TRANSISTORS: Theory and Application

Small-Signal

Part VIII

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SMALL-SIGNAL parameters are of considerable importance in specifying transistor operation. Most of the parameters discussed thus far in this series are based on small-signal operation, and large errors can result from neglect of this qualifying phrase.

Power Gain

In Part VII of this series, equations were developed which express the input and output resistance and the voltage gain of point-contact and junction transistors. Typical numerical values were given to serve as a general orientation with regard to orders of magnitude (see Table I). The power gain of an electrical device is, obviously, an important concept to the design engineer and equations will be developed for computation of this parameter of a transistor.

Figure 1 shows the familiar circuit of a generator with voltage e_g and internal resistance R_g supplying a load whose resistance is R_L. The maximum power that can be developed across R_L for a given generator may be found as follows: Since

$$E = \frac{e_g R_L}{R_g + R_L} \qquad (1)$$

$$P_o = \frac{e_g^2 R_L}{(R_g + R_L)^2} \qquad (2)$$

Differentiation with respect to R_L shows that the power delivered across the load will be a maximum if R_g is equal to R_L. Under these conditions the maximum possible power that can be drawn from generator e_g under any circumstances will be

$$P_m = \frac{e_g^2}{4R_g} \qquad (3)$$

The power output of a grounded-base transistor amplifier (Fig. 2) is in general given by $i_2^2 R_L$ where i_2 is

$$i_2 = \frac{-e_g(r_b + r_m)}{[(R_g + r_b + r_e)(R_L + r_b + r_c) - r_b(r_b + r_m)]} \qquad (4)$$

The power gain of the transistor will then be given by the ratio of $i_2^2 R_L$ to P_m, the maximum power which may possibly be drawn from the generator (generator internal resistance exactly matched by input resistance of the transistor). In general, this will not be the case and less power will be drawn from the generator than is indicated by Eq. 3. The power gain is given by

$$PG = \frac{i_2^2 R_L}{e_g^2/4R_g} = \frac{4R_L R_g i_2^2}{e_g^2} \qquad (5)$$

The power gain so determined for a given value of i_2 and R_L will be a lower limit or a minimum value since the power expression in the denominator is the maximum possible under these conditions.

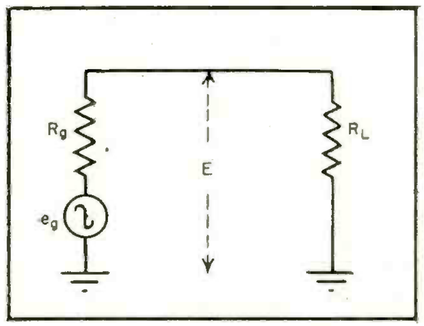

FIG. 1—Equivalent circuit illustrates condition for maximum load power

Substituting the value of i_2 from Eq. 4 into the expression for the power gain in Eq. 5

$$PG = \frac{4R_L R_g (r_b + r_m)^2}{[(R_g + r_b + r_e)(R_L + r_b + r_c) - r_b(r_b + r_m)]^2} \qquad (6)$$

This expression can be simplified somewhat since r_b is small compared to r_e or r_m (see Table I) and Eq. 6 becomes

$$PG = \frac{4R_L R_g r_m^2}{[(R_g+r_b+r_e)(R_L+r_c) - r_b r_m]^2} \qquad (7)$$

Typical Values

Using the typical values of the parameters as given in Table I for the point-contact transistor, a typical power gain is very nearly 100 representing 20 db whereas for the junction type a typical power gain is very nearly 440 representing 46 db.

Importance of Alpha

It is often desirable to know power gain as a function of the alpha of the circuit. Dividing numerator and denominator of Eq. 6 by $(r_e + r_b)^2$

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Transistor Operation

Different approaches to mathematical analysis of small-signal transistor operation are explained in detail, along with listings of advantages and disadvantages of each. Power gain and other important parameters are discussed.

