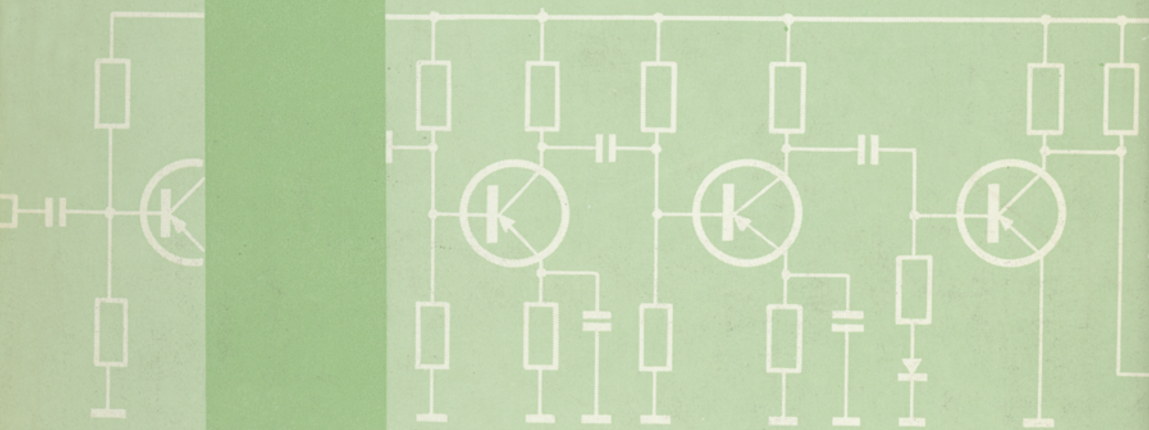
H. E. Kaden

TRANSISTORS APPLIED



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H. E. KADEN

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PREFACE

The transistor is being used nowadays as a technical amplifying element in almost every field of technology. In consequence, the valve is more and more superseded by the transistor. To the same extent as the transistor is replacing the valve, those who up to now have only worked with valves must now also become familiar with transistors. Experience has shown that the amplifying element "valve" is theoretically relatively easily mastered by students, apprentices, amateur technicians and so on. Why is it then that those very circles who successfully apply the theoretical knowledge concerning the operation of the valve in the circuit show an aversion to assimilating the theoretical fundamentals of the transistor? Is the operation of a transistor as an amplifying element really so different in principle compared with the valve? Is it necessary to replace the simple formulae of the valve by obscure four-pole equations in the case of the transistor? The answer to that is an emphatic "No!". The transistor is just as easy to understand as an amplifying element in the circuit as the valve. Granted, the internal physics of the transistor present more difficulty than with the valve. This, however, is by no means simple in the valve and what "expert" in the use of the valve has hitherto bothered himself about the internal more profound physical processes of the valve? It is important for the user of valves and transistors to know what this type of amplifier can do. Here the greatest interest lies in the characteristic values and curves. This does not mean that one should know nothing at all about the internal physical processes in a transistor. It ought to be said, however, that over-emphasis on the physical processes in the transistor has "soured" its theoretical presentation to a large extent.

This text-book contains a short essential treatise on the physical facts necessary to explain certain specific transistor properties. Its chief aim, however, is the presentation of the principles of the transistor as an amplifying element in combination with the essential circuit elements. Here we are not so much concerned with reproducing the greatest possible number of circuits built up as with arousing a basic knowledge of which suitable chosen circuits to understand and eventually to develop oneself. Pure algebra is the only mathematical science required to study this book. In order that the less experienced may be able to concentrate on transistor problems, the most important properties of the transistor compared with those of the valve have been set out in the form of an introduction. This summary may serve the more skilled reader as a quick survey. The detailed explanations mainly cover the so-called small signal amplification at low and high frequencies. The derivations are included in such a way that even those who dislike the formulae will be able to follow them without difficulty. Numerical examples given after each problem should consolidate the material and give the feeling for magnitude and dimensions.

The presentation of the transistor has been carried out in exactly the same way as that which has already become classical for the valve. It is therefore to be hoped that as a result of the uniformity of the presentation a better and clearer connection will be shown between the two amplifying elements and with it a basis for easier comprehension of the transistor.

January 1965 THE AUTHOR

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CHAPTER 1

INTRODUCTION

1.1. The transistor versus the tube

To those who are already familiar with the electron tube and wish to become acquainted with the transistor, it is doubtless intriguing to investigate the differences and similarities existing between the two amplifying devices. However, even impartial readers can frequently profit by having both devices described in uniform terms as is done in the following chapters. By way of introduction the most important facts concerning the transistor will be set out together with those of the electron tube in so far as they affect the amplification process in a circuit. The whys and wherefores will be explained in detail later. In this way, a general viewpoint is developed before we penetrate into individual problems and will not appear to be wasted effort.

1.2. The fundamental circuit of the transistor and the electron tube

The comparison between the transistor and the electron tube is first shown in the following two fundamental circuits.

1.3. Why does the transistor amplify?

The following explanation should be sufficient without having to study the physics of semiconductors.

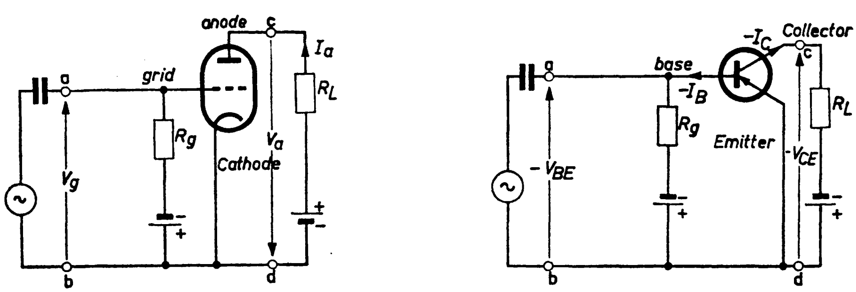


Fig. 1.1

The value of the base-emitter voltage determines how many charge carriers are drawn from the emitter into the adjoining base region, although most of these charge carriers are captured by the adjacent collector which is also attracting them. Only a small part of the charge carriers attracted by the controlling base voltage will be transferred as control current to the more distant base electrode. The transistor is therefore in essence a "current amplifier", i.e. the control of the collector current by the base-emitter voltage of varying intensity is the result of the favourable current division between base and collector. This constitutes the actual transistor effect.

If an a.c. control voltage is connected at terminals *a* and *b* via the capacitor *C*, the d.c. voltage between base and emitter is then altered by the same amount and in the same rhythm. The outcome of this is that the base draws more or less charge carriers from the emitter. The greater part of this direct current fluctuating in rhythm with the control voltage reaches the collector and at the load resistance R_L in the collector circuit, it generates an a.c. voltage which exceeds the control voltage by the amplification factor A_v, being amplified between *c* and *d*. As has already been mentioned, the controlling voltage applied between *a* and *b* also allows an alternating current, small in comparison to the emitter current, to flow from the emitter to the base. This a.c. control current is very small compared with the controlled current flowing in the collector circuit and we therefore speak of "current amplification" A_i.

The control generator, however, also allows an additional alternating current to flow via resistance R_g which is connected in series with the d.c. base voltage source but lies parallel to the transistor input. Viewed from the control source, the current amplification is apparently reduced by this. In the transistor, the base electrode and the collector electrode are both attracting, as compared with the emitter, and therefore both have the same d.c. voltage polarity (with the *p–n–p* transisitor both are negative and so attract positive charge carriers).

1.4. Why does the electron tube amplify?

In the electron tube, the electrode furthest from the cathode, i.e. the anode, draws electrons which are released by the cathode. The control grid is placed between these two electrodes in the immediate vicinity of the cathode. This control grid attains a negative voltage compared with the cathode by way of the grid resistance R_g. This reduces the attraction of the anode; the current drawn by the anode diminishes with increasing negative voltage at the grid, and becomes greater in the reverse case when the negative voltage at the grid is reduced. Since no negative electrons are passed on to the negative grid, control in the tube takes place essentially as a voltage control With external control by an a.c. voltage, a current can still flow via the necessary grid resistance R_g, or with high frequencies via the grid capacitance. This "control current", however, is not originally connected here with the control process in the electron tube.

If a controlling a.c. voltage is connected across terminals a and b via capacitor C, the d.c. voltage between grid and cathode is altered by the same amount and in the same rhythm. The result is that the attraction of the anode is more or less weakened by the grid; the current to the anode, the controlled current, fluctuates with the rhythm of the controlling voltage. This fluctuating anode current produces an a.c. voltage at the load resistance R_L, which exceeds the control voltage according to the value of the load resistance. The amplified voltage which is larger than the controlling voltage by the voltage amplification factor A_v, appears between the output terminals c and d. The control generator, however, also allows a current to flow via the grid resistance R_g which serves as a d.c. grid voltage supply. If we compare this a.c. control current at input a or b of the tube with the controlled a.c. output current in the anode circuit of the electron tube, we can also show a current amplification A_i for the electron tube.

Since in the electron tube the anode attracts charge carriers from the cathode, although the grid weakens the anode attraction, grid and anode have opposite d.c. voltage polarity (anode positive, grid negative).

