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# RANSISTOR RC AMPLIFIERS

By

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# DESIGN OF TRANSISTOR RC AMPLIFIERS

by

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Variation in transistor characteristics is one of the greatest problems encountered in the design of transistor circuits. Failure to take these variations into account will generally result in an unsatisfactory circuit design. For example, let us suppose that an unstabilized audio amplifier (Fig. 1) is assembled and that $R_B$ and $R_C$ are adjusted so as to give best performance (gain and fidelity) with a particular transistor at room temperature. Now if this amplifier is duplicated using the same values for $R_B$ and $R_C$ and the same type transistor, in many cases the performance will be unsatisfactory for one or both of the following reasons: 1) incorrect operating point due to dissimilarity of transistors of the same type, 2) incorrect operating point due to changes in transistor characteristics caused by temperature variation.

Those who are new to the transistor field sometimes underestimate the importance of the above effects until they find that a circuit may become completely inoperative when the transistor is replaced with another of the same type, or when an amplifier built in the cool of the evening just won't operate in the warmth of the afternoon.

Here we will consider two methods of dealing with the problem of variation in characteristics. The first will apply to an unstabilized

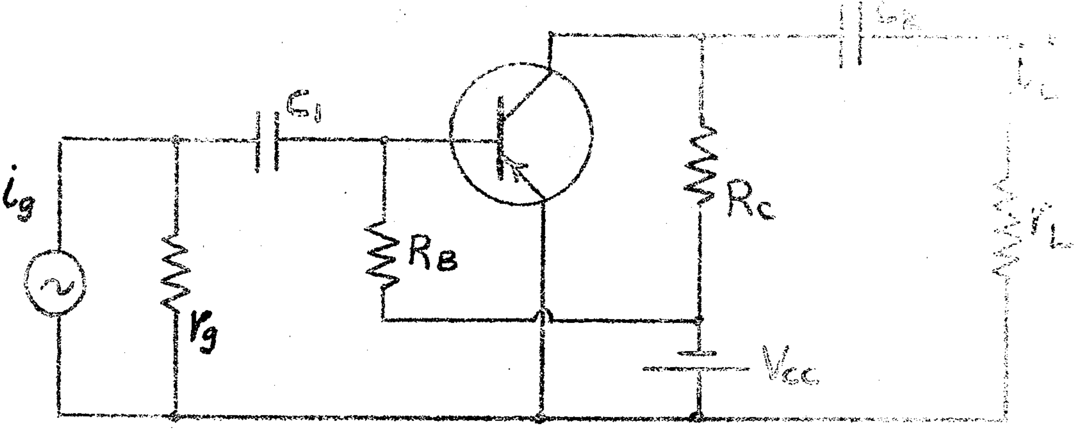

FIG. 1 UNSTABILIZED AMPLIFIER.

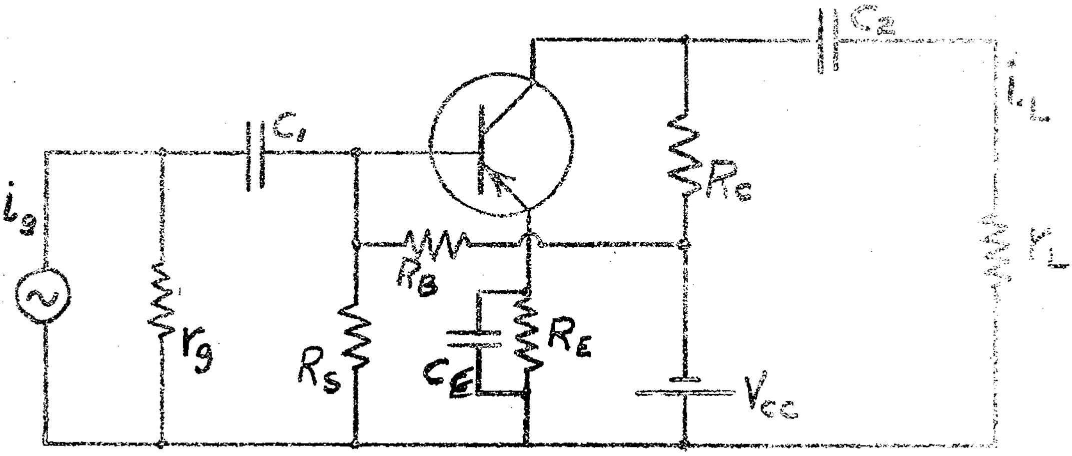

FIG. 2 STABILIZED AMPLIFIER.

amplifier (Fig. 1) in which the effects of temperature and parameter variation due to production spread are handled by proper choice of operating point and circuit resistors. The second method utilizes a stabilized circuit (Fig. 2) in which d-c negative feedback tends to make the circuit performance less sensitive to changes in transistor characteristics.

## Preliminary Considerations

In the beginning it is well to realize that it is difficult if not impossible to produce a "paper design" of a complete amplifier system in which the usual performance factors of gain, non-linear distortion, phase shift, frequency response and temperature characteristics are met with a high degree of accuracy. After initial testing, the first "paper design" may be modified considerably by adding inverse feedback circuitry to meet distortion specifications, addition of equalization circuitry to produce the desired frequency response and possibly the addition or reduction in the number of stages in accordance with gain requirements, etc. Our goal here will be the first "paper design".

Input and output stages have requirements (e.g. equalization, noise figure and impedance levels) that are peculiar to the terminal devices, and therefore we will deal with an intermediate amplifier. First one must decide whether it is to be a voltage, current or power amplifier. An inspection of the transfer characteristics of Fig. 3 shows that for low-level transistors, the collector current is more nearly proportional to the base current than to the base voltage. Thus

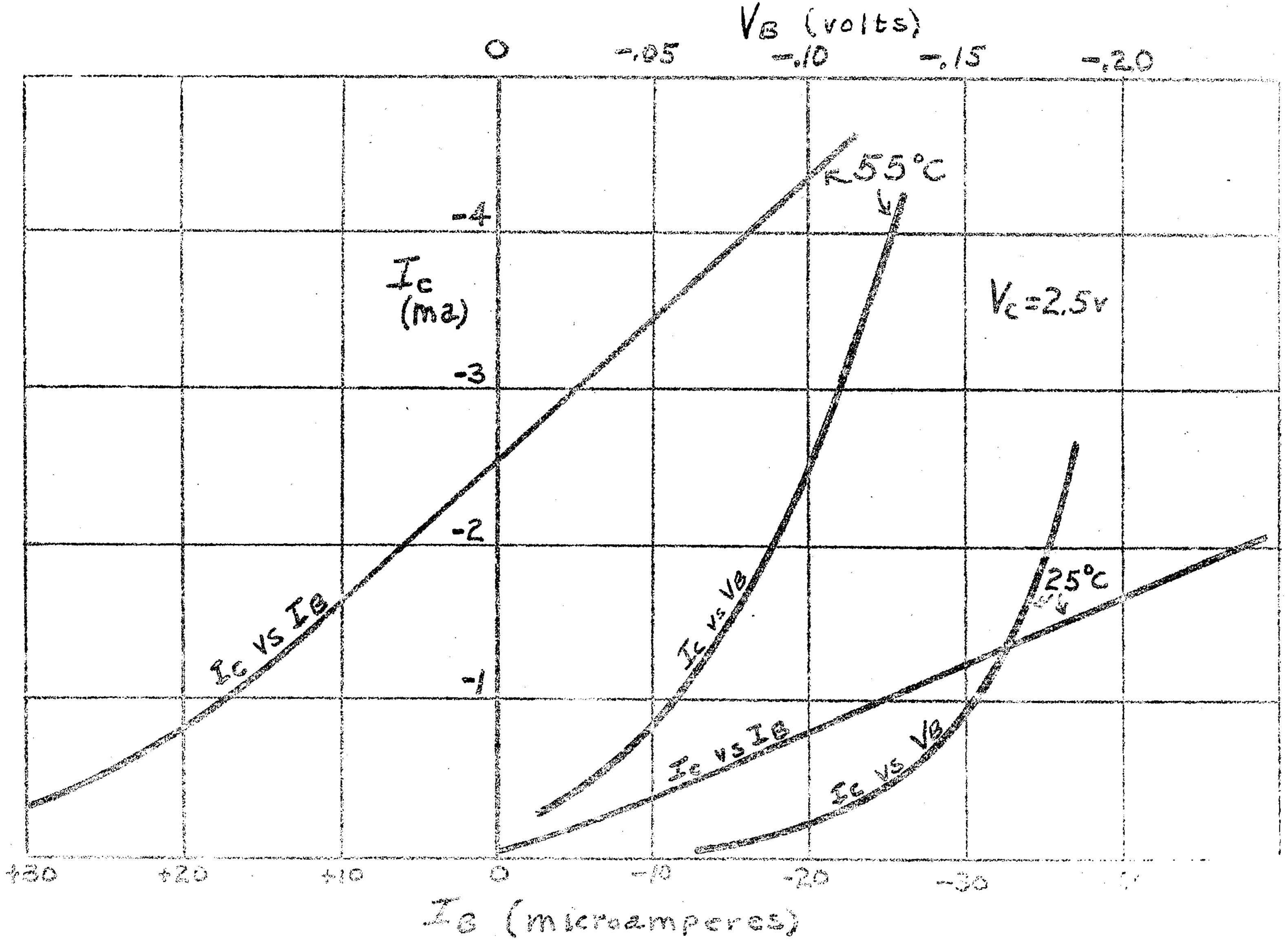

FIG. 3 COMMON-EMITTER TRANSFER CHARACTERI

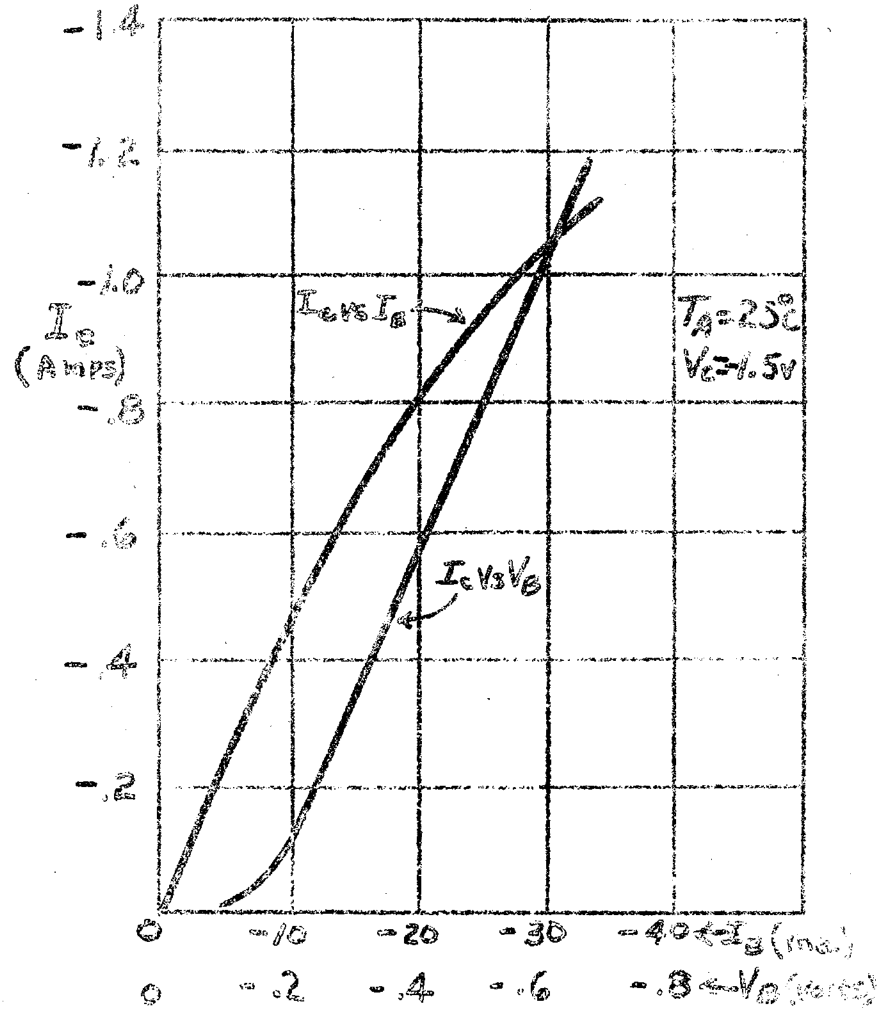

FIG. 4 POWER TRANSISTOR TRANSFER CHARACTERISTICS

we conclude that it is best to consider the low-level transistor as a current-operated device, and therefore the intermediate amplifier will be a current amplifier regardless of the system input and output. The effect of temperature on the transfer characteristics is also shown in Fig. 3. The transfer characteristics of an audio power transistor (Fig. 4) show that it can best be considered as a voltage-operated device and hence is usually driven by a low-impedance source.

