LOW-FREQUENCY COMPENSATION OF TRANSISTOR AMPLIFIERS

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I. INTRODUCTION

One of the serious disadvantages of transistor amplifiers as compared to vacuum tube amplifiers is the difficulty in obtaining good low-frequency response with reasonable sizes of bypass capacitors. Although this difficulty is somewhat lessened by the fact that transistors in general are low-voltage devices, it is frequently necessary to use capacitors of 100 μf to over 1000 μf to obtain the response required for critical video amplifiers. Applications are occasionally encountered where high voltage circuits are required to operate in the sub-audio range; in these cases, conventional emitter bypass methods are either very bulky or quite expensive. Further, the high-frequency impedance of very large capacitors does not, at the present state of the art, decrease linearly with frequency because of extraneous effects associated with their size. Thus, in any fairly wideband amplifier, it is necessary to add a smaller capacitor in parallel for high frequencies. For these and other reasons, it is important to devise some method of obtaining good low-frequency response without using excessively large capacitors.

A zener diode biasing circuit presented by Kabell and Grinich (Reference 1) presents some interesting possibilities. A modification of this circuit is analyzed from the point of view of low-frequency response; by using a pole-zero cancellation, the high-frequency gain and the low-frequency 3 db point can be traded at will with no peaking in the response. For cases where peaking has no deleterious effect, better low-frequency response can be obtained at less sacrifice in high frequency gain.

II. PROPOSED COMPENSATION METHOD

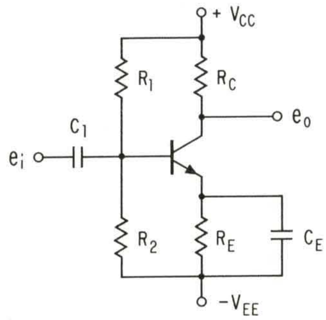

Fig. 1 (a) Conventional Transistor Biasing Circuit

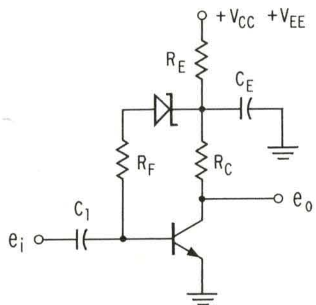

Fig. 1 (b) Zener Diode Biasing Circuit (as per reference 1)

A conventional biasing method and the zener diode biasing method presented by Kabell and Grinich is shown in Figures 1a and 1b, respectively. The latter method has many advantages, the two most important being the elimination of the large biasing current required through R_1 and R_2 and the much better low-frequency response since, in the conventional circuit, the emitter capacitance is effectively reduced by a

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factor of h_{fe}, the current gain. Two main disadvantages of zener diode bias compensation are: the feedback caused by R_F (since the d-c stability requirement and the maximum a-c gain requirement are conflicting in the choice of R_F), and for low current amplifiers, the low current through the zener diode may result in noisy and unstable operation. These disadvantages can be overcome quite simply by choosing a suitably large C and by returning the base of the transistor or the anode of the zener diode through a resistor to a negative supply. This is discussed in Reference 1.

While the low-frequency response of the Kabell and Grinich circuit has furnished some improvement, it is to be noted that at lower frequencies, the voltage which is fed back increases. Therefore, compensation is difficult, since at low frequencies the input current is reduced and the reverse of the required feedback compensation takes place. However, the load impedance increases and, therefore, net output voltage increase at low frequencies is possible with selected values of RC's.

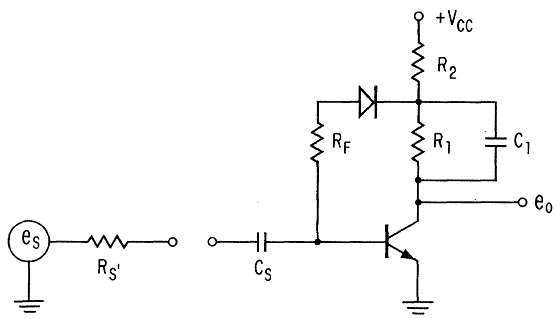

Fig. 2 An alternative method of Low-Frequency Compensation

Changing the location of C_E is one effective way of low-frequency compensation; this new circuit is shown in Figure 2 with different subscripts on the components. At low frequencies, the load impedance increases while the fed-back voltage does not increase, and good compensation is thus possible. An analysis of this method is carried out in the Appendix and is discussed below.

