A STUDY OF HIGH-FREQUENCY PEAKING BY FEEDBACK IN TRANSISTOR VIDEO AMPLIFIERS

by

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A STUDY OF HIGH-FREQUENCY PEAKING BY FEEDBACK

IN TRANSISTOR VIDEO AMPLIFIERS

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PREFACE

There has been much written on the transistor video amplifier using series and/or shunt peaking circuits to extend the frequency response; therefore, these methods were not considered in this work. The objective of this thesis was to design a transistorized video amplifier using emitter and collector to base feedback high-frequency peaking techniques to increase the frequency response. An analysis of the amplifiers considered is included in the appendices, making this an integral part of the thesis.

The author wishes to express his appreciation to Dr. Harold T. Fristoe, Professor of Electrical Engineering, for his valuable assistance, guidance, and encouragement. Also, the author is indebted to Mr. Paul A. McCollum, Associate Professor of Electrical Engineering, and Mr. Theron M. Randall for their assistance.

A special note of thanks goes to my wife, Mary, for her hard work and encouragement throughout my college career and for her help in typing and preparing this manuscript.

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SYMBOLS

A	Constant
B	Constant
C	Capacitance or Constant
f	Frequency
f_{co}	Cutoff frequency (negative 6 db/oct slope)
f_{on}	Cuton frequency (positive 6 db/oct slope)
f_e	Common emitter forward current transfer ratio cutoff frequency
f_o	Undamped resonant frequency
G_i	Current gain
G_v	Voltage gain
G·BW	Gain-Bandwidth product
h_{ie}	Mid-frequency short circuit input impedance
h_{fe}	Mid-frequency short circuit forward current transfer ratio
h_{re}	Mid-frequency open circuit reverse current transfer ratio
h_{oe}	Mid-frequency open circuit output impedance
h_{ie}'	Mid-frequency short circuit input impedance (transformed to include effects of external series emitter feedback resistor)
h_{fe}'	Mid-frequency short circuit forward current transfer ratio (transformed to include effects of external series emitter feedback resistor)
h_{re}'	Mid-frequency open circuit reverse current transfer ratio (transformed to include effects of external series emitter feedback resistor)
h_{oe}'	Mid-frequency open circuit output impedance (transformed to include effects of external series emitter feedback resistor)
h_{ie}''	Mid-frequency short circuit input impedance (transformed to include effects of collector to base external feedback impedance)
h_{fe}''	Mid-frequency short circuit forward current transfer ratio (transformed to include effects of collector to base external feedback impedance)
h_{re}''	Mid-frequency open circuit reverse current transfer ratio (transformed to include effects of collector to base external feedback impedance)
h_{oe}''	Mid-frequency open circuit output impedance (transformed to include effects of collector to base external feedback impedance)
h_{ie}^*	High-frequency short circuit input impedance
h_{fe}^*	High-frequency short circuit forward current transfer ratio
h_{re}^*	High-frequency open circuit reverse current transfer ratio
h_{oe}^*	High-frequency open circuit output impedance
i	Instantaneous current
I	D-C current
L	Inductance
p	Heaviside notation (for steady state, $p = j\omega$)
Q	Operating point

SYMBOLS

R	Resistance
R_1	Emitter series resistance
R_f	Collector to base feedback resistance
R_i	Mid-frequency transistor input impedance
R_i*	High-frequency transistor input impedance
R_l	Mid-frequency transistor load
R_l*	High-frequency transistor load
r_b	Ohmic base resistance
r_ϵ	Emitter junction resistance
v	Instantaneous voltage
V	D-C voltage
Z_f	Collector to base feedback impedance
α	Common-base forward current transfer ratio
Δ	Determinant, cofactor, or increment
Δ^{he}	Mid-frequency determinant of the matrix $[h]$
$\Delta^{he'}$	Mid-frequency determinant of the matrix $[h']$
$\Delta^{he''}$	Mid-frequency determinant of the matrix $[h'']$
Δ^{he*}	High-frequency determinant of the matrix $[h*]$
ζ	Damping factor
τ	Time constant
ω	Angular frequency
ω_e	Common-emitter forward current transfer ratio cutoff angular frequency
ω_{co}	Angular cutoff frequency (negative 6 db/oct slope)
ω_{on}	Angular cuton frequency (positive 6 db/oct slope)
ω_o	Undamped resonant angular frequency
*	Starred quantity is complex in nature

CHAPTER I

INTRODUCTION

A video amplifier may be defined as a wide-band, low pass amplifier. There are many uses for such an amplifier, as in television, radar, instrumentation, etc. The ideas and techniques employed when designing a transistor video amplifier must differ considerably from those of electron tube circuits. The purpose of this work was to develop a better design procedure for transistorized video amplifiers. Such things as internal feedback, charge carrier transit time across the base region, and other effects must now be taken into account. First order approximations of these effects on the parameters of the transistor are given in the appendices by equations (40B), (41B), and (3C).

The total charge in the barrier region is increased when a junction is biased in the reverse direction. An effective capacitance across the barrier resistance tends to short it out at the high frequencies due to the charging up of the barrier when the voltage across the junction changes. This capacity is noticeably nonlinear with respect to applied voltage up to very high frequencies, and is a function of depletion layer transit time. The collector capacitance in the amplifiers considered in this work proved to be negligibly small at all frequencies of interest as compared to the other effects.

The desired specifications are the first things to be considered when designing any amplifier, whether using tubes or transistors. The specifications for this amplifier were as follows: (1) Bandwidth — 5 MC

(10 cps to 5 MC); (2) Mid-frequency input impedance -- 600Ω; (3) Mid-frequency output impedance -- 100Ω; (4) Minimum input voltage 0.1 mv; (5) Maximum input voltage -- 1 mv; (6) Voltage gain -- 60 db $\pm$ 1 db with a 600Ω resistive load, $\pm$ 3 db with any resistive load greater than 600Ω; (7) Current stability -- less than 5. These specifications did not have to be followed too closely since the primary interest was on the design procedure for different types of feedback compensation. Amplifiers were built to verify the designs which appear in the chapters to follow. The amplifier discussed in Chapter II and analyzed in Appendix B does not satisfy specification (3) and the second part of (6), but the principles used there will apply to the amplifier of Chapter III and Appendix C which essentially meets the specifications. Only common-emitter connections of the transistors were considered.

High-frequency compensation of video amplifiers may be accomplished in many ways. The simplest means of improving the high-frequency response of a resistance-coupled amplifier is by the addition of inductances in series with the load resistor and biasing resistors of the next stage. This is known as shunt peaking and increases the impedance of the biasing network at high-frequencies. The increase in the biasing network impedance at high-frequencies tends to correct for the falling off in amplification that otherwise occurs at high-frequencies. The practical bandwidth increase which may be accomplished in this manner is in the order of two times the uncompensated bandwidth. Other variations of the shunt peaking circuit may be used, such as shunting the inductor with a capacitance. It would be necessary to use shunt peaking networks in series with all the biasing resistors if the value of all of these resistors were relatively low. Series and/or series-shunt peaking could also be employed as a means of compensation.

Another type of high-frequency compensation is made possible by high-frequency peaking through the use of negative feedback. This may be accomplished in the emitter circuit, and/or the collector to base feedback circuit of single common-emitter stages or in over-all feedback loops. The single stage types are covered in detail in Chapters II and III. The basic principle of all types of high-frequency peaking by feedback is to reduce the feedback at the point when the amplification would otherwise decrease. Stability difficulties are often encountered when using the over-all feedback loops. The choice between emitter or collector to base high-frequency peaking circuits is determined by the desired output conditions. Emitter feedback peaking would be used if it is desired to have a constant current output with collector to base feedback peaking for a constant voltage output. The compensated bandwidth of an amplifier may be ten or more times that of the uncompensated bandwidth for this type of compensation. The use of active elements as a means of compensation is obviously the reason for the relatively large increase of bandwidth in high-frequency feedback peaking as compared to series and shunt peaking values which employ only passive elements for compensation.

The h-parameters will be used almost exclusively in the analysis of the amplifiers. Appendix A gives the method used to determine the common-emitter h-parameters from output characteristics of the transistors. Four stages were used for two reasons: (1) The forward current transfer ratio ranged from 43 to 98 on the transistors available; therefore, two stages would furnish a majority of the gain at mid-frequencies; then two stages would be available for feedback peaking at the high-frequencies. (2) There would be no appreciable phase shift between input and output voltages at the mid-frequencies. The transistors used were PNP diffused-base germanium, type 2N623, made by Texas Instruments, Inc. A list of symbols used and a corresponding definition appear on pages ix and x.

CHAPTER II

HIGH-FREQUENCY EMITTER-FEEDBACK-PEAKING AMPLIFIER

General Discussion

The basic principle of an amplifier having degenerative feedback in the emitter circuit of at least one stage will be investigated. The circuits of Fig. 1 will provide peaking at any desired frequency, within the range of the transistor, by choosing the proper values of R_1, C_1, and L_1. The current gain of Fig. 1a will be of the form

$$\frac{i_2}{i_1} = K \frac{1 + p\tau_1}{\tau_2^2 p^2 + 2\zeta\tau_2 p + 1} \; ; \tau_1 = R_1 C_1 ; \; K = \text{Constant}, \qquad (1)$$

for most video amplifiers. This equation was obtained from (64B). The approximation $R_i^* \doteq h_{ie}^* + R_1(1 + h_{fe}^*)/(1 + p\tau_1)$ must hold if (1) is to be valid. The assumptions necessary for evaluating R_i^* according to this approximation are: (1) $\Delta^{he*}R_1^* \ll h^{ie\prime*}$ and (2) $h^{oe\prime*}R_1^* \ll 1$. It is also necessary that the high-frequency h-parameters satisfy the conditions given by (40B), (41B), and (3C). Further discussion and development of

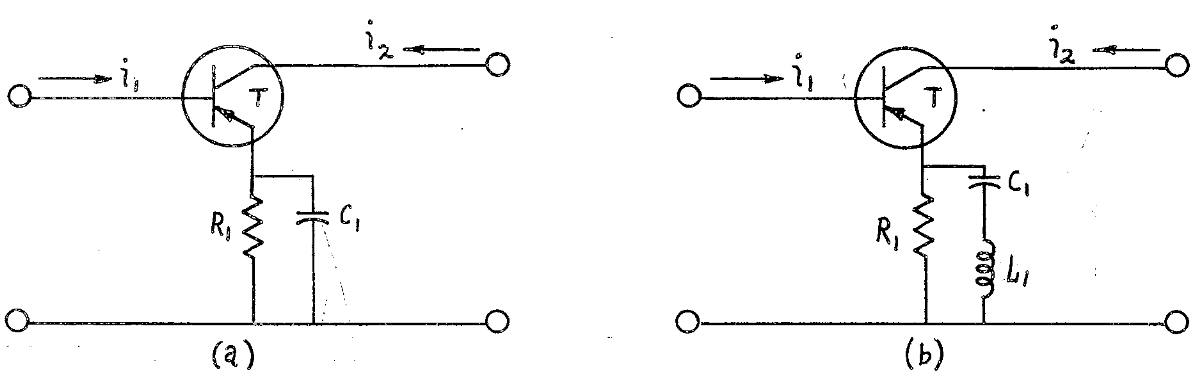

Figure 1. High-Frequency Emitter-Feedback-Peaking Circuits

(1) is contained in the last section of this chapter and (50B) through (70B). Considering the biasing network and the change in R_i* with frequency results in (71B) for the stage gain. Fig. B.9 gives the frequency response of two stages of this type. The desired value of $\omega_1^{on} = 1/\tau_1$ will normally be approximately equal to the smallest ω_i^{co} of the remaining stages when this compensation stage is used to overcome the effects of cutoff of the stage without feedback. The remainder of this chapter and Appendix B are devoted to applying this type of compensation to a specific amplifier. The type of circuit connection shown in Fig. 1b would also provide emitter-feedback peaking.[1] The maximum compensation would be at the resonant frequency of L_1 and C_1, and the shape of the response could be controlled by the L_1/C_1 ratio and the Q of the inductor.

Mid-Frequency Considerations of a Specific Amplifier

A circuit diagram of the amplifier to be considered is contained in Fig. 2 with the circuit values given in Table I. The first and fourth stage emitter resistors are adequately by-passed for the desired bandwidth, whereas the second and third stages provide the high-frequency compensation.

A complete analysis of the amplifier appears in Appendix B. The analysis is divided into three main sections: (1) Mid-frequency analysis, (2) High-frequency analysis without emitter-feedback-peaking, and (3) High-frequency analysis with emitter-feedback peaking.

The mid-frequency analysis of the amplifier consists of (16B) through (39B). The voltage gain is given by (39B) as 61 db with a load resistance of 600 Ω and 71 db with a 10K Ω load. The voltage gain with the 600 Ω load satisfies the requirement of 60 db $\pm$ 1 db, but the 10K Ω case exceeds its

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[1]Richard F. Shea, Transistor Circuit Engineering, pp. 210-211.

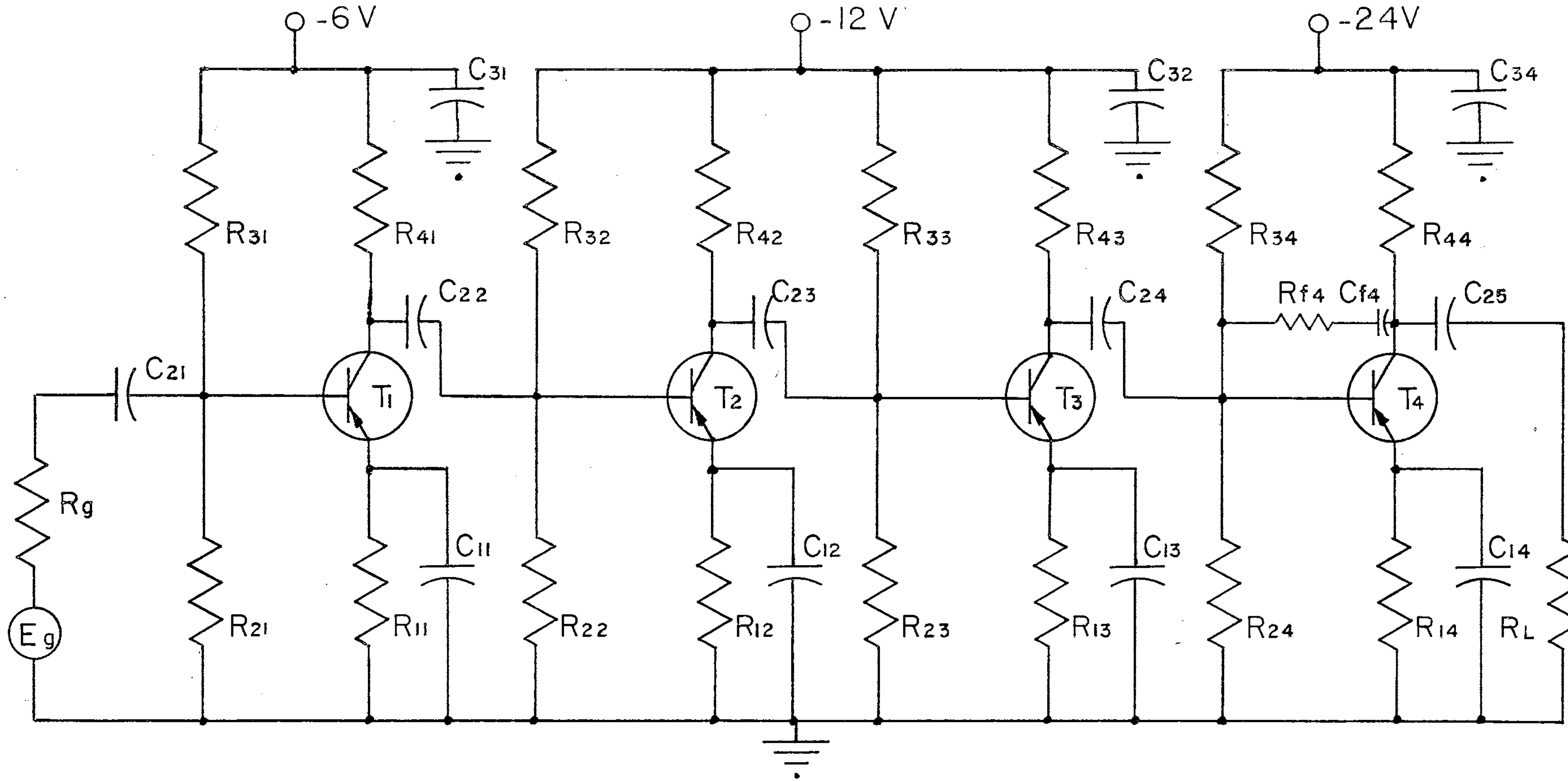

Figure 2. Circuit Diagram of Emitter-Feedback-Peaking Amplifier

TABLE I

CIRCUIT VALUES OF FIGURE 2

R_{11} = 1,200 Ω	C_{11} = 100 μf
R_{12} = 600 Ω	C_{12} = 33 μμf
R_{13} = 660 Ω	C_{13} = 185 μμf
R_{14} = 600 Ω	C_{14} = 100 μf
R_{21} = 3,000 Ω	C_{21} = 10 μf
R_{22} = 2,900 Ω	C_{22} = 30 μf
R_{23} = 2,900 Ω	C_{23} = 30 μf
R_{24} = 3,300 Ω	C_{24} = 30 μf
R_{31} = 8,600 Ω	C_{25} = 10 μf
R_{32} = 16,500 Ω	C_{31} = 500 μf
R_{33} = 12,000 Ω	C_{32} = 500 μf
R_{34} = 18,900 Ω	C_{34} = 500 μf
R_{41} = 3,000 Ω	C_{f4} = 30 μf
R_{42} = 1,500 Ω	T_1 = 2N623-15
R_{43} = 890 Ω	T_2 = 2N623-14
R_{44} = 3,000 Ω	T_3 = 2N623-12
R_{f4} = 120,000 Ω	T_4 = 2N623- 4
R_L = 600 Ω or 10,000 Ω	

requirement by 8 db. This is to be expected since the feedback of the second and third stages, as well as the transistors themselves, tend to display a constant current characteristic. The feedback in the fourth stage is insufficient to correct for all other effects, although it tends to give a constant voltage output. The specifications are already invalid for this circuit, but it will still be of interest to investigate the higher frequency effects, as much of the information gained may be applied in the second amplifier arrangement discussed in Chapter III.

