XI. TRANSISTORS

Prof. R. B. Adler   C. R. Hurtig   J. C. Pinson
S. R. Blom   D. I. Kosowsky   M. Strieby
M. Boisvert   J. W. Vorndran

A. I-F AMPLIFIER DESIGN STUDIES

Examination of the iterative analysis of the transistor amplifier using simple coupling networks (see ref. 1) has shown that a resistive active circuit with a positive definite R-matrix can be replaced by a passive circuit and an ideal flat amplifier (current and voltage amplifier) having different gains in the forward and reverse directions. This property has been proved in a more general case for nonreciprocal networks by W. Klein (2). The impedance parameters of the passive circuit are given by the matrix below, if the r's are the usual impedance parameters of the active device

$$\begin{bmatrix} r_{11} & (r_{12} \, r_{21})^{1/2} \\ (r_{12} \, r_{21})^{1/2} & r_{22} \end{bmatrix}$$

If the gain is not of interest, the amplifiers can be removed without affecting the V/I ratios anywhere, thus allowing a passive reduction of the active circuit. Rewriting the matrix with the impedance levels divided by r_{11} at the input and by (r_{22}/τ) at the output, we obtain the following equivalent for a one-stage amplifier. (See Fig. XI-1.) In a chain of amplifiers the ideal transformers can be absorbed in the interstage networks, leaving the simple resistive circuit (Fig. XI-1, dotted box) for the transistor. This circuit gives exact results for the frequency response of a cascaded amplifier, its stability (exclusive of external feedback), its input impedance (normalized to 1 ohm), its output impedance (normalized to τ ohms), and in fact for anything except the magnitude of the gain. The gain may be accounted for by adding a single, ideal, bilateral (but not necessarily reciprocal) amplifier at the end of the chain.

M. Boisvert

References

1. Quarterly Progress Report, Research Laboratory of Electronics, M.I.T. April 15, 1953
2. W. Klein: Equivalent Circuits for Non-Reciprocal, Four-Terminal Networks, Archiv der electrischen Ubertragung 6, 1952, pp. 205-209

B. BROAD-BAND I-F AMPLIFIERS

Since the use of synchronous tuning gives a marked improvement in the overall power-gain bandwidth product over that realized with stagger-tuning (Quarterly
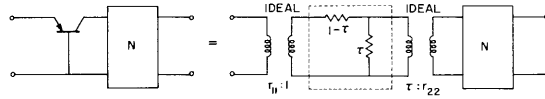
Fig. XI-1
Passive reduction of transistor stage.

Progress Report, Research Laboratory of Electronics, M.I.T. April 15, 1953), atten-
tion has been directed recently toward synchronously tuned amplifiers.

An analytical study has been made of the magnitude and phase responses obtained
with a grounded-base point-contact transistor having a load tuned to a frequency f_o com-
parable to the α cutoff frequency f_c of the transistor. It has been assumed that for any
frequency f

$$\alpha = \frac{\alpha_o}{1 + j\left(\dfrac{f}{f_c}\right)}$$

and that r_{11}, r_{12}, and r_{22} for the transistor are entirely real and constant. Moreover,
it has been possible to make the approximation

$$\tau \equiv \frac{z_{12} \, z_{21}}{z_{11} \, z_{22}} \approx \frac{r_{12} \, \alpha_o}{r_{11} \, r_{22}} = 0.8$$

Several families of normalized-gain and phase-shift characteristics have been calcu-
lated for two values of the parameter f_o/f_c ($f_o/f_c = 0.5$ and 1.0). For each family,
gain and phase shift are plotted against frequency, using an additional dimension-
less parameter $2\pi f_c \, r_{22} \, C_T$, where C_T is the total shunt capacitance of the output
circuit. Values of the ratio of bandwidth to center frequency read from the curves
are listed in Table I.

Table I

Values of the Ratio (Bandwidth/Center Frequency)

$2\pi f_c \, r_{22} \, C_T$	$f_o/f_c = 0.5$	$f_o/f_c = 1.0$
2.5	1.42	0.93
5.0	1.1	0.53
7.5	0.84	0.37
10.0	0.65	0.25

For low values of $2\pi f_c r_{22} C_T$, the phase response is fairly linear over the passband. Calculations for two coupled stages have been made for values of $f_o/f_c = 0.5$ and $2\pi f_c r_{22} C_T = 5.0$. These curves show that an interstage attenuation of 3 db provides a reasonably satisfactory degree of isolation between stages, thus confirming the experimental results quoted previously.