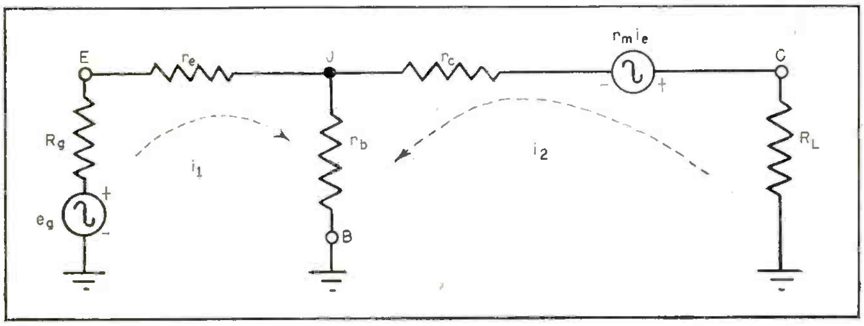

FIG. 2—Equivalent circuit of grounded-base transistor from which design equations for small-signal operation are devised

$$PG = \frac{4R_L R_g \alpha^2}{\left[(R_g+r_b+r_e)\left(1+\frac{R_L}{(r_c+r_b)}\right) - r_b\alpha\right]^2} \qquad (8)$$

Similarly, the power gain may be expressed as a function of VG and α.

$$PG = \frac{4R_g\alpha(VG)}{(R_g+r_e+r_b)\left(1+\frac{R_L}{r_b+r_c}\right) - r_b\alpha} \qquad (9)$$

Also, the expression for the VG obtained in the previous article[1] may be written as a function of α.

$$VG = \frac{\alpha R_L}{(R_g+r_e+r_b)\left(\frac{R_L}{r_b+r_c}+1\right) - r_b\alpha} \qquad (10)$$

Equations 8, 9 and 10 show why alpha is one of the important parameters for evaluation and comparison of transistors. Also, the relation $\alpha = -i_c/i_e$ shows the further usefulness of alpha as a comparison number for current gain.

Grounded-Base Equations

In the previous article of this series[1], Eq. 11 and 12 were derived to express conditions in the grounded-base transistor arrangement.

$$i_1(R_g + r_e + r_b) + i_2 r_b - e_g = 0 \qquad (11)$$

$$i_1(r_b + r_m) + i_2(R_L + r_b + r_c) = 0 \qquad (12)$$

If in Eq. 11 $r_e + r_b = r_{11}$ and $r_{12} = r_b$ according to the definitions laid down in the previous article[1], and $v_1 = e_g - i_1 R_g$, which is the net voltage acting at the transistor terminals, Eq. 11 becomes

$$v_1 = r_{11}i_1 + r_{12}i_2 \qquad (13)$$

In deriving Eq. 12 it was assumed that there was no generator voltage acting in loop 2 of Fig. 2. If there were, Eq. 12 could be written as

$$r_{21}i_1 + i_2R_L + i_2r_{22} = e_{g2} \qquad (14)$$

where again $r_m + r_b = r_{21}$ and $r_c + r_b = r_{22}$. As before, if $v_2 = e_{g2} - i_2R_L$, the net effective voltage after the i_2R_L drop in Eq. 14 becomes

$$v_2 = r_{21}i_1 + r_{22}i_2 \qquad (15)$$

Small-Signal Parameters

Consider a dependent variable y which varies with, or is a function of an independent variable x. Mathematically

$$y = y(x) \qquad (16)$$

Differentiating both sides with respect to x

$$\frac{dy}{dy} = \frac{dy(x)}{dx} \qquad (17)$$

Essentially, Eq. 17 is a tautology, or self-evident identity, and does not provide any particularly useful information. But in the form

$$dy = \frac{dy(x)}{dx}\,dx \qquad (18)$$
$$= y'\,dx$$

the equation states a very important and useful fact. Equation 16 will, in general, avoiding special curves, give a graph like in Fig. 3.

Equation 18 says that if, for a given or selected change in x, say of dx, it is desired to find the corresponding change in the value of y, when y depends on x in the manner shown in Fig. 3 and symbolized in mathematics shorthand by Eq. 16, the slope of the curve at the point in question is multiplied by the given value of dx.