1.5. How strongly does an electron tube amplify?

Let us start with the results of the following considerations: The mutual conductance (slope) S is the characteristic value of a tube which gives information concerning the controlling action of the grid on the anode current. The internal resistance R_i is a similar characteristic value which tells us about the controlling effect of the anode voltage on the anode current. With the load resistance available, the anode voltage is not constant during control. By means of its own controlling effect, it acts against the controlling effect of the grid and apparently reduces the slope of the grid. We thus have in the tube a so-called reaction of the anode on the grid. The two values, slope and internal resistance, together with the load resistance, determine the amount of voltage amplification.

Now in detail:

If we want to know the extent of the amplifying action of an electron tube, we have to establish by measurement the amount of pull the anode exerts on the anode current and how much this attraction is reduced by the grid. The values measured here are best plotted in a graph and from this we obtain the characteristic curves of this electron tube. In this way we get the so-called two-quadrant characteristic field. This consists of two diagrams with a common axis, the anode current axis I_a. We refer to the right-hand one, the anode current — anode voltage diagram, as the first quadrant. The second quadrant is the anode current — grid current diagram.

The characteristic curves plotted in the first quadrant inform us about the attracting or controlling action of the *anode voltage*. In actual measurement the grid voltage is kept constant so that only the controlling effect of the anode is expressed. If the applied grid voltage V_g amounts to -3 V, an anode current $I_a = 4$ mA will flow at an anode voltage $V_a = 80$ V. If the anode voltage is reduced to 60 V a current of only $I_a = 2$ mA flows. This controlling action of the anode voltage can be expressed in a characteristic value for this tube. The variations (Δ called "delta") were:

$$\Delta V_a = 80 - 60 = 20 \text{ V},$$
$$\Delta I_a = 4 - 2 = 2 \text{ mA}.$$

From this we obtain as a characteristic value of the *internal resistance*:

$$R_i = \frac{\Delta V_a}{\Delta I_a} = \frac{20 \text{ V}}{2 \text{ mA}} = 10 \text{ k}\Omega.$$

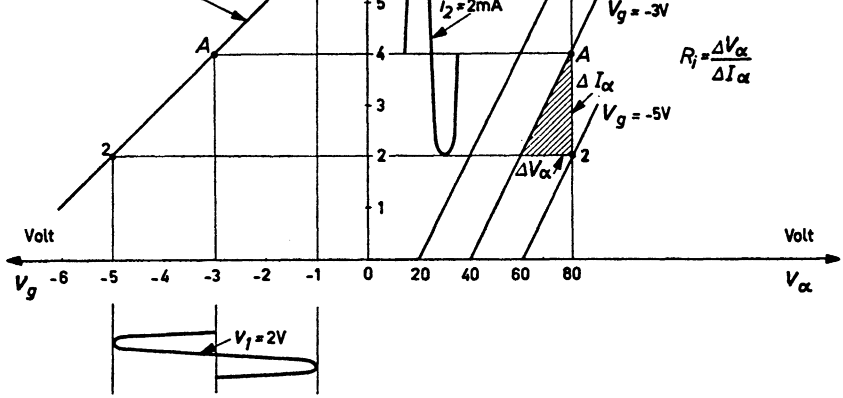


Fig. 1.2

The counter attraction, the controlling action of the grid, is expressed in the second quadrant. Each characteristic curve in the anode current-grid-voltage diagram is shown here for *constant anode voltage* in order to show the grid control action only. In Fig. 1.2, for the sake of clarity, the only characteristic plotted is for $V_a = 80$ V. We can draw this curve by transferring points 1, A_1 and 2 from the first quadrant to the second. Point 1, for example, applies for a grid voltage $V_g = -1$ V. It must therefore lie perpendicular to -1 V seen from the $-V$ axis, and with the current $I_a =$ 6 mA must give the new point of intersection in the second quadrant. If a constant voltage $V_g = -3$ V is applied at the grid, we obtain a current $I_a = 4$ mA at the "operating point" A_2. If an a.c. voltage $v_1 = 2$ V is now connected at the grid, as is shown in Fig. 1.2, the anode current then fluctuates by 2 mA, i.e. by $i_2 = 2$ mA.

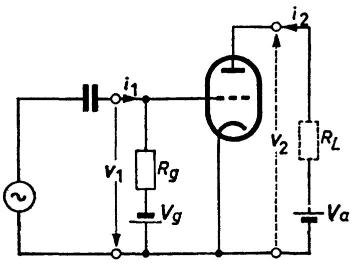


Fig. 1.3

This control action of the grid voltage on the anode current is also indicated as a characteristic value for this electron tube. This value which expresses the amount of controlling action of the grid is called the *slope* and is obtained from:

$$S = \frac{\Delta I_a}{\Delta V_g} = \frac{i_2}{v_1} = \frac{2\text{ mA}}{2\text{V}} = 1\text{ mA/V}.$$

If the grid resistance is $R_g = 1$ MΩ, an "a.c. input current" flows with the value:

$$i_1 = \frac{v_1}{R_g} = \frac{2\text{ V}}{1\text{ M}\Omega} = 2\ \mu\text{A}.$$

The *current amplification* therefore amounts to:

$$A_i = \frac{i_2}{i_1} = \frac{2\text{ mA}}{2\ \mu\text{A}} = 1000.$$

If there is also an anode resistance or load resistance R_L in the anode circuit, we also obtain voltage amplification. Because of the load resistance the anode voltage is no longer equal to the supply voltage; it is now reduced by the voltage loss at the load resistance. The characteristic field of the electron tube is shown below with a load resistance. The load resistance curve shows the anode voltage which is still present with a particular anode current. In Fig. 1.4, $R_L = 20$ kΩ so that the supply voltage V_S is 160 V.

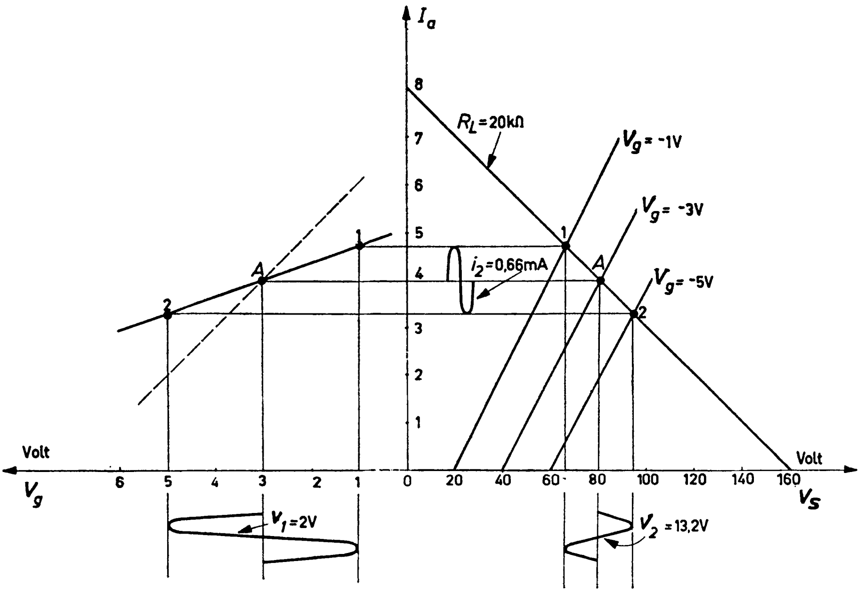


Fig. 1.4

It is now obvious in the first quadrant that with $V_g = -3$ V grid bias, we again have operating point A_1 because the load resistance line cuts the characteristic line for $V_g = -3$ V here. At this point, we therefore have an anode voltage of $V_a = 80$ V; the voltage drop at the load resistance amounts to $I_a \cdot R_L = 5$ mA $\cdot$ 20 kΩ, and $V_a = V_S - IR = 160 - 80 = 80$ V. If the grid voltage is now -1 V or -5 V, the anode voltage and anode current are adjusted according to points 1 and 2. By transferring these points to the second quadrant, we obtain the dynamic characteristic with the operational slope:

$$S_d = S\frac{R_i}{R_i + R_L} = 1 \cdot \frac{10}{10 + 20} = 0.33 \text{ mA/V}.$$

If we now apply an a.c. input current $v_1 = 2$ V, we obtain an a.c. output current:

$$i_2 = v_1 \cdot S_d = 2 \text{ V} \cdot 0.33 \text{ mA/V} = 0.66 \text{ mA}.$$

This a.c. output current i_2 produces an amplified a.c. voltage v_2 at the load resistance; according to Ohm's Law, this is derived from:

$$v_2 = i_2 \cdot R_L = v_1 \cdot S_d \cdot R_L = 0.66 \cdot 20 = 13.2 \text{ V}.$$

The voltage amplification is:

$$A_v = \frac{v_2}{v_1} = S_d \cdot R_L = S\frac{R_i \cdot R_L}{R_i + R_L} = 0.33 \text{ mA/V} \cdot 20 \text{ k}\Omega = 6.6.$$

1.6. How much does a transistor amplify?

Let us first state the result of the following considerations:

The slope S is also a characteristic value in the transistor and gives information concerning the controlling action of the control electrode and therefore about the base voltage. The amount of the controlling action of the output electrode, that is the collector voltage on the collector current, is also given in the transistor through the internal resistance R_i. This unwanted controlling action of the collector voltage occurs in two ways. It brings about a variation of the current amplification and determines the amount of base voltage necessary for a particular base current. The characteristic value "internal resistance R_i" thus comprises two control actions or two reactions in the transistor. If we only indicate one of the two control effects of the collector voltage, for example that on the current amplification, we show this internal resistance as R_i^* $(1/h_{22})$. The second reaction which influences the input voltage is also indicated through a single characteristic value known as the inverse voltage amplification D_v or the voltage reaction $D_v = h_{12} = h_{re}$. In the transistor we, therefore, obtain the internal resistance R_i from both reactions together. With the slope S, the internal resistance R_i and the connected load resistance R_L, we obtain the amount of voltage amplification A_v in the transistor, in the same way as in the electron tube.