III. ANALYSIS OF THE METHOD

In the Appendix, it is shown that

$$\frac{e_o}{e_s} = -\frac{p}{Q}\left[\frac{(M+N)+MT_1 p}{1+\left(T_1+T_S\right)p+T_1 T_s p^2}\right] \qquad (1)$$

where p = complex operator $j\omega$

$M = R_F R_2 h_{fe}$

$N = \left(R_F + R_2\right) R_1 h_{fe}$

$Q = \dfrac{R_F + h_{fe} R_2}{C_S}$

$T_1 = R_1 C_1$, $T_s = \left(R_{s'} + h_{ie}\right) C_S = R_s C_s$

The only assumptions are that $h_{re} = 0$, $h_{fe} \gg 1$ and $R_F \gg h_{ie}$.

Rearranging equation (1),

$$\frac{e_o}{e_s} = -\frac{M}{QT_s}\left[\frac{p\left(p+\frac{M+N}{MT_1}\right)}{\left(p+\frac{1}{T_1}\right)\left(p+\frac{1}{T_s}\right)}\right] \qquad (2)$$

which has zeros at p = 0 and at $p = -\frac{M+N}{MT_1}$ and poles at $p = -\frac{1}{T_1}$ and $p = -\frac{1}{T_s}$. The pole-zero diagram is shown in Figure 3.

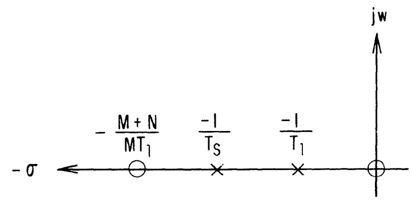

Fig. 3 Pole-Zero locations for the Amplifier

In order to improve the low-frequency response, the zero at $-\frac{M+N}{MT_1}$ can be made to cancel the pole at $-\frac{1}{T_s}$. When this is done,

$$1 + \frac{N}{M} = \frac{T_1}{T_s} \qquad (3)$$

$$N = h_{fe}\left(R_2 + R_F\right) R_1 \simeq h_{fe}\, R_F\, R_1$$

$$M = R_F\, R_2\, h_{fe}$$

Therefore, for the cancellation,

$$1 + \frac{R_1}{R_2} = \frac{T_1}{T_s} \qquad (4)$$

or, introducing a variable

$$k = \frac{1+\frac{R_1}{R_2}}{\frac{T_1}{T_s}}, \quad k = 1 \qquad (5)$$

for the cancellation.

The expression for the voltage gain then becomes

$$\frac{e_o}{e_s} = \frac{M}{QT_s} \cdot \frac{p}{p+\frac{1}{T_1}}$$

$$= -\left[\frac{R_2\, R_F\, h_{fe}\, C_s}{R_F + h_{fe}\, R_2}\right] \frac{p}{p+\frac{1}{T_1}} \qquad (6)$$

Note that this response is now the same as a series R_1C_1 and the low-frequency 3 db point will now be

$$f_1 = \frac{1}{2\pi R_1 C_1} \tag{7}$$

The essence of the method can now be seen by considering equation (4). For cases where $\frac{R_1}{R_2} \gg 1$, then, for the cancellation

$$\frac{R_1}{R_2} \simeq \frac{T_1}{T_s} \tag{8}$$

Thus, if T_1 is made larger to improve the low-frequency response, the $\frac{R_1}{R_2}$ ratio must be made correspondingly larger with the corresponding loss of available high-frequency gain. There is, therefore, for values of $\frac{R_1}{R_2}$ somewhat greater than 1, high-frequency gain and low-frequency 3 db point trading. The degree of low-frequency improvement is directly proportional to the loss of available high-frequency gain. This suggests the use of very high gain a-c amplifiers in certain applications normally using d-c techniques but which could use very low-frequency a-c amplifiers, thus eliminating drift problems associated with d-c amplifiers. This is true even when the great improvements recently made in d-c amplifiers are considered (References 2 and 3).

