave becomes more complicated. However, in wide-band amplifiers, with which this paper is concerned, this complication does not arise. In the ordinary inductance compensated amplifier, when too much inductance is placed in the plate lead, the amplifier will have excess gain at the high-frequency end of the band. This will show up in the square wave test by the presence of an initial "over shoot" of the wave and possibly by small superimposed oscillatory transients on the top of the output wave.

TESTING OF THE TRANSMISSION CHARACTERISTICS OF AMPLIFIERS BY MEANS OF SQUARE WAVES

RESULTS AND CONCLUSIONS

In figure (46) there are two photographs on the same print. The one at the top (a) is the trace on the cathode-ray tube showing the 30-cycle square wave input to the amplifier. The trace on the bottom (b) shows the output voltage resulting after passing the square wave through an amplifier which has a good deal of attenuation and positive phase shift at 30 cycles.

If the amplifier is incorrectly compensated or over compensated due to a poor design of the low-frequency compensation networks, the wave top of the square wave may have a large variety of shapes at the output of the amplifier. In figure (47) is shown two oscillograms which show the output voltage wave from an amplifier whose various phase and gain characteristics are given in figures (48) and (49). These oscillograms illustrate a variety of conditions which are undesirable. Curves (2) of figures (48) and (49) are the transmission characteristics of an amplifier which is badly over compensated and this amplifier gives the output voltage wave illustrated by (c) of figure (47). Curves (3) of figures (48) and (49) illustrate the phase shift and gain characteristics of the same amplifier

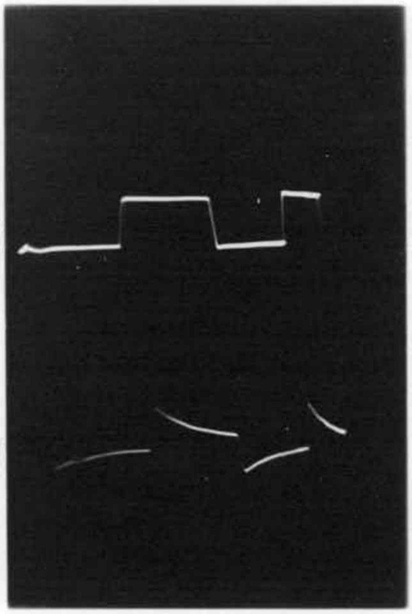


FIGURE 46

(a) 30-CYCLE SQUARE WAVE INPUT TO AMPLIFIER

(b) OUTPUT OF AMPLIFIER WITH ATTENUATION AND POSITIVE PHASE SHIFT AT 30 CYCLES

HORIZONTAL DISTORTION DUE TO NON-LINEAR TIMING WAVE

(a) (b) (c) (d)

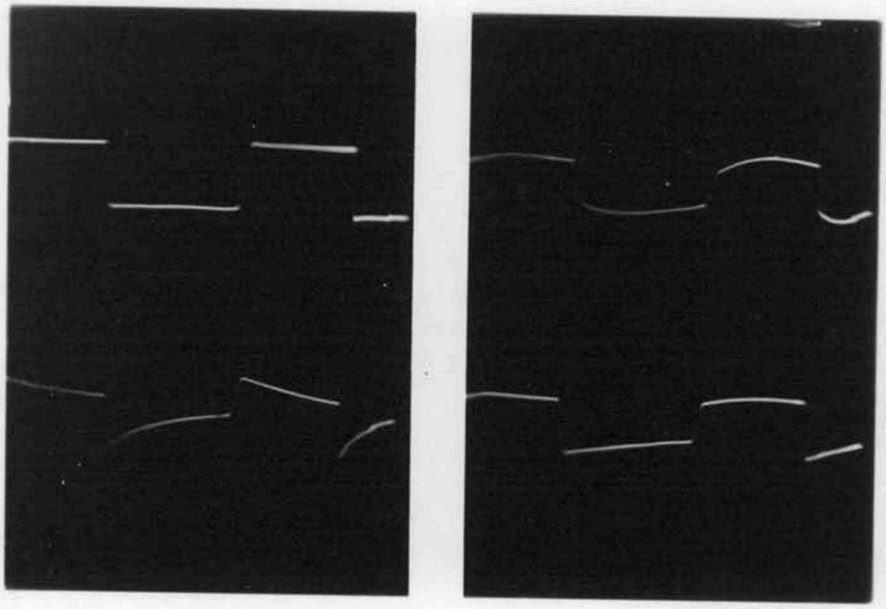


FIGURE 47

(a) 30-CYCLE SQUARE WAVE; INPUT TO AMPLIFIER

(b) OUTPUT OF AMPLIFIER. SEE CURVES (1) FIGURES (48) and (49)

(c) OUTPUT OF AMPLIFIER. SEE CURVES (2) FIGURES (48) and (49)

(d) OUTPUT OF AMPLIFIER. SEE CURVES (3) FIGURES (48) AND (49)