<div align="right">J. W. Vorndran</div>

C. RECEIVER STUDY[*]

The development of transistor circuits for the Signal Corps receiver has been continued. A satisfactory audio amplifier has been built for the receiver, using a push-pull class-AB output stage, with two p-n-p fused-junction transistors in a grounded-base connection. The grounded-base connection is used so that interchangeability of transistors in the output stage is possible. Grounded-emitter stages would require that the units be selected to have identical current gain α to prevent unbalancing of the push-pull amplifier. The push-pull output stage is driven through a transformer from a single-ended, grounded-emitter amplifier stage. Approximately 180 mw of audio output power have been obtained from the complete amplifier. In order to achieve this output, the final stage requires units having higher power-dissipation ratings than the standard commercial transistors. It has been possible to obtain units having the desired dissipation by mounting standard transistors in containers filled with either carbon tetrachloride or silicone oil. Efforts are also being made to obtain standard commercial power transistors.

A three-stage 455-kc/sec i-f amplifier has been built to study some of the problems of interstage coupling between junction transistor stages. Included in the amplifier is the Collins electromechanical filter, which will be used to obtain 2-kc/sec and 4-kc/sec bandwidths. It now appears that the wider bandwidth of 8 kc/sec can be achieved easily with conventional interstage coupling. For example, a two-stage amplifer using the special coils from the Sickles Company (see Quarterly Progress Report, Research Laboratory of Electronics, M.I.T. April 15, 1953, p. 80) has a 3-db bandwidth of approximately 10 kc/sec. The addition of further stages will reduce this bandwidth to the desired 8 kc/sec. There has, of course, been considerable difficulty in obtaining frequency responses of the desired shape because of variations in the input impedance of the various stages caused by feedback inside the transistors. This feedback reflects the output tuning-circuit impedance into the input and causes violent variations in both the real and imaginary parts of the input impedance at frequencies near the output-circuit

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resonance. An analytical study is being undertaken to determine the seriousness of this difficulty. It is possible that in order to achieve interchangeability of transistors in the i-f strip, it will be necessary to provide a certain amount of resistive padding of the input impedance of the various stages as has been required in previous studies on point-contact transistors.

An investigation of an automatic volume control system for the i-f amplifier has been undertaken. It has been possible to obtain a 700-to-1 variation in the gain of a two-stage amplifier by varying the base dc voltage of the standard grounded-base i-f amplifier. At the present time, the design of an AVC amplifier is under way.

A beat frequency oscillator for the receiver has been constructed and operates satisfactorily.

A 100-kc/sec crystal calibrator has also been constructed, and satisfactory harmonics have been obtained up to approximately 12 Mc/sec. It has been difficult to obtain satisfactorily large harmonics above this frequency because of the low α cutoff of the commercially available fused-junction transistors. Satisfactory harmonics have been obtained up to 30 Mc/sec from a special blocking-oscillator calibration circuit which employs a developmental grown-junction n-p-n transistor obtained from the Germanium Products Corporation.

As suggested in the previous Quarterly Progress Report, crystal filters are being compared with the electromechanical variety for this receiver. Therefore EDT, ADP, and DKT have been considered for use in a variable-bandwidth crystal lattice filter with a center frequency of 455 kc/sec. There is no known cut of ADP with a zero temperature-coefficient of frequency. DKT is easier to grow, cut, and handle, and is also more stable with temperature than EDT. DKT contains water of crystallization, but it is quite tightly bonded. No noticeable dehydration takes place at 80° C over a week's time.

Ten DKT crystal blanks are on order from the Brush Electronics Company, and experimental work will start as soon as they are delivered.

S. Schwartz†, W. E. Morrow, Jr.†, C. R. Miller†,
D. I. Kosowsky, J. W. Craig, Jr.†

D. FREQUENCY MULTIPLIERS

Four types of frequency quadruplers have now been studied and compared. These are

a. A full-wave diode rectifier driving the emitter of a transistor: The conduction angle of the diodes may be controlled by proper bias. The impedance match between the generator resistance and the full-wave rectifier depends upon this conduction angle.