High-Frequency Considerations of a Specific Amplifier Without Compensation ($C_{12} = C_{13} = 0$)

The performance of the amplifier at the high-frequencies without emitter-feedback-peaking is especially interesting, since it will show how much the compensation will extend the bandwidth. Equation (45B) gives the approximate gain of the first stage. It should be noted that the cutoff frequency, f_1^{co}, given by (46B) is greater than the forward current transfer ratio cutoff frequency, f_{e1}, by a factor of $\frac{(R_{x1} + h_{ie1})}{(R_{x1} + r_{b1})}$. The explanation of this is given by (40B) and (42B). From these equations it is seen that the input impedance of the transistor decreases as the frequency increases; hence, a greater amount of the current being fed to the stage is received by the transistor instead of being bled off by the biasing network. The same principle will apply to the second and third stages, but their cutoff frequency will be much greater than f_e. This is shown in (54B), since h_{iek}' is usually many times larger than any of the other resistance values in the equation. This is logical since these two stages contain 600 Ω degenerative feedback resistors. The fourth stage may be assumed to function essentially the same as the first stage since the feedback resistor, R_{f4}, is too large to have much effect on the circuit. The over-all gain of the amplifier is given by (56B) and is plotted in Fig. 3 along with the experimental results obtained. The data for the experimental results appear in Table II.

High-Frequency Considerations of a Specific Amplifier With Compensation (C_{12} = 33 $\mu\mu$f, C_{13} = 185 $\mu\mu$ f)

The frequency response of the amplifier may be extended by letting C_{12} and C_{13} take on the proper values. The first and fourth stage considerations as discussed in the previous section will hold for the amplifier when C_{12} and C_{13} take on some numerical value as well as when they are equal to zero.

The relative gain of the second or third stage is given by (71B) and the logarithmic plot of the response appears in Fig. B.9. It should be noted that each stage compensates at a rate greater than 6 db per octave for a considerable frequency range; therefore a single emitter-feedback-peaking stage could compensate for more than one stage over a limited frequency range. The corner frequency, f_k^{on}, is determined by the values of R_1 and C_1 in the emitter circuit. The maximum of the "resonant hump" is located at a frequency, f_{ok}, proportional to $(C_k/A_k)^{1/2}$ where C_k and A_k are defined by (63B). For a specific mid-frequency gain, f_{ok} will be determined by the choice of f_k. Changing the value of f_{ok} with respect to f_k may be accomplished by changing the value of the ratio R_1/C_1 in such a manner as to keep f_k^{on} constant. Changing R_1 will result in a different mid-frequency gain, but the over-all gain of the amplifier may be held constant by changing the gain of any other stage. The values of f_k^{on} and f_{ok} depend on the specific amplifier under consideration.

The gain of the complete compensated amplifier, given by (72B), is plotted in Fig. 4 along with the experimental result obtained. The data for the experimental curve are given in Table III.

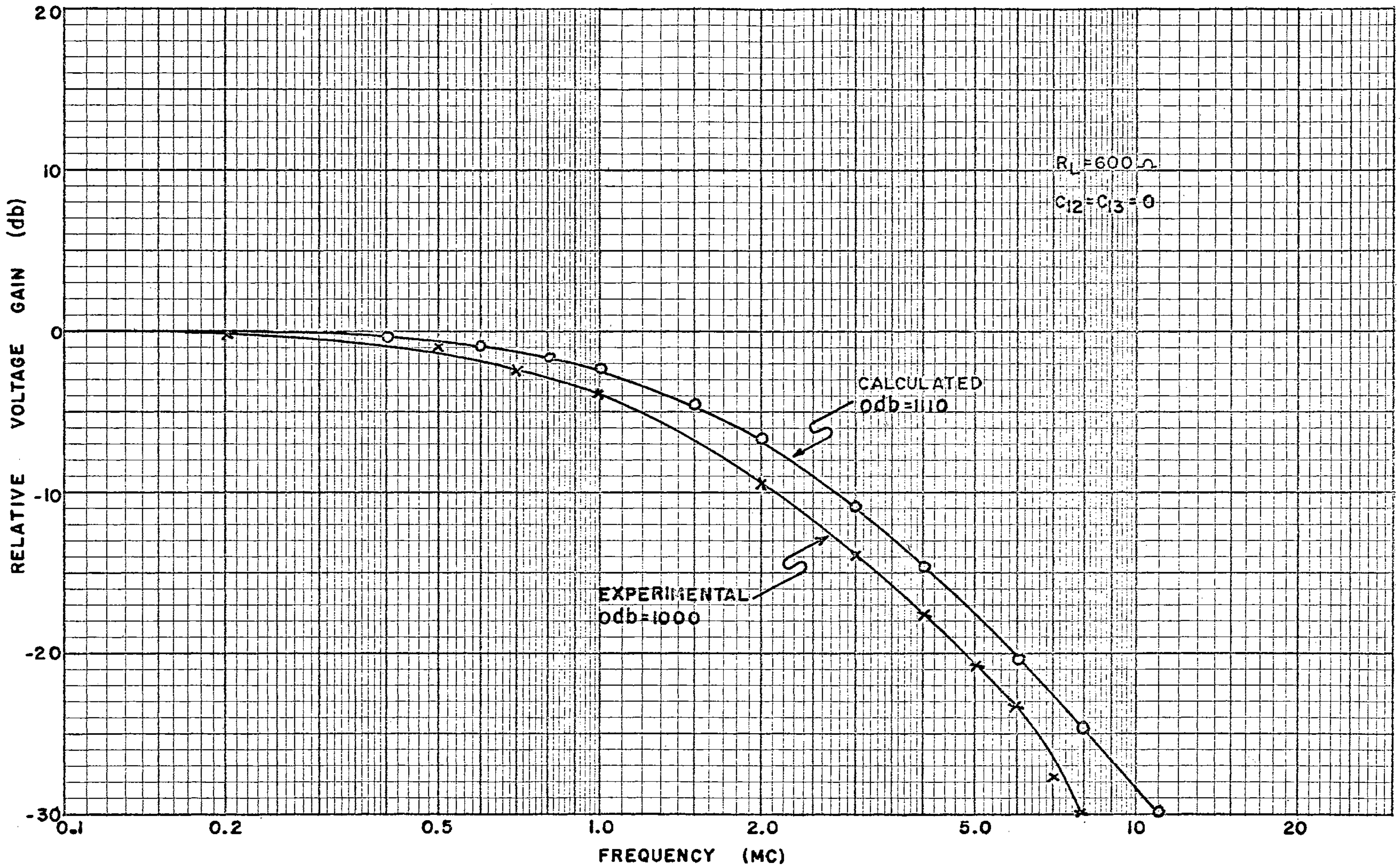

Figure 3. Frequency Response of Uncompensated Emitter-Feedback-Peaking Amplifier

TABLE II

RESPONSE OF UNCOMPENSATED EMITTER-FEEDBACK-PEAKING AMPLIFIER
$R_L = 600\Omega$

Frequency (cps)	Input Voltage (mv)	Output Voltage (mv)	$\frac{\text{Output}}{\text{Input}}$ (db)
10	1	440	52.9
20	1	720	57.2
50	1	950	59.6
100	1	1000	60.0
200	1	1000	60.0
500	1	1000	60.0
1K	1	1000	60.0
2KC	1	1000	60.0
5KC	1	1000	60.0
10KC	1	1000	60.0
20KC	1	1000	60.0
50KC	1	1000	60.0
100KC	1	1000	60.0
200KC	1	980	59.9
500KC	1	900	59.1
700KC	1	760	57.6
1MC	1	640	56.1
2MC	1	330	50.4
3MC	1	200	46.0
4MC	1	130	42.3
5MC	1	90	39.1
6MC	1	70	36.9
7MC	1	40	32.1
8MC	1	30	29.6

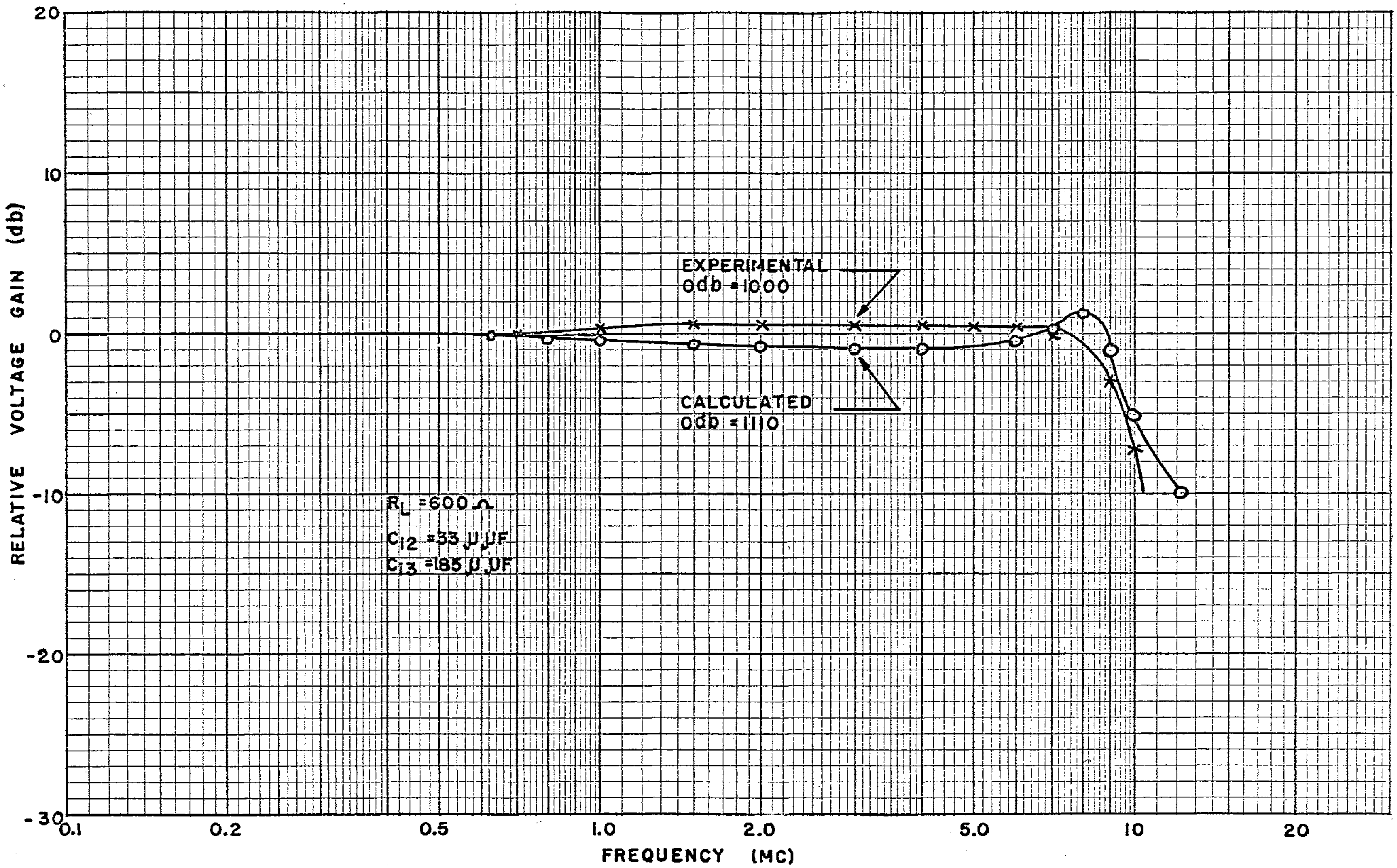

Figure 4. Frequency Response of Compensated Emitter-Feedback-Peaking Amplifier

TABLE III

RESPONSE OF COMPENSATED EMITTER-FEEDBACK-PEAKING AMPLIFIER
$R_L = 600\,\Omega$

Frequency (cps)	Input Voltage (mv)	Output Voltage (mv)	$\frac{\text{Output}}{\text{Input}}$ (db)
10	1	100	40.0
20	1	250	48.0
50	1	620	55.8
100	1	880	58.9
200	1	980	59.9
500	1	1000	60.0
1K	1	1000	60.0
2K	1	1000	60.0
5K	1	1000	60.0
10K	1	1000	60.0
20K	1	1000	60.0
50K	1	1000	60.0
100K	1	1000	60.0
200K	1	1000	60.0
500K	1	1000	60.0
700K	1	1000	60.0
1M	1	1020	60.3
2M	1	1050	60.5
3M	1	1030	60.3
4M	1	1025	60.3
5M	1	1032	60.3
6M	1	1018	60.3
7M	1	982	59.9
8.9M	1	700	57.0
10M	1	440	52.9

CHAPTER III

HIGH-FREQUENCY COLLECTOR TO BASE-FEEDBACK-PEAKING AMPLIFIER

General Discussion

The collector to base-feedback-peaking amplifier gets its name from the fact that at least one stage employs collector to base feedback as a means of increasing the bandwidth. This connection will be used in the last stage of the amplifier so that the output voltage will remain relatively independent of the load impedance.

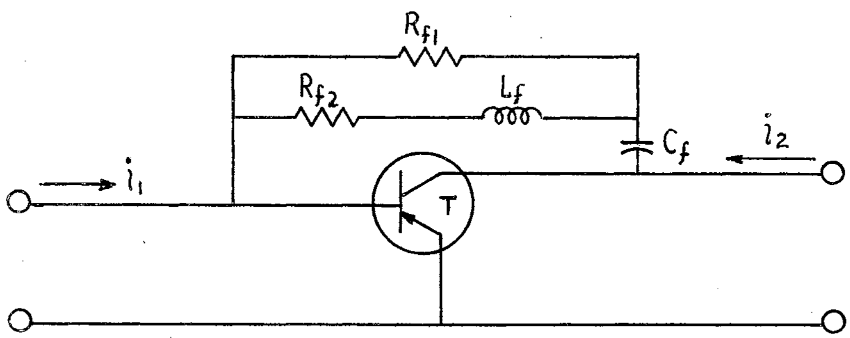

Figure 5. Collector to Base-Feedback-Peaking Stage

A collector to base-feedback-peaking stage is shown in Fig. 5. The purpose of R_{f1} is to provide a limiting frequency at which the compensation will cease to increase. This resistor may be omitted if desired. The current gain of the stage will be of the form

$$\frac{i_2}{i_1} = k \frac{1 + p\tau_1}{\tau_2^2 p^2 + 2\zeta\tau_2 p + 1} \; ; \tau_1 = \frac{L_{f2}}{R_{f2}} , \; K = \text{Constant} \qquad (2)$$

when the following assumptions, which are true at the frequencies of interest in most video amplifiers, hold.

$$\left.\begin{array}{l} h_{fe}^* Z_f \gg h_{ie}^* \\ h_{fe}^* \gg 1 \\ 1 \gg h_{oe}^* R_l^* \end{array}\right\} \quad (3)$$

The above equations were derived from (10C) through (15C). The high-frequency parameters discussed in Chapter II must also hold for these equations. It is seen that (2) is of the same form as (1) which was discussed in the previous chapter; hence, either type of compensation will give the same type of relative gain performance if the circuit values are of the correct value. The main difference between the two types is that emitter-feedback-peaking tends to give a constant current output while collector to base-feedback-peaking tends to give a constant output voltage. A specific video amplifier will now be considered.

Mid-Frequency Considerations of a Specific Amplifier

The complete circuit diagram of the amplifier to be considered is given in Fig. 6 with circuit values in Table IV. The analysis of this amplifier is given in Appendix C.

The mid-frequency analysis of this amplifier was performed basically the same as that of the emitter-feedback-peaking amplifier of Chapter II. The results are given by (1C) and (2C). The voltage gain was determined to be 61 db with a 600 Ω resistive load and 62.5 db with a 10 KΩ load. The input impedance of the amplifier is 804 Ω, and the output impedance is 100 Ω. These values are all within the specifications as set up in Chapter I with the exception of the input impedance. Since the amplifier was built primarily to investigate corrective feedback methods, no attempt will be made to improve the input impedance. The input impedance was experimentally determined to be 830 Ω.

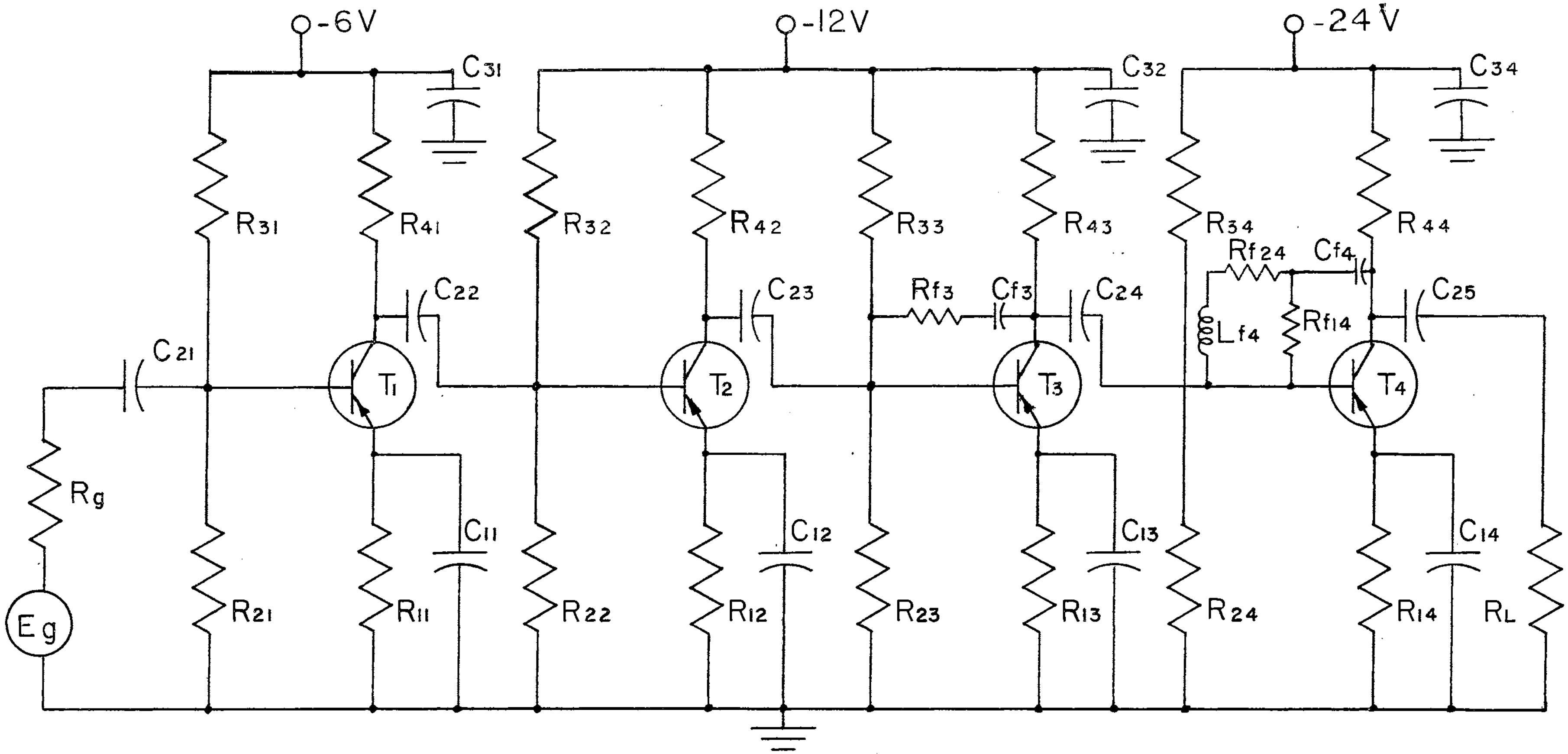

Figure 6. Circuit Diagram of Collector to Base-Feedback-Peaking Amplifier

TABLE IV

CIRCUIT VALUES OF FIGURE 6

R_{11} = 1,200 Ω	C_{11} = 250 μf
R_{12} = 600 Ω	C_{12} = 127 μμf
R_{13} = 660 Ω	C_{13} = 250 μf
R_{14} = 600 Ω	C_{14} = 250 μf
R_{21} = 3,000 Ω	C_{21} = 160 μf
R_{22} = 2,900 Ω	C_{22} = 160 μf
R_{23} = 2,900 Ω	C_{23} = 160 μf
R_{24} = 3,300 Ω	C_{24} = 160 μf
R_{31} = 8,600 Ω	C_{25} = 160 μf
R_{32} = 16,500 Ω	C_{31} = 500 μf
R_{33} = 12,000 Ω	C_{32} = 500 μf
R_{34} = 18,900 Ω	C_{34} = 500 μf
R_{41} = 3,000 Ω	C_{f3} = 0.75 μf
R_{42} = 1,500 Ω	C_{f4} = 30 μf
R_{43} = 890 Ω	L_{f4} = 30 μh
R_{44} = 3,000 Ω	T_1 = 2N623-15
R_{f3} = 3,300 Ω	T_2 = 2N623-14
R_{f41} = 3,000 Ω	T_3 = 2N623-12
R_{f42} = 1,000 Ω	T_4 = 2N623-11
R_L = 600 Ω or 10,000 Ω	

High-Frequency Considerations of a Specific Amplifier Without Compensation ($C_{12} = L_{f4} = 0$)

The gain and corner frequencies of the first and second stages of this amplifier are given by (45B), (46B), (53B) and (54B). The discussion in Chapter II, which applied to these equations, is valid for this amplifier as well.

