Amplifying Devices and Low-Pass Amplifier Design

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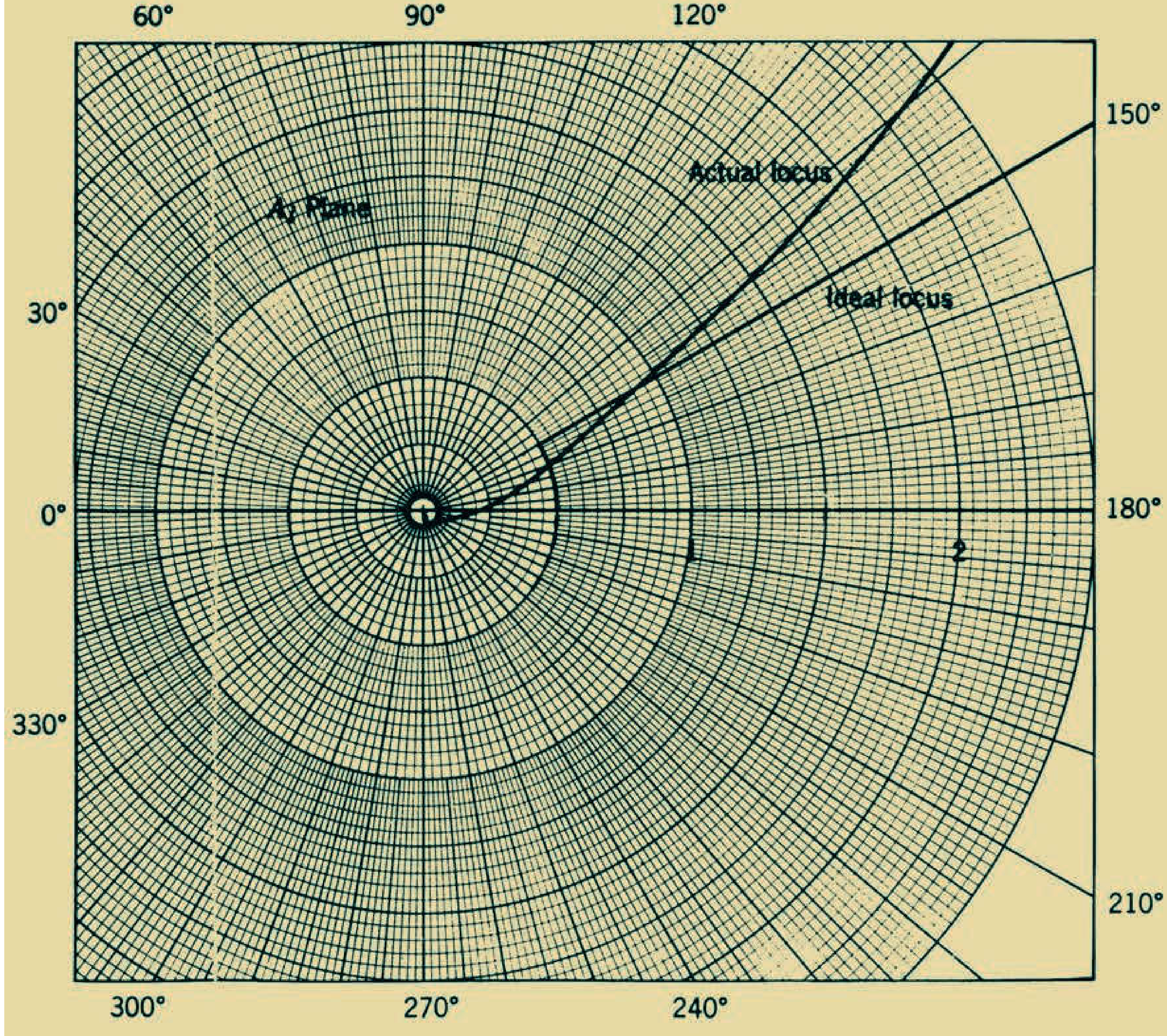

13.9.1 Other Stage Configurations

The realizable gain-bandwidth product of an active device always falls short of the intrinsic gain-bandwidth product calculated from charge-control considerations (Section 2.5.1). Losses are attributable to the stray wiring capacitances, the Miller contributions to input and output capacitances, and in the case of a transistor to the base resistance r_B. As a general rule the realizable $\mathscr{GB}$ in an amplifier circuit falls further below the intrinsic $\mathscr{GB}$ as the bandwidth is increased.