## Gain

Stage gain is a clear-cut term in vacuum tube amplifier circuitry, but with transistors the effect of one stage on another makes the use of the term "stage gain" somewhat undesirable unless one specifies insertion gain, transducer gain, etc. For our purposes here we are not so much interested in the gain itself as in the factors that affect the gain. We will consider a "working circuit" (Figs. 1 and 2) in which the single stage is only a part. Here the load transistor is represented by resistor $r_L$ , and the preceding transistor stage is represented by a constant current generator with internal resistance of $r_g$.[1] In many cases, the source resistance, $R_g$, is important. We shall define the source resistance as the resistance looking back to the left (Figs. 1 and 2) from base to ground. In Fig. 1,

$$R_g = \frac{r_g R_B}{r_g + R_B} \qquad (1)$$

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[1] $r_g$ would normally be the same as the collector coupling resistor for the preceding stage if the output resistance of the preceding transistor is much greater than the coupling resistance.

nd in Fig. 2,

$$R_g = \frac{r_g R_B R_S}{r_g R_B + r_g R_S + R_B R_S} \qquad (2)$$

We will define the "working circuit gain" to be the ratio of $i_L$ to $i_g$. The output signal current, $i_L$, may be calculated as follows:

1. Only a fraction of $i_g$ gets to the base. We will call this fraction the <u>input circuit attenuation factor</u>, $A_{in}$. For Figs. 1 and 2,

   $$A_{in} = \frac{1}{1 + \frac{r_{in}}{R_g}} \qquad (3)$$

   where $r_{in}$ is the a-c input resistance of the transistor.

2. The signal current at the collector is $\beta$ times the base signal current.

3. The fraction of the collector signal current that appears in $r_L$ we shall call the <u>output circuit attenuation factor</u>, $A_{out}$. For Figs. 1 and 2,

   $$A_{out} = \frac{1}{1 + \frac{r_L}{R_C}} \qquad (4)$$

   if the output resistance of the transistor, $r_C/(\beta+1)$, is high compared to the parallel resistance of $R_C$ and $r_L$.

4. Thus, $i_L$ is given by

   $$i_L = A_{in}\,\beta\,A_{out}\,i_g \qquad (5)$$

   and dividing by $i_g$ gives the current gain as a product of two circuit attenuation factors and the current gain of the transistor.

   $$G = A_{in}\,\beta\,A_{out} \qquad (6)$$

For Figs. 1 and 2.

$$G = \left(\frac{1}{1+\frac{r_{in}}{R_g}}\right)(\beta)\left(\frac{1}{1+\frac{r_L}{R_c}}\right) \qquad (7)$$

For high gain, both attenuation factors should be high, that is, approach one. For the input circuit, this requires that $r_g$, $R_B$ and $R_S$ (if present) be high compared to the input resistance of the transistor. For the output circuit, $R_C$ should be high compared to $r_L$. There are other requirements on the circuit resistors ($R_S$, $R_B$ and $R_C$) so compromise is necessary, but it is well to keep in mind that any reduction in resistance of any of these resistors means a reduction in gain. The circuit gain is also affected by temperature changes since $r_{in}$ and β may both change with temperature.

## Distortion

In low-level RC amplifiers, there are two main sources of severe distortion. First, the input circuit is quite non-linear as shown in Figs. 5 and 6, and secondly, distortion may occur in the output circuit as a result of the operating point being on a very non-linear portion of the characteristic. The distortion caused by non-linear input impedance can be minimized by proper choice of operating point and by driving the transistor from a high-impedance source. Thus the same circuit requirements for high gain are also desirable to swamp out the effects of the non-linear input impedance. Selection of an operating point (See Fig. 6) where the input impedance does not vary se-

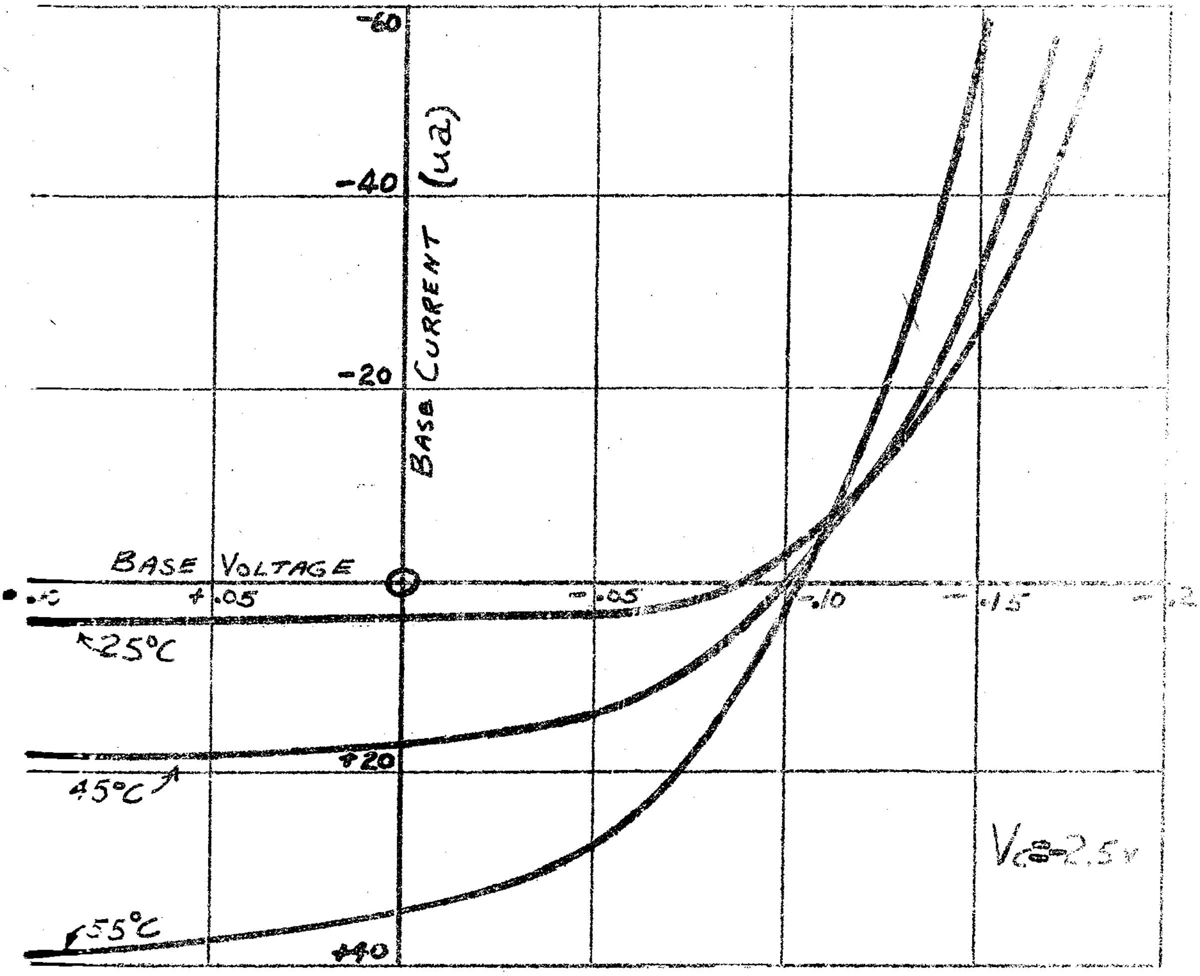

Fig. 5 COMMON-EMITTER INPUT CHARACTERISTICS.

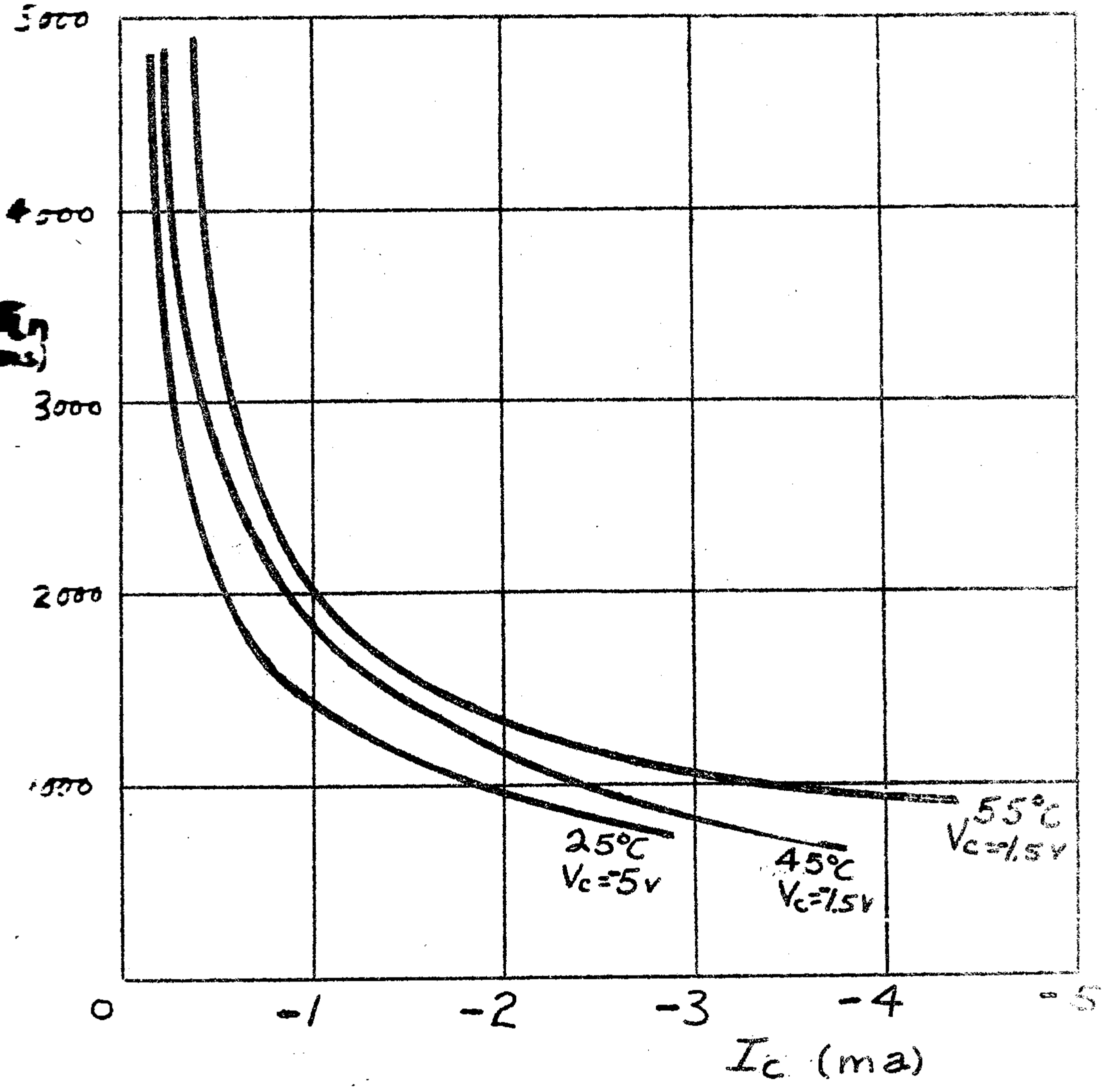

FIG.6 VARIATION OF INPUT RESISTANCE WITH COLLECTOR CURRENT. COMMON-EMITTER.

merely with collector current is desirable and quite necessary in case of a fairly low impedance source.