The relative gain and corner frequencies of the third stage are given by (23C) and (24C), respectively. The relative gain is plotted in Fig. C.2. The corner frequency f_{e3} is less than f_{33}^{co}, but it should be noted that this order would be reversed if R_{i3} were less than r_{b3}. This means that their relative positions are controlled by the mid-frequency input impedance to the transistor. Also of interest is the fact that the forward current transfer ratio cutoff frequency of the fourth stage is an important factor in the resultant response of the third stage. This was brought about since the load of the third stage is a function of the input impedance of the fourth stage. The composite curve of the third stage shows that some compensation may be possible without the use of an inductor in the feedback impedance. The maximum peaking is obtained when the input impedance of the transistor is a maximum.

The fourth stage gain is given by (33C). This equation was obtained by a slightly different procedure than that which gives (2) as a result. Equation (15C), which was used in determining the gain of (2), may be replaced by (16C) since $h_{ie4}^* + Z_{f4} << R_{l4}^* h_{fe4}^*$. The transistor gain then simply becomes the parallel combination of the feedback resistors R_{f41} and R_{f42} divided by the load, R_{l4}, seen by the transistor, which is the parallel combination of R_{44} and R_L. Considering the effects of the biasing network and the changing input impedance of the transistor gives (33C) and (34C) for the relative gain and the corner frequencies, respectively. The mid-frequency input impedance of the transistor, R_{i4}, of the fourth stage is of the same order as r_{b4}; hence the relative gain will be essentially flat. It is obvious that the frequency range for which (33C) will hold is definitely limited, but this limit is beyond the frequencies of interest for this amplifier.

The response of the complete uncompensated amplifier is given in Fig. C.3. A comparison of this result with that of the compensated amplifier of the next section gives an indication of the bandwidth increase which is possible when the amplifier is compensated.

High-Frequency Considerations of a Specific Amplifier With Compensation ($C_{12} = 127\ \mu\mu f$, $L_{f4} = 30\ \mu h$)

It should be noted that the second stage uses emitter-feedback-peaking as described in Chapter II. The discussion for this stage with compensation in Chapter II is valid for this amplifier. The discussion of the first and third stages in the previous sections of this chapter are still true since no noticeable effects are observed in these stages due to C_{12} and L_{f4} taking on numerical values other than zero. The necessary changes in the fourth stage are contained in (35C) through (39C). The relative gain and corner frequencies are given by (38C) and (39C), respectively. Fig. C.4 gives the second and fourth stage responses when compensation is provided, and Fig. 7 shows the calculated amplifier response compared to the experimental result. Table V contains the numerical data from which the experimental curves of Fig. 7 were taken.

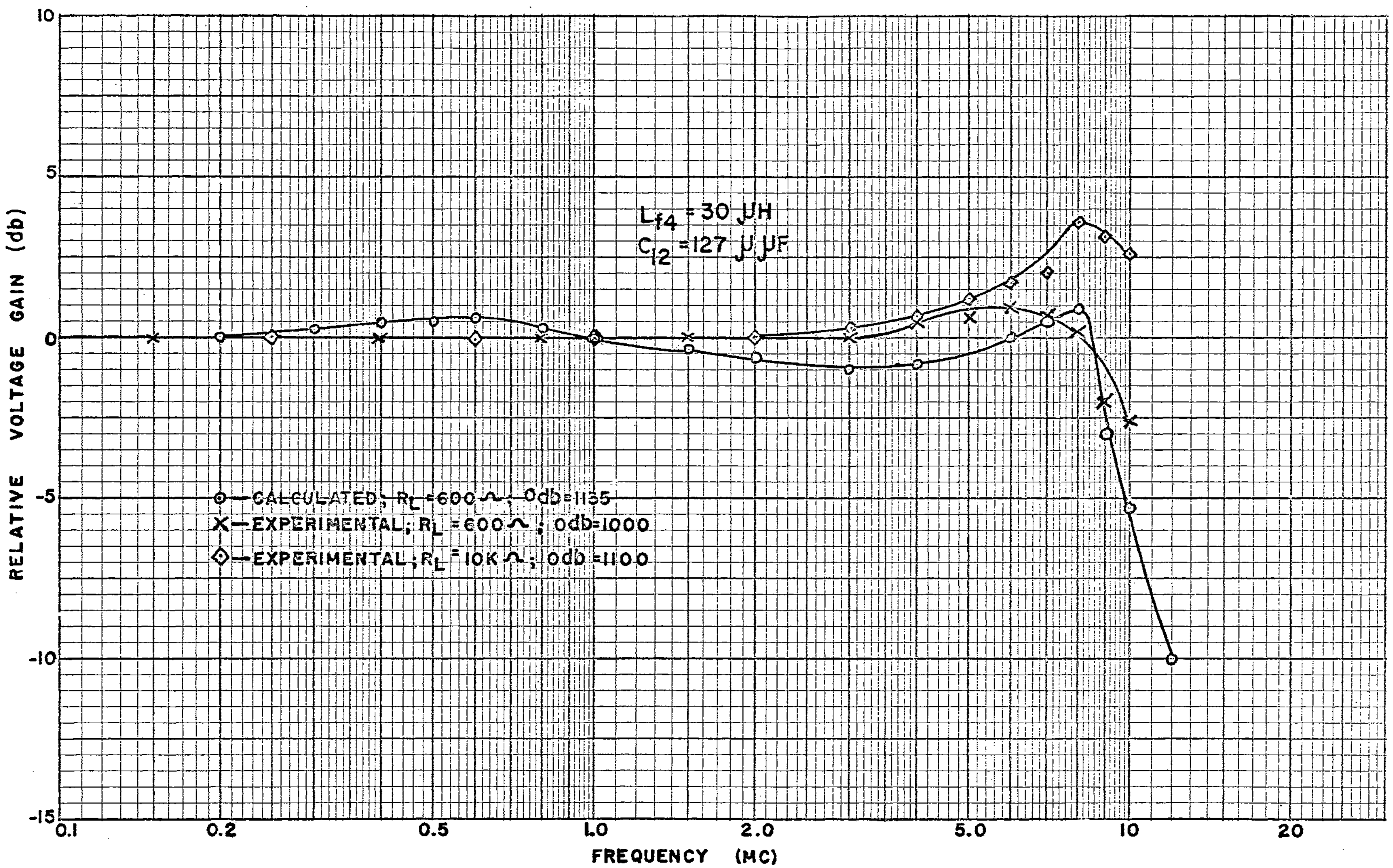

Figure 7. Frequency Response of Collector to Base-Feedback-Peaking Amplifier
With Compensation.

TABLE V

EXPERIMENTAL RESULT OF COLLECTOR TO BASE-FEEDBACK-PEAKING AMPLIFIER WITH COMPENSATION

Frequency (cps)	Input Voltage (mv)	$R_L = 600\Omega$ Output Voltage (mv)	$R_L = 600\Omega$ $\frac{\text{Output}}{\text{Input}}$ (db)	$R_L = 10K\Omega$ Output Voltage (mv)	$R_L = 10K\Omega$ $\frac{\text{Output}}{\text{Input}}$ (db)
10	1	900	59.0	950	59.5
20	1	1000	60.0	1000	60.0
50	1	1000	60.0	1050	60.5
100	1	1000	60.0	1100	60.9
200	1	1000	60.0	1100	60.9
500	1	1000	60.0	1100	60.9
1K	1	1000	60.0	1100	60.9
2K	1	1000	60.0	1100	60.9
5K	1	1000	60.0	1100	60.9
10K	1	1000	60.0	1100	60.9
20K	1	1000	60.0	1100	60.9
50K	1	1000	60.0	1100	60.9
100K	1	1000	60.0	1100	60.9
200K	1	1000	60.0	1100	60.9
500K	1	1000	60.0	1100	60.9
700K	1	1000	60.0	1100	60.9
1M	1	1000	60.0	1100	60.9
2M	1	1000	60.0	1100	60.9
3M	1	1000	60.0	1150	61.2
4M	1	1050	60.5	1200	61.5
5M	1	1070	60.6	1300	62.2
6M	1	1100	60.9	1350	62.6
7M	1	1080	60.6	1400	62.9
8M	1	1020	60.2	1700	64.6
9M	1	800	58.0	1600	64.1
10M	1	740	57.4	1500	63.5

CHAPTER IV

SUMMARY AND CONCLUSIONS

General Discussion

The input impedance of the amplifiers discussed in the previous chapters decreases at the high frequencies. Series or shunt compensation for the variation of input impedance as described by Shea[1] would provide a means of maintaining a constant input impedance. Series impedance compensation would be desired for an input from a current source and shunt impedance compensation for a voltage source.

The frequency range could be extended further by employing series and/or shunt high-frequency compensation[2] in one or more of the stages. A disadvantage of this type of compensation in R-C coupled stages is that each of the biasing resistors must be compensated. This would mean a minimum of three inductors for the simplest compensation network. The big advantage is the fact that it might be possible to eliminate the need for an active element; hence, it would usually be the most economical. This would, of course, depend on the particular elements used. Also, it should be pointed out that in a stage without feedback, the current transfer ratio between the biasing network and the transistor input impedance is usually in the order of 0.5. This would mean that the maximum compensation would be 6 db even if all of the current entered the transistor.

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[1]Richard F. Shea, Transistor Circuit Engineering, pp. 214-217.

[2]Ibid. pp. 204-210.

On the other hand, the current transfer ratio may be as low as 0.02 for the stages with emitter feedback (R_1 unby-passed); hence more compensation would apparently be possible. The input impedance would probably be decreasing at the point where high-frequency compensation is desired; therefore the limit of the series or shunt compensation would be only about 6 db, but the slope of compensating stage response could be changed from + 6 db/oct. to + 12 db/oct. A stage with considerable collector to base feedback would receive no appreciable gain by using series or shunt compensation since the current transfer ratio is already 0.9 or greater.

Emitter-Feedback-Peaking Amplifier

The Emitter-Feedback-Peaking amplifier of Chapter II did not satisfy the specifications set up in Chapter I, but there are other possible uses for this circuit. A low level voltage source may be converted to a current source by employing emitter-feedback-peaking in the fourth stage. Amplification would be obtained at the same time the conversion is being made. This type of amplifier would also be satisfactory as a voltage amplifier when it is desired to match a low impedance source to a constant high impedance load with amplification.

The equivalent input noise ranged from approximately 9 μV (R_g = 0) to 10.5 μV ($R_g = \infty$); therefore, the signal to noise ratio would be 19.6 db for minimum input voltage and 39.6 db for maximum input voltage for the worst case.

Collector to Base Feedback Peaking Amplifier

A review of the specifications and how this amplifier satisfies them will now be considered. The frequency band specified was 10 cps to 5 MC for a relative voltage gain of $\pm$ 1 db with a 600 Ω load, and the experimental result was 10 cps to 8.3 MC. With a load greater than 600 Ω, the same frequency band was desired with a ± 3 db relative gain, and the result was 9 cps to 7.0 MC. It was necessary to apply some low frequency compensation to obtain the desired response at 10 cps. This was done by selecting C_{f3} in such a manner as to decrease the feedback at the low frequencies. The desired mid-frequency input impedance was 600Ω while the experimental result was 830Ω. This is not an exact match, but mismatches of this order are often permitted in transistor work. The desired value could have been obtained by applying a small amount of collector to base feedback (large resistance) to the first stage and adjusting the gain of one of the other stages to make up for the loss in gain. All other specifications were obtained within reasonable limits.

A lower output impedance could be obtained with the same biasing and load resistor values by interchanging the second and third stages. Making this change would increase the impedance seen looking out of the input terminals of the fourth stage transistor from approximately the low output impedance of a collector to base feedback stage to essentially the value of the parallel combination of R_{24}, R_{34}, and R_{43}; thereby lowering the output impedance of the fourth stage. Making this change in the amplifier of Chapter III reduces the output impedance from 100Ω to 35Ω. Little change in output voltage would be observed as the load varied from 300Ω to infinity in this case. Still lower output impedances could be obtained by changing to a common collector connection in the fourth stage.

It would be possible to further increase the bandwidth of the amplifier by applying collector to base-feedback-peaking techniques to the third stage. This would not effect the mid-frequency response since this stage already has considerable collector to base feedback, but with no peaking present.

The equivalent input noise of this amplifier varied from approximately 7 μV ($R_g = 0$) to 10.6 μV ($R_g = \infty$). This gives a signal to noise ratio ranging from 23.1 db to 19.5 db for minimum input voltage and 43.1 db to 39.5 db for maximum input voltage. It should also be mentioned at this point that some work was done with high input impedance amplifiers (in the order of a megohm), but the equivalent input noise was even greater than the minimum input signal for the larger values of generator impedance. In general, the agreement between the calculated and experimental values was most satisfactory.

Suggested Improvements and Changes in the Final Amplifier

The input impedance decreases, and the output impedance increases as the frequency increases above 500 KC. It would be desirable to maintain both at constant values over the entire frequency band of the amplifier. Possible methods of maintaining constant input impedance were discussed previously in this chapter. If the sources did not approach true voltage or current generators, it would be necessary to make some changes in the basic amplifier to compensate for the losses which would occur in the networks necessary to give a constant input impedance. The phase shift between the input and output voltages becomes appreciable below 100 cps and above 500 KC. The networks giving constant input and output impedances would probably improve this, but, if not, it would be a big improvement in many cases to maintain zero phase shift.

This amplifier could be made into a versatile laboratory amplifier by the addition of a stepping attenuator which would give voltage gains of 20 db, 40 db, and 60 db and a vernier gain control which would vary the gain between these steps. It would be desirable in many cases to have extremely high input and low output impedances which would eliminate losses in the source impedance and still give an amplification of 1000 with any load from some low value to infinity. The output impedance of this amplifier could be made lower rather easily, but to increase the input impedance appreciably causes the noise factor to become significant. It might become necessary to use a low voltage and low power consuming vacuum tube in the first stage to overcome the noise problem. It would also be desirable in any case to improve the signal to noise ratio.

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Shea, Richard F. Transistor Circuit Engineering. New York: John Wiley & Sons, Inc., September, 1957.

Thaler, George J. and Brown, Robert G. Servomechanism Analysis. New York: McGraw-Hill Book Company, Inc., 1953.

APPENDIX A

DETERMINATION OF h-PARAMETERS FROM VISUAL DISPLAY OF OUTPUT CHARACTERISTICS

All of the common emitter h-parameters may be determined with two output characteristics. One of the output characteristics should be obtained by applying current steps to the base and the other with voltage steps to the base(see Fig. A.2). The output characteristics were obtained using a Tektronix Transistor Curve Tracer, Model 575. Since the curve tracer uses a resistor, R_b, in series with the base to limit the base current, the known value on the output characteristic will be V_b' instead of V_b.

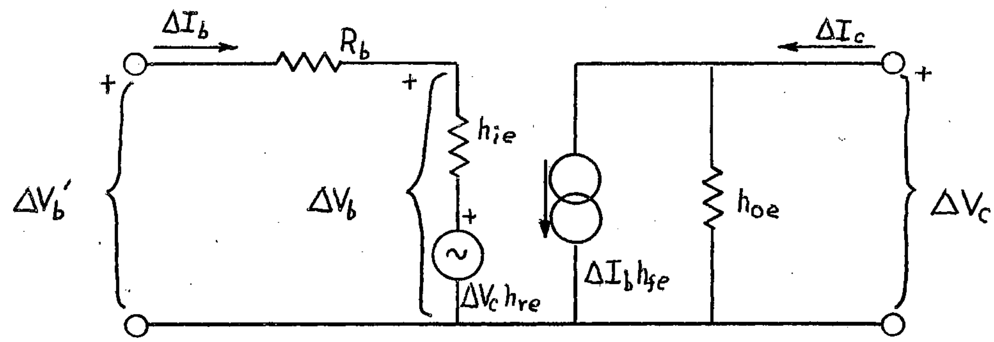

Figure A.1. Transistor Equivalent Circuit for Determining h-Parameters.