As a result of the collector voltage reaction on the input voltage the input resistance of the transistor between base and emitter is not a constant characteristic value. The input resistance r_i, therefore, depends on the amount of collector voltage variation, i.e. on the voltage amplification A_v and therefore on the value of the load resistance R_L.

Now in detail:

When we investigate the amount of amplifying action of a transistor, compared with an electron tube, we have to establish the extent of the pulling effect, the control action of the base-emitter voltage, on the collector current, and find out the strength of the attraction of the collector.

These two controlling influences on the collector current could also be expressed through a two-quadrant characteristic field, as in the electron tube. However, as the transistor is *by its nature* a current amplifying element, as has already been stated, this property is also expressed when we investigate the amount of amplification. We therefore make use of a four-quadrant characteristic field when we measure a transistor and plot all the interesting data in a graph (Fig. 1.5).

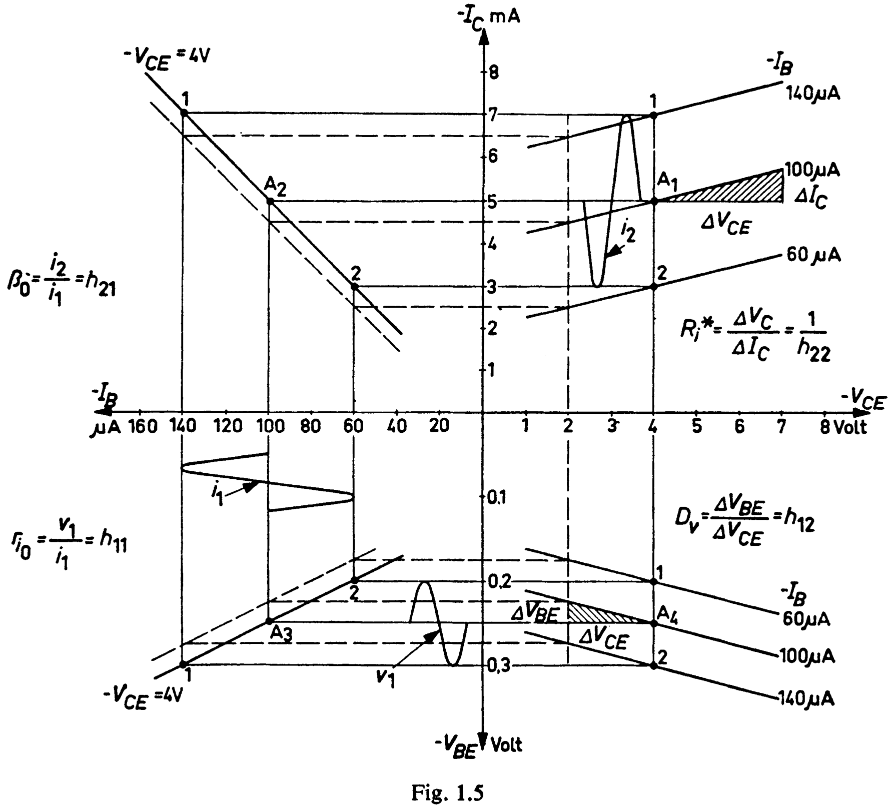


Fig. 1.5

The control effect of the collector voltage $-V_{CE}$ on the collector current $-I_C$ can be seen in *the first quadrant*. Every characteristic line which can be obtained point by point through measurement is valid for a certain constant base current, its course being very similar to that of a pentode tube characteristic. With a change of ΔV_{CE} in the collector voltage, the current varies only slightly by ΔI_C because of the flatness of the characteristic curve.

This control action can be expressed by:

$$R_i^* = \frac{\Delta V_{CE}}{\Delta I_C}$$

In the case in point, $R_i^* = 3$ V/0.6 mA $= 5$ kΩ. This characteristic value R_i^* only contains the control effect of the collector voltage on the collector current with a base current which is kept constant. The collector current, however, is additionally influenced through a reaction on the base voltage. Consequently, the internal resistance R_i of the transistor diverges somewhat from the value R_i^*, as shown in Chapter 5.

In the second quadrant which gives the connection between input current or base current $-I_B$ and output current or collector current $-I_C$, a characteristic curve can again be constructed with the help of points 1, A_1 and 2, as for the tube. The current amplification characteristic applies in this case for a constant collector voltage $-V_{CE}$ of 4 V. A new characteristic curve could be plotted for every other collector voltage.

The third quadrant shows the connection between input voltage $-V_{BE}$, the controlling base voltage, and the input current $-I_B$, the controlling base current. This kind of characteristic line is only applicable for a specific collector voltage. We can construct it with the aid of the "reaction characteristics" in the fourth quadrant. These characteristic lines tell us how large the base voltage has to be for a particular base current when the collector voltage assumes a certain value; i.e. the required base voltage depends on the collector voltage; there is a reaction on the input. With the help of points 1, A_4 and 2, points 1, A_3 and 2 can be plotted in the third quadrant for $-V_{CE} = 4$ V.

The amount of the voltage reaction can thus be observed in *the fourth quadrant*. With a variation ΔV_{CE} of the collector voltage the base voltage must be altered by ΔV_{BE} if the reaction is to produce no variation of the base current. Otherwise, ΔV_{BE} is the portion of the collector voltage variation which reacts on the base. In the fourth quadrant we obtain the voltage reaction from:

$$D_v = h_{re} = \frac{\Delta V_{BE}}{\Delta V_{CE}}.$$

In Fig. 1.5, $h_{re} = 0.025$ V/2 V $= 0.0125$.

Since the collector voltage additionally influences the collector current in this way through this second reaction, we obtain the characteristic value for the total control action of the collector, the internal resistance R_i from:

$$R_i = \frac{R_i^*}{1 - R_i^* \cdot S \cdot h_{re}}.$$

The amount and origin of the transistor slope S will be explained later.

If we apply a constant voltage $-V_{BE}$ of 0.25 V between base and emitter, a base current of 100 μA and a collector current $-I_C$ of 5 mA will flow; we obtain the operating points A_3, A_2 and A_1. If we apply to the input an a.c. voltage $v_1 = 0.05$ V, an a.c. base current i_1 of 40 μA and an a.c. collector current i_2 of 2 mA will flow.

The characteristic line in the second quadrant represents the current amplification:

$$\beta_0 = \frac{i_2}{i_1}$$

which here has the value $\beta_0 = 2$ mA/40 μA $= 50$. In the third quadrant the characteristic represents the *input resistance* $r_{i_0} = v_1/i_1$.
In this case the input resistance is:

$$r_{i_0} = 0.05 \text{ V}/40\ \mu\text{A} = 1.25\ \text{k}\Omega.$$

In the electron tube, the control effect of the grid on the anode was expressed through the slope S (mutual conductance). In the transistor also, the control action of the base on the collector current can be expressed through a similar characteristic value S. The connection between the controlling voltage v_1 and the a.c. output current i_2 extends here over quadrants 3 and 2. We obtain the transistor slope from:

$$S = \frac{i_2}{v_1} = \frac{\beta_0}{r_{i_0}}.$$

In our case, $S = 50/1.25$ kΩ $= 40$ mA/V.

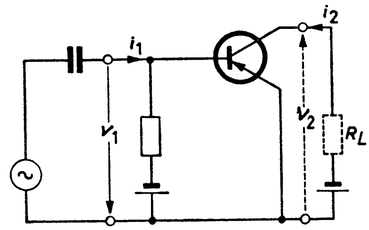


Fig. 1.6

These considerations of the control process in Fig. 1.5 were only applicable for constant collector voltage. If there is now voltage amplification as well as current amplification, a load resistance R_L has to be introduced as is indicated in Fig. 1.6. Now, however, the collector voltage is no longer constant during control and therefore contributes to the control.