It is interesting to consider values of k different from 1. Values of k greater than 1 correspond to the case where T_1 is less than the critical value; and thus, compensation begins at a frequency above that where the input capacitor becomes important so that peaking occurs in the frequency response. Values of k less than 1 correspond to the case where compensation begins at a frequency below that where the input capacitor becomes important so that low-frequency response is deteriorated; and therefore, this case is of no interest. For cases where no peak in frequency response is permissible, the value k = 1 is the optimum one, although if some peaking can be tolerated, considerable low-frequency improvement can be made by selecting k greater than 1. A practical case is shown plotted in Figure 4, and a value of k = 1.5 gives a 2 db peak with about 20% improvement in f_1 over the flat response case.

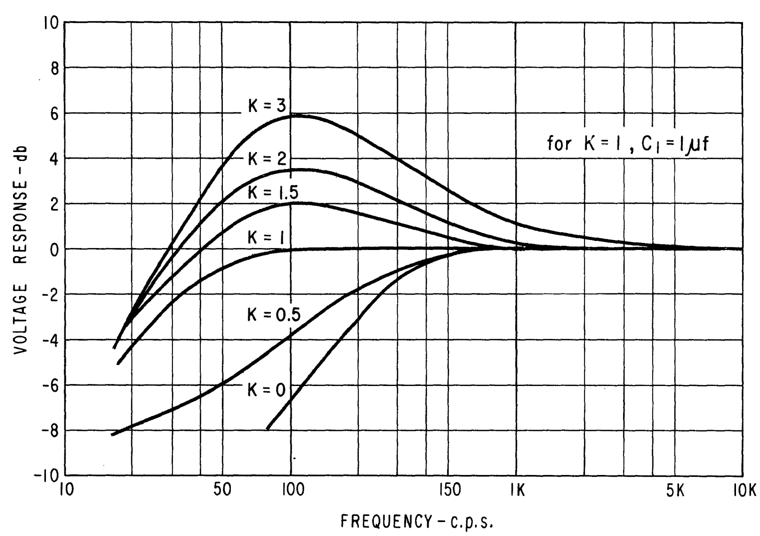

Fig. 4 Various responses showing the effect of different k's

A PRACTICAL AMPLIFIER STAGE

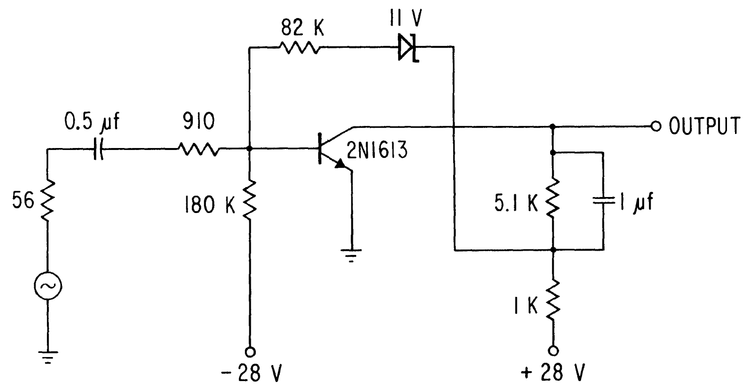

Fig. 5 A Practical Amplifier

In order to demonstrate a practical amplifier stage, the amplifier shown in Figure 5 was designed and built. A required f_1 of 30 cps was assumed, and a C_1 of 1.0 μf was accordingly chosen from equation (7). Using equation (4), T_s was calculated to be 0.8×10^{-3}. Using the measured value of h_{ie} of 690 ohms,

$$C_s = \frac{0.8 \times 10^{-3}}{(690 + 56 + 910)} = 0.5 \ \mu f$$

This circuit was built and tested giving an f_1 = 25 cps for k = 1 with no peaking. Various values of k are shown on the graph in Figure 4, including that for $k \simeq 0$ which is essentially the uncompensated case.

III. CONCLUSIONS

The frequency compensation method presented trades high-frequency gain for low-frequency response. It is felt that this method will be of some importance in the design of audio and video amplifiers requiring good low frequency response and may also be useful in the design of ultra-low frequency a-c amplifiers.