A number of useful wide-band circuits combine one or more common-control-electrode stages or common-collecting-electrode stages (Section 5.7) with each common-emitting-electrode stage. These common-control-electrode and common-collecting-electrode stages have unity current gain and voltage gain respectively and so cannot add to the total gain, but they can increase the bandwidth. The philosophy underlying their use is that it can be more efficient to increase the realizable $\mathscr{GB}$ of one common-emitting-electrode stage by means of several unity-gain auxiliary stages, rather than to achieve gain but only a small $\mathscr{GB}$ from all stages. Obviously, this will be the case only if the attainable gain per stage is small, because the increase in realizable $\mathscr{GB}$ cannot be very large. Therefore, the use of unity-gain stages cannot possibly be efficient at small bandwidths where the gain per stage can be large.

As an introduction to this type of circuit, Fig. 13.36 shows the transistor version of the *cascode*, a circuit which substantially eliminates the loss of realizable $\mathscr{GB}$ due to Miller effect in triodes and transistors. Quite apart from this application to very-wide-band circuits, the vacuum-tube cascode is extremely useful in low-noise amplifiers at moderate bandwidths (see Problem 8.1). In Fig. 13.36*a*, Q_1 is a common-emitter amplifying stage and Q_2 is an auxiliary common-base stage. The two devices in a cascode are usually of the same type; the two halves of a twin triode are often used in a vacuum-tube cascode. Moreover, the two devices are usually connected in series for dc as well as for signals (Fig. 13.36*b*) so the quiescent conditions and small-signal parameters are identical.

The load resistance for the common-emitting-electrode stage of a cascode is the input resistance of the common-control-electrode stage. Section 5.7.1 shows that this input resistance is small. Therefore the voltage gain and Miller input capacitance of the common emitting-electrode stage are small also. From Eq. 5.106 the input resistance of Q_2 in Fig. 13.36 is