For example, in Fig. 1.7 there is a supply voltage V_S of 8 V. The collector-

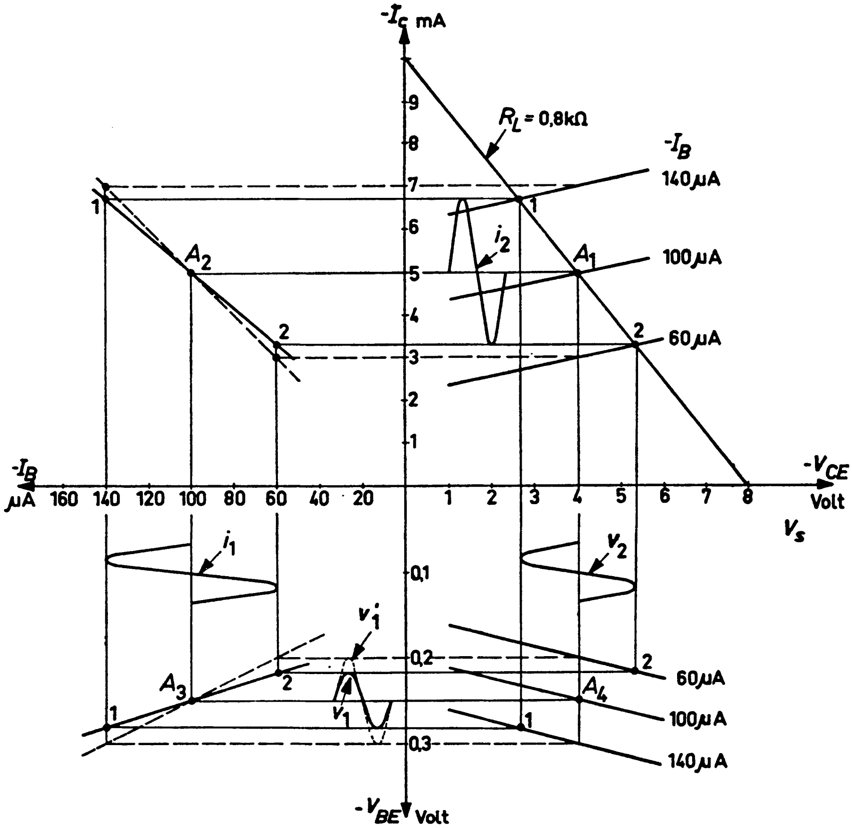


Fig. 1.7

emitter voltage — V_{CE} will only be equal to this voltage when the collector current is zero. With increasing collector current this voltage decreases as is shown by the straight line for $R_L = 0.8$ kΩ. During control we now have the intersection points 1, A_1 znd 2, or in the fourth quadrant 1, A_4 and 2. If these points are transposed, as already described, into the second or third quadrant, we obtain the dynamic characteristic for current amplification.

$$A_i = \beta_0 \cdot \frac{R_i^*}{R_i^* + R_L}.$$

In the case in point, A_i is 50 (5/(5 + 0.8)) = 43.

For the input resistance in the third quadrant we have:

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot h_{re}}.$$

With the two known characteristic values, of the electron tube, S and R_i, the voltage amplification in the transistor can also be calculated from:

$$A_v = S \cdot \frac{R_i \cdot R_L}{R_i + R_L}.$$

In our case,

$$S = 50 \text{ mA/V}.$$

$$R_i = \frac{R_i^*}{1 - R_i^* \cdot S \cdot h_{re}} = \frac{5 \text{ k}\Omega}{1 - 5 \text{ k}\Omega \cdot 40 \text{ mA/V} \cdot 0.0125} = -3.34 \text{ k}\Omega.$$

This minus sign only occurs here because the voltage reaction D_v was made much too big, for diagrammatic reasons.

The voltage amplification works out as:

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L} = 40 \text{ mA/V} \frac{-3.34 \cdot 0.8}{-3.34 + 0.8} \text{k}\Omega = 42.$$

The relation

$$A_i = A_v \frac{r_i}{R_L} = \frac{v_2}{v_1} \frac{r_i}{R_L} = \frac{i_2}{i_1}$$

already set up for the electron tube also applies for current amplification. In the case shown in Fig. 1.7, we here obtain:

$$r_i = \frac{r_{i0}}{1 + A_v \cdot h_{re}} = \frac{1.25 \text{ k}\Omega}{1 + 0.0125 \cdot 42} = 0.82 \text{ k}\Omega$$

$$A_i = A_v \frac{r_i}{R_L} = 42 \frac{0.82}{0.8} = 43.$$

The input current i_1 of 40 μA remains constant in Fig. 1.7 in contrast to Fig. 1.5. As we see, only an a.c. input voltage $v_1 = i_1 r_i = 40\ \mu\text{A} \cdot 0.82 \text{ k}\Omega = 32.8$ mV is now required to control this current because of this a.c. voltage reaction, in comparison with the 50 mV needed in the first case.

The a.c. output current is $i_2 = i_1 \cdot A_i = 40\ \mu\text{A} \cdot 43 = 1.72$ mA.

For the a.c. output voltage we obtain:

$$v_2 = i_2 R_L = 1.72 \text{ mA} \cdot 0.8 \text{ k}\Omega = 1.37 \text{ V}.$$

The characteristic curves used here to improve the graphic representation agree in their specific information with the actual characteristics; they deviate only in dimension which, however, is not important for the fundamental representation.

1.7. What simple circuit illustrates amplification in the electron tube?

When the characteristic values S and R_i of a tube are known, we can determine the voltage amplification A_v by calculation and so obtain the amplified voltage v_2 when an input voltage v_1 has been applied for control at the grid, from:

$$v_2 = v_1 \cdot A_v = v_1 \cdot S \frac{R_i \cdot R_L}{R_i + R_L}.$$

In this formula $v_1 \cdot S$ is a current which, according to Ohm's Law, produces the voltage v_2 as voltage drop in the parallel circuit:

$$R_p = \frac{R_i \cdot R_L}{R_i + R_L}.$$

Consequently the following circuit obviously applies:

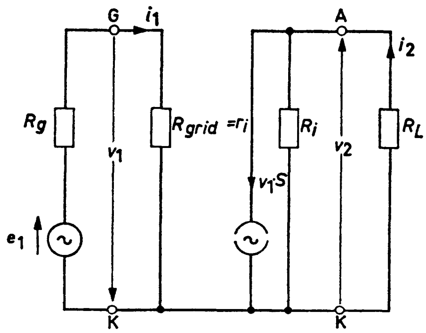


Fig. 1.8

The a.c. input voltage to be amplified is applied at the grid resistance r_i between grid and cathode. The amplified a.c. output voltage v_2 can be tapped between anode and cathode. We obtain the a.c. output voltage from the simple circuit which includes a generator yielding a current $v_1 \cdot S$. This current is imposed on the parallel circuit $R_i \parallel R_L$, i.e. this current does not depend on the total load resistance $R_i \parallel R_L$ of the generator. Consequently we have to visualise a generator with very high internal resistance, so that the current is determined by this internal resistance alone. This high internal resistance is represented by the interrupted generator switch symbol.

The advantage of this form of so-called equivalent circuit lies in the fact that we can more easily survey and calculate the operation of a tube in conjunction with, for example, other circuit elements of an amplifier. For higher frequencies we can also plot the undesirable capacitances in this type of equivalent circuit and so estimate the frequency dependence of the amplification.

1.8. What simple circuit illustrates amplification in the transistor?

As the transistor has the same characteristic values, S and R_i, as the electron tube for small signal amplification, the same equivalent circuit is valid for the transistor as for the tube.

The only difference compared with the electron tube is that the input resistance R_i of the transistor also varies with change of the load resistance R_L. However, since the amplification will always be estimated or be valid for a particular load resistance, this reaction can be ignored in such cases.

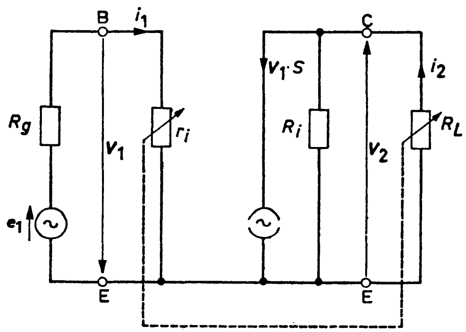


Fig. 1.9

The transistor equivalent circuit shows in a simple manner the connections with the rest of the circuit elements in a specific circuit. The matching ratios for the control generator and the load resistance are particularly easy to recognise, i.e. the greatest control voltage v_1 is obtained when the control generator internal resistance R_g is small compared with the input resistance r_i. For the largest possible voltage drop across the parallel circuit of R_i and R_L, R_L should as far as possible have the value of R_i. This is only achieved with great difficulty in transistors because R_i can be of the order of 100 kΩ, though the load resistance, for example, is formed by the low value of input resistance of the next transistor.

The behaviour of the transistor at high frequencies can also be described by introducing capacitances into the equivalent circuit, as will be shown in Chapter 16.

1.9. How is the working point set?

In the electron tube the grid bias is chiefly obtained by means of a cathode resistance.