## Frequency Response

The high frequency performance of audio RC amplifiers is generally determined primarily by the frequency characteristics of the transistor. The input impedance, being capacitive, goes down at high frequencies, tending to increase the gain. The output impedance, also capacitive, goes down at high frequencies and tends to reduce the gain. However, the amplitude and phase shift associated with β at high frequencies is usually the most significant factor in determining the high frequency performance. Thus high frequency performance is obtained by the choice of a transistor with a high cutoff frequency. Transistor specifications generally list the $\alpha$ cutoff frequency, $f_\alpha$, whereas the cutoff frequency of importance in the common-emitter amplifier is $f_\beta$. For a first approximation, the transistor selected for an RC amplifier should have an $f_\alpha$ greater than $(\beta+1)$ times the highest frequency of interest as far as the amplifier is concerned. Thus for an amplifier to be flat to 15,000 cps, a transistor with a β of 49 ought to have an $f_\alpha$ of about 1 Mc.

The low frequency performance of an RC amplifier is determined primarily by the coupling and bypass capacitors. For the emitter bypass capacitor ($C_E$ of Fig. 2), the reactance at the lowest frequency interest is given by[2]

$$X_{C_E} = \left(r_e + \frac{R_g + r_b}{\beta + 1}\right)\sqrt{\frac{1}{P^2} - 1} \qquad (8)$$

where $r_e$, $r_b$ and $\beta$ are transistor parameters, $R_g$ is the source resistance (parallel combination of $R_B$, $R_S$ and $r_g$), and P is the low frequency attenuation factor due to incomplete bypassing of $R_E$. In terms of the input resistance

$$X_{C_E} = \frac{r_{in} + R_g}{\beta + 1}\sqrt{\frac{1}{P^2} - 1} \qquad (9)$$

if $(\beta+1)r_L \ll r_C$. Since $r_{in}$ and $\beta$ are functions of temperature and operating point (see Figs. 6 and 7), a conservative design requires the use of the highest value of $\beta$ and the lowest value of $r_{in}$.

For Fig. 2, it can be shown that the reactances of $C_1$ and $C_2$ are given by

$$X_{C1} = \left(r_g + \frac{r_{in} R_S R_B}{R_S R_B + r_{in} R_S + r_{in} R_B}\right)\sqrt{\frac{1}{P^2} - 1} \qquad (10)$$

and

$$X_{C2} = (R_C + r_L)\sqrt{\frac{1}{P^2} - 1} \qquad (11)$$

If there are N such low frequency calculations in a system design, then

$$P = (P_s)^{\frac{1}{N}} \qquad (12)$$

where $P_s$ is the low frequency attenuation factor for the system.

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[2] R. P. Murray, "Emitter Bypassing in Transistor Circuits", IRE Trans. PGA, May-June, 1957.

## The Temperature Problem

The resistance of the emitter-collector path of a transistor decreases with rising temperature, so that in a transistor circuit the collector current, and/or the collector voltage, may change considerably for only a few degrees' increase in temperature. The extent of this temperature effect depends largely on the temperature of operation, being rather insignificant at low temperatures but quite severe at high temperatures. For germanium transistors, small temperature changes become noticeable at room temperature and may have radical effects for temperatures as low as 40°C. The temperature effects do not become significant for silicon transistors until the temperature has increased appreciably above room temperature.

Some of the effects of temperature on the transistor characteristics can be seen from the output characteristic curves of Fig. 7. In addition to an upward shift of all curves to higher values of $I_C$ (reduced resistance of emitter-collector path), the spacing between curves (indication of $\beta$) increases appreciably as does their slope (indicative of the output resistance).

The effect of collector dissipation ($V_C I_C$) is to raise the junction temperature above the ambient temperature, and thus the curves bend upward for high values of dissipation.[3] In Fig. 7, the range of $I_C$ and $V_C$ is limited to small values of dissipation, and thus the bending effect is not very great. At 25°C the effect of temperature increase is not nearly as great as at 55°C, so the upward bend-

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[3] For low-level audio transistors with no heat sink, the junction temperature increases approximately 0.3°C per milliwatt of junction dissipation.

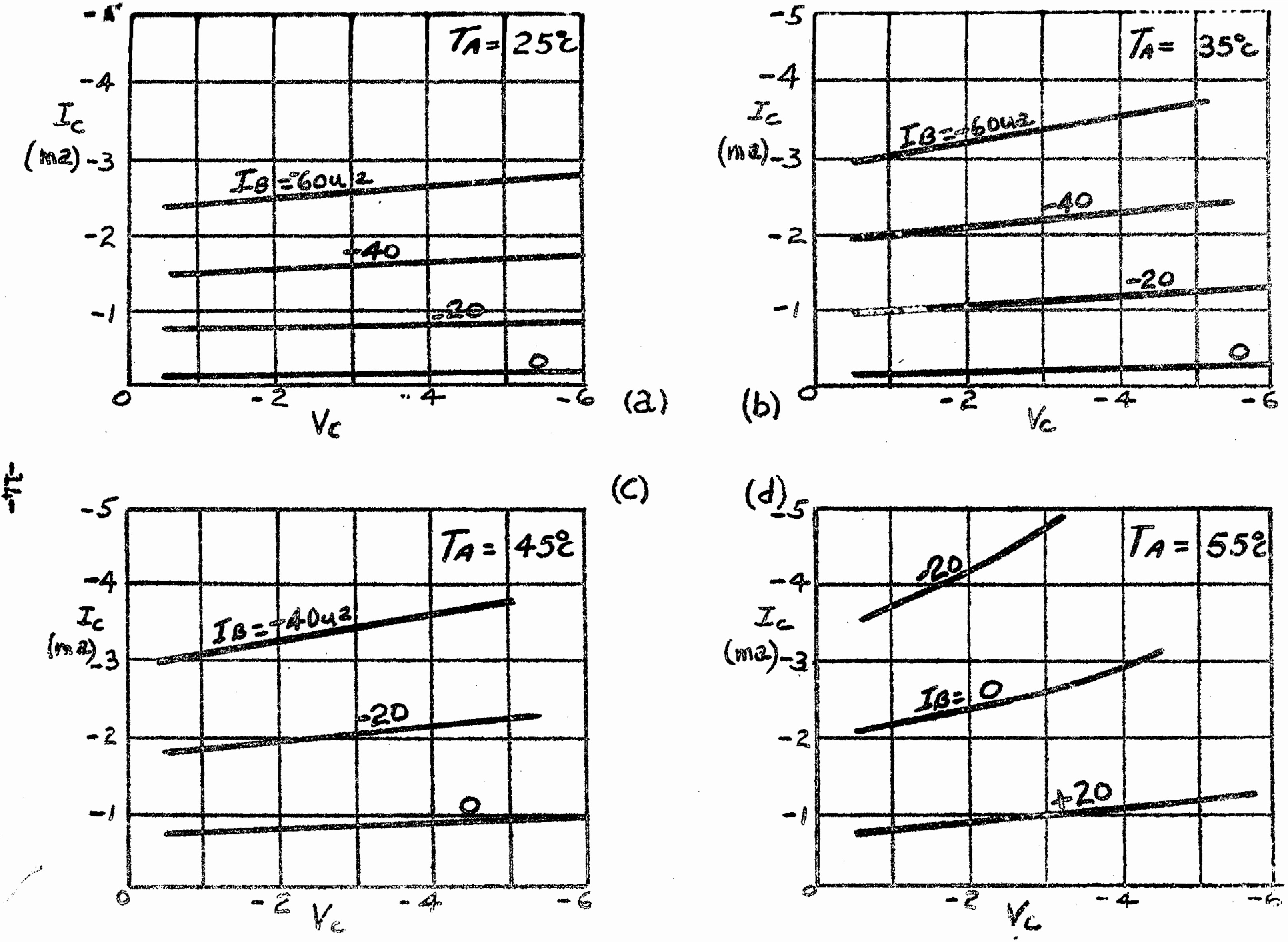

FIG. 7 EFFECT OF TEMPERATURE ON OUTPUT CHARACTERISTICS.

due to collector dissipation is more pronounced on the high temperature graph. This bending would not show on curves taken for constant junction temperature.

Although some of the reasons for the temperature effects are well known, no accurate relation is available for representing $I_C$ in terms of temperature. A relation which is satisfactory for one transistor type may not be acceptable for another. Graphs or equations which relate $I_{CO}$ to temperature are a help, but they fall short of telling the complete story.[4] It is contended here that the most satisfactory method of dealing with the temperature-dependent characteristics is to use two sets of characteristic curves, one for some low temperature (e.g. 25°C) and another set corresponding to the maximum temperature of operation. If in obtaining the two sets of graphical characteristics, we take into account, the extremes of $\beta$ and $I_{CBO}$ allowed for a certain type transistor, the problem of parameter variation due to production spread is handled along with the temperature problem.[4]

We have seen how temperature affects some of the transistor characteristics. Now let us see how the operating point of an RC amplifier is affected by the change in characteristics. Refer to the unstabilized amplifier circuit of Fig. 8 which uses a transistor with characteristic curves of Figs. 9 and 10.

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[4] $I_{CO}$ is the current that flows in the collector circuit when the emitter is opened, and $I_{CBO}$ is the collector current with the base opened.

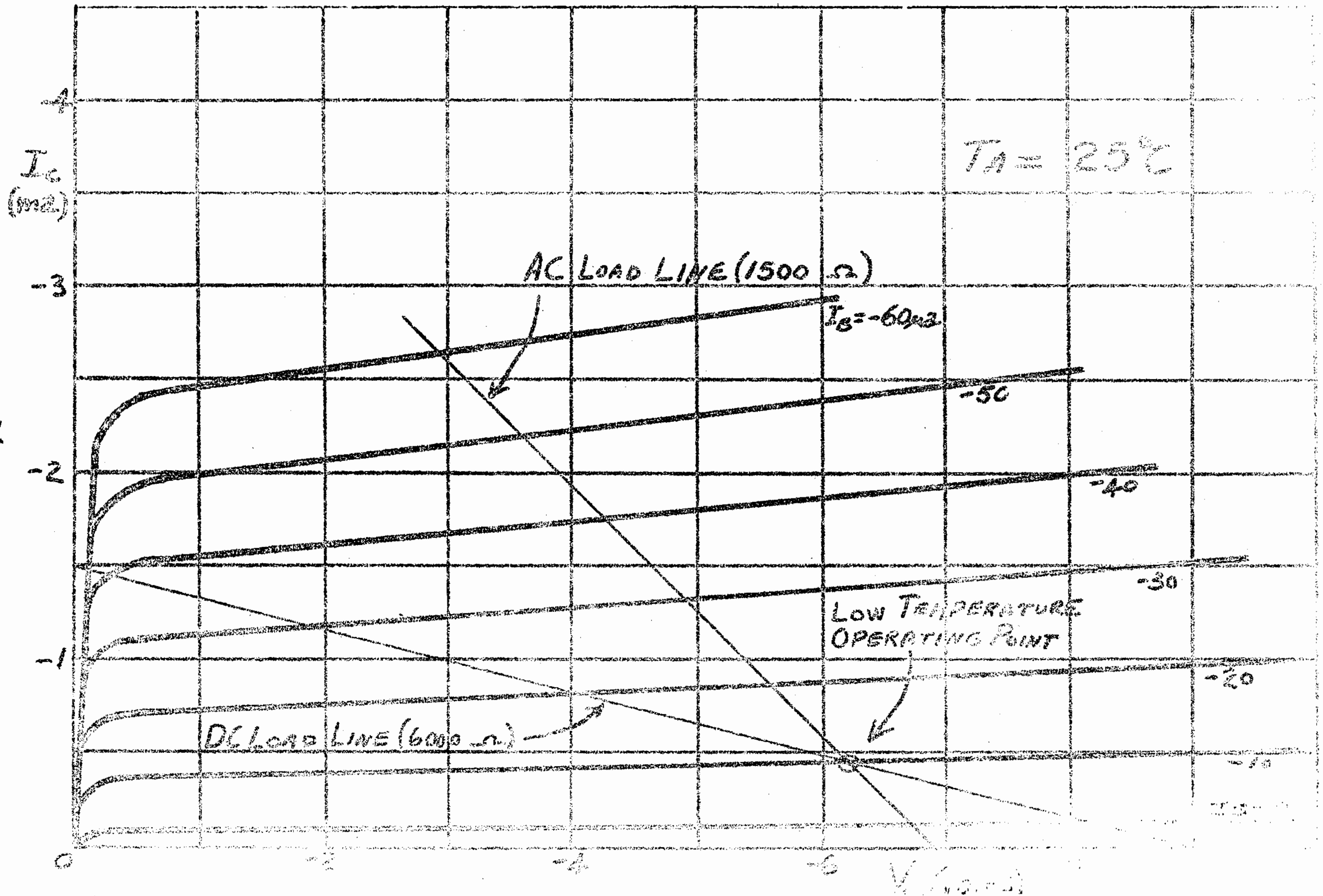

Fig. 9 Output characteristic for 25°C.