†Member of the Lincoln Laboratory.

Even with proper match, the quadrupler produces a signal power output 3 db below the signal power available from the generator.

b. A full-wave diode rectifier driven by the collector of a transistor: This circuit makes the loss associated with the rectifier diodes less significant than that of the circuit described in (a). The required additional transformer, however, adds enough new loss to make this circuit even less efficient than the first, unless a rather bulky powdered-iron core is used.

c. A high-Q ringing circuit with diode distortion and tuned transistor amplifier: Adequate output level could not be obtained without loading the ringing circuit so much that selectivity became inadequate.

d. A transistor as a nonlinear amplifier: The emitter is used as a half-wave rectifier, with the collector tuned to the fourth harmonic. Desired selectivity requires a tuned coupling transformer with an unloaded Q greater than 80. The (L/C) ratio is chosen as a compromise between gain and selectivity, the smallest ratio being 6×10^4 henrys/farad.

Two such quadruplers and a doubler have been cascaded to provide a multiplication factor of 32 (from 164 kc/sec to 5.25 Mc/sec). The over-all power gain of the chain is approximately 2 db, and it requires 200 mw of bias power.

<div align="right">J. C. Pinson</div>

E. DIRECTLY COUPLED AMPLIFIERS

Study of the limitations of junction transistors for use in directly coupled amplifiers has been continued. As previously reported (see Quarterly Progress Report, Research Laboratory of Electronics, M.I.T. Oct. 15, 1952), the major limitation is imposed by the temperature dependence of I_{co} (collector current with zero-emitter current). A few units have been tested over the range, -10° C to +55° C. The value of I_{co} may be expressed over this range as

$$I_{co} = A_o \exp\left[k_1(T - T_o)\right] \qquad (1)$$

where T_o is an arbitrary reference temperature in degrees centigrade, usually 25° C; T is the operating temperature in degrees centigrade; A_o is the measured value of I_{co} at the reference temperature T_o; and k_1 is a constant.

An average experimental value of k_1 is 0.08, with limits of 0.07 and 0.095. The over-all accuracy of the test was only approximately $\pm$ 10 percent, so the results check fairly well with the theoretical value for k_1 of 0.094 at 25° C.

In order to obtain gain with direct coupling, grounded-emitter and/or grounded-collector connections must be used. Assuming that all the drift in a complete amplifier arises from temperature variations of the I_{co} of the first stage, one can construct a

nomogram which is useful in evaluating the amplifier performance (see Fig. XI-2). Let I_o be the current caused at some reference point in the amplifier by the I_{co} of the first stage at T_o degrees C. Then according to Eq. 1

$$I_o = K A_o \qquad (2)$$

where K is some constant involving the circuit elements of the amplifier (related to the gain of the amplifier up to the reference point chosen). If the new current I_{max} at the reference point, due to I_{co} of the first stage, is desired at a new temperature T, it can be found from Eq. 1, since

$$I_{max} = K I_{co} \qquad (3)$$

at any temperature. The first three lines of Fig. XI-2 perform this computation, taking $\Delta \equiv 1$ for the moment. Thus in the example shown, an I_o of 25 μa at 25° C leads to an I_{max} of 68 μa at 37.5° C. The nomogram is based on the choice $k_1 = 0.08$, $T_o = 25°$ C in Eq. 1.

The third and fourth lines of Fig. XI-2 provide for an additional calculation. The expected maximum signal current level at the chosen reference point in the amplifier is presumably known, and the required dynamic range is also specified. These data fix the allowable drift Δ of the biasing current I_{max} at the reference point. Let I_{max} decrease by Δ as a result of a temperature decrease δT. Then by Eqs. 1 and 3

$$I_{max} = K A_o \exp\left[k_1(T - T_o)\right]$$

and $\qquad (4)$

$$I_{max} - \Delta = K A_o \exp\left[k_1(T - \delta T - T_o)\right]$$

Thus by division

$$1 - \frac{\Delta}{I_{max}} = \exp(-k_1 \delta T)$$

or $\qquad (5)$

$$\frac{I_{max}}{\Delta} = \frac{1}{1 - \exp(-k_1 \delta T)}$$

The third and fourth lines of Fig. XI-2 therefore relate δT and (I_{max}/Δ) according to Eq. 5, and the example shown illustrates that the ratio of I_{max} to Δ can be kept above 68 only by keeping δT less than 0.184° C.