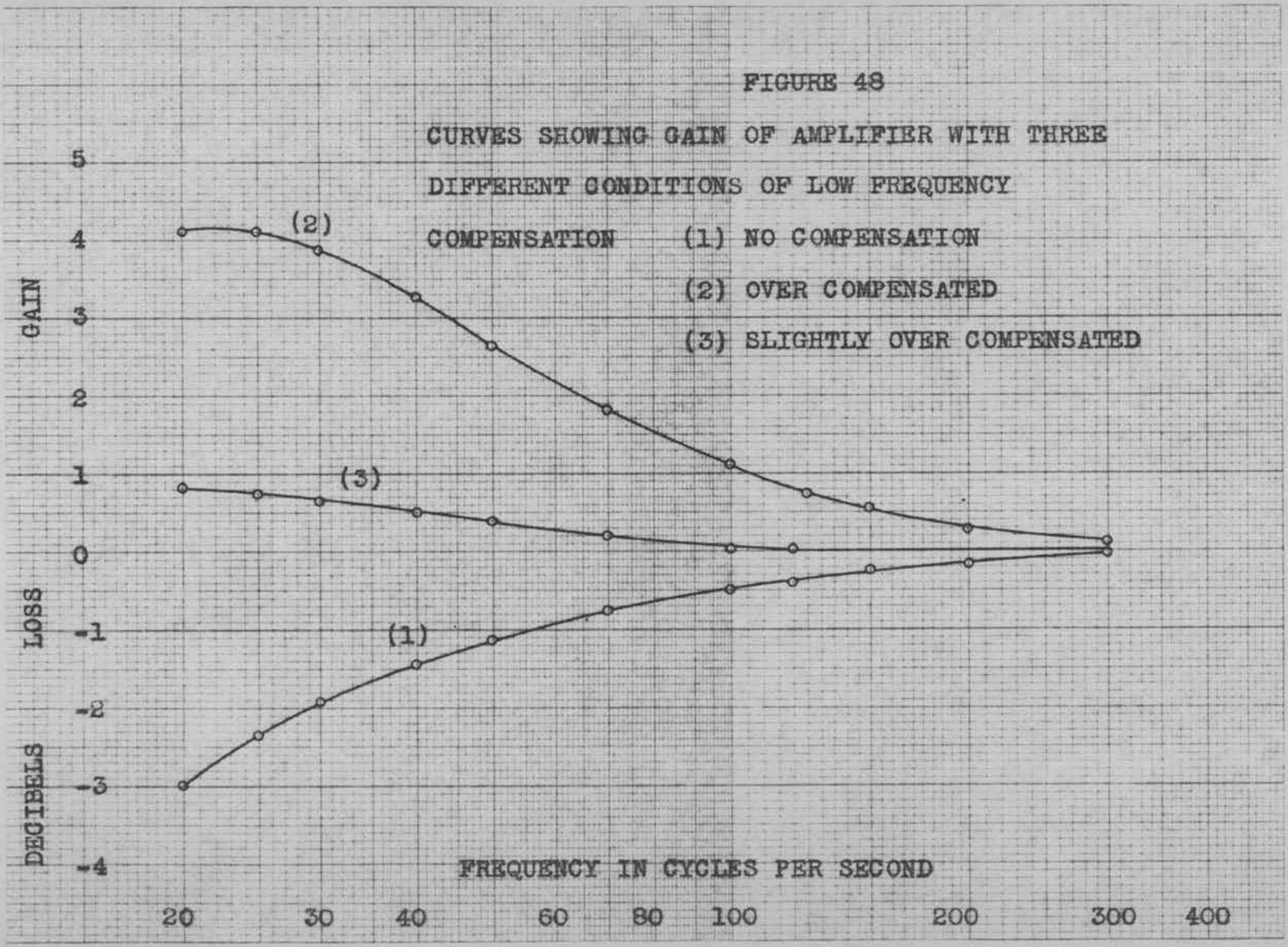


FIGURE 48
CURVES SHOWING GAIN OF AMPLIFIER WITH THREE
DIFFERENT CONDITIONS OF LOW FREQUENCY
COMPENSATION
(1) NO COMPENSATION
(2) OVER COMPENSATED
(3) SLIGHTLY OVER COMPENSATED
(1)
(2)
(3)
5
4
3
2
1
0
-1
-2
-3
-4
DECIBELS LOSS GAIN
FREQUENCY IN CYCLES PER SECOND
20
30
40
60
80
100
200
300
400