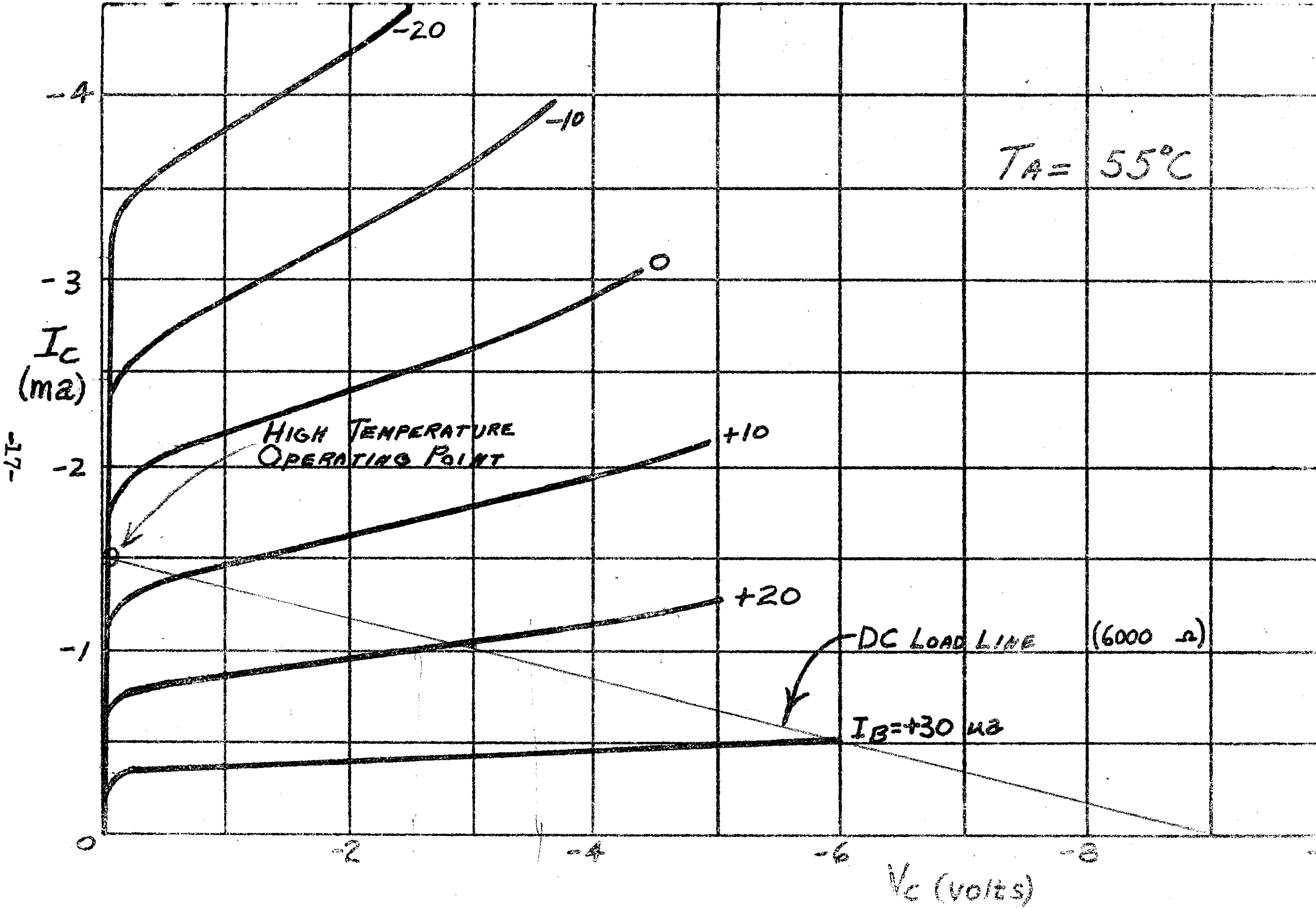

Fig. 10 Output characteristic for 55°C.

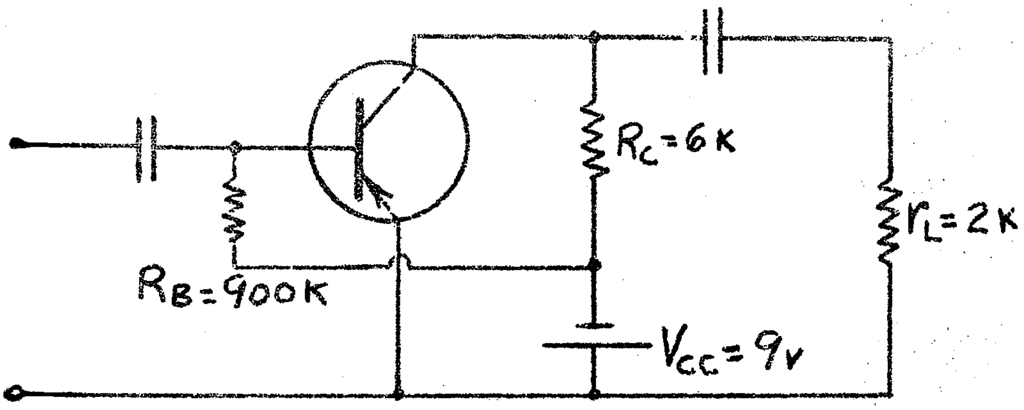

Fig. 8 Unstabilized RC amplifier in which the effect of temperature has not been taken into account.

Since the base bias current ($I_B \cong V_{CC}/R_B = -10$ microamperes) is not affected by temperature, the operating point for any temperature will be at the intersection of the d-c load line and the transistor curve corresponding to a base bias of -10 microamperes. Note the low temperature operating point in Fig. 9. As the temperature rises, the $I_B = -10$ microamperes bias curve rises, and thus the operating point shifts up and to the left along the d-c load line. As shown in Fig. 10, the high temperature operating point is in such a position that small changes in base current produce no change in collector current. The amplifier is inoperative (gain equals zero) at 55°C.

## The Unstabilized RC Amplifier

From the standpoint of battery drain and collector dissipation, it is desirable to select an operating point corresponding to the smallest permissible values of collector voltage and current. A higher level operating point may be chosen for many reasons, some of which are: linearity, gain, noise figure, impedance levels and temperature stabilization. Here we will choose the lowest level operating point that will permit proper operation over the specified temperature range. Thus the signal to be considered in the design will correspond to the maximum signal that the amplifier is expected to handle.

The most significant specifications for the intermediate amplifier are signal level and input and output impedance levels. Let us assume the maximum values of the signal voltage and current in $r_L$ to be $V_{sig}$ and $I_{sig}$. Now if $r_L$ is small compared to $R_C$, the variations in collector voltage and current will also be $V_{sig}$ and $I_{sig}$. This condition on $r_L$ and $R_C$ can generally be met if the temperature variation is not severe and if a high value of $V_{CC}$ is available; however, it will not be met for wide temperature variations, and a modification of the first design will be required or else the amplifier will not handle the specified signal level.

In the unstabilized RC amplifier, the operating point is allowed to shift (with temperature) along the d-c load line. See Fig. 11. The factors that limit the extent of operating-point shift are the minimum allowable collector voltage at the left end of the d-c load line (high temperature graph) and the minimum allowable collector current at the right end (low temperature graph).

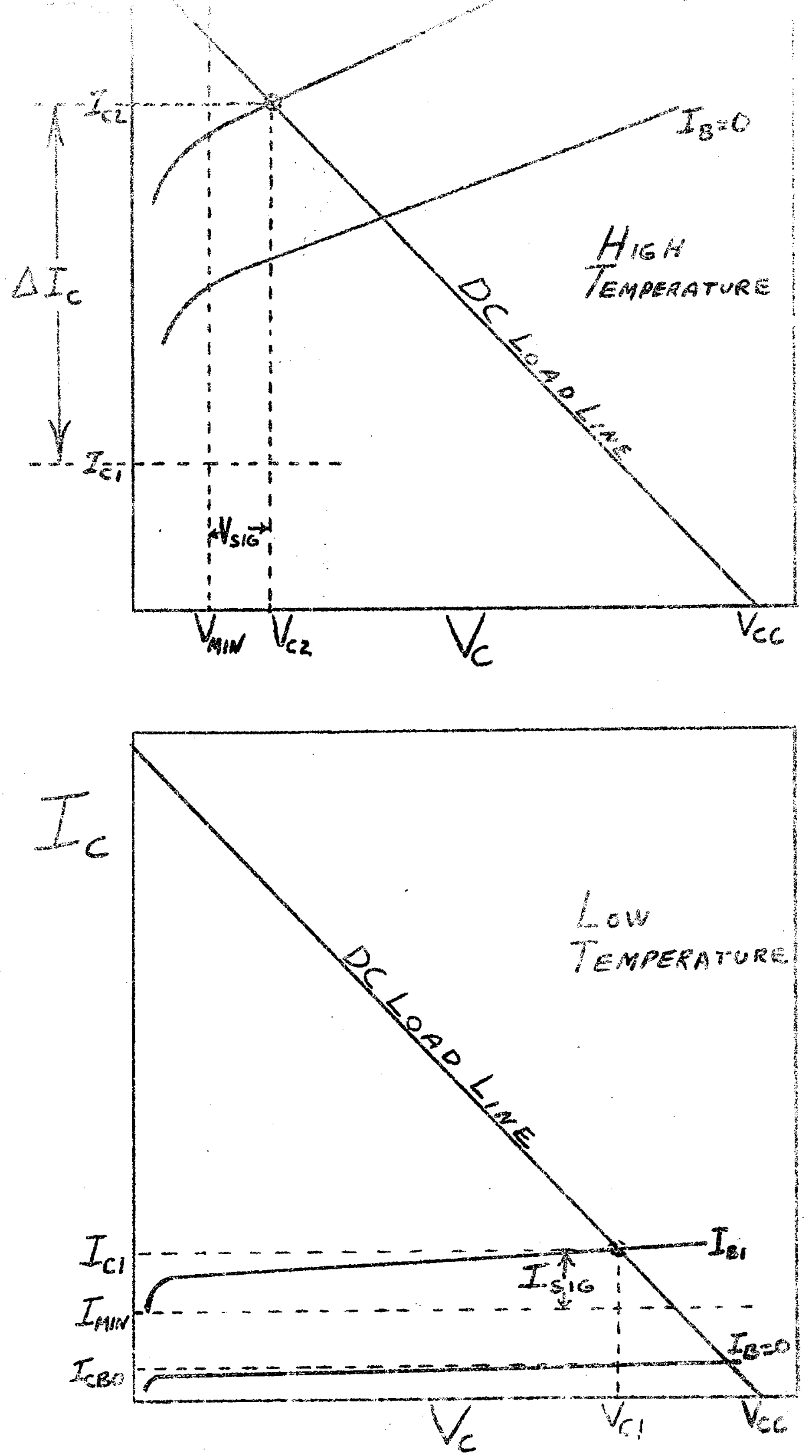

FIG. 11 SHIFT OF OPERATING POINT IN A PROPERLY DESIGNED UNSTABILIZED RC AMPLIFIER.

The smallest allowable operating point collector current will occur at the low temperature.

$$I_{C1} = I_{min} + I_{sig} \tag{13}$$

where $I_{C1}$ is the low temperature operating point collector current and $I_{min}$ is the lowest permitted value of $I_C$. To choose $I_{min}$ requires an inspection of the low temperature output, transfer and input characteristic curves. From the output curves (Fig. 11) we see that $I_{min}$ must be at least equal to $I_{CBO}$ if we are to allow for operation below 25°C. From the transfer curve (Fig. 3) we see that non-linearity at low values of $I_C$ is not great and frequently may be neglected. However, the non-linearity of the input characteristics (Figs. 5 and 6) must not be overlooked as severe distortion may result from this non-linearity. As mentioned previously, the effects of the input circuit non-linearity may be swamped out by driving the amplifier from a high-impedance source, i.e., high impedance with respect to $r_{in}$. The lower limit on the collector current can be determined only by considering the input characteristics along with the source resistance, $R_g$, and the amount of allowable distortion. If the source resistance is comparatively low, it may be necessary to employ an unbypassed resistor in series with either the emitter or the base to swamp out the non-linearity of the emitter junction.