The seriousness of this dynamic range restriction becomes more apparent when it is realized that all the temperatures to which we referred apply to the collector junction itself and not necessarily to the ambient temperature. Therefore in a directly coupled amplifier, the signal itself may cause appreciable temperature variations. As reported
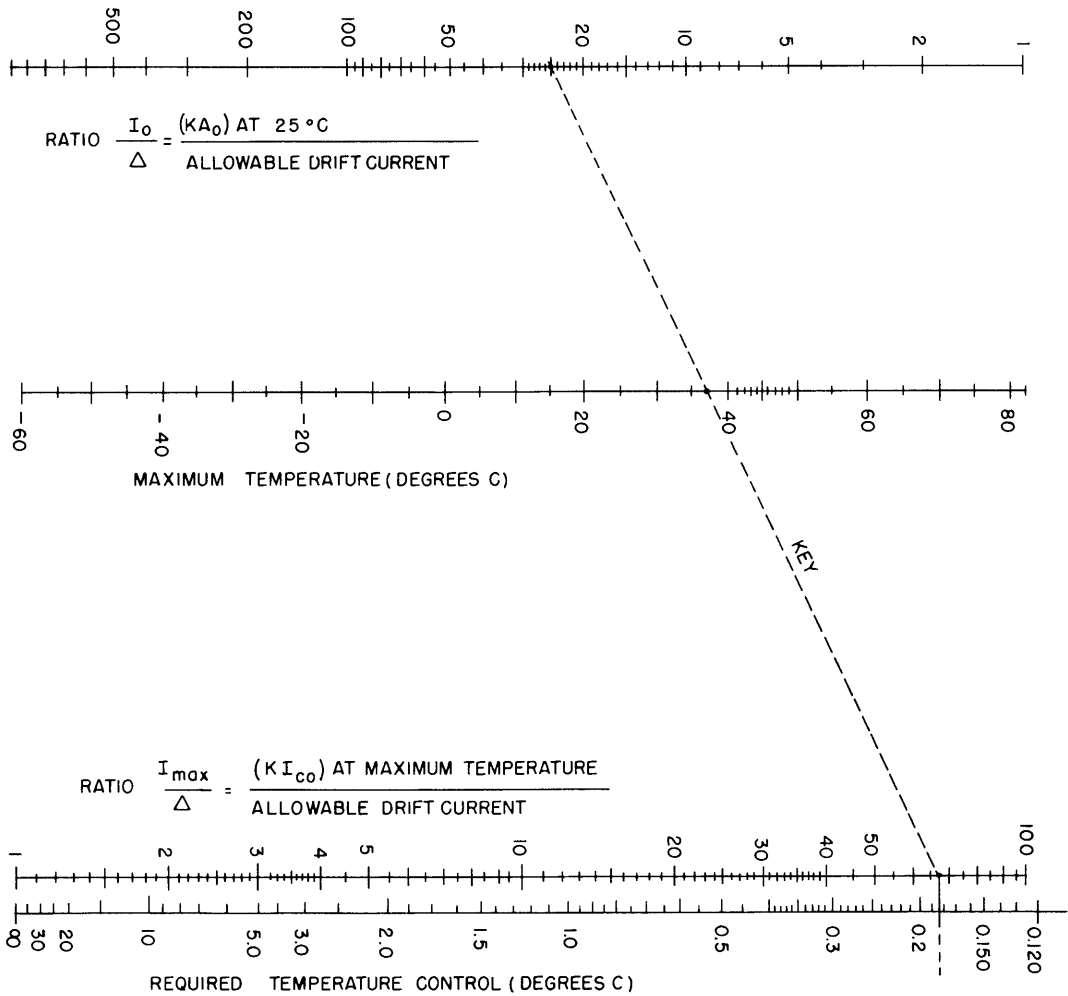
Fig. XI-2
Drift nomogram.