If y is a dependent variable which varies with either or both of two parameters, x and z, Eq. 16

Table I—Typical Transistor Parameters

Parameter	Point Contact	Junction
r_e	150 ohms	25 ohms
r_b	120 ohms	500 ohms
r_m	35 kilohms	0.96 megohm
r_c	15 kilohms	1.0 megohm
R_g	500 ohms	500 ohms
R_L	20 kilohms	100 kilohms
α	2.3	0.96
e_g	0.01 volt	0.001 volt

will then become

$$y = y(x,z) \qquad (19)$$

Since a change in y, dy, may now be due to a change in either or both of x and z, an equation entirely analogous to Eq. 17 would be Eq. 20

$$dy = \left.\frac{dy}{dx}\right]_{z-k} dx + \left.\frac{dy}{dz}\right]_{x-k} dz \qquad (20)$$

This is usually written

$$dy = \left.\frac{\partial y}{\partial x}\right]_z dx + \left.\frac{\partial y}{\partial z}\right]_x dz$$

The dimension of each term on the right-hand side of Eq. 20 is dy, since dx and dz cancel—at least dimensionally. This must be so for any physically valid interpretation. Also dy/dx, the first term right-hand side, is not quite the same as in Eq. 17 because the change in dy is measured for a unit change in dx, while z is kept constant. A similar remark applies to dy/dz for x = constant.

In general, V_e, the d-c emitter voltage, will depend on the emitter and collector currents or on the operating point. Thus

$$V_e = V_e(I_e, I_c) \qquad (21)$$

Applied to Eq. 19, $y \approx V_e$, $I_e \approx x$, $I_c \approx z$. The dependent variable is voltage V_e and the independent variables are I_e and I_c. This shows that the transistor is a current-operated device and the currents are the independent variables. In transistor work the currents are adjusted to the correct operating point, and the voltages appearing are then fixed by the current values.

To illustrate further the effect of this concept in practice, the static characteristics of transistors are plotted, by common consent, with current as abcissa and voltage as ordinate, in keeping with the mathematical convention of plotting the independent variable along the x direction.

Analogous to Eq. 20, obtained from 19, is Eq. 22 for Eq. 21.

$$dV_e = \left.\frac{dV_e}{dI_e}\right]_{Ic-k} dI_e + \left.\frac{dV_e}{dI_c}\right]_{Ie-k} dI_c \qquad (22)$$

The d-c collector voltage also depends on emitter and collector currents, and

$$V_c = V_c(I_e, I_c) \qquad (23)$$

Again on differentiating

$$dV_c = \left.\frac{dV_c}{dI_e}\right]_{Ic-k} dI_e + \left.\frac{dV_c}{dI_c}\right]_{Ie-k} dI_c \qquad (24)$$

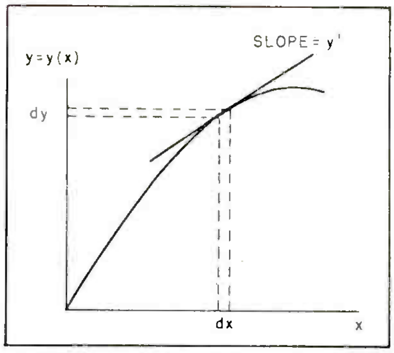

FIG. 3—Typical curve showing y as a function of x. Slope is $y'=dy(x)/dx$

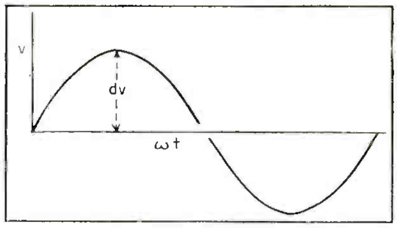

FIG. 4—Sinusoidal voltage (or current) used to simulate differential increments

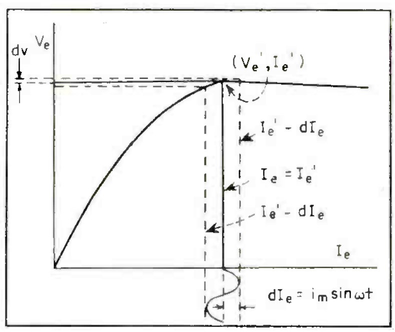

FIG. 5—Plot of emitter voltage versus emitter current

Equations 22 and 24 are extremely important in transistor analysis.