Once $I_{min}$ and $I_{C1}$ have been determined, we go to the low temperature output characteristic curves and determine the value of the base bias current required to give $I_{C1}$. Since $I_C$ is not completely independent of $V_C$, the choice of $I_{B1}$ is not exact, but if chosen near $V_C = 2/3\, V_{CC}$, the approximation will be quite accurate at least for transistors with high collector resistance.

Making use of the fact that the voltage between emitter and base is very small compared to $V_{CC}$, the base bias resistor may be calculated as

$$R_B = \frac{V_{CC}}{I_{B1}} \tag{14}$$

Now for the high temperature operating point. The smallest value of the operating point collector voltage is

$$V_{C2} = V_{min} + V_{sig} \tag{15}$$

where $V_{min}$ is determined by the non-linearity of the high temperature output characteristics at low values of collector voltage. See Fig. 11. The high temperature operating point is on the curve $I_B = I_{B2} = I_{B1}$ at $V_{C2}$. The value of $I_{C2}$ can now be read from the graph and the d-c load resistance (equal to $R_C$) calculated as

$$R_C = R_{dc} = \frac{V_{CC} - V_{C2}}{I_{C2}} \tag{16}$$

and the d-c load line drawn in on the low and high temperature graphs.

The actual value of signal voltage and current in $r_L$ will be less than the original value assumed for $V_{sig}$ and $I_{sig}$ by a factor that depends on the relative values of $R_C$ and $r_L$. A redesign based on the first can be used to increase the signal level that can be handled. The increase in signal-handling capacity will be at the expense of reduced gain and higher power consumption.

## Illustrative Example No. 1

Purpose: 1. To illustrate the design of an unstabilized RC amplifier that can be made to perform properly over a specified temperature range.

2. To observe the effects of signal level and input nonlinearity on distortion.

Given: Transistor with characteristics of Figs. 3, 5, 6, 12 and 13; circuit diagram of Fig. 1; maximum temperature of operation of 45°C; and

$r_L$ = 1000 ohms $\qquad V_{sig}$ = 0.5 volt

$r_g$ = 5000 ohms $\qquad I_{sig}$ = 0.5 milliampere

$V_{CC}$ = -9 volts

Solution: 1. From Fig. 12, we see that $I_{CBO}$ = -0.14 milliampere near $V_C$ = -6 volts, but the 25°C curve of Fig. 6 shows that the input resistance rises sharply for small values of $I_C$. If $I_C$ is not allowed to go below about one milliampere, the variation in $r_{in}$ is not great, and the input circuit distortion would be held to a low value. However, here we will deliberately choose $I_{min}$ a little lower than dictated by fidelity so as to illustrate the effect of input circuit distortion. Choose $I_{min}$ = -0.5 milliampere and then,

$$I_{C1} = I_{min} + I_{sig} = -0.5 - 0.5 = -1 \text{ milliampere}$$

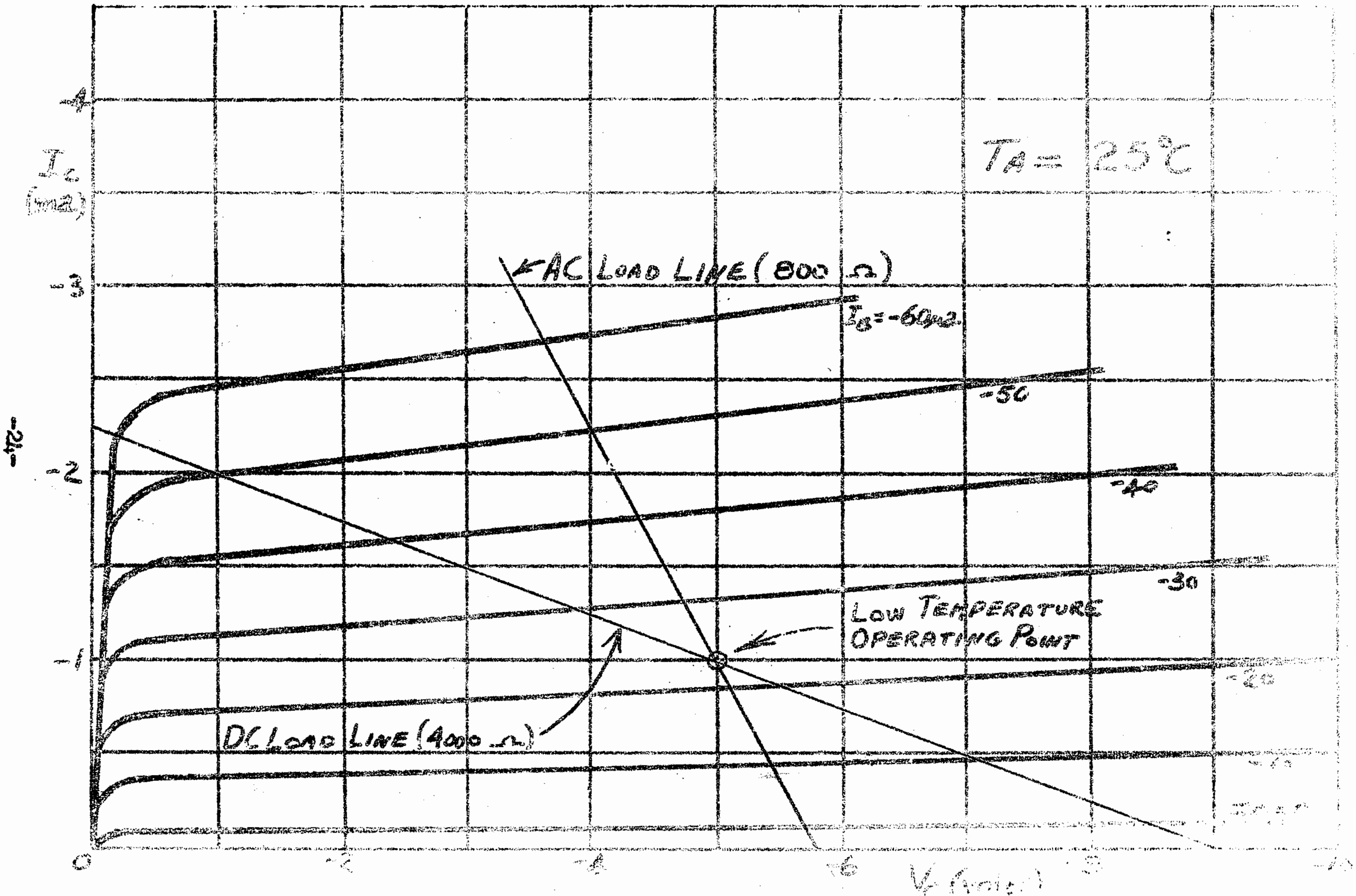

Fig. 12 Output characteristic for [illegible]

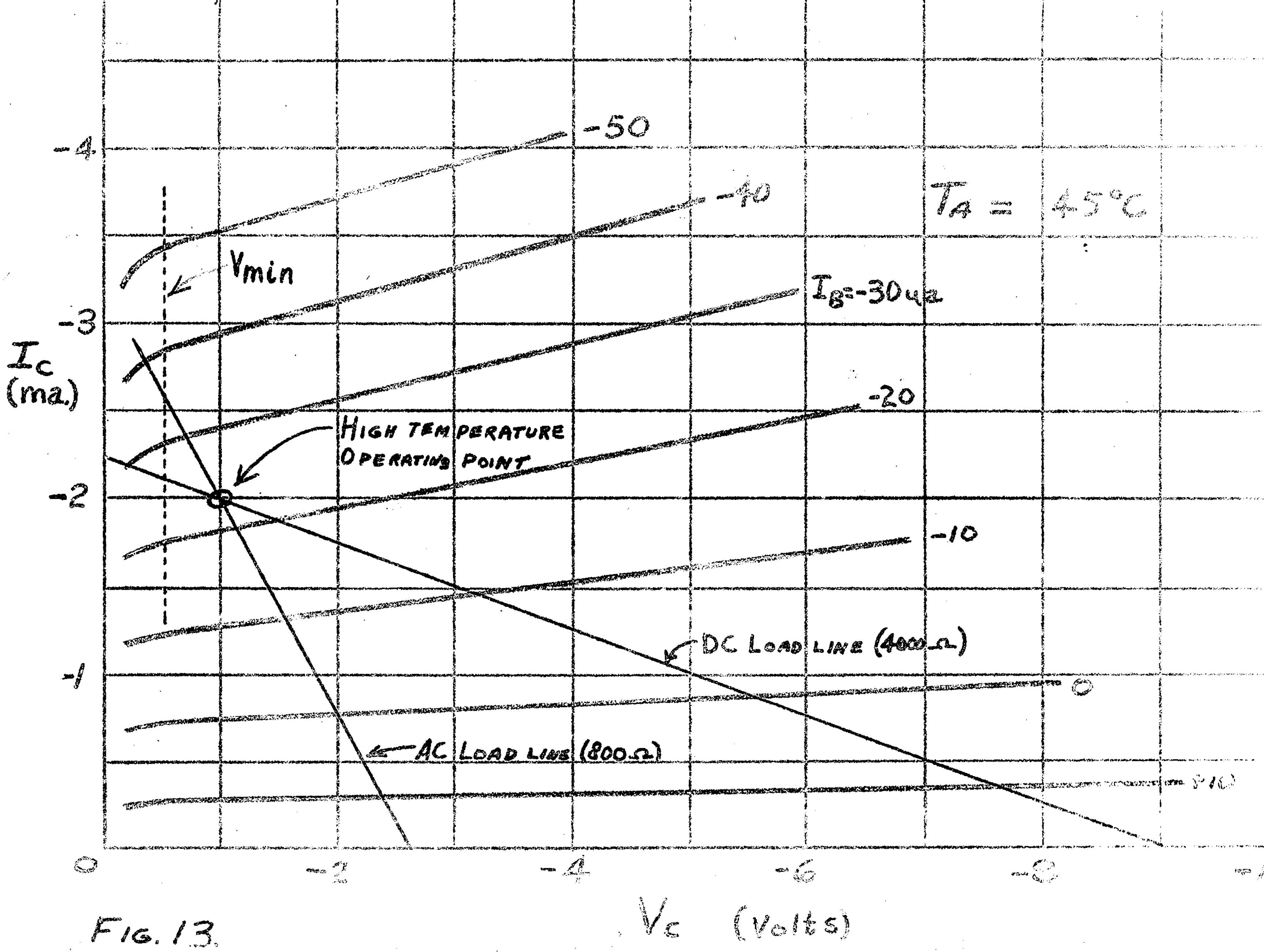
-4
-3
-2
-1
0
$I_c$ (ma.)
-2
-4
-6
-8
$V_c$ (Volts)
-50
-40
$I_B$=-30 μa
-20
-10
0
$T_A$ = 45°C
Vmin
HIGH TEMPERATURE OPERATING POINT
DC LOAD LINE (4000 Ω)
AC LOAD LINE (800 Ω)

FIG. 13.

2. From Fig. 12, we see that a base current of about -23 microamperes is required for an $I_C$ of one milliampere.

Thus

$$R_B = \frac{V_{CC}}{I_B} = \frac{-9}{-23 \times 10^{-6}} = 391{,}000 \text{ ohms}$$

3. From Fig. 13, we observe $V_{min}$ to be about -0.5 volt, and thus

$$V_{C2} = V_{min} + V_{sig} = -0.5 - 0.5 = -1 \text{ volt}$$

At the point, $V_{C2} = -1$ volt and $I_B = -23$ microamperes, $I_{C2} = -2$ milliamperes and therefore,

$$R_C = \frac{V_{CC} - V_{C2}}{I_{C2}} = \frac{-9 + 1}{-0.002} = 4{,}000 \text{ ohms}$$

4. The d-c and a-c load lines may then be drawn on Figs. 12 and 13 and the low and high temperature operating points indicated.