The resulting voltage drop across the cathode resistance acts with negative polarity on the control grid and so represents the negative grid to cathode bias for setting the anode current working point. The resistance R_K is capacitively bridged and thus short-circuited for a.c. voltage, so that the bias remains constant during control of the tube. In the transistor an emitter resistance R_E would correspond to the cathode resistance of the electron tube. Unfortunately, this resistance in the transistor is not in a position to produce the bias for the base. The reason is that in the transistor the control

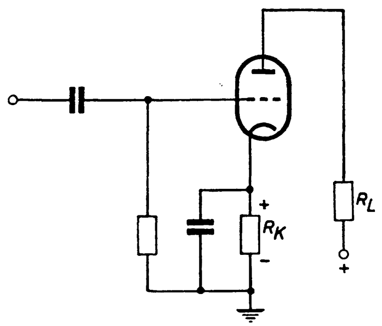


Fig. 1.10

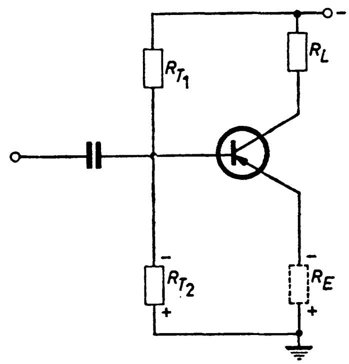


Fig. 1.11

electrode and the output electrode are both attracting and so have the same negative polarity. The voltage drop across an emitter resistance has positive polarity towards the base and can thus produce no negative bias. To generate a base bias we therefore generally make use of a voltage divider, R_{T_1}, R_{T_2}, which reduces the negative working voltage to the voltage required for adjusting the collector current. If this divider is of high resistance, it then becomes a pure series resistor. This type of divider is used especially in output stages and is partly made up of temperature-dependent resistances. The effects of temperature on the transistor can be compensated in this way. This process is called "stabilisation". An emitter resistance which serves as negative feedback is also used for stabilisation.

1.10. What is the effect of feedback?

Feedback means leading a part of the amplified a.c. output voltage back to the input so that the controlling a.c. input voltage (the voltage to be amplified) is weakened in its control action. Feedback is used, for example, to stabilise the amplifying properties of the electron tube and the transistor. Therefore, if the amplification becomes weaker for any reason, for instance, through ageing, the feedback automatically becomes less and a greater part of the input voltage is able to control. In this way the output voltage only varies slightly in spite of reduced amplification; the amplification is "stabilized". We usually speak of "current feedback" and "voltage feedback" in the electron tube while the corresponding circuit in the transistor is known as series or parallel feedback. It will be shown later that the feedback has the same effect in both amplifying devices, but with the transistor, the resistance of the control generator has to be studied more closely because of its low value.

The action of feedback in the transistor and the electron tube can be most easily compared if we consider that the voltage v'_1 to be amplified lies at a series connection of the input resistance r_i and the feedback resistance R_{fb}. This resistance R_{fb} appears greater for the applied voltage v'_1 than the resistance R_E or R_K because the current through this resistance, corresponding to the current amplification $A_i = S \cdot r_i$, is greater than the current i_1 which is actually driven through v'_1. Therefore only 1/51 part of the input voltage $v'_1 = 51$ mV actually operates as controlling voltage $v_1 = 1$ mV, in accordance with the voltage divider ratio in the two cases shown as an example in Fig. 1.12.

The feedback with a transistor and an electron tube taken as example in Fig. 1.12 can thus operate equally in both cases as is indicated by the numerical values chosen. All the same, there is still a basic factor to be considered which is apt to be forgotten in the tube, that is the internal resistance of the control generator. The following important principle applies:

The effect of the feedback depends on the value of the internal resistance of the control generator and it can happen that the influence of the feedback can be completely eliminated by this. As Fig. 1.12 shows, the internal resistance of the generator R_i must be equal to 1 MΩ in both cases. We could consequently believe that equal ratios are found as was evident for the other values. This, however, is not the case. On the contrary – and this has to be watched with both the tube and the transistor – the feedback action is entirely eliminated in the transistor. Why?

When the amplification in the transistor, that is to say, the slope, becomes less, the feedback resistance $R_{fb} = r_i \cdot S \cdot R_E$ is also reduced. With constant input voltage v'_1, v_1 would now increase and compensate the amplification loss by a higher control voltage. Unfortunately, however, in our case with $R_g = 1\ \text{M}\Omega$, the input voltage v'_1 decreases if R_{fb} becomes smaller. This is because the control current i_1 in this control circuit is virtually only determined by R_g because the total input resistance in the transistor only

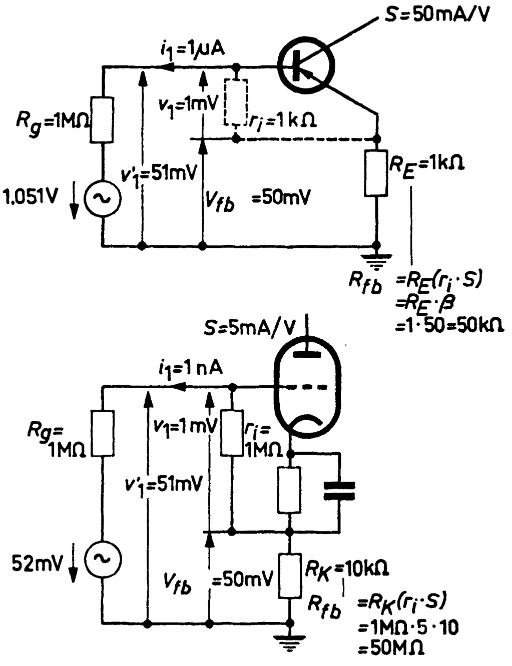


Fig. 1.12

amounts to $r_i + R_{fb} = 51\ \text{k}\Omega$. Therefore ,if the feedback resistance R_{fb} varies with reduced amplification, practically the same control current flows through r_i; the control voltage v_1 does not increase. There is thus no stabilisation and the effect of the feedback is nil.

If we now compare the corresponding values for the electron tube we find that with a variation of the feedback resistance $R_{fb} = S \cdot r_i \cdot R_K = 50\ \text{M}\Omega$, the input voltage will almost remain constant because here the generator with $R_g = 1\ \text{M}\Omega$ is loaded by 51 MΩ. The stabilisation is still fully effective with the tube. If we had only given the tube a grid resistance r_i of 100 kΩ and connected in a cathode resistance R_K of 1 kΩ, the feedback resistance R_{fb} would be $S \cdot r_i \cdot R_K = 5 \cdot 100 \cdot 1 = 500\ \text{k}\Omega$, and so the total input resistance would be 600 kΩ. We can see here that the stabilising effect is greatly reduced with $R_g = 1\ \text{M}\Omega$ and with a control generator with $R'_g = 10\ \text{M}\Omega$ would become zero. Again this is because a variation in the feedback resistance cannot change the control current if the internal resistance of the control generator is large in proportion to the feedback resistance. Therefore in unfavourable cases the control generator resistance can cancel the stabilising effect of the feedback in the electron tube also. This danger is greater in the transistor than in the electron tube because of the transistor's low ohmic quality.

We can say in both cases that the series or current feedback, is only effective with voltage control, i.e. when the control generator resistance R_g is small in comparison with the total input resistance $r_i + R_{fb}$.

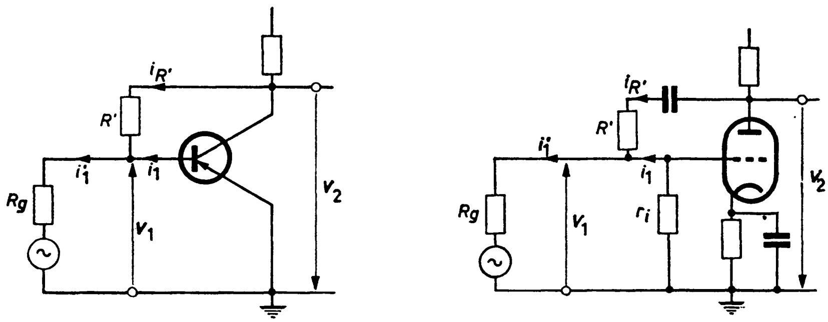


Fig. 1.13

This form of feedback in Fig. 1.13 is based on the fact that the resistance R' with a value $R'/(1 + A_v)$ is in parallel with the input terminals, i.e. at the input voltage v_1. The total input current i'_1 is formed from the normal control current i_1 and the current i'_R. Since the current i'_R is driven in the resistance R' by the voltage $v_1 + v_2 = v_1(1 + A_v)$ this resistance appears divided by $1 + A_v$ when we consider it as connected directly at the input to v_1, and is thus reduced to $R'/(1 + A_v)$. The stabilisation is then effective because with a variation, for instance, a decrease, of A_v, this resistance appears greater. The stabilisation, however, only functions here if both amplifying elements are controlled with a constant current, i.e. if the control generator resistance R_g is large compared with the total input resistance $r_i \parallel R'/(1 + A_v)$. We call this "current control". With decreasing voltage amplification A_v, the resistance of the parallel circuit will increase and the constant inflowing control current will cause a greater voltage drop, thus producing a greater control voltage v_1. The decrease in amplification is again compensated for here by a higher control voltage. It is important to realise here that with voltage control, that is, with a low resistance control generator, the feedback does not function in either case because then the input voltage which at the same time is the control voltage at r_i, cannot be altered by the variable resistance $R'/(1 + A_v)$.

The parallel feedback in the transistor and the voltage feedback in the electron tube are thus only fully effective with current control.