REFERENCES

1. "Zener Diode Circuits for Stable Transistor Biasing", L. J. Kabell and V. H. Grinich, Semiconductor Products, May 1961

2. "A New D.C. Transistor Differential Amplifier", D. F. Hilbiber, Presented at 1961 Philadelphia Solid State Circuits Conference

3. "The Design of High Stability D.C. Amplifiers", P. J. Bénéteau, Semiconductor Products, February 1961, p. 27

APPENDIX

Analysis of the amplifier:

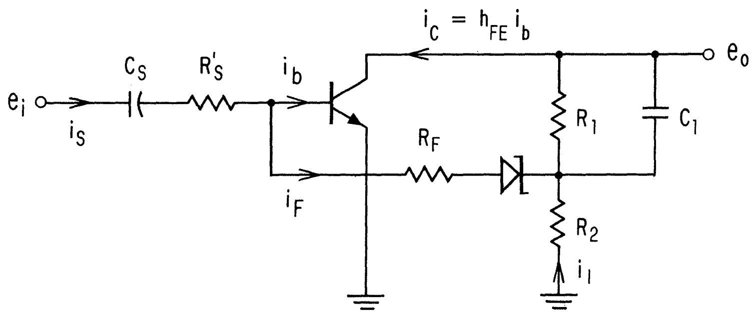

Fig. 6 Amplifier to be analyzed

The equations for the circuit of Figure 6 can be written as follows, the only assumptions being that $h_{12e} = 0$ and that the a-c resistance of the zener is trivial compared to R_F.

$$i_s = i_b + i_F \qquad \text{(A 1)}$$

$$i_1 + i_F = h_{FE} i_b \qquad \text{(A 2)}$$

$$i_b h_{ie} + i_1 R_2 = i_F R_F \qquad \text{(A 3)}$$

$$e_s - i_b h_{ie} = i_s Z'_s \qquad \text{(A 4)}$$

$$\text{where } Z'_s = R'_s + \frac{1}{C_s p}$$

There are four equations in four unknowns (i_s, i_b, i_F, i_1). Solving by determinants or by some other method, we get for $R_F \gg h_{ie}$ and $h_{fe} \gg 1$,

$$\left.\begin{aligned} i_1 &= \frac{e_s h_{fe} R_F}{Z'_s \left(R_F + h_{FE} R_2\right) + h_{ie} R_F} \\ \text{and } i_c &= \frac{e_s h_{fe} \left(R_2 + R_F\right)}{Z'_s \left(R_F + h_{FE} R_2\right) + h_{ie} R_F} \end{aligned}\right\} \qquad \text{(A 5)}$$

Now, $e_o = -i_1 R_2 - i_c Z_1$

which gives, using equation (A 5),

$$\frac{e_o}{e_s} = - \frac{h_{fe} R_F R_2 + h_{fe} \left(R_2 + R_F\right) Z_1}{Z'_s \left(R_F + h_{fe} R_2\right) + h_{ie} R_F} \qquad \text{(A 6)}$$

$$\text{where } Z_1 = R_1 \left(R_1 C_1 p + 1\right)$$

Defining $Z_s = Z'_s + h_{ie}$

$$= \left(R'_s + \frac{1}{C_s p}\right) + h_{ie}$$

$$= R_s + \frac{1}{C_s p},$$

and noting that at low frequencies $Z'_s \simeq Z_s$ due to the large reactance of C_s, equation (A 6) can then be written as

$$\frac{e_o}{e_s} \simeq - \frac{h_{fe} R_F R_2 + h_{fe} \left(R_2 + R_F\right) Z_1}{Z_s \left(R_F + h_{fe} R_2\right)} \qquad \text{(A 7)}$$

Putting $M = h_{fe} R_F R_2$,

$N = h_{fe} \left(R_2 + R_F\right) R_1$,

$Q = \frac{1}{C_s} \left(R_F + h_{fe} R_2\right)$,

$T_1 = R_1 C_1$ and $T_s = R_s C_s$,

yields after a little manipulation

$$\frac{e_o}{e_s} = - \frac{M}{T_s} \left[\frac{p \left(p + \frac{M + N}{MT_1}\right)}{\left(p + \frac{1}{T_1}\right) \left(p + \frac{1}{T_s}\right)} \right] \qquad \text{(A 8)}$$

which is the expression quoted in equation (2) in the paper.