The first h-parameter, h_{ie}, may be determined by using Fig. A.1, Fig. A.2 and the definition of h_{ie},

$$h_{ie} = \frac{\partial V_b}{\partial I_b}; \qquad V_c = \text{Constant} \tag{1A}$$

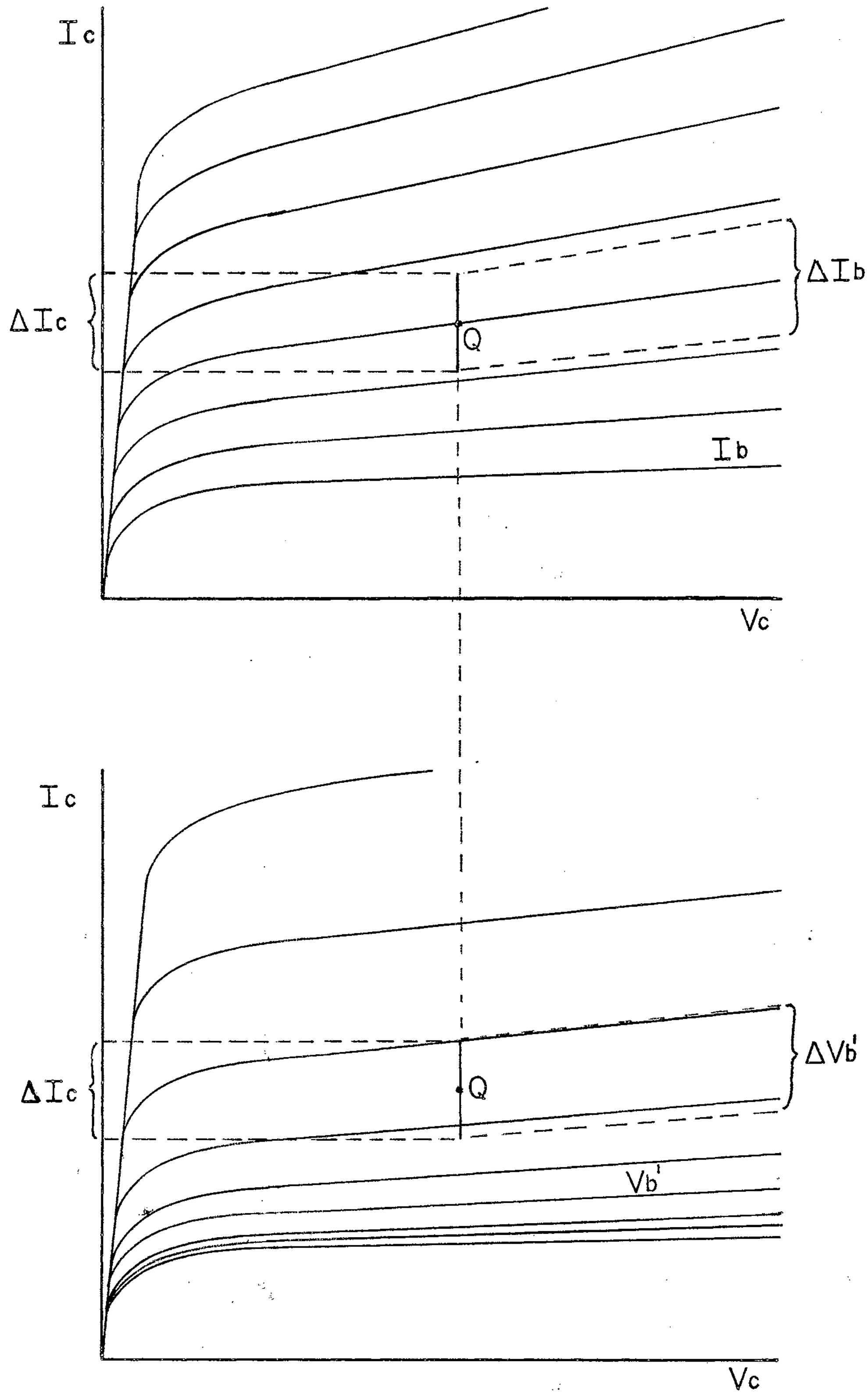

Figure A.2. Determination of h_{ie} and h_{fe} from Output Characteristics.

Summing the voltages around the first loop of Fig. A.1 gives

$$\Delta V_b' = \Delta I_b(R_b + h_{ie}) + \Delta V_c h_{re} \tag{2A}$$

but, $\Delta V_c = 0$; hence,

$$h_{ie} \doteq \frac{\Delta V_b' - \Delta I_b R_b}{\Delta I_b}. \tag{3A}$$

Equation (3A) will give a good approximation of h_{ie} at the low- and mid-frequencies provided that the Q point is the one used when the transistor is placed in the circuit and if ΔI_b is the approximate swing that is expected in the application.

The first characteristic of Fig. A.2 will also give h_{fe}, which is defined as

$$h_{fe} = \frac{\partial I_c}{\partial I_b} \doteq \frac{\Delta I_c}{\Delta I_b}\ ; \quad V_c = \text{Constant}. \tag{4A}$$

Equation (2A) and Fig. A.3 give a method of determining h_{re}. By definition,

$$h_{re} = \frac{\partial V_b}{\partial V_c}\ ; \quad I_b = \text{Constant}. \tag{5A}$$

Since I_b is equal to a constant, $\Delta I_b = 0$, (2A) becomes

$$\Delta V_b' = \Delta V_c h_{re}, \tag{6A}$$

but $\Delta V_b' = \Delta V_b$ when $\Delta I_b = 0$; hence,

$$h_{re} \doteq \frac{\Delta V_b'}{\Delta I_b}. \tag{7A}$$

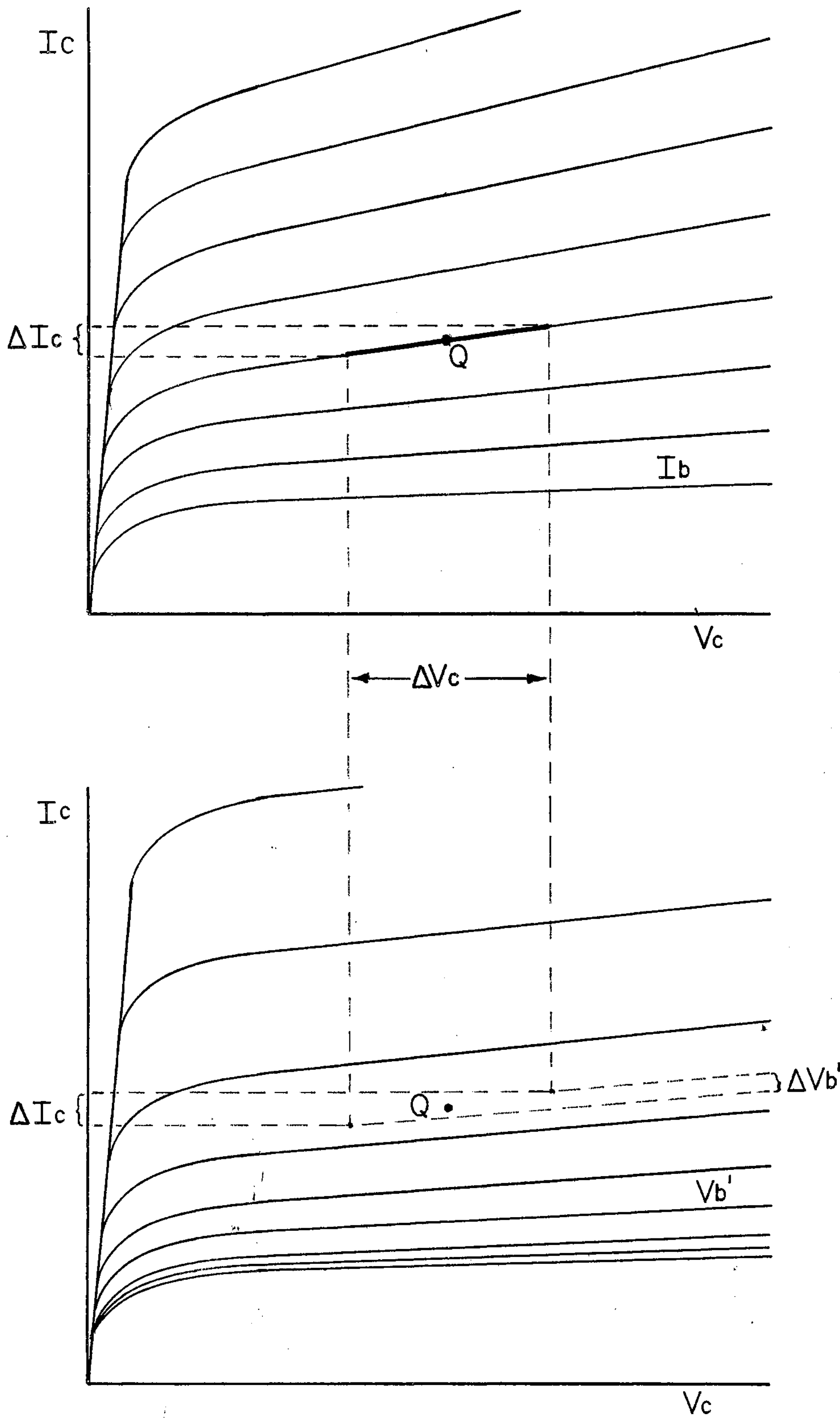

Figure A.3. Determination of h_{re} and h_{oe} from Output Characteristics.

The first characteristic of Fig. A.3 also gives h_{oe}.

$$h_{oe} = \frac{\partial I_c}{\partial V_c} \doteq \frac{\Delta I_c}{\Delta V_c} ; \quad I_b = \text{Constant} . \tag{8A}$$

APPENDIX B

ANALYSIS OF EMITTER-FEEDBACK-PEAKING AMPLIFIER

General Discussion

Before considering the direct analysis of the circuit, some transformations will be made which will be used later. Fig. B.1 may be transformed into that of Fig. B.2 using the following equations.[1]

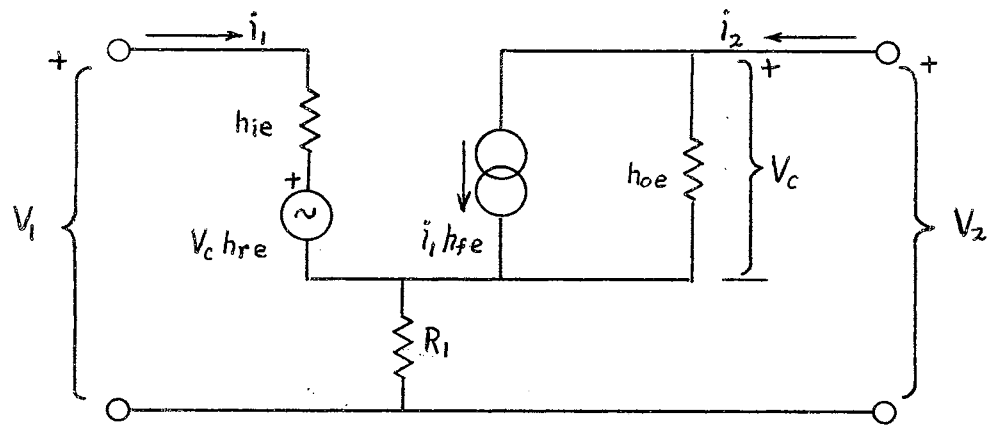

Figure B.1. Equivalent Circuit of Emitter Feedback Stage.

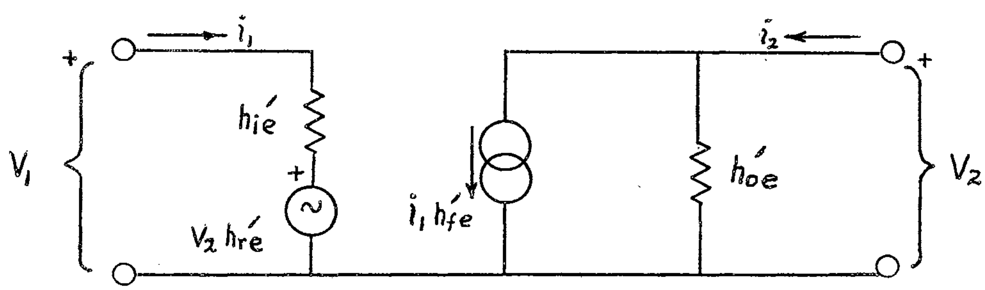

Figure B.2. Transformed Equivalent Circuit of Emitter Feedback Stage.

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[1]Lloyd P. Hunter, Handbook of Semiconductor Electronics, pp. 11-26.

$$h_{ie}' = h_{ie} + \frac{(1 + h_{fe})R_1}{1 + h_{oe}R_1}, \tag{1B}$$

$$h_{re}' = \frac{h_{re} + h_{oe}R_1}{1 + h_{oe}R_1}, \tag{2B}$$

$$h_{fe}' = \frac{h_{fe} - h_{oe}R_1}{1 + h_{oe}R_1}, \tag{3B}$$

$$h_{oe}' = \frac{h_{oe}}{1 + h_{oe}R_1}. \tag{4B}$$

Fig. B.3 may be transformed into that of Fig. B.4 using (5B) through (8B).[1]

$$h_{ie}'' = \frac{h_{ie}R_f}{h_{ie} + R_f}, \tag{5B}$$

$$h_{re}'' = h_{re} + \frac{h_{ie}(1 - h_{re})}{h_{ie} + R_f}, \tag{6B}$$

$$h_{fe}'' = \frac{h_{fe}R_f - h_{ie}}{h_{ie} + R_f} \doteq \frac{h_{fe}R_f}{h_{ie} + R_f}; \quad h_{fe}R_f >> h_{ie}, \tag{7B}$$

$$h_{oe}'' = \frac{(1 - h_{re})(1 + h_{fe})}{h_{ie} + R_f} + h_{oe} = h_{oe} + \frac{1 + h_{fe}}{h_{ie} + R_f}; \quad 1 >> h_{re}. \tag{8B}$$

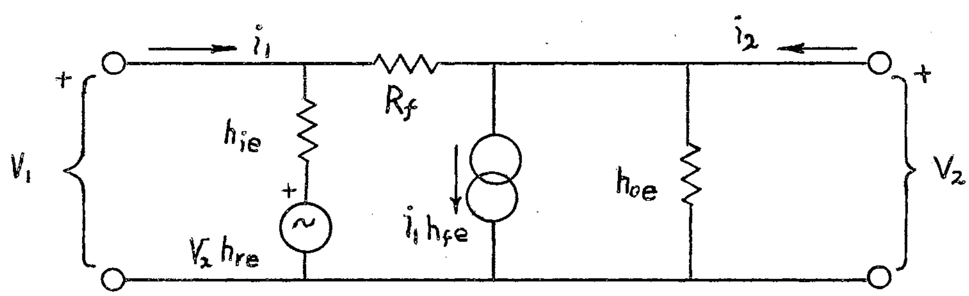

Figure B.3. Equivalent Circuit of Collector to Base Feedback Stage.

[1]Lloyd P. Hunter, Handbook of Semiconductor Electronics, pp. 11-27.

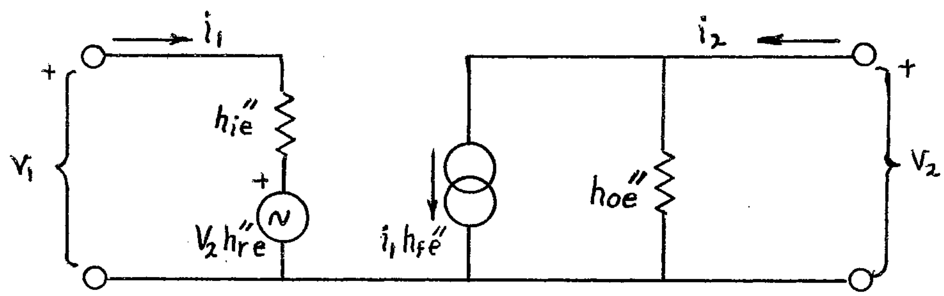

Figure B.4. Transformed Equivalent Circuit of Collector to Base Feedback Stage.

Using the method of Appendix A to determine the h-parameters and (1B) through (8B), the mid-frequency equivalent circuit of Fig. 2 becomes that of Fig. B.5 with the numerical values given in Table B.I where,

$$\frac{1}{R_{x1}} = \frac{1}{R_{21}} + \frac{1}{R_{31}} ; \qquad R_{x1} = 2.22 K\Omega , \tag{9B}$$

$$\frac{1}{R_{x2}} = \frac{1}{R_{22}} + \frac{1}{R_{32}} + \frac{1}{R_{41}} ; \quad R_{x2} = 1.36\ K\Omega , \tag{10B}$$

$$\frac{1}{R_{x3}} = \frac{1}{R_{23}} + \frac{1}{R_{33}} + \frac{1}{R_{42}} ; \quad R_{x3} = 916\ \Omega , \tag{11B}$$

$$\frac{1}{R_{x4}} = \frac{1}{R_{24}} + \frac{1}{R_{34}} + \frac{1}{R_{43}} ; \quad R_{x4} = 676\ \Omega , \tag{12B}$$

$$\frac{1}{R_{x5}} = \frac{1}{R_L} + \frac{1}{R_{44}} ; \qquad \begin{aligned} R_{x5} &= 471\ \Omega \ (R_L = 600\ \Omega) \\ &= 2.3 K\Omega \ (R_L = 10 K\Omega), \end{aligned} \tag{13B}$$

and[1]

$$R_{ij} = \frac{\Delta^{h_{e_j}} R_{1j} + h_{iej}}{1 + h_{oej} R_{1j}} ; \qquad j = 1, 2, 3, 4, \tag{14B}$$

with

[1]Richard F. Shea, Transistor Circuit Engineering, p. 441.

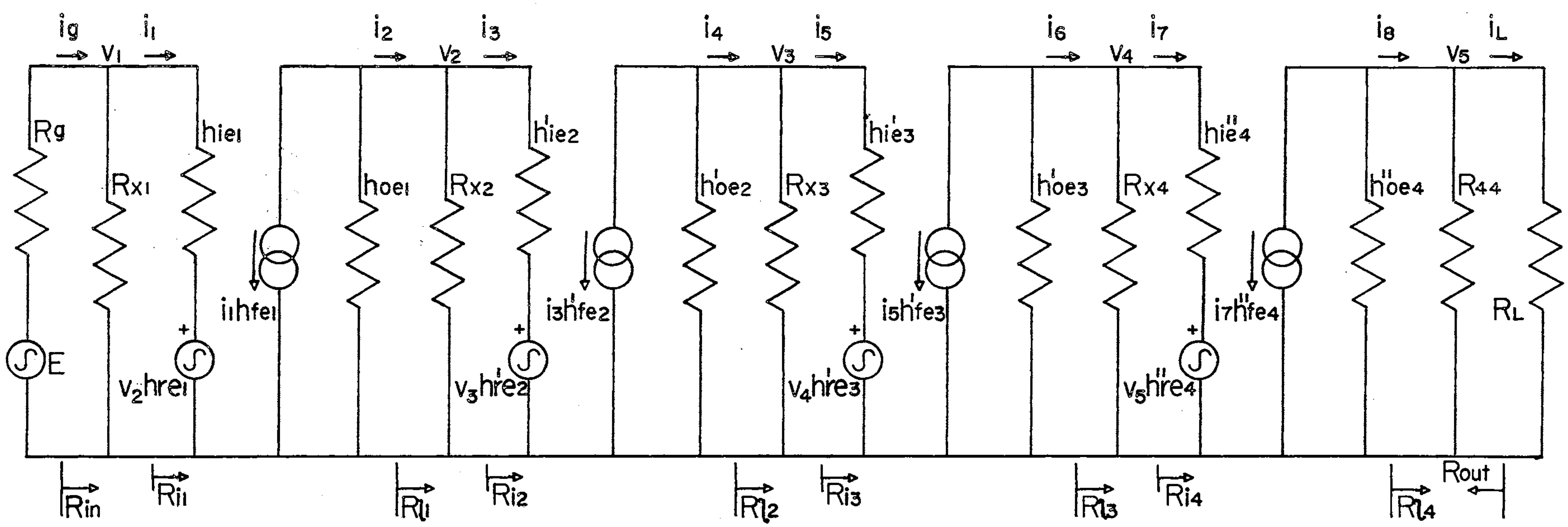

Figure B.5. Transformed Equivalent Circuit of Emitter-Feedback-Peaking Amplifier.