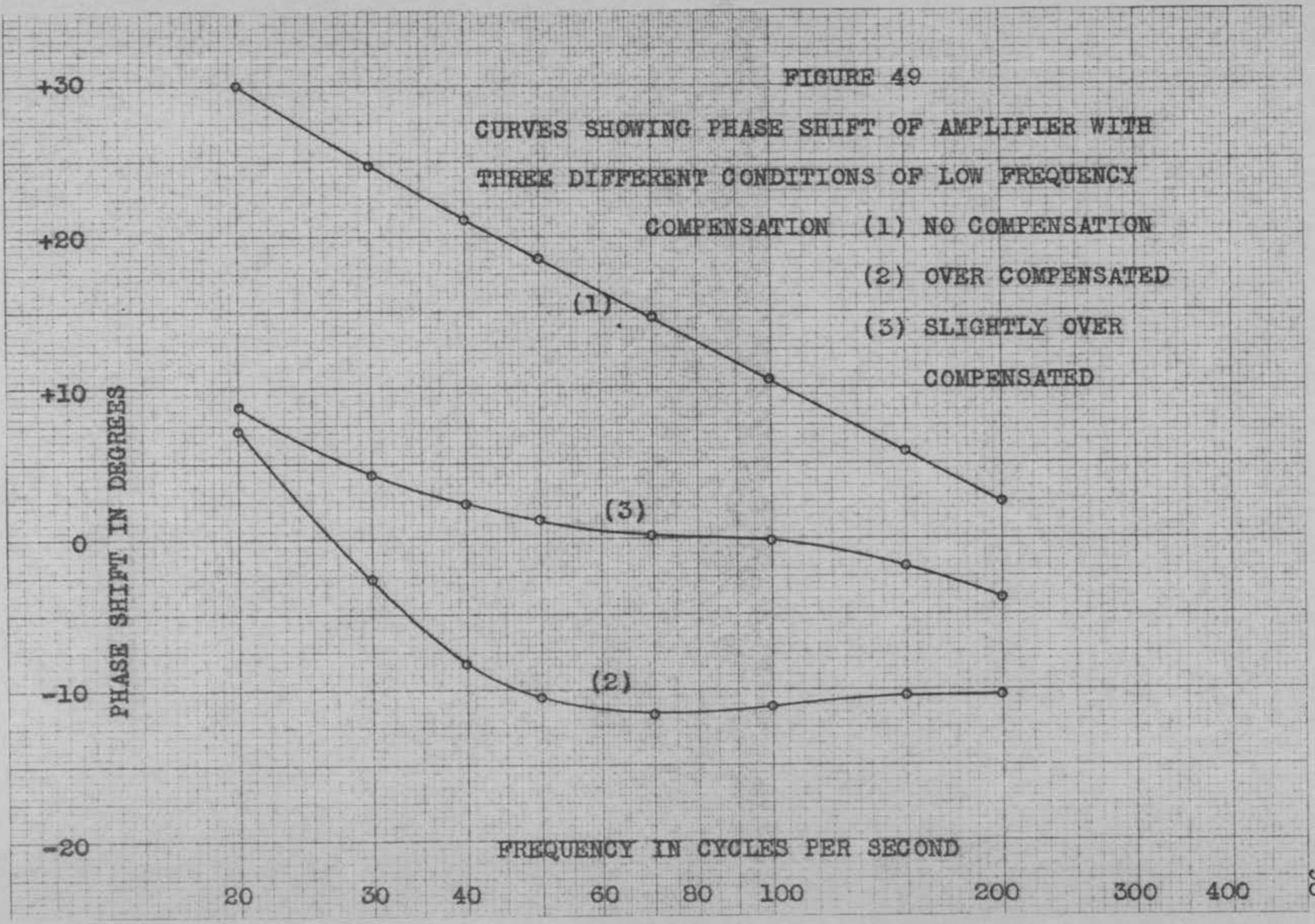


FIGURE 49

CURVES SHOWING PHASE SHIFT OF AMPLIFIER WITH THREE DIFFERENT CONDITIONS OF LOW FREQUENCY COMPENSATION (1) NO COMPENSATION (2) OVER COMPENSATED (3) SLIGHTLY OVER COMPENSATED with a much better low-frequency compensation network. However, this still leaves much to be desired since the phase shift is negative at about 100 cycles and the gain is still rising at 20 cycles. This amplifier gives the output waveform (d) of figure (47).

When the frequency of the input square wave is high enough to be affected by the high-frequency response of the amplifier, the first thing that is noticed is a rounding of the diagonal corners of the output wave. See (b) of figure (50). As the frequency becomes higher, in the ordinary uncompensated amplifier, the corners become more and more rounded and the wave is greatly distorted. See (a) of figure (50).

For the testing of amplifiers to be used as video amplifiers in television, square waves of two frequencies are sufficient. Since the gain and phase characteristics of these amplifiers should be good down to 30 cycles per second, the low-frequency square wave should be 30 cycles per second. The amplifier should not appreciably distort this wave if it is properly designed and adjusted. For the high-frequency end of the band, it is necessary to use a square wave of a fundamental frequency of 100 kilocycles per second or more. The amplifier should be able to pass a 100 kilocycle square wave with absolutely no distortion

(a)

(b)

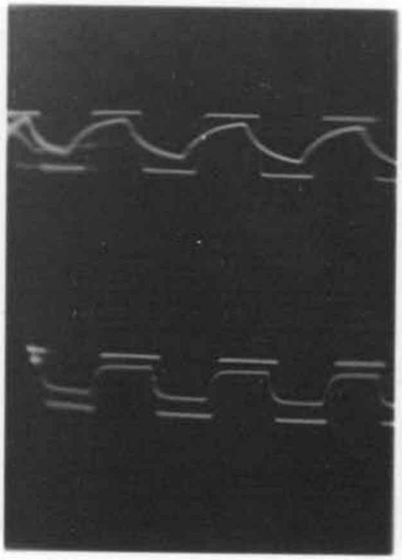


FIGURE 50

(a) SQUARE WAVE INPUT SUPERIMPOSED ON THE OUTPUT OF AN AMPLIFIER WITH HIGH ATTENUATION AND LARGE NEGATIVE PHASE SHIFT AT HIGH FREQUENCIES

(b) SQUARE WAVE INPUT SUPERIMPOSED ON THE OUTPUT WAVE OF AN AMPLIFIER WITH SOME ATTENUATION AND NEGATIVE PHASE SHIFT AT HIGH FREQUENCIES if it is to be any good for high definition television picture reception. No distortion of a 100 kilocycle wave means that the amplifier is flat and has constant delay to about 2 megacycles which is 20 times the fundamental frequency of the testing square wave.