Small Changes

A simple explanation of Eq. 24 might be based on an a-c voltage and current wave as shown in Fig. 4, where the sinusoidal voltage wave $v = v_m \sin \omega t$ where $v_m = dv$.

In an interval $\omega t = \pi/2$, the amplitude changes from 0 to dv, and in the next interval of time from dv to 0, so that, disregarding the sign, the amplitude changes by dv for each ¼ cycle. In general, Eq. 22 and 24 apply properly if dv is actually infinitesimal, or extremely small, but in actual application a practical value of dv may be used.

In Fig. 5 is shown a possible plot of d-c values of voltage vs current, say of the emitter voltage V_e and emitter current I_e for a transistor. Assuming the operating point of the transistor is at V_e', I_e', if the emitter current is modulated sinusoidally by a small signal so that I_e varies at most by dI_e, then I_e will vary from I_e' to $I_e' + dI_e$, from $I_e' + dI_e$ to I_e', from I_e' to $I_e' - dI_e$, and so forth. The quantitative relation between the change dI_e in I_e' and the change dV_e in V_e, assuming that V_e depends on I_e alone, is

$$dV_e = \frac{dV_e}{dI_e} dI_e$$

as in Eq. 18.

Just as $dy/dx = y'$ is the slope of the $y = y\ (x)$ curve, so dV_e/dI_e is the slope of the characteristic curve of Fig. 5. (Note that dV_e/dI_e has the dimensions of a resistance, so that in this case the slope is a resistance, hence designated by R.) The slope of the curve must be taken virtually at a point if the relation is to be true.

$$dV_e = R dI_e \qquad (25)$$

Signal Limitations

If in using Eq. 25 the slope over a large region is used the condition shown in Fig. 6 occurs. The slope at point (V_e', I_e') is actually S_1 as shown, but using a large swing about (V_e', I_e') gives a slope S_2 which is in general not equal to S_1. Thus, using a large distance about the point in question to compute the slope may introduce large errors; the larger the distance used, the larger may be the error, particularly as the curve departs more and more from a straight line. If a sinusoidal current is used to simulate dI_e, it must be a very small current or signal, or else large errors may be introduced.

In view of these remarks, the differential quantity dV_e is representable, for purposes of a laboratory experiment, by an a-c voltage, and dI_e by an a-c current, *provided they are suitably small.*

Regardless of whether V_e is a function of one or two variables, the remarks about the size of the signal are still valid.

Equations 22 and 24 may be written (Eq. 26 and 27 respectively) in terms of small a-c v's and i's by applying the foregoing principles.

$$v_e = \frac{v_e}{i_e}\Big]_{I_c=k} i_e + \frac{v_e}{i_c}\Big]_{I_e=k} i_c \quad (26)$$

$$v_c = \frac{v_c}{i_e}\Big]_{I_c=k} i_e + \frac{v_c}{i_c}\Big]_{I_e=k} i_c \quad (27)$$

There are several comments which must be made about these two equations: (1) Note the use of small letters for a-c values, and capitals for d-c values, in keeping with the convention adopted by the authors for this series, and which is a proposed standard.