TABLE B.I

h-PARAMETER VALUES FOR FIGURE B.5

Transistor No.	Stage No.	h_{ie} (Ω)	h_{fe} (—)	h_{re} (x10^{-6})	h_{oe} (x10^{-6})	Δ^{he} [1] (x10^{-3})
2N623-15	1	1.26K	55.4	39.0	10	10.40
2N623-14	2	1.03K	98.0	59.5	30	25.10
2N623-12	3	0.50K	47.5	106.0	50	20.10
2N623- 4	4	0.51K	42.8	340.0	120	46.65

Transistor No.	Stage No.	h_{ie}' (Ω)	h_{fe}' (—)	h_{re}' (x10^{-3})	h_{oe}' (x10^{-6})	$\Delta^{he}{}'$ (x10^{-3})
2N623-14	2	59.6K	96.3	17.75	29.5	52
2N623-12	3	31.4K	45.9	32.00	48.3	53

Transistor No.	Stage No.	h_{ie}'' (Ω)	h_{fe}'' (—)	h_{re}'' (x10^{-3})	h_{oe}'' (x10^{-6})	$\Delta^{he}{}''$ (x10^{-3})
2N623- 4	4	0.51K	42.7	4.57	483	50

[1] $\Delta^{he} = h_{ie}\, h_{oe} - h_{re}\, h_{fe}$

$$R_{lj} = \frac{R_{x(j+1)} R_{i(j+1)}}{R_{x(j+1)} + R_{i(j+1)}} ; \quad j = 1,2,3,4. \tag{15B}$$

Mid-Frequency Analysis

The mid-frequency analysis of the amplifier may now be completed as follows with values given when $R_L = 600\Omega$ immediately following each equation and the values when $R_L = 10k\Omega$ in parenthesis.

$$\frac{i_L}{i_8} = \frac{R_{44}}{R_{44} + R_L} = 0.843 \quad (0.231) \tag{16B}$$

$$R_{l4} = R_{x5} = 471\Omega \quad (2.31k\Omega) \tag{17B}$$

$$\frac{i_8}{i_7} = \frac{h_{fe4}''}{1 + h_{oe4}'' R_{l4}} = 34.8 \quad (20.2) \tag{18B}$$

$$R_{i4} = \frac{\Delta^{he_4''} R_{l4} + h_{ie4}''}{1 + h_{oe4}'' R_{l4}} = 434\Omega \quad (295\Omega) \tag{19B}$$

$$\frac{i_7}{i_6} = \frac{R_{x4}}{R_{x4} + R_{i4}} = 0.609 \quad (0.697) \tag{20B}$$

$$R_{l3} = \frac{R_{x4} R_{i4}}{R_{x4} + R_{i4}} = 264\Omega \quad (206\Omega) \tag{21B}$$

$$\frac{i_6}{i_5} = \frac{h_{fe3}'}{1 + h_{oe3}' R_{l3}} = 45.3 \quad (45.4) \tag{22B}$$

$$R_{i3} = \frac{\Delta^{he_3'} R_{l3} + h_{ie3}'}{1 + h_{oe3}' R_{l3}} = 31.0k\Omega \quad (31.1k\Omega) \tag{23B}$$

$$\frac{i_5}{i_4} = \frac{R_{x3}}{R_{x3} + R_{i3}} = 0.02875 \quad (.0286) \tag{24B}$$

$$R_{l2} = \frac{R_{x3}\,R_{i3}}{R_{x3} + R_{i3}} = 890\Omega \quad (890\Omega) \qquad (26B)$$

$$\frac{i_4}{i_3} = \frac{h'_{fe2}}{1 + h'_{oe2} R_{l2}} = 94.0 \quad (94.0) \qquad (27B)$$

$$R_{i2} = \frac{\Delta^{h'e}_{2} R_{l2} + h'_{ie2}}{1 + h'_{oe2} R_{l2}} = 58.1\text{k}\Omega \quad (58.1\text{k}\Omega) \qquad (28B)$$

$$\frac{i_3}{i_2} = \frac{R_{x2}}{R_{x2} + R_{i2}} = 0.0229 \quad (0.0229) \qquad (29B)$$

$$R_{l1} = \frac{R_{x2}\,R_{i2}}{R_{x2} + R_{i2}} = 1.33\text{k}\Omega \quad (1.33\text{k}\Omega) \qquad (30B)$$

$$\frac{i_2}{i_1} = \frac{h_{fe1}}{1 + h_{oe1} R_{l1}} = 54.7 \quad (54.7) \qquad (31B)$$

$$R_{i1} = \frac{\Delta^{he}_{1} R_{l1} + h_{ie1}}{1 + h_{oe1} R_{l1}} = 1.38\text{k}\Omega \quad (1.38\text{k}\Omega) \qquad (32B)$$

$$\frac{i_1}{i_g} = \frac{R_{x1}}{R_{x1} + R_{i1}} = 0.617 \quad (0.617) \qquad (33B)$$

$$R_{in} = \frac{R_{x1}\,R_{i1}}{R_{x1} + R_{i1}} = 854\Omega \quad (854\Omega) \qquad (34B)$$

$$R_{out} = \frac{(h''_{ie4} + R_{x4})\,R_{44}}{h''_{ie4} + R_{x4} + R_{44}(\Delta^{he''}_{4} + h''_{oe4} R_{x4})} = 1.62\text{k}\Omega \quad (1.62\text{k}\Omega) \qquad (35B)$$

$$G_i = \frac{i_L}{i_g} = \frac{i_L}{i_8}\cdot\frac{i_8}{i_7}\cdot\frac{i_7}{i_6}\cdot\frac{i_6}{i_5}\cdot\frac{i_5}{i_4}\cdot\frac{i_4}{i_3}\cdot\frac{i_3}{i_2}\cdot\frac{i_2}{i_1}\cdot\frac{i_1}{i_g}$$

$$= 1688 \quad (303) \qquad (36B)$$

$$G_v = \frac{G_i R_L}{R_{in}} = 1110 \qquad (3550) \tag{37B}$$

$$20 \text{ Log } G_i = 64.5 \text{ db} \qquad (49.7 \text{ db}) \tag{38B}$$

$$20 \text{ Log } G_v = 61.0 \text{ db} \qquad (71.0 \text{ db}) \tag{39B}$$

High-Frequency Analysis Without Compensation

The high frequency analysis will be made using Bode plots and the following high frequency h-parameter approximations.[1]

$$h_{ie}^* \doteq r_b + \frac{r_\epsilon\,(1 + h_{fe})}{1 + j\left(\frac{f}{f_e}\right)} = h_{ie}\left[\frac{1 + j\left(\frac{f}{f_e}\right)\left(\frac{r_b}{h_{ie}}\right)}{1 + j\left(\frac{f}{f_e}\right)}\right], \tag{40B}$$

Where $h_{ie} = r_b + r_\epsilon\,(1 + h_{fe})$ and f_e = common-emitter beta cutoff frequency.

$$h_{fe}^* = \frac{h_{fe}}{1 + j\left(\frac{f}{f_e}\right)}. \tag{41B}$$

When $\Delta^{h}e1 \cdot R_{l1} << h_{ie1}$ and $h_{oe1}\, R_{l1} << 1$, (33B) becomes

$$\frac{i_1}{i_g} = \frac{R_{x1}}{R_{x1} + R_{i1}^*}\,; \quad R_{i1}^* \doteq h_{ie1}^*. \tag{42B}$$

Substituting (40B) in (42B) for R_{i1}^* gives

$$\frac{i_1}{i_g} \doteq \frac{R_{x1}}{R_{x1} + h_{ie}\left[\frac{1 + j\left(\frac{f}{f_{e1}}\right)\left(\frac{r_{b1}}{h_{ie1}}\right)}{1 + j\left(\frac{f}{f_{e1}}\right)}\right]} = \frac{\left[\frac{R_{x1}}{R_{x1} + h_{ie1}}\right]\left[1 + j\left(\frac{f}{f_{e1}}\right)\right]}{1 + j\left(\frac{f}{f_{e1}}\right)\left(\frac{R_{x1} + r_{b1}}{R_{x1} + h_{ie1}}\right)}. \tag{43B}$$

Substituting (41B) in (31B) and assuming $h_{oe1}\, R_{l1} << 1$ gives

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[1]Richard F. Shea, Transistor Circuit Engineering, p. 39.

$$\frac{i_2}{i_1} = \frac{h_{fe1}}{1 + j\left(\frac{f}{f_{e1}}\right)} . \tag{44B}$$

Equations (43B) and (44B) give the over-all gain of the first stage as

$$G_1 \doteq \frac{R_{x1}\, h_{fe1}}{R_{x1} + h_{ie1}} \cdot \frac{1}{1 + j\left(\frac{f}{f_{e1}}\right)\left(\frac{R_{x1} + r_{b1}}{R_{x1} + h_{ie1}}\right)} , \tag{45B}$$

assuming that the transistor collector capacitance cutoff frequency is high compared to the common-emitter beta cutoff frequency. Equation (45B) gives the equation for the cutoff frequency of the first stage as

$$f_1^{co} = \frac{f_{e1}\,(R_{x1} + h_{ie1})}{R_{x1} + r_{b1}} , \tag{46B}$$

Where f_1^{co} is defined by Fig. B.6.

Figure B.6. Bode Plot of (45B).

The second and third stages both have emitter resistors which are not by-passed at all frequencies; therefore, their high frequency analysis differs somewhat from that of the first stage. When $C_{12} = C_{13} = 0$, the frequency dependent parameters become

$$h_{ie}^{\prime *} \doteq h_{ie}^{*} + (1 + h_{fe}^{*})\, R_1 ; \quad h_{oe} R_1 << 1 . \tag{47B}$$

Substituting (40B) and (41B) in (47B) gives

$$h_{ie}^{\prime *} = h_{ie}\left[\frac{1 + j\left(\frac{f}{f_e}\right)\left(\frac{r_b}{h_{ie}}\right)}{1 + j\left(\frac{f}{f_e}\right)}\right] + \left[1 + \frac{h_{fe}}{1 + j\left(\frac{f}{f_e}\right)}\right]$$

$$= \frac{h_{ie} + (1+h_{fe})R_1 + j\left(\frac{f}{f_e}\right)(r_b + R_1)}{1 + j\left(\frac{f}{f_e}\right)}$$

$$= h_{ie}' \left[\frac{1 + j\left(\frac{f}{f_e}\right)\left(\frac{r_b + R_1}{h_{ie}'}\right)}{1 + j\left(\frac{f}{f_e}\right)} \right]. \tag{48B}$$

When $h_{oe}\, R_l << 1 << h_{fe}$,

$$h_{fe}'^{*} = h_{fe}^{*}. \tag{49B}$$

Which is given by (41B). Equations (24B) and (29B) now are of the form

$$\frac{\dot{i}_k}{\dot{i}_j} \doteq \frac{R_x}{R_x + R_i^{*}} \; ; \; R_i^{*} \doteq h_{ie}'^{*} \, , \tag{50B}$$

When $h_{oe}\, R_l << 1$ and $\Delta^{he}\, R_l << h_{ie}$. Substituting (48B) in (50B) for R_i^{*} gives

$$\frac{\dot{i}_k}{\dot{i}_j} \doteq \frac{R_x}{R_x + h_{ie}' \left[\frac{1 + j\left(\frac{f}{f_e}\right)\left(\frac{r_b + R_1}{h_{ie}'}\right)}{1 + j\left(\frac{f}{f_e}\right)} \right]}$$

$$= \frac{R_x}{R_x + h_{ie}'} \cdot \frac{1 + j\left(\frac{f}{f_e}\right)}{1 + j\left(\frac{f}{f_e}\right)\left(\frac{R_x + R_1 + r_b}{R_x + h_{ie}'}\right)} \, . \tag{51B}$$

Equations (22B) and (27B) have the form

$$\frac{\dot{i}_n}{\dot{i}_k} \doteq h_{fe}'^{*} \; ; \quad h_{oe}' R_l << 1 \, , \tag{52B}$$

Where $h_{fe}'^{*} \doteq h_{fe}^{*} \doteq \frac{h_{fe}}{1 + j\frac{f}{f_e}}$. The gain of the second or third stages is the product of (51B) and (52B).

$$G_{1K} \doteq \frac{R_{xK}\, h_{feK}}{R_{xK} + h'_{ieK}} \cdot \frac{1}{1 + j\left(\frac{f}{f_{eK}}\right)\left(\frac{R_{xK} + R_{1K} + r_{bK}}{R_{xK} + h'_{ieK}}\right)}\,; \quad k = 2, 3. \tag{53B}$$

It should be noted again that (47B) through (53B) apply only when the emitter resistor is not shunted by a capacitor or when the frequency is too low for the capacitor to have any noticeable effect. Equation (53B) gives the corner frequency of the stages to be

$$f_K^{co} = \frac{f_{eK}(R_{xK} + h'_{ieK})}{R_{xK} + R_{1K} + r_{bK}}\,; \quad K = 2, 3 \tag{54B}$$

Where f_k^{co} is defined by Fig. B.6 by substituting f_k^{co} for f_1^{co}.

The fourth stage high frequency gain will approximate that of the first stage equations since the collector to base feedback resistance R_{f4} is so large. This means that (42B) through (46B) will apply to the fourth stage by changing the numbers of the subscripts.

Equation (45B) and (53B) will then give the high frequency gain when the amplifier has no high frequency compensation.

$$G_{(high)} \doteq \frac{G_{(mid)}}{\left\{\left[1 + j\frac{f(R_{x1} + r_{b1})}{f_{e1}(R_{x1} + h_{ie1})}\right]\left[1 + j\frac{f(R_{x2} + R_{12} + r_{b1})}{f_{e2}(R_{x2} + h'_{ie2})}\right]\left[1 + j\frac{f(R_{x3} + R_{13} + r_{b3})}{f_{e3}(R_{x3} + h'_{ie3})}\right]\left[1 + j\frac{f(R_{x4} + r_{b4})}{f_{e4}(R_{x4} + h_{ie4})}\right]\right\}} \tag{55B}$$

but, $R_{xk} + R_{1k} + r_{bk} \ll R_{xk} + h_{ie}{'}_k$; $k = 2, 3$; therefore, in the desired frequency range

$$G_{(high)} \doteq \frac{G_{(mid)}}{\left[1 + j\frac{f(R_{x1} + r_{b1})}{f_{e1}(R_{x1} + h_{ie1})}\right]\left[1 + j\frac{f(R_{x4} + r_{b4})}{f_{e4}(R_{x4} + h_{ie4})}\right]}. \tag{56B}$$

The gain-bandwidth product of the transistor used in the first stage was found to be 50×10^6, and that of the last three stages was found to be

75 x 10^6; therefore,

$$f_{ek} = \frac{GBW_K}{h_{fe_K}} \quad ; \; f_{e_1} = 0.904 \text{ MC}, \quad f_{e4} = 1.755 \text{ MC} \tag{57B}$$

and r_b of all the stages was taken as 75Ω as given by the manufacturer. Equation (46B) gives f_1^{co} = 1.37 MC and f_4^{co} = 2.77 MC. The frequency response for (56B) is shown in Fig. B.7.

High-Frequency Analysis with Compensation

If the bandwidth of the amplifier is increased by emitter feedback compensation, C_{12} and C_{13} take on values other than zero, (47B) through (57B) will have to be revised as follows:

$$R_{iK}^* \doteq h_{ie_K}^* + \frac{(1 + h_{fe_K}^*) R_{1K}}{1 + j\omega R_{1K} C_{1K}} \quad ; \quad K = 2, 3. \tag{58B}$$

Substituting (40B) and (41B) in (58B) gives

$$R_{iK}^* \doteq h_{ie_K} \left[\frac{1 + j\dfrac{f \, r_{bK}}{f_{eK} h_{ie_K}}}{1 + j\dfrac{f}{f_{eK}}} \right] + \left[\frac{R_{1K}}{1 + j\dfrac{\omega}{\omega_K}} \right] \left[1 + \frac{h_{fe_K}}{1 + j\dfrac{f}{f_{eK}}} \right]$$

$$= \frac{h_{ie_K}\left[1 + j\dfrac{\omega \, r_{bK}}{\omega_{eK} h_{ie_K}}\right]\left[1 + j\dfrac{\omega}{\omega_K}\right] + \left[1 + h_{fe_K} + j\dfrac{\omega}{\omega_{eK}}\right] R_{1K}}{\left[1 + j\dfrac{\omega}{\omega_{eK}}\right]\left[1 + j\dfrac{\omega}{\omega_K}\right]} \tag{59B}$$

Where, $\omega_k = \dfrac{1}{R_{1k} \, C_{1k}}$. Making the substitution $p = j\omega$ in (59B) and simplifying gives

$$R_{iK}^* \doteq \frac{r_{bK}\left[p + \dfrac{\omega_{eK} h_{ie_K}}{r_{bK}}\right]\left[p + \omega_K\right] + R_{1K}\,\omega_K \left[p + (1 + h_{fe_K})\,\omega_{eK}\right]}{(p + \omega_{eK})(p + \omega_K)} \tag{60B}$$