$$R_{i2} = \frac{r_B}{\beta_N} + r_E.$$

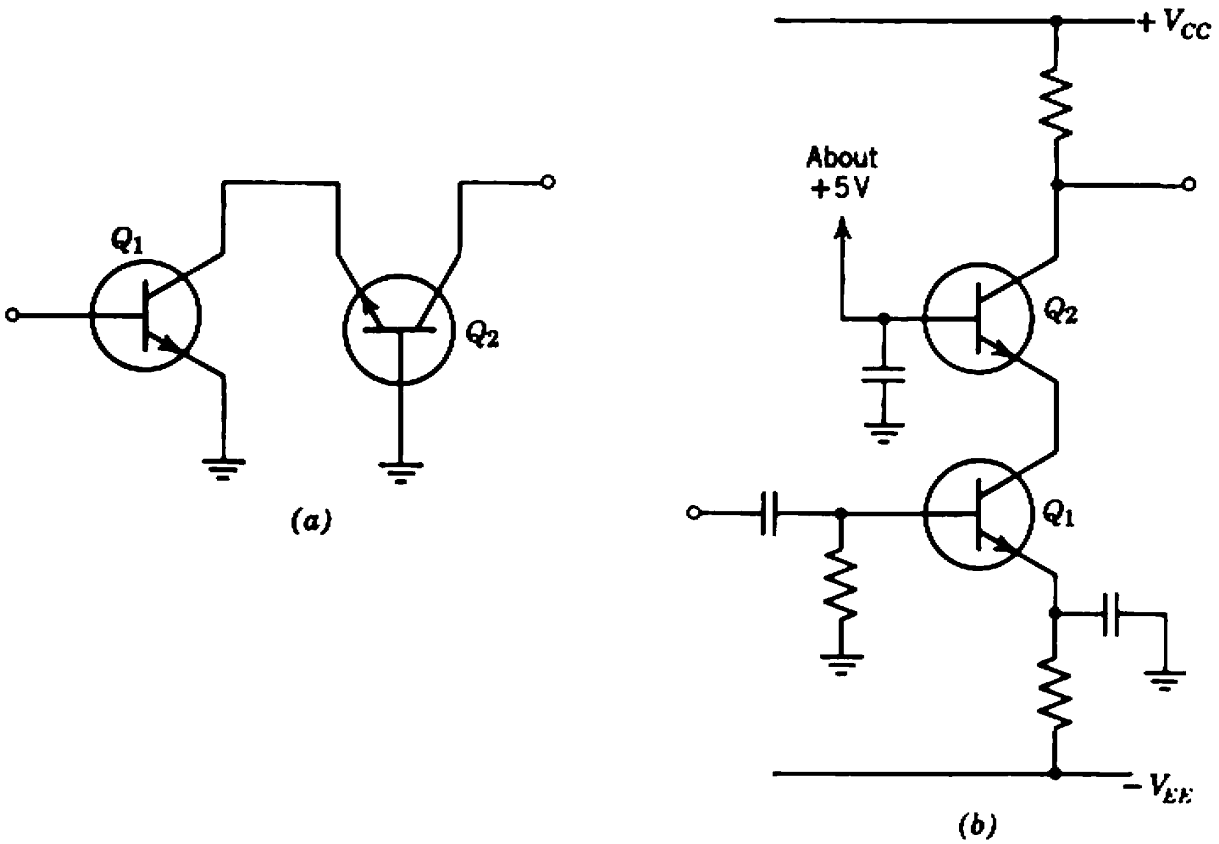

Fig. 13.36 The cascode: (*a*) elemental circuit diagram; (*b*) complete circuit, showing the usual biasing arrangement.

The voltage gain of Q_1 is

$$A_{V1} = -G_{T1} \times R_L,$$

and substituting

$$A_{V1} = -\left(\frac{\alpha_N}{r_B/\beta_N + r_E}\right)(r_B/\beta_N + r_E) = -\alpha_N \approx -1.$$

Hence the Miller input capacitance is

$$c_M = c_{tC}(1 - A_V) \approx 2c_{tC}.$$

The transfer conductance of the complete cascode is

$$G_{T(\text{tot})} = G_{T1} \times A_{I2} = \frac{\alpha_N{}^2}{r_B/\beta_N + r_E},$$

which is almost equal to G_T of a simple common-emitter stage; there is no significant loss of gain through using the cascode.

Many circuits using common-base stages and emitter followers have appeared in the literature. Possibly the most ingenious is Rush's feedback current amplifier* shown in Fig. 13.37. In outline, the design philosophy and operation are:

Ideally, a transistor has a short-circuit current gain-bandwidth product ω_T, with losses occurring due to r_B, C_{tC}, and stray wiring capacitances.