(2) Consider the $I_c = k$, $I_e = k$ factors of Eq. 26 and 27. If a change dI is representable by a small a-c current, $i \sin \omega t$, as has been shown, of amplitude i, then 0 change in d-c current, $I_c = k$, or $I_e = k$, will be represented by an a-c current of 0 amplitude. Using this fact we may properly write Eq. 26 and 27 as

$$v_e = \frac{v_e}{i_e}\Big]_{i_c=0} i_e + \frac{v_e}{i_c}\Big]_{i_e=0} i_c \quad (28)$$

$$v_c = \frac{v_c}{i_e}\Big]_{i_c=0} i_e + \frac{v_c}{i_c}\Big]_{i_e=0} i_c \quad (29)$$

(3) At $i_c = 0$, v_e/i_e is r_{11} as described in Part VII of this series.

$$r_{11} = \frac{v_e}{i_e}\Big]_{i_c=0} \quad (30)$$

Using the definitions laid down in Part VII

$$\frac{v_e}{i_c}\Big]_{i_e=0} = r_{12}$$

$$\frac{v_c}{i_e}\Big]_{i_c=0} = r_{21}$$

$$\frac{v_c}{i_c}\Big]_{i_e=0} = r_{22}$$

Hence Eq. 28 and 29 can be rewritten

$$v_e = r_{11}i_e + r_{12}i_c \quad (31)$$

$$v_c = r_{21}i_e + r_{22}i_c \quad (32)$$

Equation 31 states that a potential v_e is acting in a circuit and produces the two potential drops $r_{11}\,i_e$ and $r_{12}i_c$ which together make up v_e. Similar remarks apply to Eq. 32. Replacing subscript e in the emitter circuit by subscript 1, and subscript c by 2, Eq. 31 and 32 become

$$v_1 = r_{11}i_1 + r_{12}i_2 \quad (33)$$

$$v_2 = r_{21}i_1 + r_{22}i_2 \quad (34)$$

Equations 13 and 15 are identical to Eq. 33 and 34 respectively, but were arrived at by entirely different methods of reasoning. Equations 13 and 15 were obtained from a circuit analysis using Kirchhoff's law for the voltage loops; Eq. 33 and 34 were derived on a mathematical basis from the simple concept that V_e and V_c are dependent on both I_c and I_e in small-signal operation.

As has been mentioned, the ultimate fact mathematically expressed by these equations is that, given all the other data, it is possible to find two numbers, i_1 and i_2, that satisfy these equations. The fact that they happen to represent a current, in milliamperes perhaps, does not influence the mathematical description of how to determine these two numbers. Since Eq. 33 and 34 are linear equations (that is, they contain no products like $i_1 \times i_2$, or i raised to some power) the delineation is unique; that is, the equations will define only one pair of values, i_1 and i_2, and no other.

Consequently, whatever assumptions were made to find Eq. 33 and 34 are also binding on Eq. 13 and 15 because they represent the same numbers.

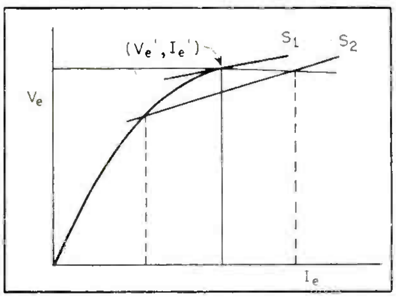

FIG. 6—Sketch shows error in determining slope with large signal input

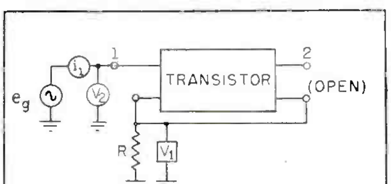

FIG. 7—Connections for measuring r_{11}

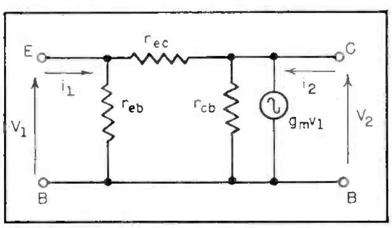

FIG. 8—Equivalent pi circuit of transistor

Assumptions

First, from the discussion in connection with Eq. 31 and 32, the parameters are open-circuit values, that is, $i_c = 0$, when measuring r_{11} and r_{21} and $i_e = 0$ when measuring r_{12} and r_{22}. This condition has already been met for Eq. 13 and 15[1].

Second, from the discussion in connection with Eq. 26 and 27, the parameters are measured using ratios of a-c values. This condition has also been met[1].