Then

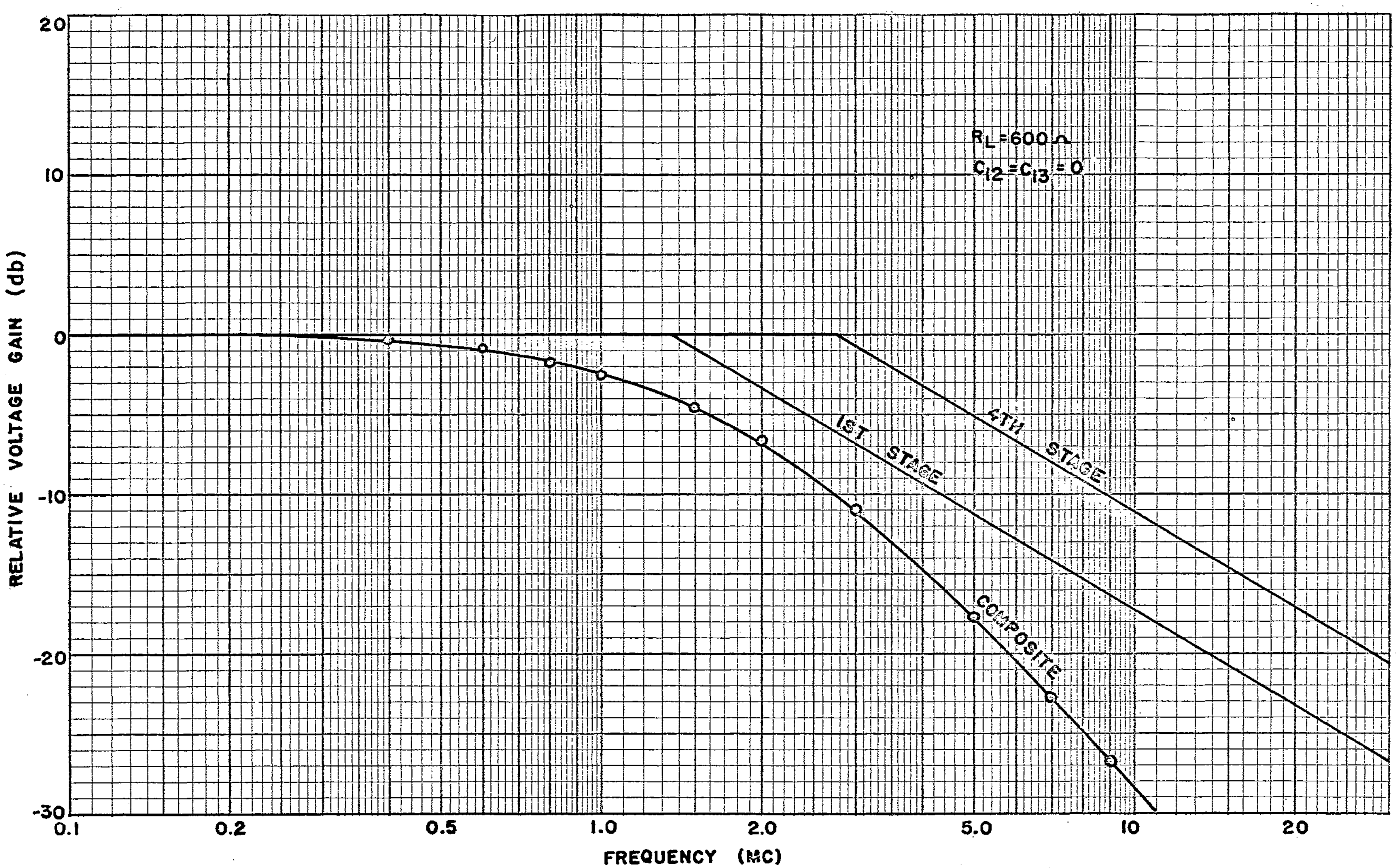

Figure B.7. Frequency Response of Emitter-Feedback-Peaking Amplifier When $C_{12} = C_{13} = 0$.

$$\frac{i_n}{i_m} = \frac{R_{xk}}{R_{xk} + R_{ik}^{*}} \; ; \quad k=2,3 \; , \quad m=2,4 \; , \quad n=3,5. \tag{61B}$$

Substituting (60B) in (61B) results in

$$\frac{i_n}{i_m} \doteq \frac{R_{xk}(p+\omega_{ek})(p+\omega_k)}{\left\{\begin{array}{c} p^2\left[R_{xk}+r_{bk}\right] + p\left[R_{xk}(\omega_k+\omega_{ek}) + r_{bk}\left(\omega_k + \frac{\omega_{ek} h_{iek}}{r_{bk}}\right) + R_{ik}\omega_{ik}\right] + \\ \omega_k \omega_{ek}\left[R_{xk} + h_{iek} + R_{ik}(1+h_{fek})\right] \end{array}\right\}} . \tag{62B}$$

Let

$$\left.\begin{aligned} A_k &= R_{xk} + r_{bk} \\ B_k &= R_{xk}(\omega_k+\omega_{ek}) + r_{bk}\left(\omega_k + \frac{\omega_{ek} h_{iek}}{r_{bk}}\right) + R_{ik}\omega_{ik} \\ C_k &= \omega_k \omega_{ek}\left[R_{xk} + h_{iek} + R_{ik}(1+h_{fek})\right] \doteq \omega_k \omega_{ek}\left[R_{xk} + h_{iek}\right] \end{aligned}\right\} \tag{63B}$$

Substituting (63B) in (62B) gives

$$\frac{i_n}{i_m} = \frac{\omega_k \omega_{ek} R_{xk}}{C_k} \cdot \frac{\left(p\frac{1}{\omega_{ek}}+1\right)\left(p\frac{1}{\omega_k}+1\right)}{\frac{A_k}{C_k}p^2 + \frac{B_k}{C_k}p + 1} \tag{64B}$$

The characteristic equation of (64B) may be plotted using the following equations when the roots are complex.[1] Fig. B.8 defines the terms.

$$\omega_{0k} = \sqrt{C_k / A_k} \tag{65B}$$

$$\zeta_k = \frac{B_k \omega_{0k}}{2\, C_k} \tag{66B}$$

$$G_k = \frac{1}{\left[1 - \left(\frac{\omega}{\omega_{0k}}\right)^2\right] + j\left[2\zeta_k \frac{\omega}{\omega_{0k}}\right]} \tag{67B}$$

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[1]George J. Thaler and Robert G. Brown, Servomechanism Analysis, pp. 243-246.

$$G_{(max)K} = 1/2\zeta_K\sqrt{1-\zeta_K^2} \tag{68B}$$

$$G_{(\omega_o)K} = 1/2\zeta_K \tag{69B}$$

$$\omega_{(G_{max})K} = \omega_{0K}\sqrt{1-\zeta_K^2} \tag{70B}$$

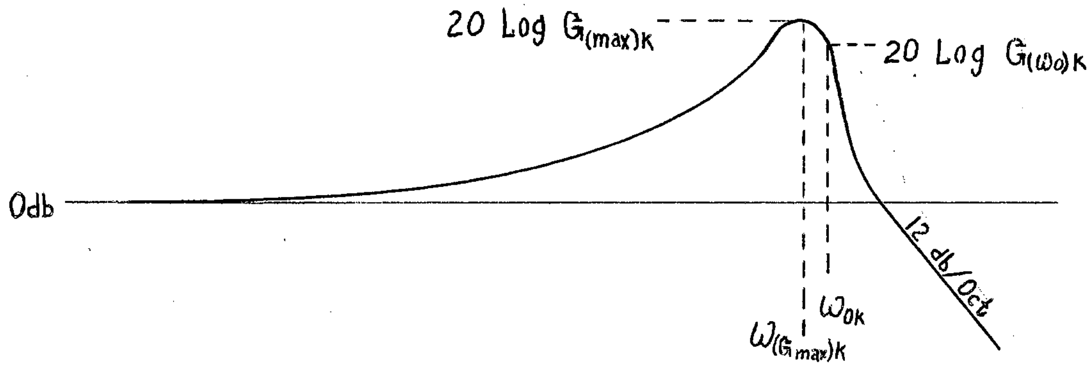

Figure B.8. Characteristic Equation Plot of (64B).

The gain of the second or third stage is the product of (52B) and (64B) if (67B) is substituted for the denominator of (64B).

$$G_{(high)_K} \doteq \frac{G_{(mid)K}\left(1+j\frac{\omega}{\omega_K}\right)}{\left[1-\left(\frac{\omega}{\omega_{0K}}\right)^2\right] + j\left[2\zeta_K\frac{\omega}{\omega_{0K}}\right]} \tag{71B}$$

Equation (71B) is plotted in Fig. B.9 for the second and third stage. The gain of the amplifier is given by (45B) and (71B).

$$G_{(high)} \doteq \frac{G_{(mid)}\left(1+j\frac{f}{f_2}\right)\left(1+j\frac{f}{f_3}\right)}{\left[\left\{1+j\frac{f(R_{x1}+r_{b1})}{f_{e1}(R_{x1}+h_{ie1})}\right\}\left\{\left[1-\left(\frac{f}{f_{02}}\right)^2\right]+j\left[2\zeta_2\frac{f}{f_{02}}\right]\right\} \left\{\left[1-\left(\frac{f}{f_{03}}\right)^2\right]+j\left[2\zeta_3\frac{f}{f_{03}}\right]\right\}\left\{1+j\frac{f(R_{x4}+r_{b4})}{f_{e4}(R_{x4}+h_{ie4})}\right\}\right]} \tag{72B}$$

The corner frequencies of the portions of (72B) due to the first and fourth stage were previously determined to be

$$f_1^{co} = 1.37\,\mathrm{MC} \quad \text{and} \quad f_4^{co} = 2.77\,\mathrm{MC} \tag{73B}$$

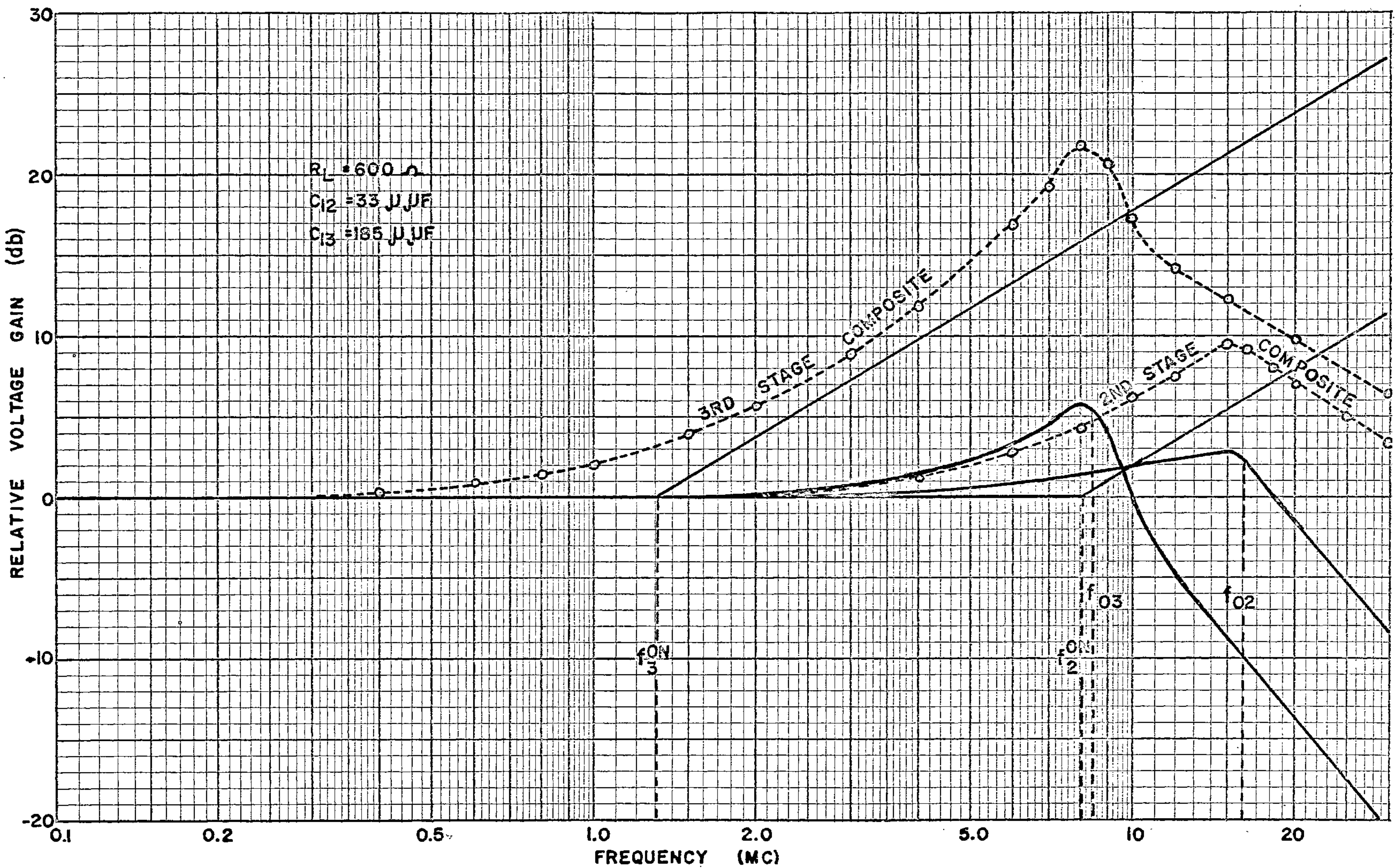

Figure B.9. Second and Third Stage Response With Emitter-Feedback-Peaking.

Substituting circuit values in (63B), (65B), (66B), (70B) and $\omega_k = \frac{1}{R_{1k}C_{1k}}$ gives

$$\left.\begin{array}{llll} f_2^{ON} = 8.03 \text{ MC}; & & f_3^{ON} = 1.31 \text{ MC} \\ f_{02} = 16.32 \text{ MC}; & & f_{03} = 8.35 \text{ MC} \\ f_{G(max)2} = 15 \text{ MC}; & & f_{G(max)3} = 8.05 \text{ MC} \\ \zeta_2 = 0.392; & & \zeta_3 = 0.267 \end{array}\right\} . \qquad (74B)$$

Using the values found in (73B) and (74B), (72B) is plotted in Fig. B. 10.

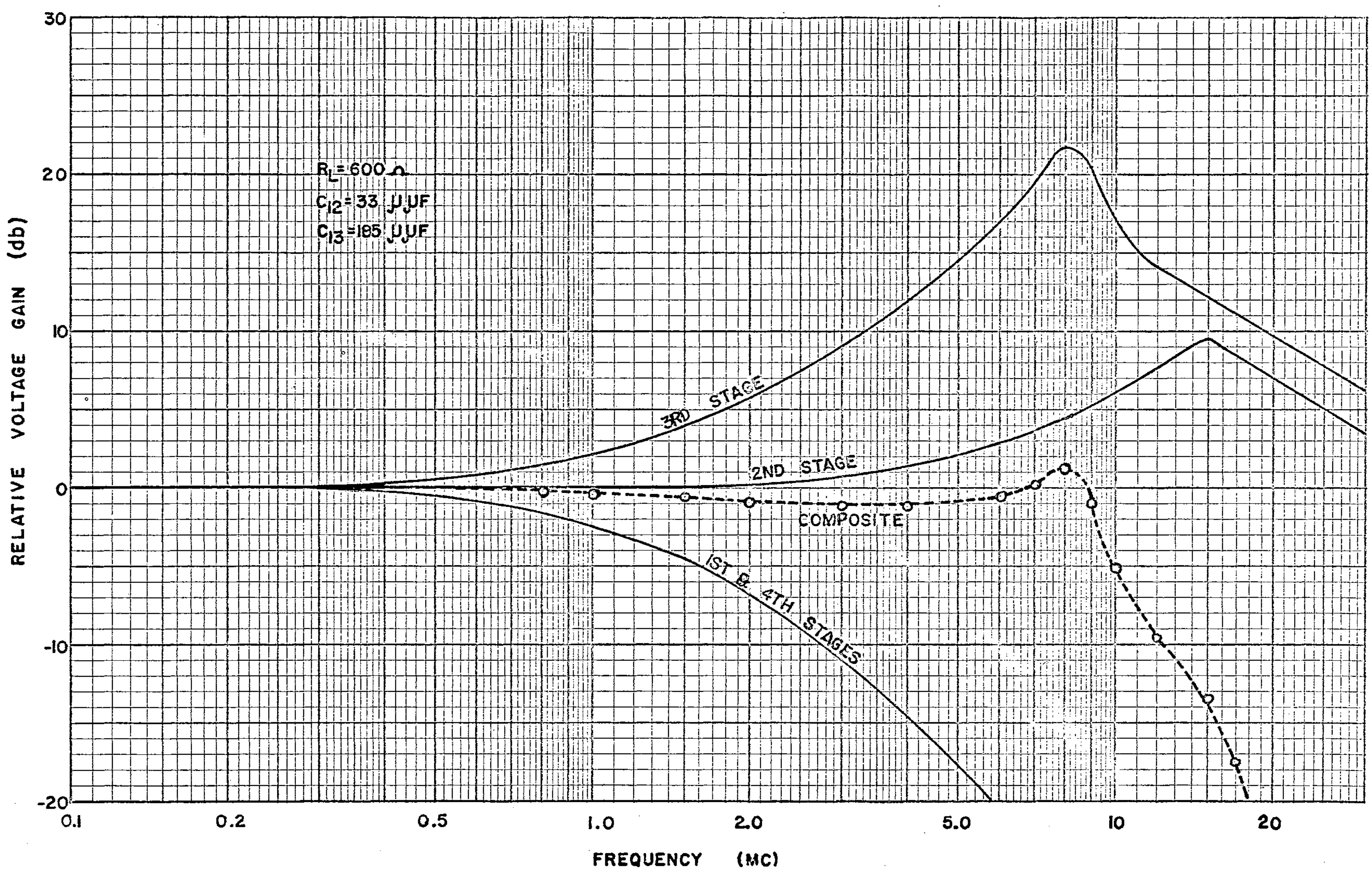

Figure B.10. Frequency Response of Emitter-Feedback-Peaking Amplifier With Compensation.

APPENDIX C

ANALYSIS OF COLLECTOR TO BASE-FEEDBACK-PEAKING AMPLIFIER

General Discussion:

The h-parameters and their transformed values for this amplifier are given in Table C.I. The values in the table were obtained using the method of Appendix A and (1B) through (8B). Fig. B.5 may be used as the equivalent circuit of this amplifier changing the prime parameters of the third stage to double-prime parameters. Equation (9B) through (15B) are applicable to this amplifier because the two amplifiers use the same biasing networks.

Mid-Frequency Analysis

The mid-frequency analysis of this amplifier may be performed using (16B) through (39B) by changing the prime parameters of (22B), (23B), and (24B) to double-prime parameters and using the numerical values of Table C.I. The result obtained using the new values are listed below with the value when $R_L = 600\,\Omega$ listed first and the value when $R_L = 10\ K\Omega$ in parenthesis.

$$\left.\begin{array}{llllll}
i_L/i_8 = & 0.842 & (0.231) & R_{l4} = & 471\,\Omega & (2.31\ K\Omega) \\
i_8/i_7 = & 1.453 & (0.309) & R_{i4} = & 26.2\,\Omega & (14.43\,\Omega) \\
i_7/i_6 = & 0.963 & (0.980) & R_{l3} = & 25.2\,\Omega & (14.17\,\Omega) \\
i_6/i_5 = & 38.4 & (44.0) & R_{i3} = & 381\,\Omega & (434\,\Omega) \\
i_5/i_4 = & 0.707 & (0.678) & R_{l2} = & 269\,\Omega & (294\,\Omega) \\
i_4/i_3 = & 95.6 & (95.50) & R_{i2} = & 59.2\ K\Omega & (59.2\ K\Omega)
\end{array}\right\} \quad (1C)$$

Cont'd.