The amplifier described in the foregoing example was constructed and tested with the following results:

| $T_A = 25°C$ | $T_A = 45°C$ |
|---|---|
| $I_{C1} = -1$ milliampere | $I_{C2} = -2$ milliamperes |
| $V_{C1} = -5$ volts | $V_{C2} = -1$ volt |
| $I_{B1} = -23$ microamperes | $I_{B2} = -23$ microamperes |
| $I_g = 18$ microamperes (max) | $I_g = 15.6$ microamperes (max) |
| $I_L = 500$ microamperes (max) | $I_L = 500$ microamperes (max) |
| $G = 28$ | $G = 32$ |
| Distortion = 5.8% | Distortion = less than 1% |
| Battery drain = 1 milliampere | Battery drain = 2 milliamperes |

Note the 5.8% distortion at 25°C as compared to less than 1% at 45°C. The reason for this is obvious when one inspects the input characteristic of Fig. 6. The input resistance for $T = 45^{\circ}C$ and $I_C = -2$ milliamperes is more nearly constant than for $T = 25^{\circ}C$ and $I_C = -1$ milliampere. To demonstrate this input circuit non-linearity, the generator resistance, $r_g$, was increased to 500,000 ohms and the distortion at 25°C decreased to 3.6%.

Recall that the design was based on the approximation that the collector signal current and voltage were the same as the signal current and voltage in the load, $r_L$. It can be determined from the low temperature graph (Fig. 12) that the maximum output signals to be handled according to the stated value of $I_{min}$ are: $V_{sig} \angle 0.4$ volt and $I_{sig} \angle 0.4$ milliampere. Accordingly, the distortion at 25°C was measured with $V_{sig} = 0.3$ volt and found to be 1.6% with $r_g =$ 500,000 ohms and 3.4% with $r_g = 5000$ ohms. The remaining 1.6% distortion may be attributed to the non-linearity of the transfer and output characteristics.

It is possible to achieve satisfactory performance from an unstabilized amplifier circuit. However, if the gain, distortion or battery drain are of primary importance, then the signal level, temperature variation, and production spread of parameters must not be great.

## The Stabilized RC Amplifier

In the unstabilized RC amplifier, no attempt is made to check

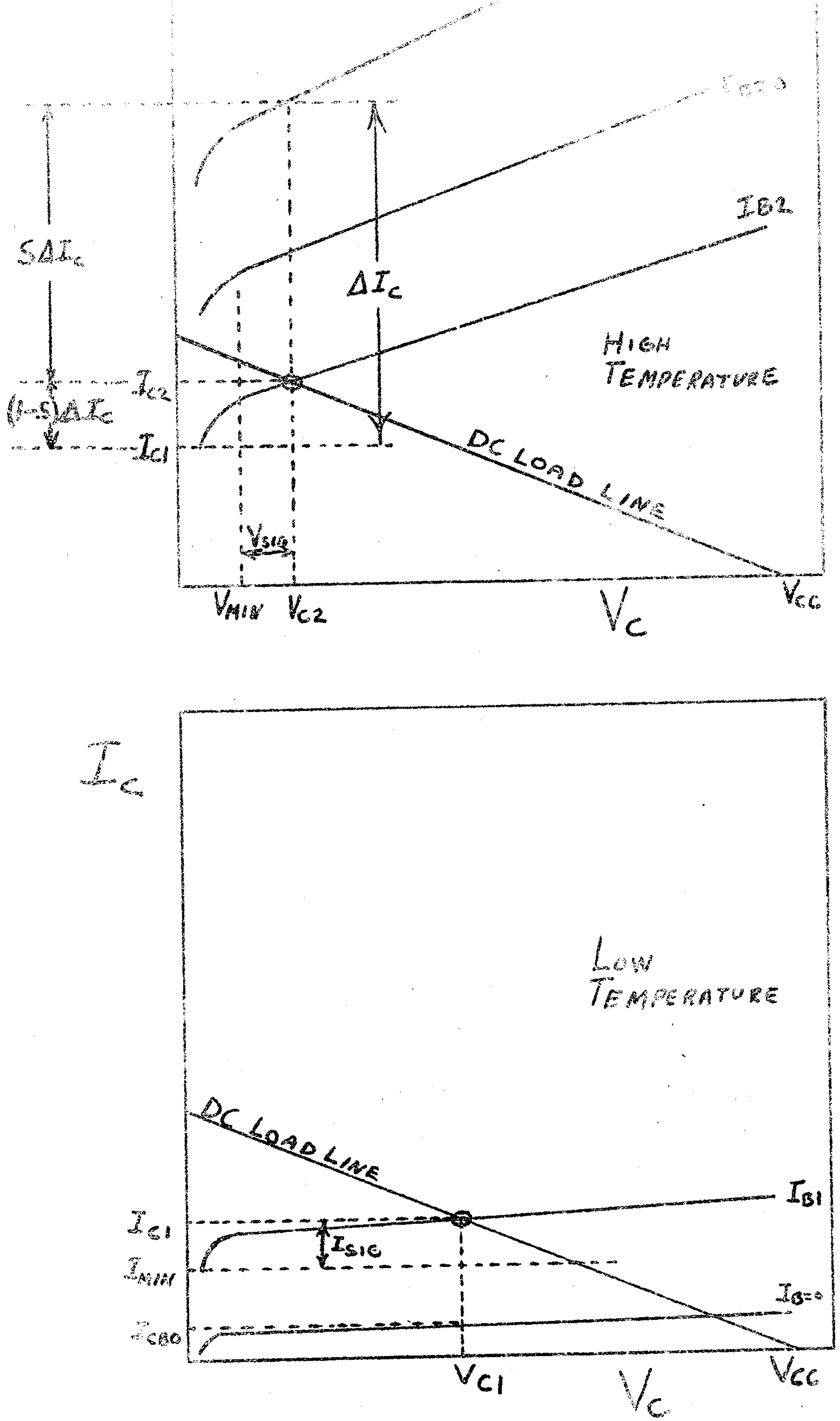

FIG.14 SHIFT OF OPERATING POINT IN A STABILIZED RC AMPLIFIER.

the increasing collector current as the temperature rises. The circuit is merely designed to take the variation in operating point into account at the expense of reduced gain (due to a low value of $R_C$) and high battery drain at the high temperature. The change in $I_C$ in the unstabilized amplifier is designated as $\Delta I_C$ in Fig. 11.

In the stabilized circuit, the collector current is allowed to rise only a fraction of $\Delta I_C$ . In Fig. 14, $\Delta I_C$ has been divided into two parts: $S\Delta I_C$ and $(1-S)\Delta I_C$, where S is a circuit stabilization factor.[5] Zero stabilization corresponds to the unstabilized amplifier, and unity or 100% stabilization indicates that $I_C$ would not change as the temperature rises to its maximum value. From the standpoints of gain and battery drain there may be an optimum value of S for a particular circuit.

The most commonly employed stabilization circuitry is shown in Figs. 2 and 15. Here d-c negative feedback from the emitter resistor causes the base bias current to be dependent on the collector current. Note that the effect of $I_C$ in $R_E$ is to produce a "source" voltage in the base circuit which subtracts from $V_{CC}$. Thus as $I_C$ increases with temperature, the voltage across $R_E$ increases and reduces the base bias current. The bias current may even reverse direction at high temperatures.

It can be shown that the stabilization factor for the circuit of Fig. 15 is given by [5]

$$S = \frac{\beta}{\beta + 1 + \left[R_E\left(\frac{1}{R_S} + \frac{1}{R_B}\right)\right]^{-1}} \qquad (16)$$

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[5] R. P. Murray, "Systematic Design of Transistor Bias Circuits", *Electronic Industries and Tele-Tech*, November 1957.

So if it is desired that the variation in $I_C$ be held to a small value, a high degree of stabilization requires a large value of $R_E$ and small values of $R_B$ and $R_S$. But both of these requirements have some adverse effects: small values for $R_B$ and $R_S$ reduce the gain, increase the battery drain and increase the distortion caused by the non-linearity of the input circuit. A high value of $R_E$ means that for a given $V_{CC}$, $R_C$ will have to be reduced accordingly since

$$R_{dc} = R_C + R_E \tag{17}$$

if we make the approximation that $I_E \cong I_C$. A reduced value for $R_C$ means less gain.

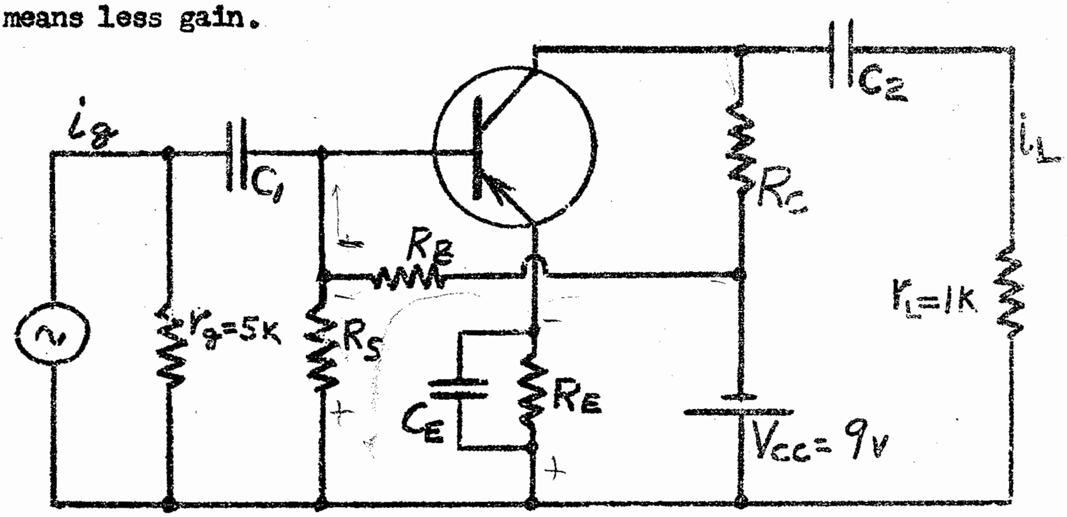


Fig. 15 Stabilized Common-emitter amplifier.

Fig. 16 shows the computed battery drain as a function of S for the circuit of Fig. 15. Note that the battery drain at the low temperature is essentially equal to $I_{C1}$ ($I_{C1}$ does not change with S) until the effect of the bleeder current in $R_S$ becomes significant with respect to $I_{C1}$. On the other hand, the high temperature battery drain