APPENDIX

APPENDIX I

DERIVATION OF THE GENERAL EQUATION FOR THE PHASE SHIFT AND GAIN CHARACTERISTICS OF PLATE CIRCUIT LOW-FREQUENCY COMPENSATION NETWORKS

Figure (51) shows the equivalent circuit for the derivation of the phase and gain characteristics of a low-frequency compensation circuit inserted in series with the plate load resistor of high transconductance pentode tubes.

Let E_o = output voltage at low frequencies,

E'_o = output voltage at high frequencies,

E_i = input voltage,

R_L = plate load resistance,

R_p = dynamic plate resistance,

R_c = low frequency compensation network resistance,

C_c = low frequency compensation network capacitance,

$T_c = R_cC_c$ time constant of the low frequency compensation network,

$w = 2\pi f$,

f = frequency in cycles per second, and

$Z = R_c/(1 + jwT_c)$ impedance of the low frequency compensation network.

The impedance of the condenser C_c is zero at very high frequencies and the output voltage can be expressed

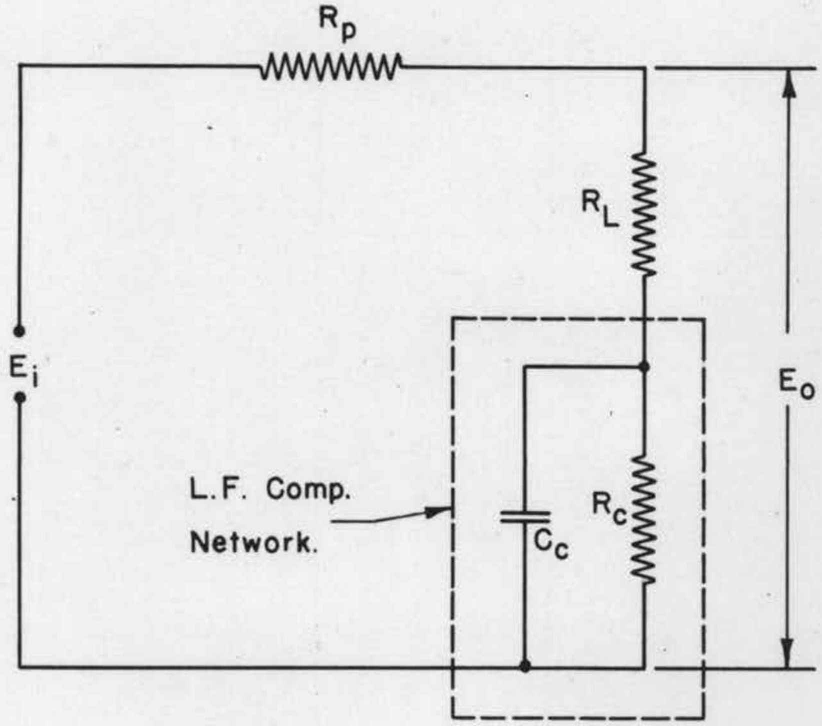


Figure 51.

CIRCUIT DIAGRAM FOR THE MATHEMATICAL DERIVATION OF PHASE AND GAIN CHARACTERISTICS OF A LOW FREQUENCY COMPENSATION NETWORK by the equation

$$E_o' = E_i R_L/(R_p + R_L). \qquad (1)$$

The output voltage at low frequencies can be expressed by the equation

$$E_o = \frac{E_i(R_L + Z)}{R_p + R_L + Z} \qquad (2)$$

The response of the circuit may be expressed as the ratio between the output voltage at low frequencies and the output voltage at high frequencies.

$$\text{Response ratio} = E_o/E_o' = \frac{(Z + R_L)(R_p + R_L)}{(Z + R_p + R_L)(R_L)} \qquad (3)$$

Making the assumption that R_p is very much larger than R_L (3) can be written

$$\text{Response ratio} = \frac{(Z + R_L)(R_p)}{(Z + R_p)(R_L)} \qquad (4)$$

Substituting the value of Z in the above equation

$$\text{Response ratio} = \frac{R_p(R_c + R_L(1 + jwT_c))}{R_L(R_c + R_p(1 + jwT_c))} \qquad (5)$$

Making the assumption that R_p is very much larger than R_c in equation (5) and dividing out R_p gives the equation

$$\text{Response ratio} = \frac{R_c/R_L + (1 + jwT_c)}{1 + jwT_c} \qquad (6)$$

Letting the ratio $R_c/R_L = n$ (a parameter) the final equation is

$$\text{Response ratio} = \frac{(n + 1) + jwT_c}{1 + jwT_c} \qquad (7)$$

This is the final equation from which the curves of figures (5) and (6) were plotted. The assumptions made in the derivation are valid when dealing with high transconductance pentodes used in television amplifiers, where R_L and R_c are always only a small fraction of the value of R_p.