TABLE C.I

h-PARAMETER VALUES FOR APPENDIX C

Transistor No.	Stage No.	h_{ie} (Ω)	h_{fe} (—)	h_{re} ($x10^{-6}$)	h_{oe} ($x10^{-6}$)	Δh_e ($x10^{-3}$)
2N623-15	1	1.26K	55.4	39.0	10	10.4
2N623-14	2	1.03K	98.0	59.5	30	25.1
2N623-11	3	640	66.0	205	80	34.7
2N623-12	4	500	47.5	106	50	20.1

Transistor No.	Stage No.	h_{ie}' (Ω)	h_{fe}' (—)	h_{re}' ($x10^{-3}$)	h_{oe}' ($x10^{-6}$)	$\Delta h_e'$ ($x10^{-3}$)
2N623-14	2	59.63K	96.3	17.75	29.5	52

Transistor No.	Stage No.	h_{ie}'' (Ω)	h_{fe}'' (—)	h_{re}'' ($x10^{-3}$)	h_{oe}'' ($x10^{-6}$)	$\Delta h_e''$ ($x10^{-3}$)
2N623-11	3	536	54.7	163	17.1	260
2N623-12	4	299	28.1	399	39.0	440

$$\left.\begin{array}{llllll} i_3/i_2 = & 0.01525 & (0.01525) & R_{l1} = & 904\,\Omega & (904\,\Omega) \\ i_2/i_1 = & 54.9 & (54.9) & R_{i1} = & 1.26\ K\Omega & (1.26\ K\Omega) \\ i_1/i_g = & 0.638 & (0.638) & R_{in} = & 804\,\Omega & (804\,\Omega) \\ & & & R_{out} = & 100\,\Omega & (100\,\Omega) \end{array}\right\} \quad (1C)$$

$$\left.\begin{array}{llll} G_i = 1630 & (106.2) & 20 \text{ Log } G_i = 64.4 \text{ db} & (40.5 \text{ db}) \\ G_v = 1135 & (1320) & 20 \text{ Log } G_v = 61.1 \text{ db} & (62.5 \text{ db}) \end{array}\right\} \quad (2C)$$

High-Frequency Analysis Without Compensation

The high-frequency analysis will make use of (40B) and (41B) with the addition of [1]

$$h_{oe}^{*} \doteq \frac{h_{oe}}{1 + j\frac{\omega}{\omega_e}} . \tag{3C}$$

Equation (42B) through (54B) are still applicable to the high-frequency analysis of the first and second stages when the second stage is uncompensated, $C_{12} = 0$. For convenience the equations for the gain and corner frequencies of the two stages are rewritten as

$$G_{(high)_1} = \frac{i_2}{i_g} \doteq \frac{G_{(mid)1}}{1 + j\frac{f(R_{x1} + r_{b1})}{f_{e1}(R_{x1} + h_{ie_1})}} . \tag{45B}$$

$$f_1^{co} \doteq \frac{f_{e1}(R_{x1} + h_{ie1})}{R_{x1} + r_{b1}} . \tag{46B}$$

$$G_{(high)_2} = \frac{i_4}{i_2} \doteq \frac{G_{(mid)2}}{1 + j\frac{f(R_{x2} + R_{12} + r_{b2})}{f_{e2}(R_{x2} + h_{ie_2}')}} . \tag{53B}$$

$$f_2^{co} \doteq \frac{f_{e2}(R_{x2} + h_{ie2}')}{R_{x2} + R_{12} + r_{b2}} . \tag{54B}$$

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[1]Richard F. Shea, Transistor Circuit Engineering, p. 39.

The frequency response of the first and second stages is shown in Fig. C.1.

The values of R_{i3} and R_{l3} as defined by (14B) and (15B) must be known to make a high-frequency analysis of the third stage, but they now become the frequency dependent quantities,

$$R_{i3}^{*} = \frac{\Delta^{he}{}_{3}^{\prime\prime *} R_{l3}^{*} + h_{ie_3}^{\prime\prime *}}{1 + h_{oe_3}^{\prime\prime *} R_{l3}^{*}} \tag{4C}$$

and

$$R_{l3}^{*} = \frac{R_{x4} R_{i4}^{*}}{R_{x4} + R_{i4}^{*}}. \tag{5C}$$

These equations become too complicated to be of much practical use. R_{i4} in (1C) was determined to be 26.2Ω for $R_L = 600\Omega$. Equation (40B) gives $R_{i4}^{*} \doteq r_{b4}$; $f \gg f_e$, hence if we make the assumption that $R_{i4}^{*} = R_{i4}$; $f < f_e$ and $R_{i4}^{*} = r_{b4}$; $f \gg f_e$, the following equation will result.

$$R_{i4}^{*} \doteq R_{i4}\left[\frac{1 + p\frac{1}{\omega_{e4}}}{1 + p\frac{R_{i4}}{\omega_{e4} r_{b4}}}\right] \quad ; R_{i4} < r_{b4}. \tag{6C}$$

Making the same assumptions for R_{i3}^{*} ($R_{i3} = 381\Omega$) gives

$$R_{i3}^{*} \doteq R_{i3}\left[\frac{1 + p\frac{r_{b3}}{\omega_{e3} R_{i3}}}{1 + p\frac{1}{\omega_{e3}}}\right] \quad ; R_{i3} > r_{b3}. \tag{7C}$$

Proceeding with the analysis,

$$\frac{i_5}{i_4} = \frac{R_{x3}}{R_{x3} + R_{i3}^{*}}. \tag{8C}$$

Substituting (7C) in (8C) and simplifying results in

$$\frac{i_5}{i_4} \doteq \frac{R_{x3}}{R_{x3} + R_{i3}} \cdot \left[\frac{1 + p\frac{1}{\omega_{e3}}}{1 + p\frac{R_{x3} + r_{b3}}{\omega_{e3}(R_{x3} + R_{i3})}}\right]. \tag{9C}$$

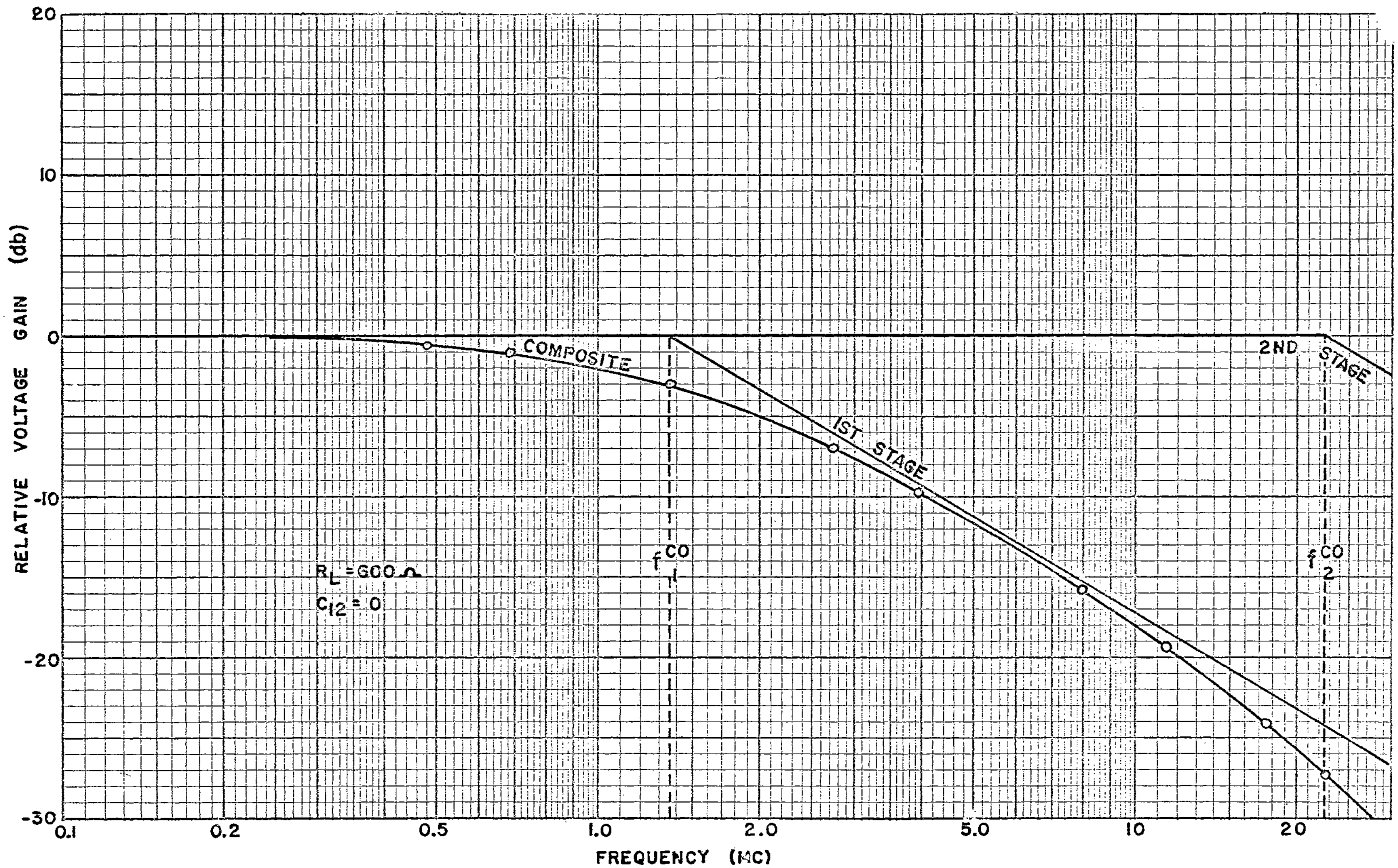

Figure C.1. Frequency Response of First and Second Stages of Collector to Base-Feedback-Peaking Amplifier Without Compensation.

Then

$$\frac{i_6}{i_5} = \frac{h_{fe3}''^{*}}{1 + h_{oe3}''^{*} R_{l3}^{*}} . \tag{10C}$$

Equation (7B) gives

$$h_{fe}''^{*} \doteq \frac{h_{fe}^{*} Z_f}{h_{ie}^{*} + Z_f} ; \quad h_{fe} Z_f >> h_{ie} . \tag{11C}$$

Further, (8B) gives

$$h_{oe}''^{*} \doteq h_{oe}^{*} + \frac{1 + h_{fe}^{*}}{h_{ie}^{*} + Z_f} . \tag{12C}$$

Substituting (11C) and (12C) in (10C) and making four consecutive approximations give

$$\frac{i_6}{i_5} \doteq \frac{\dfrac{h_{fe3}^{*} Z_{f3}}{h_{ie3}^{*} + Z_{f3}}}{1 + \dfrac{[h_{oe3}^{*}(h_{ie3}^{*} + Z_{f3}) + 1 + h_{fe3}^{*}] R_{l3}^{*}}{h_{ie3}^{*} + Z_{f3}}} ; h_{fe3}^{*} Z_{f3} >> h_{ie3}^{*} \tag{13C}$$

$$\frac{i_6}{i_5} \doteq \frac{h_{fe3}^{*} Z_{f3}}{(h_{ie3}^{*} + Z_{f3})(1 + h_{oe3}^{*} R_{l3}^{*}) + R_{l3}^{*} h_{fe3}^{*}} ; \quad h_{fe3}^{*} >> 1 \tag{14C}$$

$$\frac{i_6}{i_5} \doteq \frac{h_{fe3}^{*} Z_{f3}}{h_{ie3}^{*} + Z_{f3} + R_{l3}^{*} h_{fe3}^{*}} ; \quad 1 >> h_{oe3}^{*} R_{l3}^{*} \tag{15C}$$

$$\frac{i_6}{i_5} \doteq \frac{Z_{f3}}{R_{l3}^{*}} ; \quad h_{ie3}^{*} + Z_{f3} << R_{l3}^{*} h_{fe3}^{*} \tag{16C}$$

The approximation made in (16C) is not valid for the third stage but is valid for the fourth stage. Substituting (6C) in (5C)

$$R_{l3}^{*} \doteq \frac{R_{x4} R_{i4}}{R_{x4} + R_{i4}\left[\dfrac{1 + p\dfrac{1}{\omega_{e4}}}{1 + p\dfrac{R_{i4}}{\omega_{e4} r_{b4}}}\right]} , \tag{17C}$$

then simplifying (17C)

$$R_{l3}^* \doteq R_{l3}\left[\frac{1+p\dfrac{1}{\omega_{e4}}}{1+p\dfrac{R_{i4}(r_{b4}+R_{x4})}{\omega_{e4}r_{b4}(R_{x4}+R_{i4})}}\right]. \tag{18C}$$

Substituting (40B), (41B), (18C), and $Z_{f3} = R_{f3}$ in (15C) gives

$$\frac{i_6}{i_5} \doteq \frac{\left[\dfrac{R_{f3}h_{fe3}}{1+p\dfrac{1}{\omega_{e3}}}\right]}{h_{ie3}\left[\dfrac{1+p\dfrac{r_{b3}}{\omega_{e3}h_{ie3}}}{1+p\dfrac{1}{\omega_{e3}}}\right]+R_{f3}+R_{l3}\left[\dfrac{1+p\dfrac{1}{\omega_{e4}}}{1+p\dfrac{R_{i4}(r_{b4}+R_{x4})}{\omega_{e4}r_{b4}(R_{x4}+R_{i4})}}\right]\left[\dfrac{h_{fe3}}{1+p\dfrac{1}{\omega_{e3}}}\right]}. \tag{19C}$$

Simplifying (19C) results in

$$\frac{i_6}{i_5} \doteq \frac{h_{fe3}R_{f3}\left[1+p\dfrac{y}{\omega_{e4}}\right]}{\left\{p^2\left[\dfrac{y(r_{b3}+R_{f3})}{\omega_{e3}\omega_{e4}}\right]+p\left[\dfrac{r_{b3}+R_{f3}}{\omega_{e3}}+\dfrac{y(h_{ie3}+R_{f3})+R_{l3}h_{fe3}}{\omega_{e4}}\right]+\left[h_{ie3}+R_{f3}+R_{l3}h_{fe3}\right]\right\}}, \tag{20C}$$

where, $y = \dfrac{R_{i4}(r_{b4}+R_{x4})}{r_b(R_{x4}+R_{i4})}$.

Let

$$\left.\begin{aligned} A &= \frac{y(r_{b3}+R_{f3})}{\omega_{e3}\omega_{e4}} \\ B &= \frac{r_{b3}+R_{f3}}{\omega_{e3}}+\frac{y(h_{ie3}+R_{f3})+R_{l3}h_{fe3}}{\omega_{e4}} \\ C &= h_{ie3}+R_{f3}+R_{l3}h_{fe3} \end{aligned}\right\} \tag{21C}$$

then,

$$\frac{i_6}{i_5} \doteq \frac{h_{fe3}R_{f3}}{C}\cdot\frac{1+p\dfrac{y}{\omega_{e4}}}{\dfrac{A}{C}p^2+\dfrac{B}{C}p+1}. \tag{22C}$$

The characteristic equation of (22C) may be factored, without complex root, using the numerical values of this particular amplifier. The product of (9C) and (22C) is the gain of the third stage.

$$G_{(high)3} \doteq G_{(mid)3} \frac{\left[1 + p\frac{1}{\omega_{e3}}\right]\left[1 + p\frac{y}{\omega_{e4}}\right]}{\left[1 + p\frac{R_{x3} + r_{b3}}{\omega_{e3}(R_{x3} + R_{i3})}\right]\left[\frac{A}{C}p^2 + \frac{B}{C}p + 1\right]}. \tag{23C}$$

From (23C), the following corner frequency equations are obtained.

$$\left.\begin{aligned} f_{13}^{on} &= f_{e3} & f_{43}^{co} &= \frac{B + \sqrt{B^2 - 4AC}}{4\pi A} \\ f_{23}^{on} &= f_{e4}/y & f_{53}^{co} &= \frac{B - \sqrt{B^2 - 4AC}}{4\pi A} \\ f_{33}^{co} &= \frac{f_{e3}(R_{x3} + R_{i3})}{R_{x3} + r_{b3}} \end{aligned}\right\} \tag{24C}$$

Another method of obtaining i_6/i_5 should be pointed out at this time. First assume

$$h_{fe}^{\prime\prime *} \doteq \frac{h_{fe}^{\prime\prime}}{1 + p\frac{1}{\omega_e}} \tag{25C}$$

and

$$h_{oe}^{\prime\prime *} \doteq \frac{h_{oe}^{\prime\prime}}{1 + p\frac{1}{\omega_e}} \tag{26C}$$

Substituting (18C), (25C), and (26C) in (10C) and simplifying, the following result is obtained.

$$\frac{\dot{i}_6}{i_5} \doteq \frac{h_{fe3}^{\prime\prime}}{1 + h_{oe3}^{\prime\prime} R_{l3}} \cdot \frac{1 + p\frac{y}{\omega_{e4}}}{p^2\left[\frac{y}{\omega_{e3}\,\omega_{e4}(1 + h_{oe3}^{\prime\prime} R_{l3})}\right] + p\left[\frac{\frac{1}{\omega_{e3}} + \frac{y + h_{oe3}^{\prime\prime} R_{l3}}{\omega_{e4}}}{1 + h_{oe3}^{\prime\prime} R_{l3}}\right] + 1} \tag{27C}$$

This equation gives approximately the same result as (22C). The response curve of (23C) is shown in Fig. C.2.