1.11. What is a residual current?

The current drawn from the base voltage out of the emitter is a dynamic current, that is, the current flowing with a base-emitter diode connected in the forward direction. This current flows for the most part via the base-collector diode connected in the inverse direction because there is no blocking effect for this current on account of the polarity of its charge carriers. However, the working voltage also allows an additional current to flow in the inverse direction which would even flow on its own when the emitter diode is not connected.

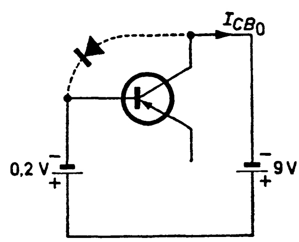


Fig. 1.14

This kind of current is called residual current. Here it is the collector-base residual current, I_{CB_0}. This residual current is of particular importance since, among other possible residual currents, it is the only one flowing in the circuit during transistor operation because even then the base-collector diode is connected in the inverse direction. The value of such a residual current is only slightly dependent on the driving voltage, here $9 - 0.2 = 8.8$ V (saturation), as is obvious from every diode characteristic for the blocking range.

1.12. How does temperature affect the transistor?

The currents in the transistor are not constant in spite of constant applied voltages if the temperature in the transistor varies. Temperature variation can occur through the external temperature or the internal heating due to dissipation. The dynamic current thus increases with rising temperature as though the base-emitter voltage has altered by about 2 mV for 1°C temperature rise. In this way the collector current rises to double its value with about 20 to 30°C temperature increase. The residual current, for instance, I_{CB_0} is considerably more dependent on temperature and already reaches twice its value at a variation of 9°C. Depending on the circuit and the amount of normal base current in proportion to the residual current I_{CB_0}, this increased residual current can lead to a still greater variation of the collector current which can be dangerous for the transistor. These influences therefore depend largely on the circuit and will be dealt with in detail later.

1.13. What is "stabilisation"?

Every amplifying device, whether electron tube or transistor, should retain its amplifying properties unchanged infinitely, as far as possible. There are ageing phenomena in the electron tube and temperature influences in the transistor which result in variation. In both cases measures for stabilisation are taken, mainly through feedback arrangements. In the transistor, for example, there is the frequently used emitter resistance which through a d.c. feedback ensures stabilisation of the emitter current and therefore of the collector current.

The working point, i.e. the collector static current, can therefore be only very slightly displaced by a variation of the temperature. Changes in the characteristic values for a.c. voltage amplification are also prevented as a result. Direct stabilisation of the a.c. voltage can be achieved through an uncapacitively bridged emitter resistance or other special voltage feedbacks.

1.14. What peculiarities have the cathode follower stage and the common collector circuit?

The cathode follower stage is used in electron tube circuits for matching. The input resistance of such stages can be made very large and the effective internal resistance at the output can be made very small. In this way we can provide a very high ohmic design with a low ohmic output if we connect this kind of stage after it.

The collector-base stage also has the property of this form of resistance transformation. However, such high values of input resistance as with electron tubes are not reached. On the other hand, the output resistance of such a stage is greater in comparison with the electron tube because it is dependent on the amount of the control generator resistance.

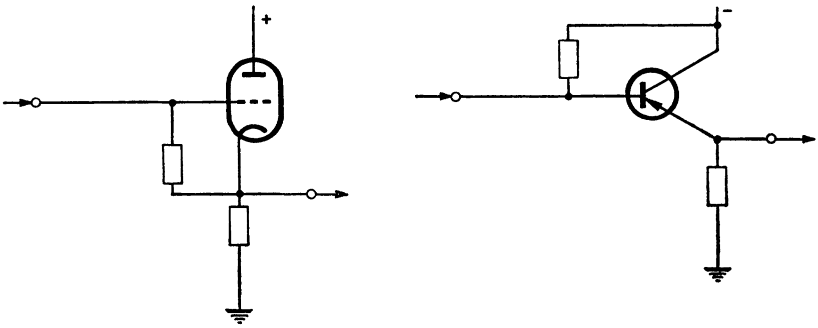


Fig. 1.15

1.15. Are the earthed grid stage and the common base stage equal to each other?

The small input resistance for control is common to both stages. Also, considering the amplification action at high frequency, both stages show advantages over the fundamental circuit. The common base circuit in the transistor is regarded as having a very high limiting frequency. That is correct for the current amplification, but unfortunately this is less than unity in the common base circuit and so we can gain nothing from it. Consequently, in the common base circuit only the voltage amplification and its limiting

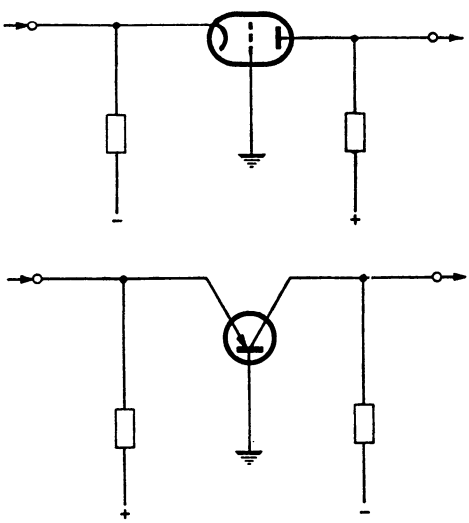


Fig. 1.16

frequency are of interest. Depending on the component values it can be somewhat higher than in the fundamental circuit, the common emitter circuit.

1.16. A word about current and voltage directions

In principle, the technical current or voltage direction is used here in all considerations concerning circuits. Every voltage thus drives a current from the positive pole to the negative pole. This fact makes us say, for example, that in the electron tube a current flows from the anode to the cathode although we know that from the physical standpoint electrons are liberated in the cathode which then fly to the anode. This determination of current direction is therefore not always linked with the true physical current direction. It is only meant to serve as a means of explaining, comparing and calculating the regularity of the laws of flow in any network. If, for example, we assume that electron tubes and transistors are both present in a circuit and that all the currents and voltages have been calculated, we also know then that the direction of current in the electron tube is actually reversed while in the transistor the assumed technical direction of current is correct because (with a *p–n–p* transistor) positive charge carriers travel.

If every current and voltage is now in a particular direction, we have to indicate these also in a circuit diagram. By agreement, this indication of direction is shown by arrows or double indices or by both together.

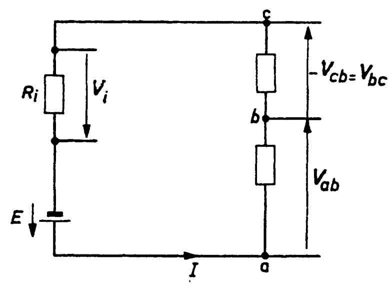


Fig. 1.17

In Fig. 1.17 an initial voltage E drives a current I in its own direction. This current is driven from a to b so that the voltage drop V_{ab} drives between a and b. The voltage between b and c also operates from b to c. A voltage V_{cb} would operate corresponding to its index from c to b. If, however, we wish to indicate that it does not do this, we write a minus sign before it.

Now $-V_{cb}$ indicates a voltage which does not operate from c to b but in the opposite direction. A voltage indication $-V_{BE}$ thus also means that a voltage acts here from E to B, i.e. E is positive and B is negative. These symbols are found with transistor d.c. voltages.

The current direction is usually only indicated by the arrow. With the transistor it has become customary for all currents which flow from the transistor to be denoted with a minus sign. As the base current flows out of the transistor, it is therefore indicated as $-I_C$.

In spite of all this, however, the principle remains that all currents and voltages are based on the technical current direction. This is also valid for a.c. currents and voltages for which the momentary relative directions are given.

CHAPTER 2

PHYSICAL PRINCIPLES OF THE TRANSISTOR

The physics of the semiconductor play an important part in the consideration of internal transistor processes. A precise knowledge of all the physical processes is not necessary for technical use of transistor amplifying devices. Here the measured characteristic curves and values are of the greatest interest. All the same, some knowledge of the physical "inner life" of the transistor is useful for interpreting specific transistor effects. The following investigations are based on the processes in the germanium semi-conductor but also apply fundamentally for the behaviour of silicon semiconductors.

2.1. The conduction mechanism in germanium

Germanium has the atomic number 32 in the periodic system of elements. It therefore has 32 electrons. The external 4th electron shell whose electrons have the highest energy content, is occupied by 4 electrons.

2.1.1. CRYSTAL STRUCTURE

Only the four electrons of the external shell play any part in the consideration of the crystal lattice bonds between the germanium atoms. In an "equivalent atom" for germanium, therefore, four electrons and four positive nuclear charges are assumed (Fig. 2.1). All four electrons contribute to the bond, i.e. they act with a certain force on another adjacent atomic nucleus. This dynamic effect takes place because these electrons not only circulate about their own atomic nucleus but interchange with an electron of an adjacent nucleus around which they themselves now revolve. Because of these combination functions of the external valency electrons, germanium is chiefly (at low temperatures) an insulator. Through external influences in the form of energy supply by means of heat or light, however, individual electrons can lose their valency bond (semiconductor). We therefore speak of intrinsic conduction and call these charge carriers minority carriers.