APPENDIX II

DERIVATION OF THE GENERAL PARAMETRIC EQUATION FOR THE PHASE AND GAIN CHARACTERISTICS OF A NEGATIVE FEED BACK AMPLIFIER AT LOW FREQUENCIES

Figure (22) shows the circuit diagram for the mathematical analysis of the phase and gain characteristics of a negative feedback amplifier.

Let A_o = mid-band amplification of amplifier with no feedback,

A = complex amplification factor of the amplifier at low frequencies without feedback,

β_o = feedback at mid-band frequencies,

β = complex expression for feedback at low frequencies,

$T_g = C_g R_g$ = time constant of grid coupling circuit,

$T_\beta = C_\beta (R_k + R_\beta)$ time constant of feedback circuit,

R_k = cathode resistor of first tube (part of feedback circuit),

E_i = input voltage of amplifier,

E_o = output voltage of amplifier,

$p = T_\beta / T_g$ (a parameter),

$d = A_o \beta_o$ feedback factor (a parameter), and

$w = 2\pi f$.

It may easily be shown that the complex amplification factor of the amplifier without feedback shown in figure

(22) is

$$A = A_o/(1 - j/wT_g). \qquad (1)$$

It may also be seen that the feedback ratio at low frequencies is given by the expression

$$\beta = \frac{R_k}{R_k + R_\beta - j/wC_\beta} \qquad (2)$$

Dividing through by the quantity $(R_k + R_\beta)$ gives

$$\beta = \frac{R_k/(R_k + R_\beta)}{1 - j/wT_\beta} \qquad (3)$$

But, since $\beta_o = R_k/(R_k + R_\beta)$ and substituting this in the above expression gives

$$\beta = \frac{\beta_o}{1 - j/wT_\beta} \quad . \qquad (4)$$

Substituting equation (1) and (4) in the familiar feedback equation (4) which is

$$\text{Gain} = E_o/E_i = \frac{A}{1 - A\beta} \qquad (5)$$

gives the expression

$$\text{Gain} = \frac{A_o/(1 - j/wT_g)}{1 + \dfrac{A_o \qquad \beta_o}{(1-j/wT_g)(1-j/wT_\beta)}}. \qquad (6)$$

The sign is positive since β is negative for negative feed back. Simplifying the above equation gives the equation

$$\text{Gain} = \frac{A_o wT_g(wT_\beta - j1)}{T_gT_\beta(w)^2(1 + A_o\beta_o) - 1 - jw(T_g + T_\beta)} \qquad (7)$$

The gain of the amplifier at mid-band frequency is given by the equation

$$\text{Gain(mid-band)} = A_o/(1 + A_o\beta_o). \qquad (8)$$

The relative response of the amplifier is given by the ratio between the gain at low frequencies and the gain at mid-band frequencies.

$$\text{Response ratio} = \frac{\text{Equation (7)}}{\text{Equation (8)}} \qquad (9)$$

Simplification results in the expression

$$\text{Response ratio} = \frac{wT_g(wT_\beta - j1)(1 + A_o\beta_o)}{T_gT_\beta(w)^2(1 + A_o\beta_o) - 1 - jw(T_g + T_\beta)} \qquad (10)$$

Dividing through by T_g and multiplying through by T_β and substituting the parameters

$$p = T_\beta/T_g \qquad \text{and} \qquad d = A_o\beta_o$$

gives the equation

$$\text{Response ratio} = \frac{wT_\beta(wT_\beta - j1)(1 + d)}{(wT_\beta)^2(1 + d) - p - j(p+1)wT_\beta} \qquad (11)$$

Dividing through by the quantity (1+d) gives the final equation

$$\text{Response ratio} = \frac{wT_\beta(wT_\beta - j1)}{(wT_\beta)^2 - p/(1+d) - jwT_\beta(1+p)/(1+d)} \qquad (12)$$

APPENDIX III

DERIVATION OF THE EQUATION FOR THE PHASE SHIFT OF A PARALLEL TYPE NEGATIVE PHASE SHIFTING BRIDGE.

Figure (10) shows the circuit diagram for the parallel type negative phase shifting bridge used in this investigation. The following gives the derivation of the expression for the phase shift of this bridge.

Let E_i = input voltage (reference vector),

E_o = output voltage,

R_1 = series resistance,

R_2 = resistance in parallel with the capacitance in the bridge,

C = parallel capacitance,

Z = impedance of parallel circuit consisting of C and R_2,

T = CR_2 = time constant of parallel circuit,

w = $2\pi f$, and

f = frequency in cycles per second.