Analysis of the uncompensated fourth stage, $L_{f4} = 0$, begins as follows:

$$\frac{i_7}{i_6} = \frac{R_{x4}}{R_{x4} + R_{i4}^*} \tag{28C}$$

In this stage, $R_{i4} < r_b$; hence, (6C) applies. Substituting this equation in (28C) gives

$$\frac{i_7}{i_6} \doteq \frac{R_{x4}}{R_{x4} + R_{i4}\left[\dfrac{1 + p\dfrac{1}{\omega_{e4}}}{1 + p\dfrac{R_{i4}}{\omega_{e4} r_{b4}}}\right]} \,. \tag{29C}$$

Simplifying (29C),

$$\frac{i_7}{i_6} \doteq \frac{R_{x4}}{R_{x4} + R_{i4}} \cdot \frac{1 + p\dfrac{R_{i4}}{\omega_{e4} r_{b4}}}{1 + p\dfrac{R_{i4}(R_{x4} + r_{b4})}{\omega_{e4} r_{b4}(R_{x4} + R_{i4})}} \,. \tag{30C}$$

When the fourth stage is uncompensated ($L_{f4} = 0$), (16C) becomes

$$\frac{i_8}{i_7} \doteq \frac{R_{f4}}{R_{l4}^*} \quad ; \quad R_{f4} = \frac{R_{f14} R_{f24}}{R_{f14} + R_{f24}} \,, \tag{31C}$$

but, $R_{l4}^* = R_{l4}$ so (31C) becomes simply

$$\frac{i_8}{i_7} \doteq \frac{R_{f4}}{R_{l4}} \,. \tag{32C}$$

The product of (30C) and (32C) then gives the uncompensated gain of the fourth stage to be

$$G_{(high)4} \doteq G_{(mid)4} \, \frac{1 + p\dfrac{R_{i4}}{\omega_{e4} r_{b4}}}{1 + p\dfrac{R_{i4}(R_{x4} + r_{b4})}{\omega_{e4} r_{b4}(R_{x4} + R_{i4})}} \,. \tag{33C}$$

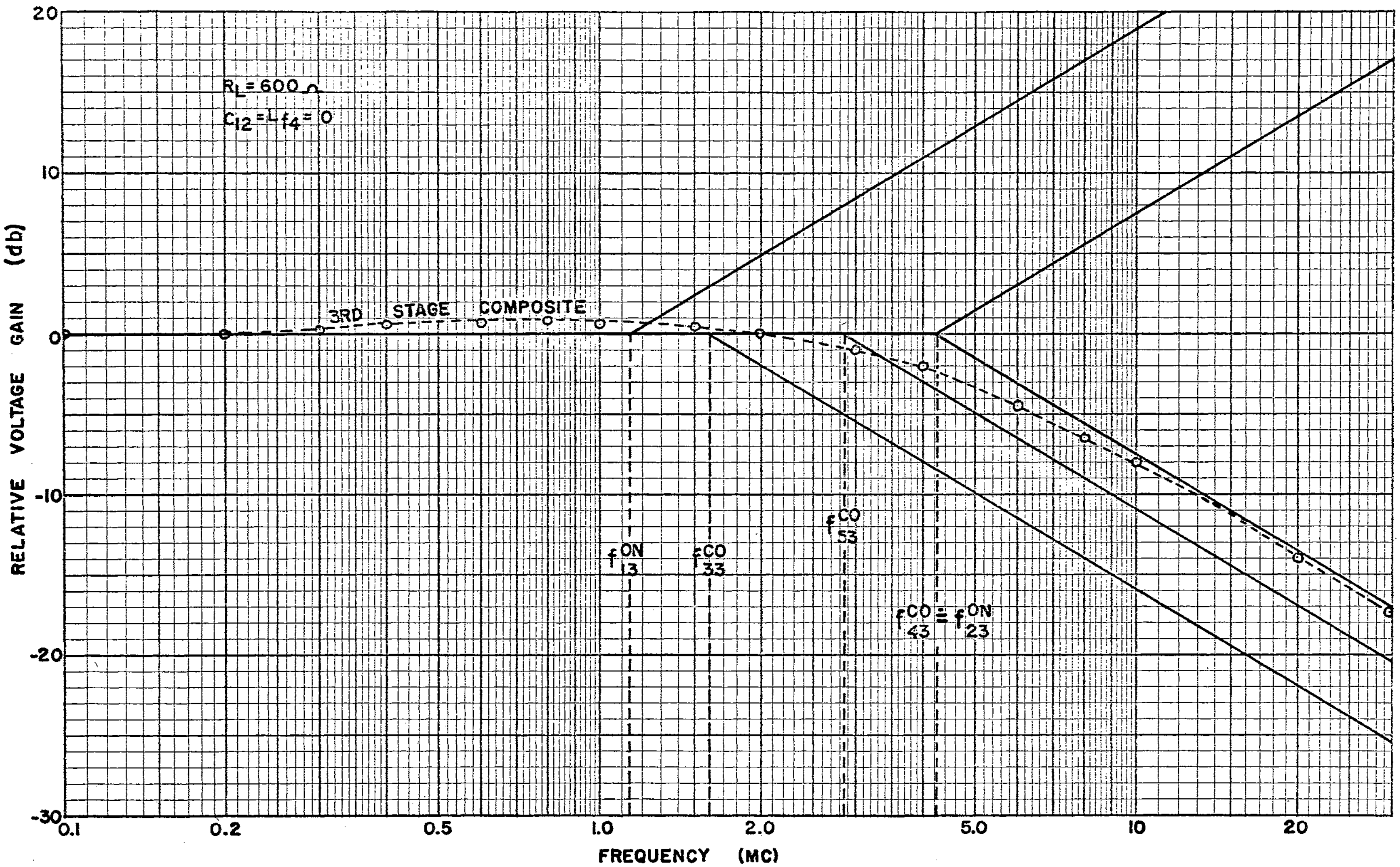

Figure C.2. Third Stage Frequency Response of the Collector to Base-Feedback-Peaking Amplifier.

The following corner frequency equations are obtained from (33C)

$$f_{14}^{on} = \frac{f_{e4}\, r_{b4}}{R_{i4}} \quad ; \quad f_{24}^{co} = \frac{f_{e4}\, r_{b4}(R_{x4} + R_{i4})}{R_{i4}(R_{x4} + r_{b4})} . \tag{34C}$$

In this case, $\frac{R_{x4} + r_{b4}}{R_{x4} + R_{i4}} \doteq 1$; therefore, $G_{(high)} \doteq G_{(mid)}$ for this stage. It is seen that the response remains at the same level as frequency becomes increasingly larger, i.e., the power of P in the numerator is the same as that of the denominator. This would not be true if (16C) were replaced by either of the three equations preceding it when the frequency became large enough to prevent the approximation made to become invalid. The response is not of interest at these frequencies for this particular case.

The current gain of the complete amplifier, when uncompensated ($L_{f4} = 0$, $C_{12} = 0$), is the product of (45B), (53B), (23C), (33C) and $i_8/i_L = R_{44}/(R_{44} + R_L)$. The frequency response of the complete amplifier is shown in Fig. C.3.

High-Frequency Analysis With Compensation

The desired frequency compensation may be obtained by selecting the proper values of C_{12} and L_{f4}. Changing these values causes (53B) and (33C) for the second and fourth stages, respectively, to become invalid. The first and third stage equations, (45B) and (23C), are still valid. Equations (58B) through (71B) now apply to the second stage, with (71B) giving the relative gain. Equation (30C) is still valid, but (32C) must now be replaced by

$$\frac{i_8}{i_7} \doteq \frac{Z_{f4}}{R_{l4}} . \tag{35C}$$

Substituting

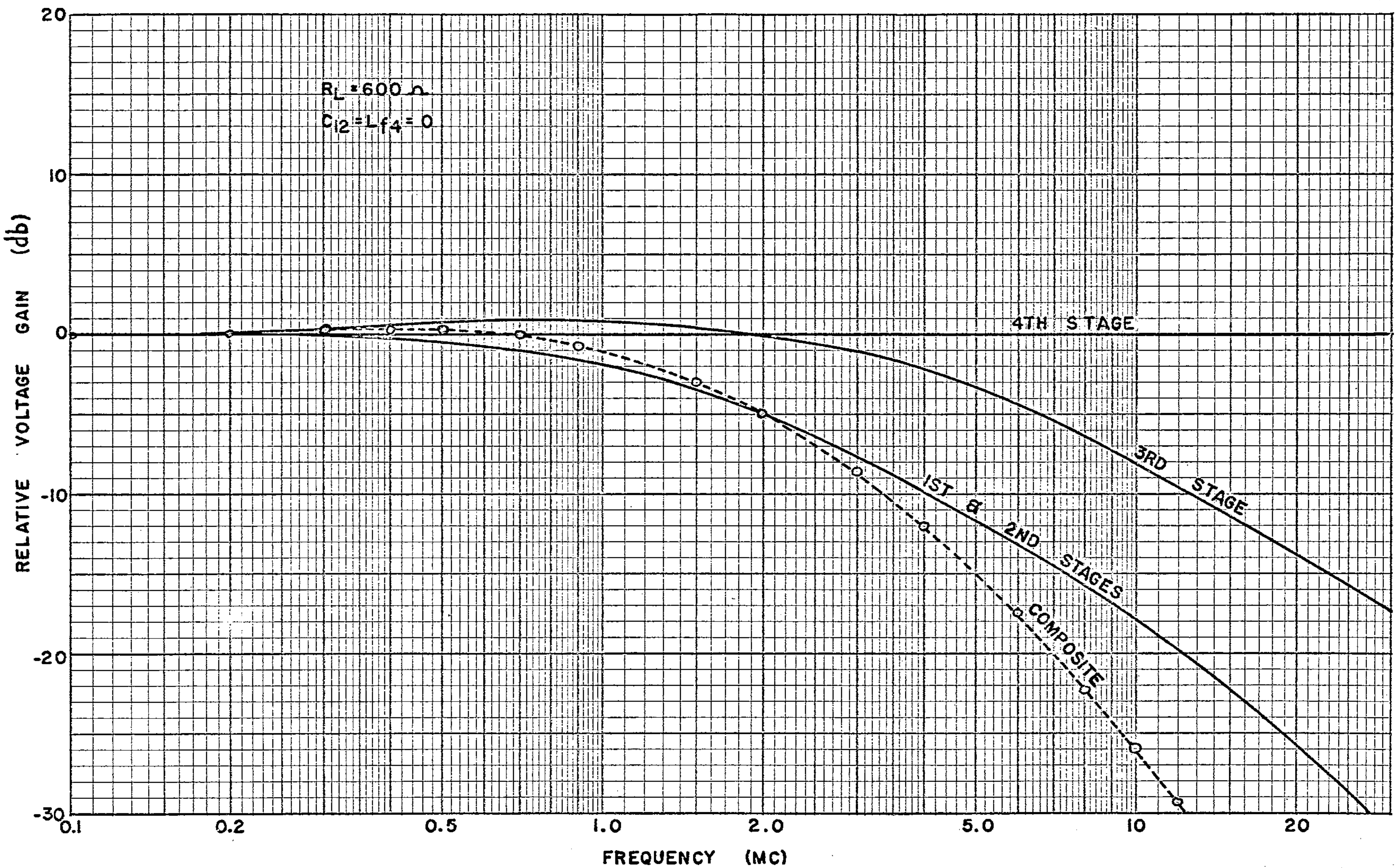

Figure C.3. Frequency Response of Uncompensated Collector to Base-Feedback-Peaking Amplifier.

$$Z_{f4} = R_{f4}\left[\frac{1 + p\frac{L_{f4}}{R_{f24}}}{1 + p\frac{L_{f4}}{R_{f14} + R_{f24}}}\right] \; ; \quad R_{f4} = \frac{R_{f14}\,R_{f24}}{R_{f14} + R_{f24}} \tag{36C}$$

in (35C) results in

$$\frac{i_8}{i_7} \doteq \frac{R_{f4}}{R_{l4}}\left[\frac{1 + p\frac{L_{f4}}{R_{f24}}}{1 + p\frac{L_{f4}}{R_{f14} + R_{f24}}}\right] \tag{37C}$$

The current gain of the fourth stage is the product of (30C) and (37C)

$$G_{(high)4} \doteq G_{(mid)4}\frac{\left[1 + p\frac{R_{i4}}{\omega_{e4} r_{b4}}\right]\left[1 + p\frac{L_{f4}}{R_{f24}}\right]}{\left[1 + p\frac{R_{i4}(R_{x4} + r_{b4})}{\omega_{e4} r_{b4}(R_{x4} + R_{i4})}\right]\left[1 + p\frac{L_{f4}}{R_{f14} + R_{f24}}\right]} \tag{38C}$$

where, $G_{(mid)4} \doteq \frac{R_{x4}\quad R_{f4}}{R_{l4}(R_{x4} + R_{i4})}$. The corner frequencies obtained from (38C) are

$$\left.\begin{aligned} f_{14}^{on} &= \frac{f_{e4} r_{b4}}{R_{i4}} \; ; & f_{24}^{co} &= \frac{f_{e4} r_{b4}(R_{x4} + R_{i4})}{R_{i4}(R_{x4} + r_{b4})} \\ f_{34}^{on} &= \frac{R_{f24}}{2\pi L_{f4}} \; ; & f_{44}^{co} &= \frac{R_{f14} + R_{f24}}{2\pi L_{f4}} \end{aligned}\right\} \tag{39C}$$

The over-all current gain of the amplifier is the product of (45B), (71B), (23C), (38C) and $i_L/i_8 = R_{44}/(R_{44} + R_L)$. The second and fourth stage frequency response given by (71B) and (38C), respectively, is plotted in Fig. C.4. The over-all frequency response ($R_L = 600\,\Omega$) of the amplifier with the second and fourth stages compensated is plotted in Fig. C.5.

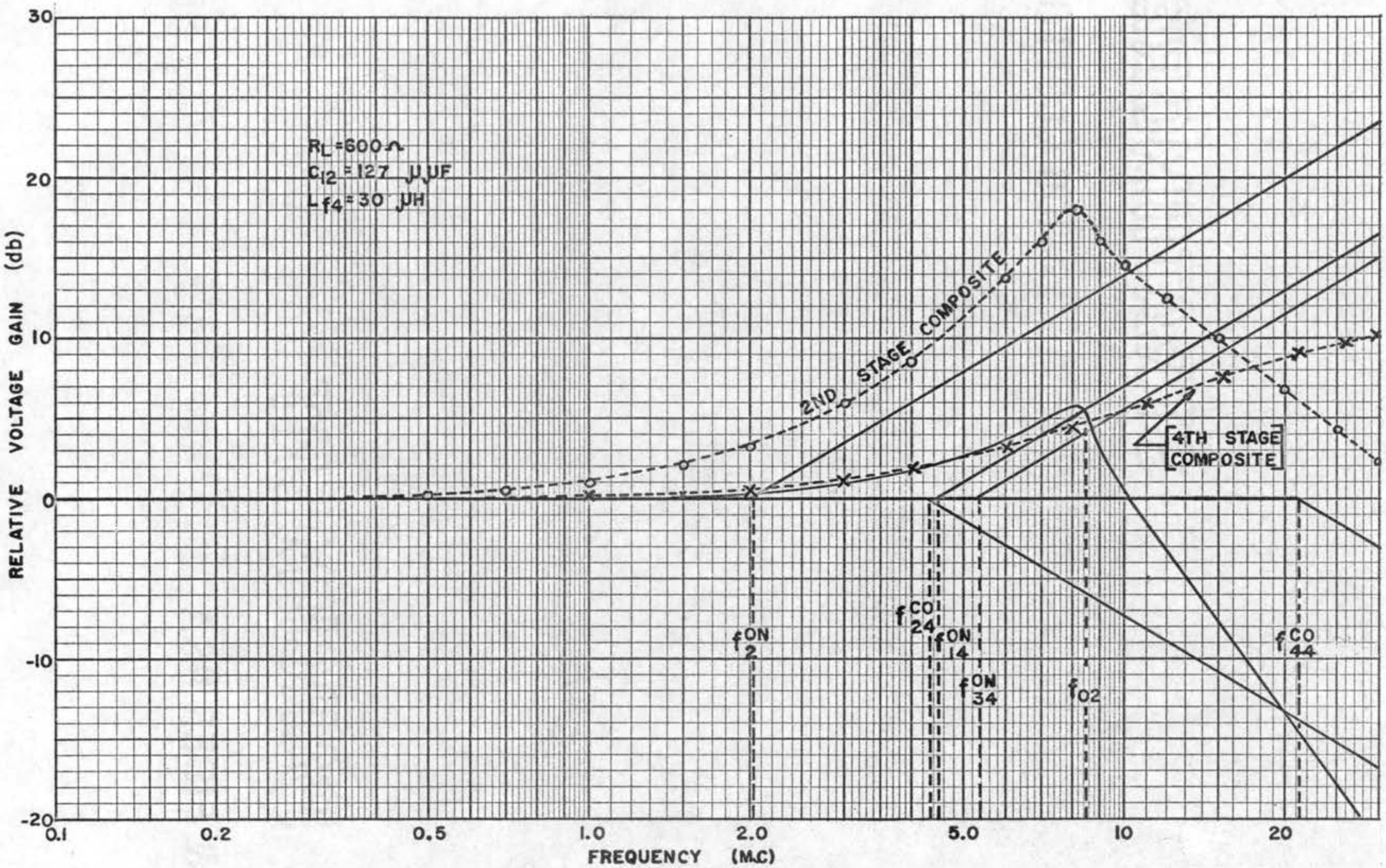

Figure C.4. Frequency Response of the Second and Fourth Stages in the Collector to Base-Feedback Peaking Amplifier.

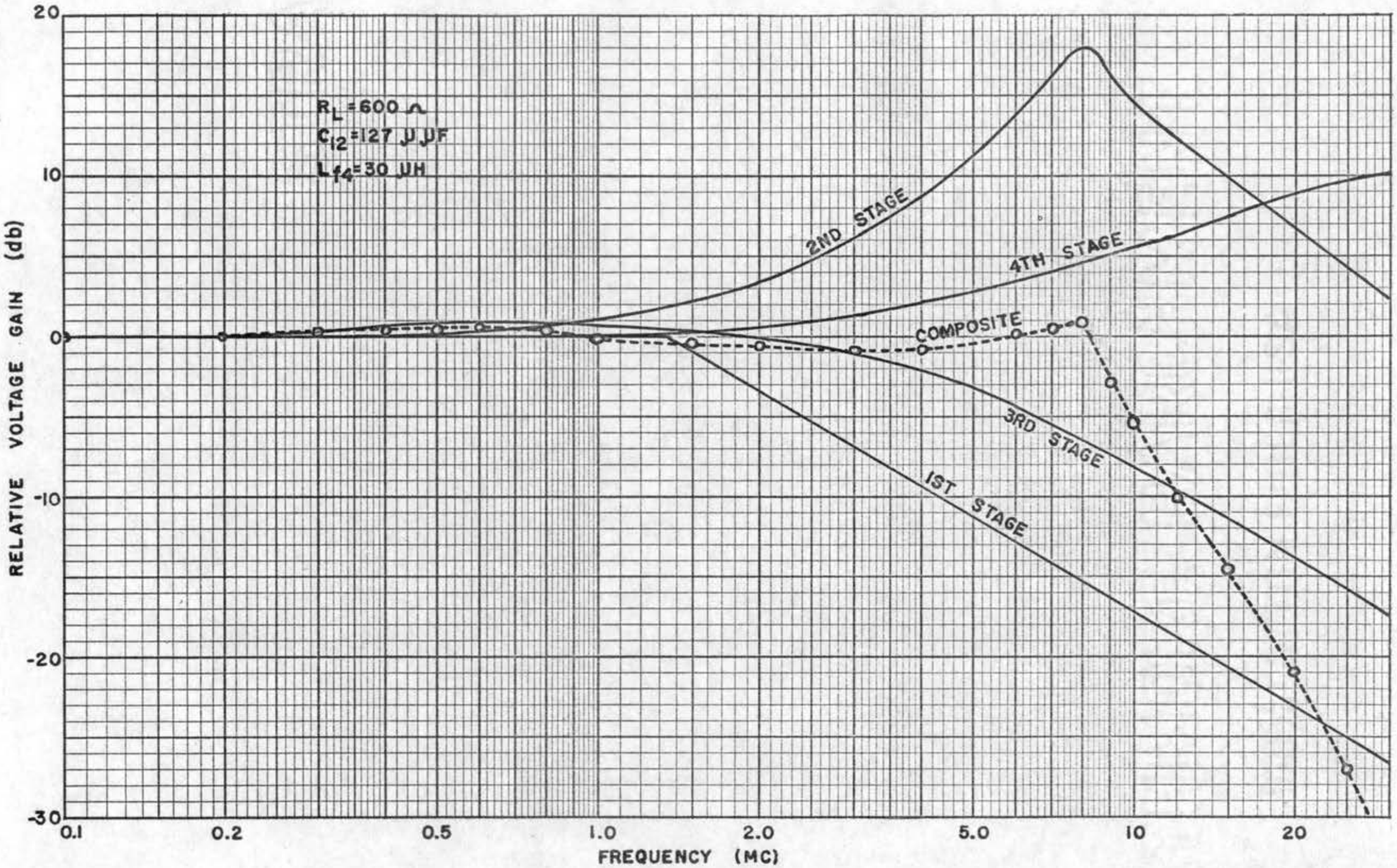

Figure C.5. Frequency Response of the Complete Collector to Base-Feedback-Peaking Amplifier.

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