2.1.2. INTRINSIC CONDUCTION

When an electron leaves its valency shell and so breaks open a lattice bond, a "hole" is left behind. This hole acts as a positive charge carrier because

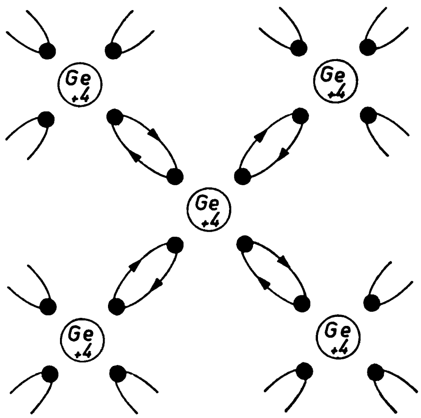


Fig. 2.1

the positive nuclear charge with + 4 predominates over the charge of −3 electrons. The hole will therefore endeavour to recapture an electron (recombination) which re-establishes the lattice bond. This tendency is not found in metals so we do not refer to "holes" in that case. It is immaterial in recombination whether the electron comes from an adjacent stream or is another free electron. If an impulse is exerted by an electrical field, electrons travel in one direction and holes in the other. We also call the positive charge carriers "defective electrons". If we want to raise the number of free carrier pairs, we have to increase the energy supply (heat or light) (high temperature conductor, photoconductor). There are then more carrier pairs in the medium than will recombine.

There is, however, still another method of producing carrier pairs, namely by alloying the semiconductor with other materials (extrinsic conduction). To sum up we can say that a semiconductor is a material which according to its lattice structure represents a non-conductor. Free charge carriers are produced through heat or light so that in germanium there is an increase of about the factor 3 per 15°C rise in temperature.

2.1.3. EXTRINSIC CONDUCTION

In order to raise the conductivity of germanium we add an alloy of antimony or arsenic atoms in the ratio of about 1 : 10 millions. Both are quinquevalent atoms, i.e. they have five electrons in their outer shell. These spurious atoms are absorbed into the lattice bond in place of a germanium atom (Fig. 2.2). The result is that only 4 of the 5 valency electrons of arsenic, for example, contribute to the electron bond in the germanium crystal lattice. The fifth

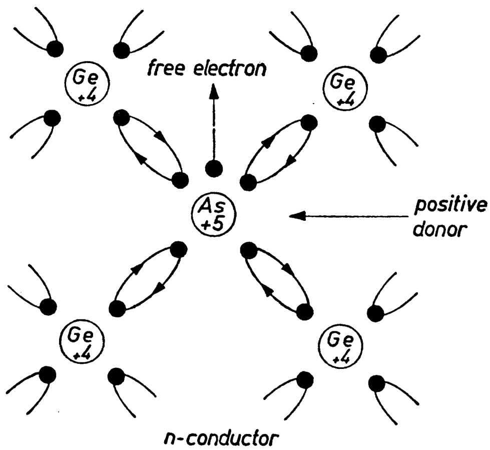


Fig. 2.2

valency electron is so loosely attached that it can move as a free conduction electron. The positive atomic residue does not show the tendency to capture an electron and there are therefore no holes. Disturbing materials like arsenic and antimony are known as donors; this kind of imperfect semiconductor is of the *n*-type.

If we alloy germanium with trivalent material, that is one with three valency electrons, the following situation is the outcome: If gallium or indium, for instance, is used, these atoms work themselves into the crystal structure of the germanium. However, there is now one electron missing

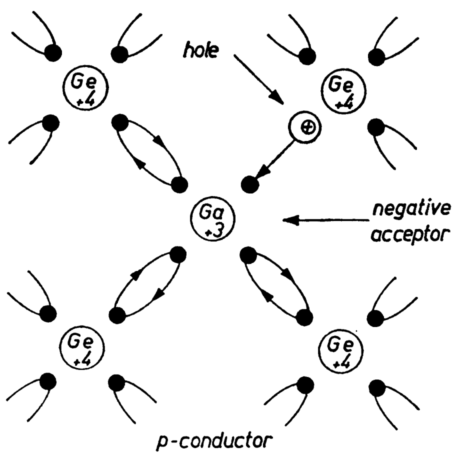


Fig. 2.3

from each gallium atom for the lattice bond. Through the combination exchange of the electrons from atom to atom, this missing electron is eventually replaced by an electron from a germanium atom (Fig. 2.3). There is now a hole in each germanium atom concerned. The conductivity is therefore raised through positive charge carriers. The gallium atom securely installed in the lattice bond now has 4 electrons and is thus negatively charged. It does not show a tendency to give up this electron. Such disturbing materials which combine with electrons are called acceptors; the semiconductor is of the *p*-type. Since extrinsic conduction far exceeds intrinsic conduction, the charge carriers formed here are known as majority carriers.

2.2. The *p–n* boundary layer

If we allow close impact between *p*- and *n*-germanium through alloying, a boundary layer is produced between the two in the dimension of 20 μm, resulting in the following transformations. The *n*- and *p*-germanium are each considered electrically neutral, and therefore uncharged. The sum of all the charges is nil.

Positive charge carriers can move in the *p*-germanium and negative ones in the *n*-germanium. If two electrically neutral *p*- and *n*-materials strike one another, there is no electrical impulse (electric field) which could lead

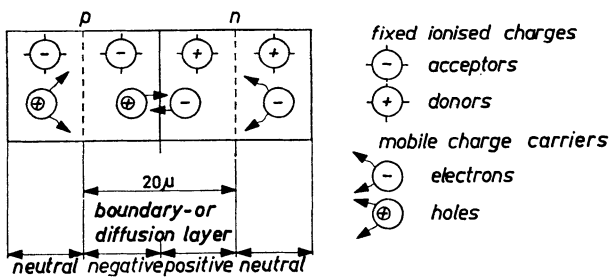


Fig. 2.4

to a spontaneous union of the positive and negative charge carriers. As a result of the natural mobility (thermal agitation), however, holes from the *p*-germanium diffuse in the boundary layer into the *n*-germanium and recombine there with the electrons. Conversely there is also a diffusion of electrons from the *n*-layer into the *p*-layer. This process is not kept up long enough for all the free electrons to be combined with holes or vice

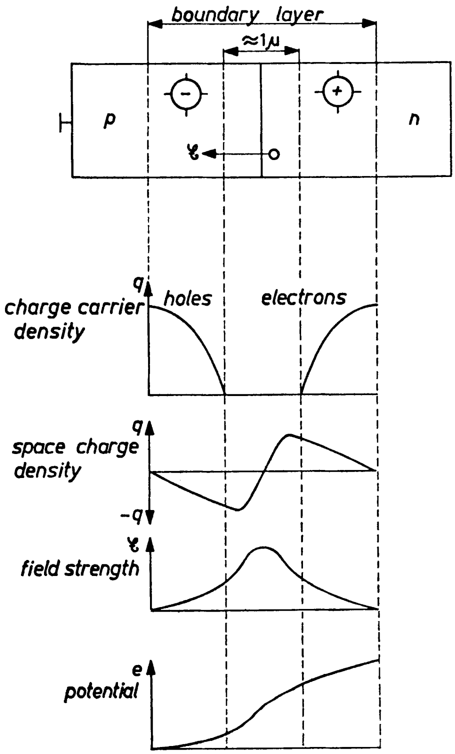


Fig. 2.5

versa, as we might at first assume. As far as the diffusion does succeed, there is now an electric charge field between the positive donors and the negative acceptors which are no longer neutralised by increasing diffusion (Fig. 2.4.)

The resulting charge field is now so directed, however, that it counteracts the combination, that is, the straying of the electrons towards the p-germanium and the movement of the holes towards the n-germanium. Therefore all the holes cannot combine with the electrons. Equilibrium is produced between the effect of diffusion and the reversing energy of the space charge field (Fig. 2.5). Along the space charge field the field strength is dependent on the charge density of the space charge. In a narrow limiting zone all the electrons and holes are recombined; there the acceptor- or donor-density, the space charge density, is at its greatest. The possibility of recombination becomes gradually less with growing boundary layer thickness, while the density of charge carriers increases and the space charge density decreases (more acceptors or donors are neutralised). The sum of all $E \cdot \Delta l$ or the line integral of the field strength gives the voltage gradient or the potential threshold along the boundary layer. The potential gradient in relation to the p-layer is shown in Fig. 2.5. The expression "potential threshold" results from the fact that the electrons or holes remaining cannot jump over this potential threshold unless energy is supplied in the form of heat or radiation.