The current which flows in the bridge is given by the equation

$$i = E_i/(R_1 + Z). \qquad (1)$$

The output voltage is given by

$$E_o = iZ = E_iZ/(R_1 + Z). \qquad (2)$$

But

$$Z = R_2/(1+jwT). \tag{3}$$

Substituting equation (3) in equation (2) gives, after simplification,

$$E_o = \frac{E_i R_2}{(R_1 + R_2) + jR_1 wT} \tag{4}$$

If E_i is chosen to the reference vector with a phase angle of zero, then it immediately apparent from the equation (4) that the phase angle of E_o is

$$\theta = (-) \arctan R_1 wT/(R_1+R_2), \tag{5}$$

or sutbstituting for T its value R_2C gives the final equation

$$\theta = (-) \arctan wCR_1R_2/(R_1+R_2). \tag{6}$$

APPENDIX IV

DERIVATION OF THE GENERAL PARAMETRIC EQUATION FOR HIGH-FREQUENCY COMPENSATION BY MEANS OF CATHODE IMPEDANCES.

Figure (52) shows the circuit diagram to be used in this derivation.

Let E_i = input voltage,

E_o = output voltage,

E_g = actual voltage on the grid,

E = voltage of equivalent generator in equivalent circuit,

u = amplification factor of tube,

$w = 2\pi f$,

f = frequency in cycles per second,

R_p = dynamic plate resistance of the tube,

R_k = cathode resistance,

C_k = cathode bypass capacitance,

$T_k = R_k C_k$ = time constant of cathode impedance,

R_L = plate load resistance,

C_L = stray capacitance in plate circuit,

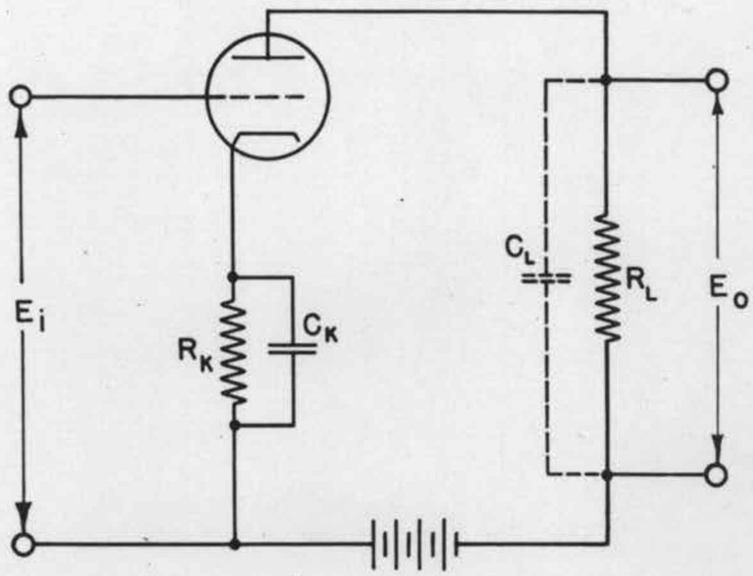
$T_L = C_L R_L$ = time constant of plate load circuit,

$Z_k = R_k/(1+jwT_k)$ cathode impedance,

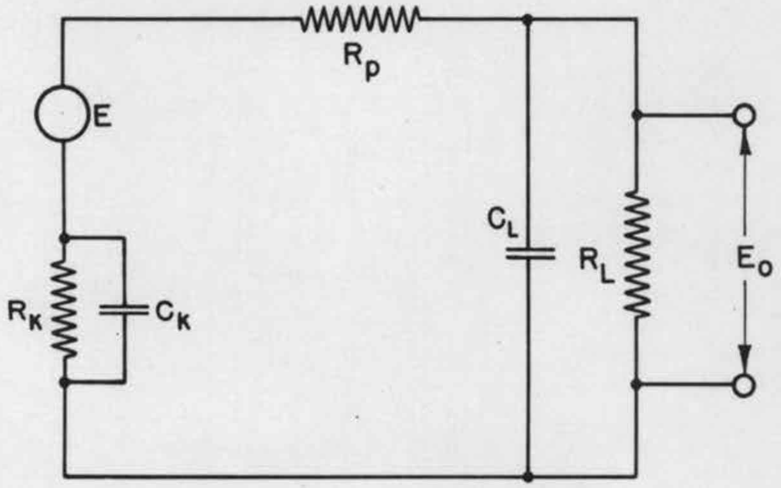
$Z_L = R_L/(1+jwT_L)$ plate load impedance,

j = operator $(-1)^{1/2}$, and

i = current in equivalent circuit.



(a) Actual Circuit.



(b) Equivalent Circuit.

Figure 52

CIRCUIT DIAGRAM SHOWING HIGH FREQUENCY

COMPENSATION BY CATHODE IMPEDANCE

The voltage of the equivalent generator is given by the equation (see figure(52))

$$E = u(E_g - iZ_k) = i(R_p + Z_L + Z_k). \qquad (1)$$

Solving the above expression for "i" and multiplying this value of i by Z_L to get the output voltage of the circuit at high frequencies

$$iZ_L = E_o = \frac{uE_gZ_L}{R_p + Z_L + Z_k(1+u)} \qquad (2)$$

The output voltage of the circuit at low frequencies where the impedance of the capacitances of the circuit are infinite is

$$E'_o = \frac{uE_gR_L}{R_p + R_L + R_k(1+u)}. \qquad (3)$$

The ratio between the output voltage of the circuit at high frequencies and the output voltage at low frequencies is given by

$$\text{Response ratio} = E_o/E'_o = \frac{\text{Equation (2)}}{\text{Equation (3)}}$$

or

$$\text{Response ratio} = \frac{Z_L\ (R_p + Z_L + Z_k(1+u)\ )}{R_L\ (R_p + R_L + R_k(1+u)\ )} \qquad (4)$$

