Optimization of the Amplified-Diode Bias Circuit for Audio Amplifiers*

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An economic enhancement to the conventional "amplified diode" bias circuit is presented for use in power amplifier circuit topologies which do not allow precise, temperature invariant control of the operating current of the bias circuit. In essence, the modification minimizes the sensitivity of the derived bias voltage to changes in operating current without compromising the desirable temperature tracking properties when thermally bonded to the complementary follower output cell.

0 BACKGROUND

The output stage of power amplifiers and some operational amplifiers requires circuitry to allow precision control of the output bias current. The classical approach is to use a bias network that consists of either a series of diodes [1] or a transistor with local feedback in a circuit called an amplified diode [2], [3]. In Fig. 1 we illustrate the basic output-cell topology for the complementary follower configuration.

The circuit objective is to produce a dc offset V_B between the base connections of the output devices to compensate for the ON bias voltage that is required to establish the output-device bias quiescent current I_Q.

Although in practice emitter resistors R_e are used as local series feedback elements to help stabilize changes in I_Q with changes in device temperature and circuit parameters, their use only degenerates performance in other areas by increasing the output resistance of the stage and by making the output resistance a nonlinear function of output current (especially in class AB stages).

Consequently, as is well known, the use of a bias network that in principle can track changes in output-device temperature is necessary to control I_Q within reasonable bounds and thus allows low or zero values of R_e to be employed. Such networks are generally of the type shown in Fig. 1, where the bias devices (diodes or transistor) should be in close thermal contact with the output cell for good temperature tracking.

However, one area of bias network design that has been given little attention is the variability of V_B with changes in operating current I (see Fig. 1). We define a function $S_I^{V_B}$, which is a measure of this dependency,

$$S_I^{V_B} = \frac{\partial V_B}{\partial I} . \tag{1}$$

In many amplifier circuits the quiescent value of I can change as a function of temperature, where in general the trend is for a positive temperature coefficient, that is, I increases with temperature. To some extent this can be compensated by a suitable choice of feedback structure to the input stage, but this may well compromise other areas of performance and prove impractical to implement in a low-feedback amplifier.

This communication addresses the optimization of $S_I^{V_B}$ and suggests a simple modification to the design of the amplified diode which allows $S_I^{V_B}$ to be zero or indeed negative, thus reducing the tendency for increased output device bias current I_Q with temperature due to changes within the input stage of the amplifier.

1 THE MODIFIED AMPLIFIED DIODE

The modification to the basic amplified diode that enables optimization of $S_I^{V_B}$ is shown in Fig. 2 where

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I_E is the emitter current and V_{BE} the base emitter voltage.

The addition to the circuit is the resistor R_3. To investigate the operation of the modified circuit, we calculate the parameter $S_I^{V_B}$. We proceed by choosing resistors R_1 and R_2 such that the current in R_1 is

$$I_{R1} = \sqrt{I_C I_B} = \frac{I_C}{\sqrt{\beta}} \tag{2}$$

where β is the current gain, I_C the collector current, and I_B the base current of T_1. Thus as an example, if $\beta = 100$, then $I_{R1} \cong 10 I_B$ and $I_C \cong 10 I_{R1}$. This will therefore realize good bias stability within the amplified diode. Consequently we may assume (for high B) that

$$I_C \cong I \; . \tag{3}$$

Hence,

$$(V_B + IR_3)\frac{R_1}{R_1 + R_2} = V_{BE}$$

and thus,

$$V_B = \left(1 + \frac{R_2}{R_1}\right)V_{BE} - IR_3 \; . \tag{4}$$

Differentiating V_B with respect to I to determine $S_I^{V_B}$, we have

$$S_I^{V_B} = \left(1 + \frac{R_2}{R_1}\right)\frac{\partial V_{BE}}{\partial I} - R_3 \; . \tag{5}$$

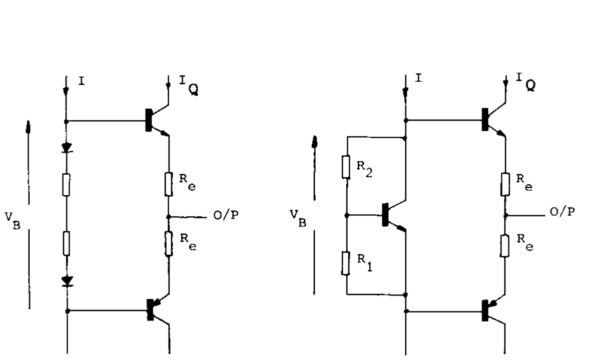
Since $I_E \cong I$, then from the diode equation,

$$I_E = I_S e^{qV_{BE}/KT} \cong I$$

where I_S is the transistor saturation current, q is the charge on an electron, K is Boltzmann's constant, and T is the junction temperature.