Third, the symbol convention that

$$v_1 = e_{g1} - i_1R_g$$

$$v_2 = e_{g2} - i_2R_L$$

assumed to obtain Eq. 13 and 15, is binding on Eq. 33 and 34.

Last, and probably most important, from the discussion in connection with Eq. 26 and 27, the use of small signals is assumed. This means that the four-pole parameters r_{11}, r_{12}, r_{21} and r_{22} are small-signal parameters.

In selecting an amplitude for the measurement of the four-pole parameters, several factors must be considered.

If the signal is not truly small, different signal amplitudes will give different values of parameters.

In practice, a signal of some convenient size is selected, and the parameter, say r_{11}, measured. Then the signal is reduced slightly, and the parameter measured again. If the value of r_{11} obtained in the second trial is close to the r_{11} originally obtained, within the accuracy desired, the signal is a small signal and conversely.

It might appear that there is no objection to using a very small amplitude signal originally, so that there is no question about the signal being small. If the V-I characteristic is very curved, it is very difficult to guess what constitutes a truly small signal. The test suggested above should therefore always be made. Too small a signal introduces measuring problems such as noise and the general difficulty of measuring fractions of a microvolt at a-c.

When feasible, the parameter may be measured by the volt-ammeter method tacitly assumed here, and the resulting value compared to that obtained by using an entirely different measuring scheme. Other possible methods include the bridge method, scope presentation, and variations and combinations of these. The results must compare within the accuracy desired if the signal is truly small.

Open Circuit

Let us now review in a general way the data that has been assembled regarding r_{11}, r_{12}, r_{21} and r_{22}. Because these parameters are so important in the specification of transistor characteristics, much experimental and theoretical work has been done and is continuing regarding these and other suitable parameters.

These parameters are called small-signal open-circuit grounded-base four-pole parameters. In this and previous articles the terms small-signal, grounded-base, and four-pole parameters have been explained. To understand the open-circuit term, consider Fig. 7 which shows the circuit arrangement for measurement of r_{11}.

Both V_1 and V_2 are vacuum-tube voltmeters, and R is a small resistance of known value placed in the signal generator return to measure the current $i_1 = V_1/R$. Voltmeter V_2 measures the voltage across the input of the transistor, and $r_{11} = V_2/i_1$. There is not a true open circuit across the output or collector circuit since whatever scheme is used to bias the collector, a d-c return path to the base is essential, and thus a d-c path is always present to act as a closed circuit across the collector. The internal resistance of the d-c bias supply may be made very high by using a choke of several hundred henries but not infinite. In most tests $f = 270$ cps.

Also, there is an a-c shunting path always present due to internal transistor and stray capacitance. The impedance of this shunting path decreases as the frequency increases.

A typical value of C_c, the collector capacitance, is 50 $\mu\mu$f and a typical circuit may have an additional stray capacitance of 10 $\mu\mu$f. At 270 cps, $X_c = 9.8$ megohms. If r_{22} were three megohms, at 270 cps the 9.8 megohms would represent a shunt path of some importance, and it is not true that measurements are being made under open-circuit conditions.

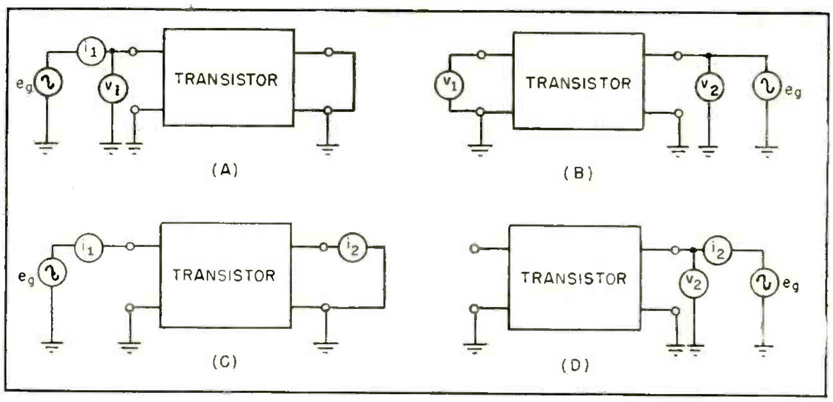

FIG. 9—Sketches show connections for measuring g's or short-circuit conductances

Two principal approaches to this problem have been adopted.