Making the assumptions that (1+u) is practically equal to u and that $(R_p + R_L)$ is practically equal to R_p, equation (4) reduces to the equation

$$\text{Response ratio} = \frac{Z_L(R_p + Z_L + Z_k u)}{R_L(R_p + R_k u)} \tag{5}$$

Substituting the values for Z_L and Z_k in equation (5) and simplifying

$$\text{Response ratio} = \frac{(R_p + uR_k)(1 + jwT_k)}{R_p(1+jwT_L)(1+jwT_k)+R_L(1+jwT_k)+uR_k(1+jwT_L)} \tag{6}$$

Dividing through by R_p and letting the parameter $M = uR_k/R_p$,

$$\text{Response ratio} = \frac{(1 + M)(1 + jwT_k)}{(1+jwT_L)(1+jwT_k)+M(1+jwT_L)+(1+jwT_k)R_L/R_p} \tag{7}$$

The error in the final results introduced by the neglect of the factor which is multiplied by the ratio R_L/R_p is negligible when the plate resistance is at least one hundred times the plate load resistance. Dropping this factor the response is

$$\text{Response ratio} = \frac{(1 + M)(1 + jwT_k)}{(1+jwT_L)(1+jwT_k)+M(1+jwT_L)} \tag{8}$$

Dividing through by T_k and multiplying through by T_L and letting the parameters

$$P = T_L/T_k \qquad \text{and} \qquad M = uR_k/R_p$$

The final equation is

$$\text{Response ratio} = \frac{(1 + M)(P + jwT_L)}{(1+jwT_L)(P + PM + jwT_L)} \tag{9}$$

APPENDIX V

A SQUARE-WAVE GENERATOR

The circuit diagram of the square-wave generator used in this investigation is shown in figure (53). A photograph of the apparatus is shown in figure (54). It is essentially a two stage wide-band amplifier with good low and high frequency characteristics. The high-frequency response is attained by means of small plate load resistances. The low-frequency response is good because of the large value of the time constant of the grid to plate coupling circuit.

The generation of the square wave is brought about by applying a very large driving voltage to the grid of the first tube. This grid has a one megohm resistor connected in series with the driving oscillator. This gives the oscillator a very high effective impedance. When the grid is driven positive by the excessive driving voltage, the grid draws current and the voltage drop across the high resistance becomes very large.

The negative half of the oscillator wave drives the input tube of the square wave generator far beyond cut-off and this cuts off the other half of the wave top. The first tube applies this flat top wave to the grid of the second tube and it, in turn, is driven positive and beyond