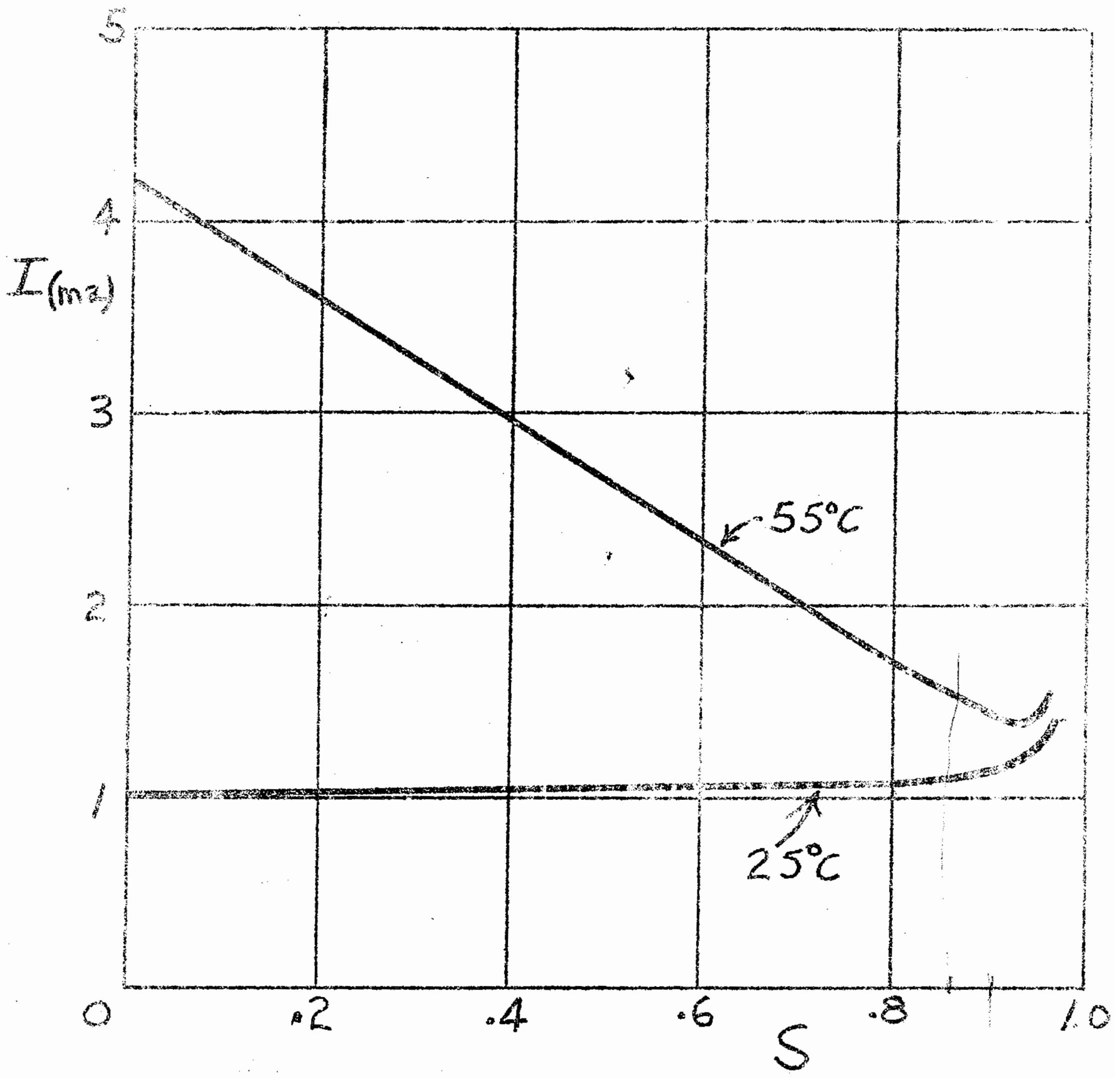


FIG. EFFECT OF "S" ON BATTERY DRAIN.
APPLIES TO EXAMPLE No. 2.

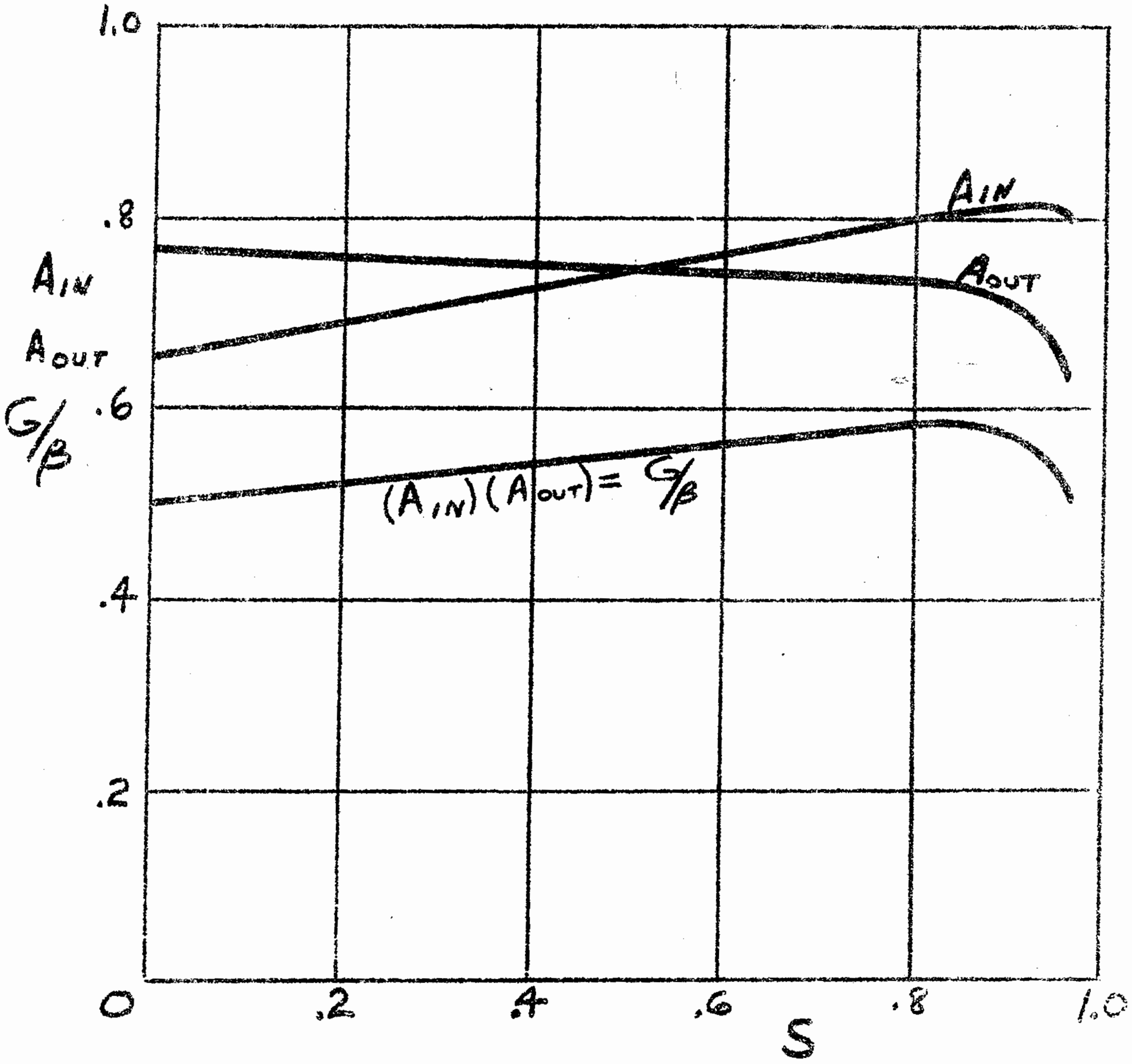


FIG. 17 EFFECT OF "S" ON GAIN. APPLIES TO EXAMPLE NO. 2.

is greatly affected by S since low values of S indicate that $I_{C2}$ will be high. A minimum in high temperature battery drain occurs when the decreasing collector current (due to increasing S) is offset by an increasing bleeder current in $R_S$. Thus from the standpoint of battery drain, S should be chosen slightly less than that value at which the minimum in high temperature battery drain occurs.

From the standpoint of circuit gain, the optimum value for S occurs when the increase in gain due to the use of a high $R_C$ and low $R_E$ is offset by the decrease in gain due to the shunting effect of $R_B$ and $R_S$. Fig. 17 shows the two attenuation factors and the total gain (in terms of $\beta$) for the amplifier of Fig. 15.

Another factor to be taken into account in the choice of S is the collector dissipation. Low values of S mean high dissipation at high temperatures.

We now turn to the circuit design. Choose $V_{C2}$ and $I_{C1}$ in accordance with the nature of the transistor characteristics and signal level as before. Next, in accordance with the foregoing discussion, we select a stabilization factor or operating-point current on the high temperature graph. Then

$$R_{dc} = \frac{V_{CC} - V_{C2}}{I_{C2}} \qquad (18)$$

We must now decide how much of $R_{dc}$ to allot to $R_E$. If $R_E$ is chosen low, then the output attenuation factor will be high due to a high $R_C$, but both $R_S$ and $R_B$ will be low and reduce the input attenuation factor. The proper value for $R_E$ can be found by maximizing the circuit gain with respect to $R_E$. Using the approximation that $V_B \cong 0$, the equations for the emitter-base circuit (Fig. 15) are

$$V_{CC} = I_S(R_B + R_S) + I_B R_B \qquad V_{CC} = (I_S + I_B) R_B + I_S R_B \tag{19}$$

$$0 = I_C R_E + I_B R_E - I_S R_S \tag{20}$$

Eliminating $I_S$ from Eqs. 19 and 20 and solving for $R_B$ gives

$$R_B = \frac{V_{CC} - (I_B + I_C)R_E}{\frac{R_E}{R_S}(I_B + I_C) + I_B} \tag{21}$$

Making two equations from Eq. 21 for the two conditions of temperature, and eliminating $R_B$ we get

$$R_S = \frac{V_{CC}R_E\left[(I_{C2} - I_{C1}) - (I_{B1} - I_{B2})\right]}{V_{CC}(I_{B1} - I_{B2}) - R_E(I_{B1}I_{C2} - I_{B2}I_{C1})} \tag{22}$$

and in terms of S,

$$R_S = \frac{V_{CC}R_E\left[(1 - S)\Delta I_C - \Delta I_B\right]}{V_{CC}\,\Delta I_B - R_E(I_{B1}I_{C2} - I_{B2}I_{C1})} \tag{23}$$

where $(1 - S)\Delta I_C$ and $\Delta I_B$ are the changes in $I_C$ and $I_B$ as the temperature goes from low to high (see Fig. 14). Solving for $R_B$

$$R_B = \frac{V_{CC}\left[(1 - S)\Delta I_C - \Delta I_B\right]}{I_{B1}I_{C2} - I_{B2}I_{C1}} \tag{24}$$

The gain of the circuit of Fig. 15 is given by Eq. 7. Substituting Eqs. 17, 23, and 24 into Eq. 7 gives the gain as

$$G = \frac{\beta}{\left(1 + \frac{r_L}{R_{dc} - R_E}\right)\left[1 + r_{in}\left(\frac{1}{r_g} + \frac{\Delta I_B}{R_E\left[(1-S)\Delta I_C - \Delta I_B\right]}\right)\right]} \tag{25}$$

Now assuming $R_{dc}$ to be constant (determined by choice of S), taking the derivative of the gain (Eq. 25) with respect to $R_E$ and equating to zero we get

$$R_E = \frac{R_{dc} + r_L - \sqrt{(R_{dc} + r_L)\big[r_L + \left(\frac{(1-S)\Delta I_c}{\Delta I_B} - 1\right)\left(\frac{1}{r_{in}} + \frac{1}{r_g}\right)(R_{dc}\, r_L)}}{1 - r_L\left(\frac{(1-S)\Delta I_c}{\Delta I_B} - 1\right)\left(\frac{1}{r_{in}} + \frac{1}{r_g}\right)} \qquad (26)$$

We now have relations for the four circuit resistors and three capacitors as follows:

$R_E$ from Eq. 26
$R_C$ from Eq. 17
$R_S$ from Eq. 23
$R_B$ from Eq. 24

$C_E$ from Eq. 9
$C_1$ from Eq. 10
$C_2$ from Eq. 11

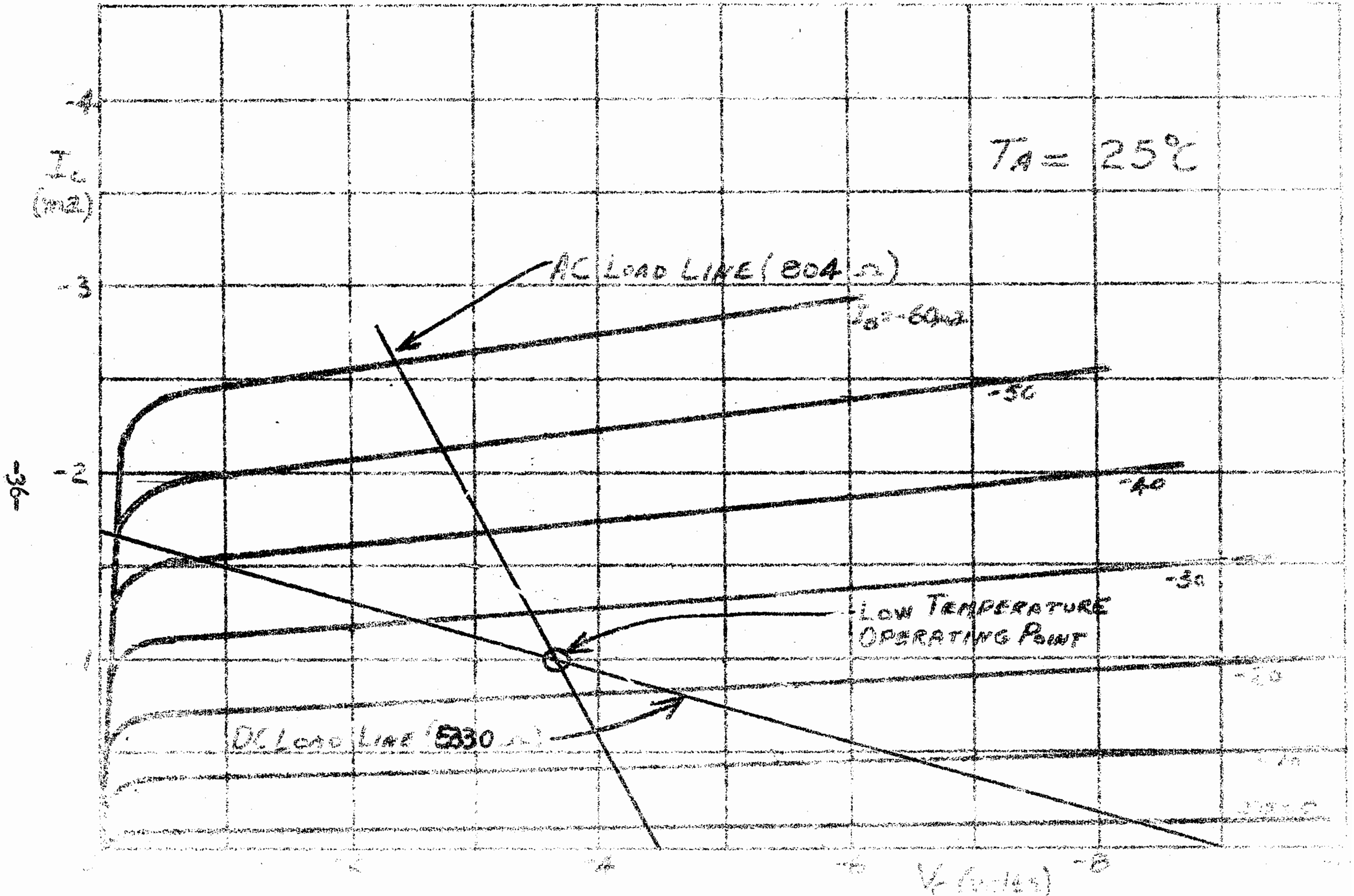
The actual resistances of the circuit resistors will not be these calculated values, but the nearest available values. Some of these resistance values are more critical than others. For example it would be better to choose $R_E$ nearest to its calculated value and take up the slack in $R_C$.

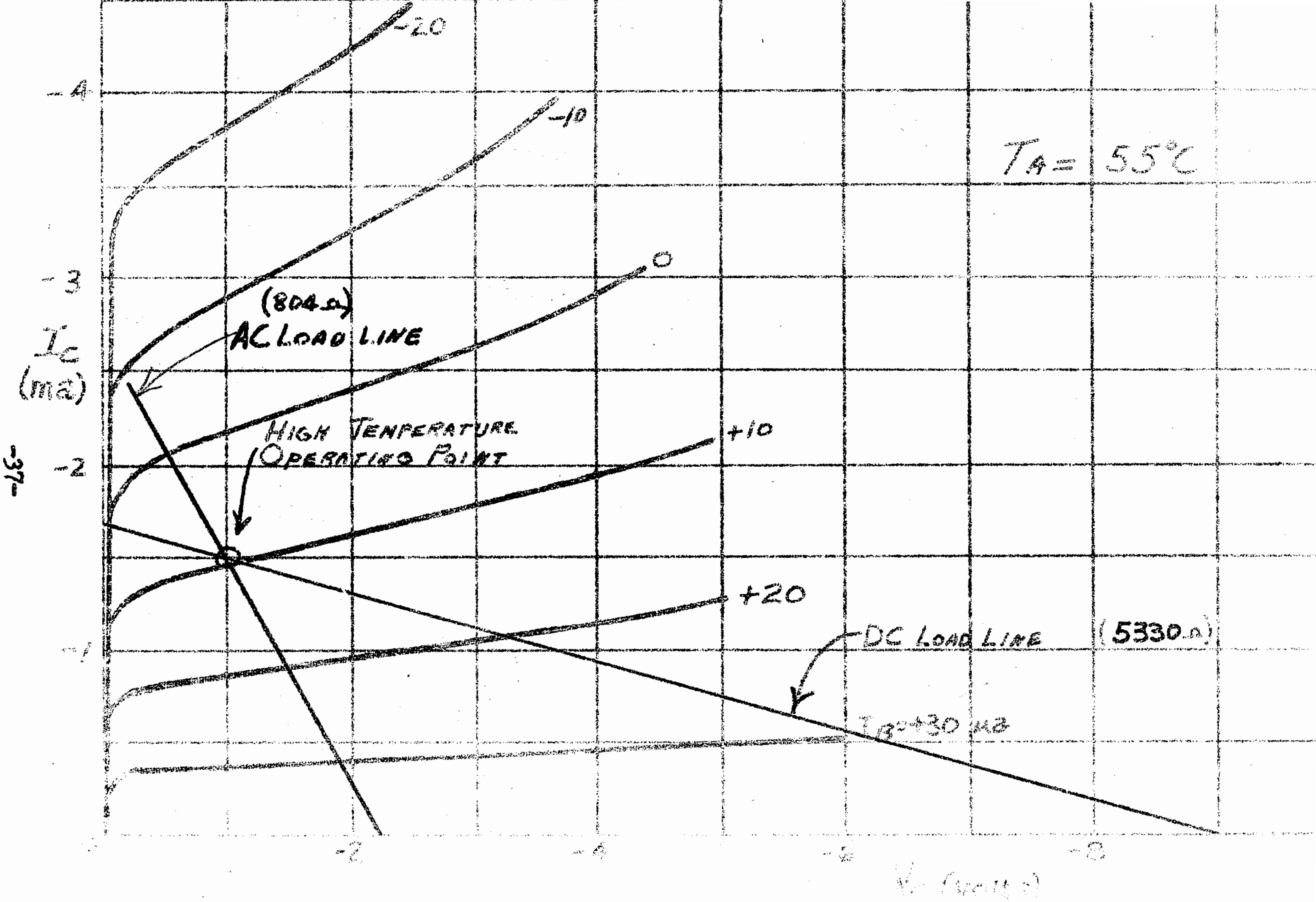
## Illustrative Example No. 2

Given: Circuit of Fig. 15, using transistor described by curves of Figs. 3, 5, 6, 18, and 19. Maximum temperature of 55°C and 30 cps for the low frequency at which the response is down 3 db.

$r_L$ = 1000 ohms
$r_g$ = 5000 ohms
$V_{CC}$ = − 9 volts

$V_{sig}$ = 0.5 volt
$I_{sig}$ = 0.5 milliampere


$T_A = 25°C$
$I_c$ (ma)
-4
-3
-2
-1
AC LOAD LINE (804 Ω)
$I_B$ = -60μa
-50
-40
-30
-20
-10
LOW TEMPERATURE OPERATING POINT
DC LOAD LINE (5330 Ω)
-2
-4
-6
-8
$V_c$ (volts)


-20
-10
0
+10
+20
$I_B$=+30 μa
$T_A$ = 55°C
-4
-3
-2
-1
$I_C$
(ma)
(804Ω)
AC LOAD LINE
HIGH TEMPERATURE
OPERATING POINT
DC LOAD LINE
(5330Ω)
-2
-4
-6
-8
$V_C$ (volts)

Solution: 1. As in parts 1, 2, and 3 of Example 1, we have

$$I_{C1} = -1 \text{ milliampere}$$

$$I_{B1} = -23 \text{ microamperes}$$

$$V_{C2} = -1 \text{ volt}$$

2. From Figs. 16 and 17, choose $S = 0.843$ or $I_{C2} = -1.5$ milliamperes.

$$R_{dc} = \frac{V_{CC} - V_{C2}}{I_{C2}} = \frac{-9 + 1}{-0.0015} = 5330 \text{ ohms}$$

3. From Eqs. 26, 17, 23 and 24 and using $r_{in} = 1500$ ohms, we have

$$R_E = 1245 \text{ ohms}$$

$$R_C = 4085 \text{ ohms}$$

$$R_S = 22{,}000 \text{ ohms}$$

$$R_B = 95{,}600 \text{ ohms}$$

4. If the gain is to be 3 db down at 30 cps, this corresponds to $P_s = 0.707$. If there are a total of three low frequency calculations in the entire circuit, we have from Eq. 12, $P = 0.89$. From Eqs. 9, 10 and 11 we get

$$X_{CE} = 38.1 \text{ ohms} \quad (\text{From Fig. 19, } \beta = 71)$$

$$X_{C1} = 3270 \text{ ohms}$$

$$X_{C2} = 2600 \text{ ohms}$$

and at 30 cps, these reactances correspond to capacitances of

$$C_E = 139 \text{ uf}$$

$$C_1 = 1.62 \text{ uf}$$

$$C_2 = 2.04 \text{ uf}$$

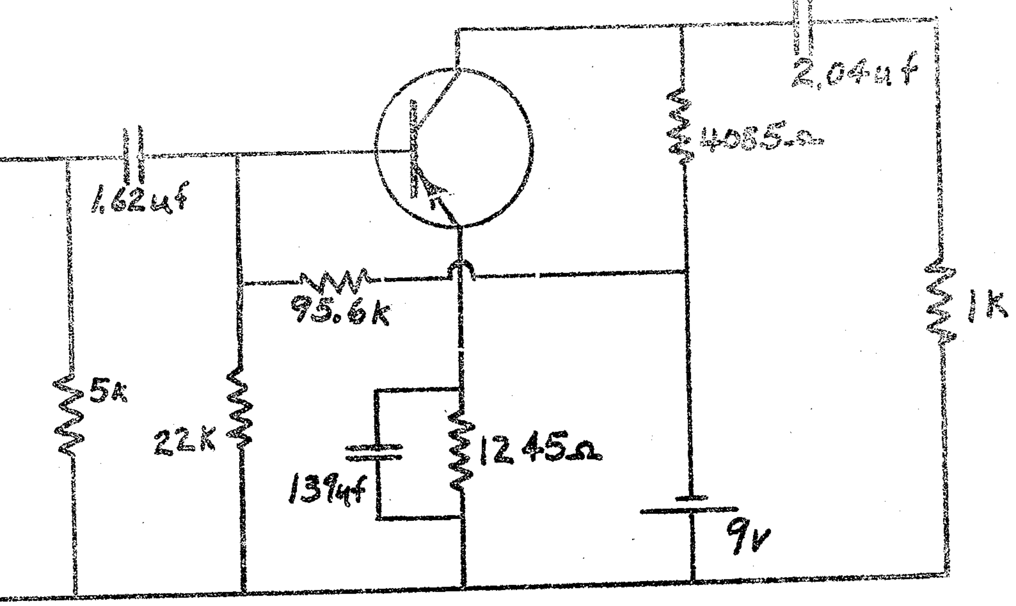


Fig. 20 Stabilized amplifier of example 2.

The amplifier described in the foregoing example was constructed and tested with the following results:

| $T_A = 25^\circ C$ | $T_A = 55^\circ C$ |
|---|---|
| $I_{C1}$ = -0.93 milliampere | $I_{C2}$ = - 1.44 milliamperes |
| $V_{C1}$ = - 4 volts | $V_{C2}$ = - 1.2 volts |
| $I_{B1}$ = -22 microamperes | $I_{B2}$ = + 9 microamperes |
| $I_g$ = 13 microamperes (max) | $I_g$ = 9.38 microamperes (max) |
| $I_L$ = 300 microamperes (max) | $I_L$ = 300 microamperes (max) |
| G = 23 | G = 32 |
| Distortion = 3.4% | Distortion = 1.86% |
| Battery drain = 1 milliampere | Battery drain = 1.5 milliamperes |

As in example 1, the distortion at 25°C may be reduced by choosing a slightly higher value for $I_{C1}$ , and the distortion at both temperatures may be reduced by increasing the resistance in the base-emitter circuit. The signal handling capability may be increased by a redesign in which $V_{C2}$ and $I_{C1}$ are increased.

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