(1) The arbitrary convention that the circuit is open if the shunt resistance is fifty times the internal transistor resistance. Thus when measuring r_{11} and r_{21}, where the collector circuit must be open, the shunt path shall be $50 \times r_{22}$; and when measuring r_{12} or r_{22} the shunt path shall be $50 \times r_{11}$. Or, where feasible, the collector-to-base resistance (or emitter-to-base resistance) may be doubled. If the value of the parameter measured remains invariant within the accuracy desired, the circuit is truly an open circuit.

(2) Measurement of other parameters such as the g's and h's described below to characterize transistor operation, whose measurement does not require the use of an open circuit across high impedance elements. Then, if desired, parameters such as the r's may be derived mathematically from the quantities measured. Or, the performance of the transistor may be expressed in terms of these new parameters directly, it being assumed that with experience engineers will readily compare transistor performance in terms of these parameters.

Conductances

It was mentioned that the equivalent T as characterized by Eq. 13 and 15 is not the only equivalent circuit for the specification of the allegorical black box. The black box or transistor is equally well represented by an equivalent circuit which is a π network,[2] as in Fig. 8. This circuit is analyzed by means of Eq. 35 and 36 using conductances.

$$i_1 = g_{11}v_1 + g_{12}v_2 \quad (35)$$

$$i_2 = g_{21}v_1 + g_{22}v_2 \quad (36)$$

Using Eq. 35 and 36 all the four g's can be measured. For illustration, setting $v_2 = 0$, $g_{11} = i_1/v_1$ and similarly one can find g_{12}, g_{21} and g_{22}; see Fig. 9. These g's are sometimes called the short-circuit conductances.

While it is true that resistances and conductances are reciprocals, the r_{11} which appears in Eq. 13 is not the reciprocal of g_{11}. In Eq. 13 r_{11} was obtained using an open-circuited collector, but g_{11} is obtained by using a short-circuited collector. Hence, r_{11} is not $1/g_{11}$ and is, in fact, given by

$$r_{11} = \frac{g_{22}}{g_{11}g_{22} - g_{12}g_{21}} \quad (37)$$

To obtain g_{11}, v_2 is set to 0. In other words, the output is short circuited. This does away with the need for establishing an open circuit. Like any other good idea in engineering, it has its drawbacks. To find g_{12} the emitter circuit may be shorted, making $v_1 = 0$, and so on for all the other parameters.

The disadvantage of the g method is that some point-contact transistors may be short-circuit unstable. If the input or the output is a-c short-circuited, the units may break into parasitic oscillations. Fortunately, the applicability of this method is greatest for junction types, where this instabil-

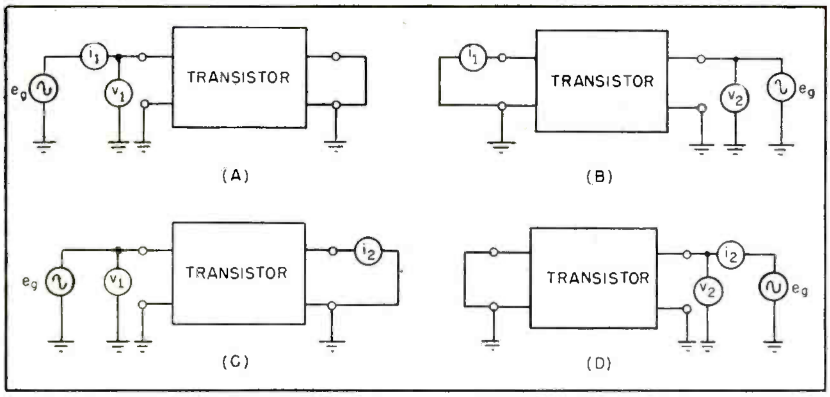

FIG. 10—Circuits for measuring *h*'s or hybrid transistor parameters

ity is not normally encountered. The inference is that for point-contact units, the r's may be suitable, and with junction units, the g's may be preferable.