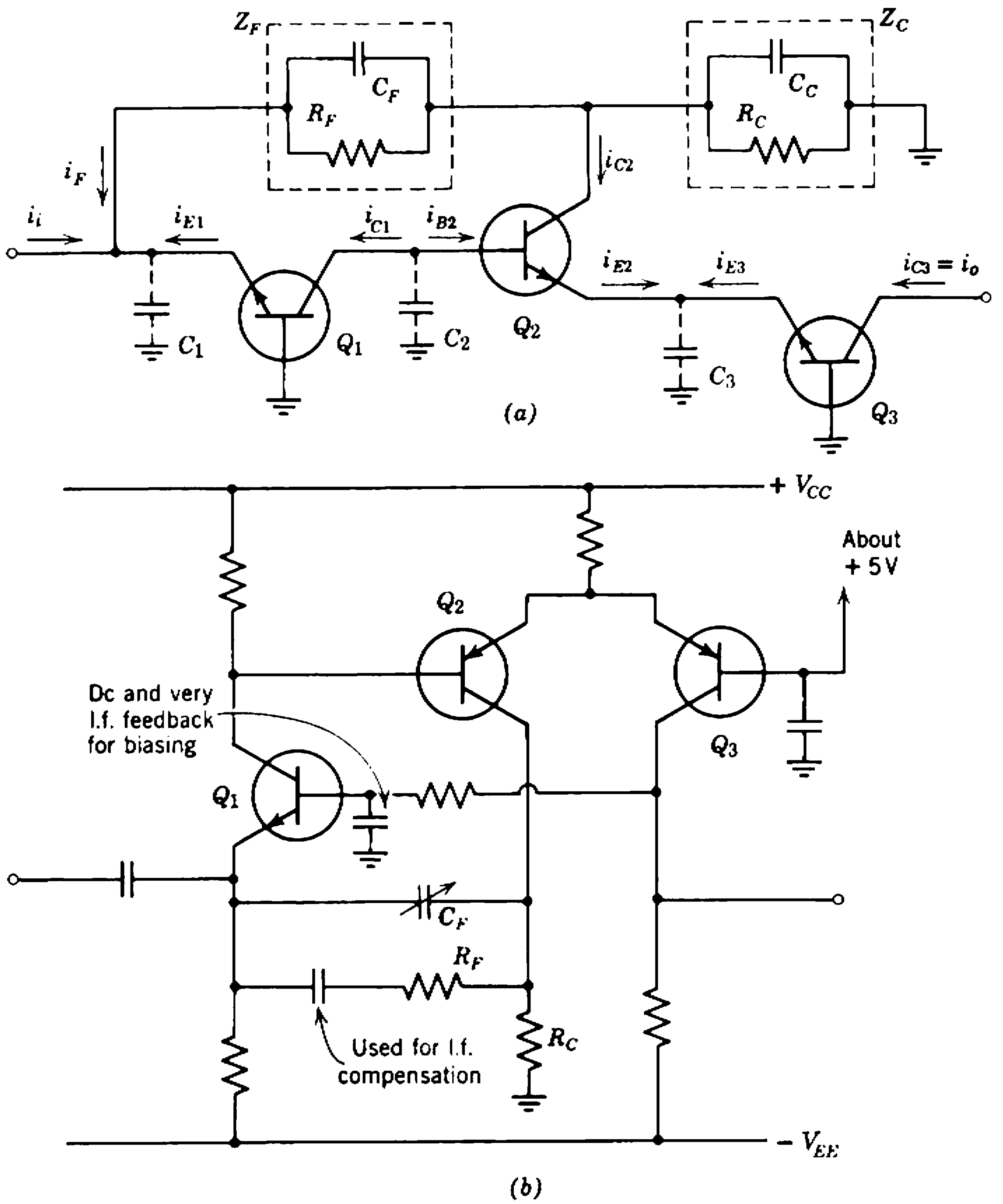

Fig. 13.37 Rush's feedback current amplifier: (*a*) elemental circuit diagram; (*b*) one convenient practical realization.

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* C. J. Rush, "New technique for designing fast-rise transistor pulse amplifiers," *Rev. Sci. Instr.*, **35**, 149, February 1964.

The current amplifying stage in Fig. 13.37 is Q_2 operating in the common-collector configuration for which the mid-band gain is

$$\frac{i_{E2}}{i_{B2}} = \beta_N + 1.$$

However, Q_2 has a low output resistance at its emitter (Eq. 5.121) and is therefore not a good current source; Q_3 is added as a near-unity-current-gain common-base stage to raise the output resistance. Assuming that Q_2 and Q_3 are identical, their collector currents are equal but opposite; the current gain from the base of Q_2 to either collector is therefore