RATIO $\dfrac{I_0}{\Delta} = \dfrac{(KA_0) \text{ AT } 25\,^\circ C}{\text{ALLOWABLE DRIFT CURRENT}}$

MAXIMUM   TEMPERATURE (DEGREES C)

KEY

RATIO $\dfrac{I_{max}}{\Delta} = \dfrac{(K I_{co}) \text{ AT MAXIMUM TEMPERATURE}}{\text{ALLOWABLE DRIFT CURRENT}}$

REQUIRED   TEMPERATURE CONTROL (DEGREES C)

previously, efforts to employ balancing schemes to compensate the change in I_{co} of one stage by using that of another have been relatively unsuccessful. The trouble seems to arise from differences in the thermal time constants of the packaging, as well as the more obvious differences in k_1 and/or channel effect from unit to unit.

An additional simple example may point up both the use of Fig. XI-2 and the problem of dynamic range. Suppose a transistor having $a = 0.99$ is used in the most elementary grounded-emitter connection as the first stage of a dc amplifier which must operate from a high-impedance signal source of a maximum level of 10 μa. The signal current at the collector will then be 1000 μa. A reasonable dynamic range of 30 db is required, which limits Δ to approximately 33 μa. Assume that the temperature will vary between 25° C and 30° C so that $\delta T = 5°$ C. From Fig. XI-2, therefore, $(I_{max}/\Delta) = 3$ at 30° C, which leads to $(I_o/\Delta) = 2$ at 25° C. Thus $I_o = 66$ μa must characterize the collector current of the first stage in the absence of signal. This current is, however, $I_{co}/(1-a)$, that is, $K = 1/(1-a)$, for the grounded-emitter connection with an open circuited base. Hence a transistor with $I_{co} = 0.66$ μa at 25° C is required, even though the temperature rise is restricted to only 5° C. Note that bias-stabilization schemes to avoid the treacherous open-base bias condition in this example are of little help in the dc amplifier problem because they reduce the signal gain as much as they reduce the drift. The difficulty is merely postponed until the next stage.

It is possible to envision four solutions to these problems: (a) Small-area, low-resistivity germanium junction transistors might be designed to reduce I_{co}. (b) Unipolar transistors might be considered, if specially designed. (c) Silicon junction transistors might be employed, if they are available. (d) Refrigeration of germanium transistors might be included in the circuit design.

C. R. Hurtig

F. CIRCUIT NOISE PROBLEMS

1. According to simple energy considerations, the validity of the 1/f law for the power density spectrum of the excess noise from a crystal should be limited to a finite band of frequencies. With the thought that the limits of validity of the law might be simply indicated by the behavior of the autocorrelation function of the noise, we hypothesized a power density spectrum of the form shown in Fig. XI-3(a). With ω_o and ω_1 as parameters, the autocorrelation functions corresponding to various spectra of this form were calculated. A typical autocorrelation function is approximated in Fig. XI-3(b), where A depends only upon ω_1, and R(o) depends in a simple way on both ω_o and ω_1.

It appears from this result that a measurement of the autocorrelation function of crystal noise might indicate an approximate range of validity of the 1/f law.

2. The possible implications for transistor circuits of the modulation noise from

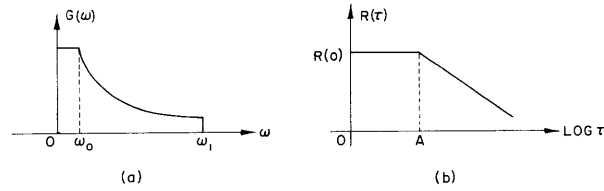Fig. XI-3

Idealized power spectrum, $G(\omega)$, and corresponding auto-correlation function for excess crystal noise.

a crystal diode, as observed by C. Dean (1), have led to some interesting speculations, which are presented below.

According to Dean's ideas, the volt-ampere characteristic of the diode may fluctuate with time in a random way, so the instantaneous voltage e across it is related to the instantaneous current i through it by a relation of the form

$$e = f(i, t) \tag{1}$$

Suppose a current i is impressed on the diode, such that

$$i = I_o + s(t) \tag{2}$$

where I_o is a constant dc reverse bias, and s(t) is a much smaller signal of arbitrary spectral density. Then a Taylor expansion of e about I_o gives

$$e = f(I_o, t) + \left(\frac{\partial f}{\partial i}\right)_{I_o} s + \ldots \tag{3}$$

provided s(t) is small enough.