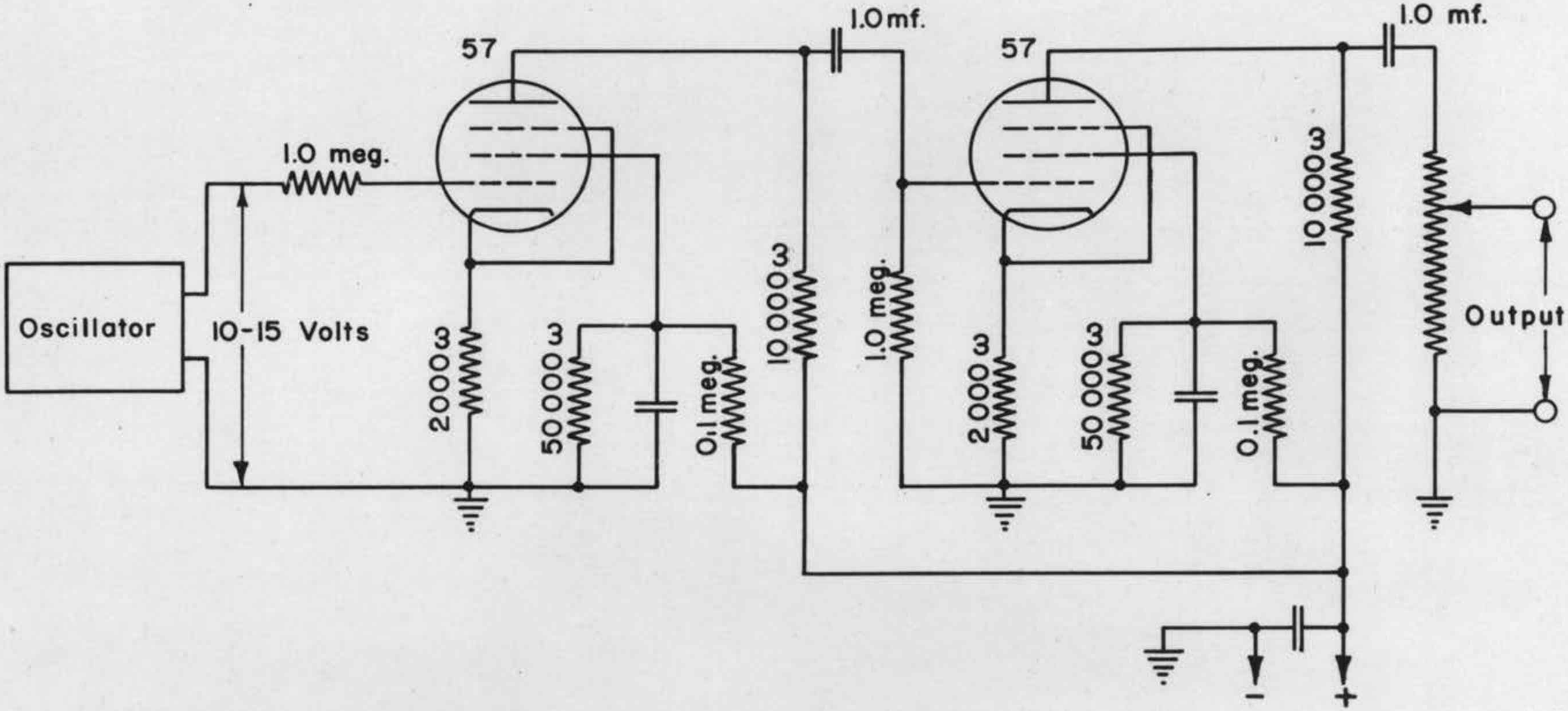


Figure 53

CIRCUIT DIAGRAM OF SQUARE WAVE GENERATOR

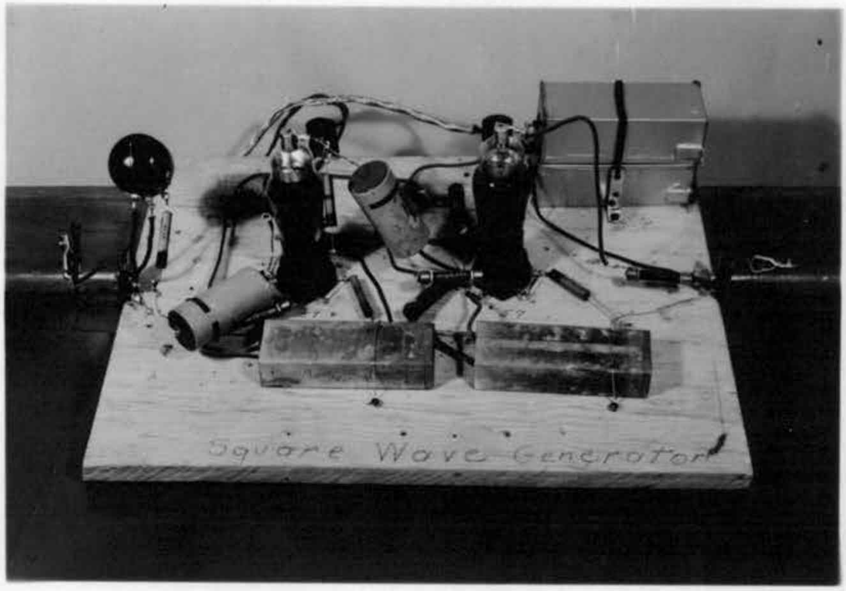


FIGURE 54

SQUARE WAVE GENERATOR USED

IN THIS INVESTIGATION

cutoff. The output of the second tube is a fair approximation of a square wave.

This apparatus is capable of much improvement. A third or even a fourth stage would be desirable to get a perfectly straight sided square top wave in the output.

LIST OF SYMBOLS

C_g = coupling capacitance in grid circuit in farads

R_g = coupling resistance in grid circuit in ohms

$T_g = C_gR_g$ time constant of grid circuit

C_k = cathode resistor bypass capacitance in farads

R_k = cathode resistance in ohms

$T_k = C_kR_k$ time constant of cathode circuit

C_c = low-frequency compensation network capacitance in farads

R_c = low-frequency compensation network resistance in ohms

$T_c = C_cR_c$ time constant of low-frequency compensation network

R_p = dynamic plate resistance of tube in ohms

G_m = transconductance of tube in mhos

R_L = plate load resistance in ohms

C_L = stray capacitance in shunt with R_L in farads

T_L = time constant of plate load circuit $= C_LR_L$

β = complex feedback factor

β_o = mid-band feedback factor

A_o = amplification of amplifier at mid-band

A = complex amplification factor

$w = 2\pi f$

f = frequency in cycles per second

C_β = capacitance in feedback circuit in farads

LIST OF SYMBOLS (continued)

R_β = resistance in feedback circuit in ohms

$T_\beta = C_\beta(R_\beta + R_k)$ time constant of feedback circuit

u = amplification factor of tube

p $= T_\beta/T_g$ parameter used in negative feedback curves

d $= A_o\beta_o$ parameter used in negative feedback curves

n $= R_c/R_L$ parameter used in study of low-frequency compensation networks

M $= G_mR_k$ parameter used in cathode impedance curves and also in the study of high-frequency compensation by means of cathode impedances

P $= T_L/T_k$ parameter used in the study of high-frequency compensation by means of cathode impedances

j = complex vector operator $(-1)^{\frac{1}{2}}$

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