$$\frac{i_{C2}}{i_{B2}} = -\frac{i_{C3}}{i_{B2}} = \beta_N.$$

Loss of realizable $\mathscr{GB}$ due to r_B of Q_2 is minimized by feeding Q_2 from a high-resistance source; Q_1 is a common-base buffer stage with current gain α_N. Loss of $\mathscr{GB}$ due to Miller effect is minimized by operating Q_2 with a small collector load resistance (R_F and R_C in shunt). Loss of $\mathscr{GB}$ due to stray wiring capacitances is small because there is a low impedance to ground at all points; stray capacitances C_1, C_2, and C_3 are in shunt with the relatively large input capacitances of Q_1, Q_2, and Q_3, respectively. Over-all, the current gain from the emitter of Q_1 to the collector of either Q_2 or Q_3 is

$$\frac{i_{C2}}{i_{E1}} = -\frac{i_{C3}}{i_{E1}} = -\alpha_N\beta_N,$$

and the gain-bandwidth product is

$$\mathscr{GB} \approx \omega_T.$$

In addition to the small losses in $\mathscr{GB}$ mentioned above, there are losses due to the α cutoff frequencies of Q_1 and Q_3; these losses are small also, because ω_α is nearly an octave above ω_T for a high-frequency transistor (Sections 4.3.2.1 and 4.7.2.1).

Current gain can be exchanged for bandwidth by means of negative feedback derived from the collector of Q_2 and returned to the emitter of Q_1. The feedback factor is

$$\boldsymbol{\beta} = \frac{i_F}{i_o} = -\frac{i_F}{i_{C2}} = +\frac{Z_C}{Z_F + Z_C}.$$

If the loop gain is large, the mid-band closed-loop current gain is $-1/\boldsymbol{\beta}_m$:

$$A_{Im} = \frac{i_o}{i_i} = -\frac{R_F + R_C}{R_C}.$$

In Fig. 13.37*a* C_C represents the stray capacitance from the collector of Q_2 to ground, where it contributes a lagging phase shift to the feedback loop. C_F is added in shunt with R_F to contribute a leading phase shift to the loop and modify the damping ratio of the closed-loop poles.

Figure 13.37*b* shows a convenient practical form of the circuit using *p-n-p* and *n-p-n* transistors in combination. In his original paper Rush describes a stage with a gain of 4.5 and a rise time (without overshoot) of 1.2 nsec.

In concluding this outline of other useful stage configurations it is worth enlarging on a statement made in the introduction. The use of common-base stages or emitter followers at moderate bandwidths cannot result in a more efficient amplifier than one designed in accordance with Sections 13.3 to 13.5. Indeed there is almost always a loss of efficiency. There is also a loss of designability—designability being a measure of the extent to which a circuit can be designed on paper and its performance predicted. These facts are not at all widely recognized. Notice, however, that common-base stages and emitter followers are useful in integrated-circuit amplifiers; transistors occupy less chip area than resistors and capacitors, and it follows that an inefficient circuit which uses many transistors but few passive components is more economical than an efficient circuit with few transistors but elaborate *RC* peaking.

13.9.2 Distributed Amplifiers

Section 13.6.1.2 shows that there is an upper limit to the bandwidth which can be obtained from a conventional type of amplifier having a specified gain, even if the number of stages can be increased indefinitely. This limit is not of much practical importance, because transistors with gain-bandwidth products of 1 GHz are available readily, and at the time of writing (1967) 3 GHz is available on a limited production basis. Nevertheless, it is an interesting philosophical point whether the $\mathscr{G}\mathscr{B}$ limit is truly fundamental. Percival's* distributed amplifier is a major invention because it does allow the realizable $\mathscr{G}\mathscr{B}$ to be increased without limit. The operation depends on the propagation of electromagnetic waves along a transmission line.

Ideally, the lumped *LC* transmission line shown in Fig. 13.38*a* has a resistive characteristic impedance

$$R_0 = \sqrt{\frac{L}{C}}, \qquad (13.136)$$

* W. C. PERCIVAL, British patent 460, 562, July 24, 1935.