Now $(\partial f/\partial i)_{I_o}$ is a function of time, as well as of I_o, and it presumably has a time-average value R_o which is the normal incremental back resistance of the diode. Splitting $(\partial f/\partial i)_{I_o}$ into its time-average and time-dependent parts, we have

$$\left(\frac{\partial f}{\partial i}\right)_{I_o} = R_o + r_o(t) \tag{4}$$

Similarly, writing the voltage $f(I_o, t)$ as $V_o + v_{N_o}(t)$ and using Eq. 4, we find that Eq. 3 becomes

$$e = V_o + v_{N_o}(t) + R_o s(t) + r_o(t) s(t) \tag{5}$$

where presumably $v_{N_o}(t)$ and $r_o(t)$ are random functions having zero time averages. The incremental equivalent circuit of the diode, according to Eq. 5, would be that shown in Fig. XI-4. From these results, we are led to conclude immediately that:
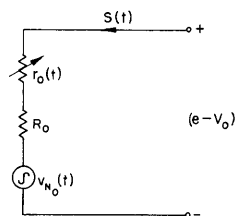
Fig. XI-4
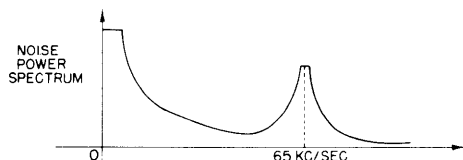
Incremental equivalent circuit
of a crystal diode.

Fig. XI-5

Possible noise power spectrum
(not to scale).

a. If $s(t) = 0$, the diode noise voltage is simply $v_{N_0}(t)$. Hence, the usual open-circuit noise measurement is unaffected by, and does not bring out, the fluctuation of the incremental resistance $r_0(t)$.

b. If an incremental signal current $s(t)$ is applied to a biased diode, one component of the total noise voltage is proportional to the signal amplitude. For this component, the signal-to-noise ratio is independent of the signal amplitude (within the limits of Eq. 3, of course) and depends only upon the bias and the diode characteristics. This signal-to-noise ratio might be characterized, for example, by the ratio of the rms value of $r_0(t)$ to the incremental resistance R_0.

Many measurements have been made of the power spectrum of $v_{N_0}(t)$. These suggest that in the range $\omega_0 < \omega < \omega_1$ this spectrum is something like that shown in Fig. XI-3(a), with $\omega_0 < 10^{-3}$ radians per second (2) and ω_1 somewhere above 10^5 radians per second. Moreover, Dean's work suggests that the power spectrum of $r_0(t)$ is similar to that of $v_{N_0}(t)$, at least in the frequency band from 0 cps to approximately 1 cps. He employed for $s(t)$ a 65-kc/sec carrier, so that the total spectrum of the noise voltage presumably appeared like that shown in Fig. XI-5. Dean's results show that the noise spectrum between 0 cps and 1 cps is nearly identical to that in the one-cycle band above 65 kc/sec. The latter was, however, very much smaller in magnitude than the former, even when the amplitude of $s(t)$ was made comparable to I_0.

The point of interest here, on the other hand, is that (on the basis of this model) a limiting signal-to-noise ratio is encountered when the frequency of operation is moved up until the spectrum of $v_{N_0}(t)$ alone has dropped to a negligible value. This limiting ratio may or may not be serious, depending upon the relative sizes of $r_0(t)$ and R_0. Crude direct measurements have been attempted using a wave analyzer to examine the spectrum about a 15-kc/sec carrier, but the results were inconclusive because of carrier interference. The demodulation scheme used by Dean would not suffer from this drawback.

M. Strieby

References

1. C. Dean: Lincoln Laboratory Quarterly Progress Report of Group 35, Transistors and Solid State, Oct. 1, 1952 (unclassified)

2. B. V. Rollin: Noise in Semiconductors at Very Low Frequencies, Proc. Phys. Soc. 66 B, 259, 1953 (work performed on carbon filaments)