Hybrid Parameters

Because the use of the g's suggests that two different parameters are needed to take care of point-contact and junction units, an entirely different set has been suggested.[3] These are the hybrid parameters, or h's, characterizing the transistor by the following equations

$$v_1 = h_{11}i_1 + h_{12}v_2 \qquad (38)$$

$$i_2 = h_{21}i_1 + h_{22}v_2 \qquad (39)$$

To find h_{11}, the output is shorted, (see Fig. 10) $v_2 = 0$, and

$$h_{11} = \frac{v_1}{i_1} \qquad (40)$$

The r_{11} of Eq. 13 was obtained with open-circuited output, hence h_{11} is not the same as r_{11}. As g_{11} was obtained with $v_2 = 0$, $h_{11} = 1/g_{11}$, so that the g_{11} of Eq. 35 is the reciprocal of h_{11} in Eq. 35; and since the reciprocal of a conductance is a resistance, h_{11} is of the nature of a resistance, dimensionally.

To find h_{12}, i_1 is made 0 by making the input an open circuit, placing the generator in the collector circuit, and measuring v_1 and v_2. Then

$$h_{12} = \frac{v_1}{v_2} \qquad (41)$$

As this is the ratio of two voltages, h_{12} has no dimensions. A similar analysis will show, for Eq. 39, that $h_{21} = i_2/i_1$ for a short-circuited collector. This is α_{ce} by definition; $h_{22} = i_2/v_2$ is conductance but not identical with g_{22} of Eq. 36 since all the parameters of Eq. 36 are obtained under short-circuit conditions. Note that h_{22} is obtained under open-circuit conditions.

Thus the hybrid parameters contain two pure numerics, h_{12} and h_{21}, a resistance h_{11} and a conductance h_{22}. They possess, however, some of the advantages of both the r's and the g's:

(1) In the matter of avoiding the necessity for maintaining an open circuit in the high-resistance collector, the h method shares the advantage of the g system in that h_{11} and h_{21} are made with collector shorted.

In the matter of avoiding the necessity for maintaining a short circuit in the low-resistance emitter circuit, the h system shares the advantage of the r system, since both h_{12} and h_{22} are made with emitter open circuited. Since the input resistance of some transistors may be quite low, of the order of tens of ohms, difficulties are encountered in effectively a-c short circuiting such impedances. In the h system, the input is open circuited for such measurements (h_{12} and h_{22}).

(2) Measurements for alpha in the g and r methods are indirect. In the r method, r_{21} and r_{22} are found; their ratio is alpha $= r_{21}/r_{22}$. In the g method, alpha $= g_{21}/g_{22}$; but in the hybrid parameter method, alpha $= -h_{21}$ directly.

The principal disadvantages of the h's is that they are not directly amenable to circuit analysis. Whereas most engineers are acquainted with resistances and conductances and use them readily in circuit analysis, few are prepared to use h's directly. If transformations are necessary, and these are quite cumbersome, the advantages mentioned above may well be overshadowed. Engineers are currently studying the relative merits of each set of parameters. The indications are that considerable additional experience is needed before a single set of parameters to characterize transistors will be generally adopted.

Resume

In resume, the following are the salient points:

(1) Typical power gains that are possible with the grounded-base transistor connection are: point-contact, 20 db; junction, 46 db.

(2) In transistor terminology r_{11}, r_{12}, r_{21} and r_{22} are called small-signal, grounded-base, open-circuit, four-pole, equivalent-circuit parameters. Each part of the name has significance.

(3) A test to determine whether a signal is truly a small signal is to measure the parameters using a selected amplitude of signal, decrease the signal by approximately 50 percent and remeasure the parameters. The measured values must compare within the orders of accuracy required.

(4) Either the r's, or the g's or h's may be used to characterize the transistor equivalent circuit. Considerations such as common usage, short-circuit stability, and effective open and short circuit are involved in selecting the parameters for a given analysis.

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