Differentiating, $\partial V_{BE}/\partial I \cong KT/qI$, and thus,

$$S_I = \left(1 + \frac{R_2}{R_1}\right)\frac{KT}{qI} - R_3 \; . \tag{6}$$



(a) (b)

Fig. 1. Basic output-cell biasing circuit. (a) Series diode. (b) Amplified diode.

At room temperature $KT/q \cong 0.025$ V, and

$$S_I^{V_B} = \frac{0.025}{I}\left(1 + \frac{R_2}{R_1}\right) - R_3 \; . \tag{7}$$

Eq. (7) allows R_3 to be selected to minimize $S_I^{V_B}$. Hence for zero S_I, where optimum $R_3 = R_{3\,\text{opt}}$,

$$R_{3\,\text{opt}} = \frac{0.025}{I}\left(1 + \frac{R_2}{R_1}\right) . \tag{8}$$

Under this condition the bias voltage V_B is to a good first-order approximation independent of I. However, in practice it is suggested that $R_3 > R_{3\,\text{opt}}$, thus implying a negative S_I. This will counteract tendencies for thermal runaway with increasing ambient temperature. Also, as $R_{3\,\text{opt}}$ is temperature dependent, the value of R_3 should be calculated at the maximum device temperature. Thus for lower temperatures S_I will be slightly negative.

In circuit applications requiring a bias voltage V_B of several V_{BE}, such as where Darlington output devices are used, two transistors may be used as shown in Fig. 3.

2 CONCLUSIONS

This communication has discussed a modification to the basic amplified–diode circuit which will minimize the dependency of the bias voltage V_B on the magnitude of the amplified-diode operating current. The modification is simple, yet has proved to be extremely effective in operation. It is important on two counts. First, it will minimize changes in output-cell bias current due to changes in the driving circuit, which will generally be temperature dependent. Also in circuits that use a differential drive current to the amplified diode, it will

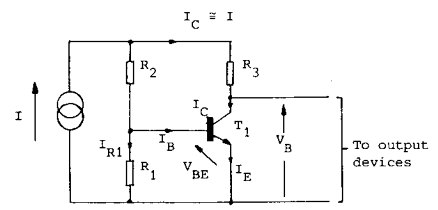


Fig. 2. Modified amplified-diode circuit.

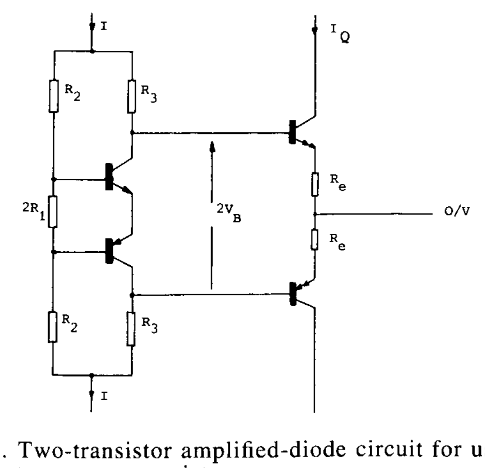


Fig. 3. Two-transistor amplified-diode circuit for use with Darlington output transistor.

minimize common-mode current variations with this drive configuration. Finally it should be noted that the modification in no way compromises the performance of the amplified diode with respect to thermal tracking of the output devices since for constant I, the voltage IR_3 is almost independent of temperature where, from Eq. (4),

$$\frac{\partial V_B}{\partial T} = \left(1 + \frac{R_2}{R_1}\right) \frac{\partial V_{BE}}{\partial T} . \quad (9)$$

3 REFERENCES

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THE AUTHOR

Malcolm Hawksford was educated at the University of Aston in Birmingham, England, from 1965 to 1971. In 1968 he obtained a first class honors degree in electrical engineering, and that same year was awarded a BBC research scholarship to investigate the application of deltamodulation to color television. In 1972 he obtained a Ph.D. degree. In 1971 Dr. Hawksford became a lecturer at the University of Essex, England, in the Department of Electrical Engineering Science. During his time at Essex he has actively pursued research projects within the field of audio engineering, where projects on power amplifier design, loudspeaker crossover design, analog-to-digital conversion, and music synthesis have been undertaken. He has presented papers at conventions of the Audio Engineering Society. He is currently a member of the AES, IEE, and RTS.