2.3. The p–n-combination as rectifier

The rectifying effect of a p–n-layer is based on the fact that by applying an external voltage the potential threshold in the boundary layer is strengthened or weakened according to the polarity (Fig. 2.6). In case a, the negative

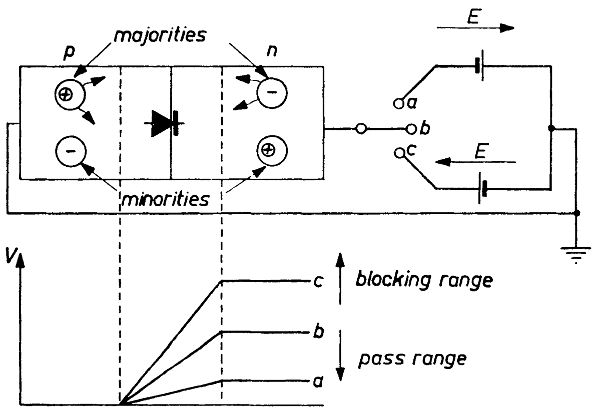


Fig. 2.6

pole of the voltage source with the initial voltage E is at the n-layer. Electrons of the n-layer are driven through this towards p and the holes of the p-layer move towards n. The initial voltage E causes a voltage drop from p to n at the boundary layer, i.e. n becomes more negative and the potential threshold is lowered from b to a. The rectifier operates in the passband; the reduced potential threshold can no longer prevent the charge exchange (transmission current). It is different in case c with polarity in the blocking direction. Here, through the + pole at the n-layer the electrons of this zone are removed from the boundary layer while holes in the p-layer are also removed from the boundary layer. The space charge is therefore increased and the potential threshold becomes greater (c). Nevertheless a slight current still flows in the blocking direction. It is not formed through the majority carriers (extrinsic conduction) but through minority carriers (intrinsic conduction). The potential threshold cannot detain these minority carriers but even forces them on. By applying higher blocking voltages the small blocking current can rise sharply and a flood of new charge carriers is produced (Zener zone).

2.4. The junction transistor

In the junction transistor two *p–n*-layers are arranged one behind the other. If one *n*-layer forms the common layer for two boundary layers, we obtain a *p–n–p* transistor. We can regard this transistor as a counter circuit containing two diodes. The centre electrode is called the base and two outer ones are known as emitter and collector respectively. A voltage in the blocking direction is always applied at the collector-base section. The diode operating in the transmission is the base-emitter section (Fig. 2.7).

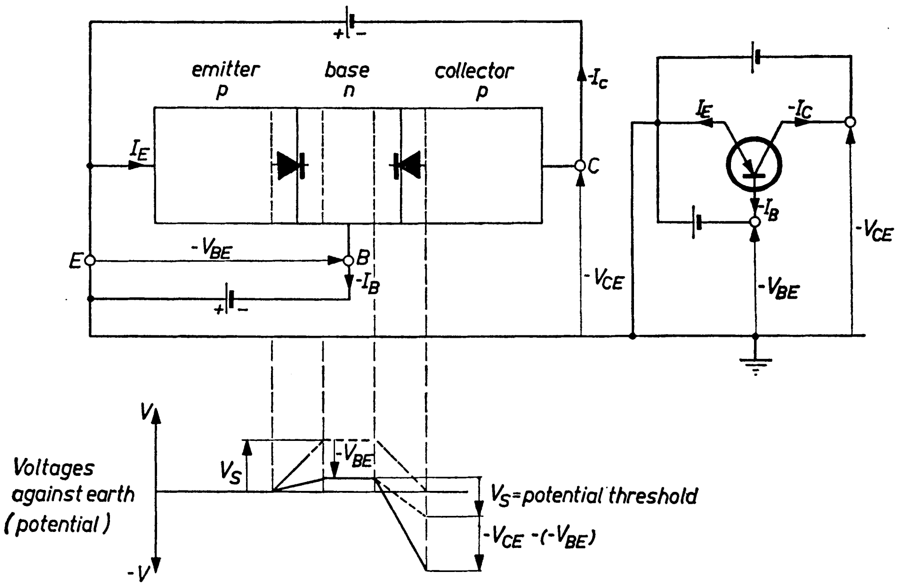


Fig. 2.7

If at first only the base-emitter voltage $-V_{BE}$ is applied, holes from the *p*-layer travel into the base space because the potential threshold $V_{S_{BE}}$ between emitter or base has been reduced ($V_{S_{BE}} - V_{BE}$). With equal proportioning or imperfection density in the *p*- and *n*-germanium, the holes would recombine with the electrons in the base space. Actually, however, the imperfection density in the base space is made about 500 times less than in the emitter zone. At first, therefore, nearly all the holes from the emitter injected into the base travel as far as the base electrode and unite there with conduction electrons. The holes in the base space are produced because the combination electrons stray into the emitter layer. If a voltage $-V_{CE} + V_{BC}$ is now applied at the base-collector section, only a very small current flows to the base electrode. The holes drawn through the base voltage and injected into the base zone now move for the most part into the collector zone. This *transistor effect* takes place for the following reasons. The applied voltages mainly occur as voltage drops at the boundary layers. There is consequently practically no force on the holes through external voltages in the base space. The holes move only as a result of the diffusion effect or, in the drift transistor are driven through a drift charge field in the direction of the collector. This drift field is produced through a concentration drop i.e. the donor density decreases in the collector direction inside the base. Since in addition to this the base zone is made very narrow ($10 + 100\,\mu$m) the holes drawn out of the emitter easily reach the base-collector boundary layer. The potential threshold between the negative base charge carriers and the positive collector charge carriers is further increased by the applied collector voltage $-V_{CE}$ ($-V_{S_{BE}}$ $-V_{CE} + V_{BE}$). This potential threshold, however, is no longer a barrier to the holes injected into the base, on the contrary, they "rush" down at this potential drop. Almost all the holes from the emitter injected into the base are gathered in, so to speak, by the collector. The remainder goes as base current to the base electrode. The collector current is only slightly dependent on the collector voltage. The ratio between the collector current and the small controlling base current is the current amplification factor B, (direct current amplification, h_{FE}).

CHAPTER 3

SYMBOLS, SIGNS, FUNDAMENTAL CIRCUIT

The technical symbol for the *p–n–p* transistor has already been shown in the fundamental circuit, the common emitter circuit, in Fig. 2.7. In this circuit the emitter is the common connecting point for all applied transistor voltages. All the current and voltage directions are based on the technical current direction from positive to negative. The voltage indication $-V_{BE}$ signifies that the voltage does not operate according to the given index

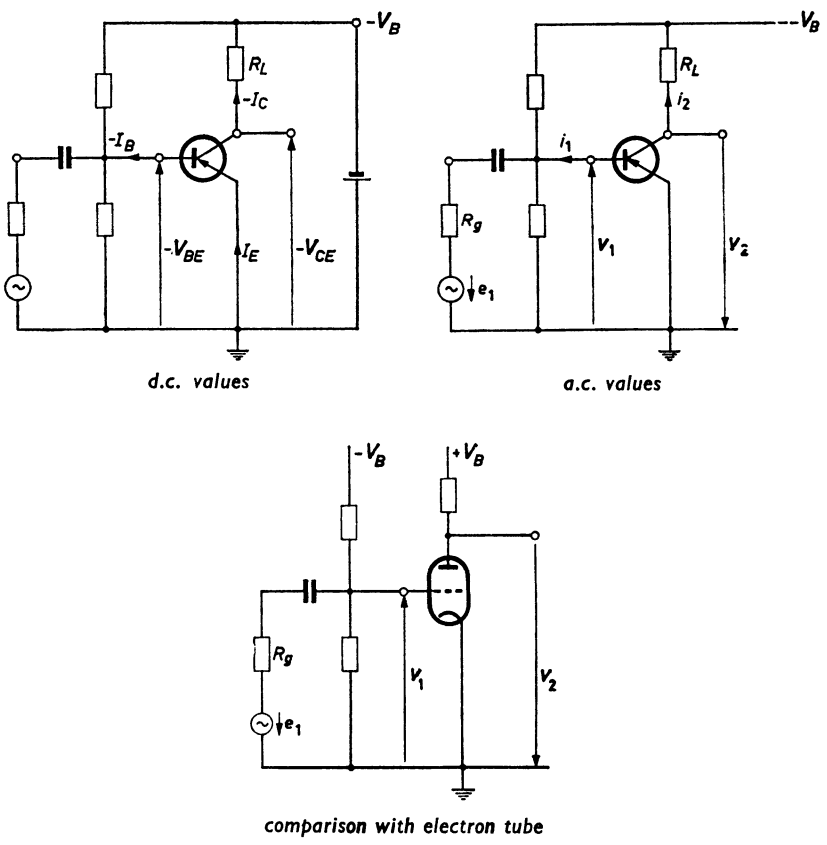


Fig. 3.1

BE, from the base to the emitter, but in the opposite direction and is therefore negative. All direct currents which flow to the interior of the transistor are indicated with the positive sign. A current $-I_C$ thus means that the current flows away from the transistor. These direction signs are customary in published characteristic fields. In the *n–p–n* transistor the emitter current and the collector current are reversed in direction compared with the *p–n–p* transistor. The transistor symbol therefore shows a reversed arrow for the emitter.

We insert small letters for control with alternating current values. Thus v_1 indicates the a.c. input voltage, v_2 the a.c. output voltage, i_1 the a.c. input current and i_2 the a.c. output current. The a.c. current values are superimposed on the d.c. operating currents and voltages. The actual relative directions of the a.c. current values are always used in the following statements. The d.c. and a.c. values are shown separately in two circuit diagrams in Fig. 3.1. It can be seen, for example, that the output voltage is rotated in phase by 180 ° in relation to the input voltage.

By describing the a.c. values alone and entirely omitting the d.c. supply, we finally reach the transistor equivalent circuit diagram which will be shown later.

If we contrast the electron tube fundamental circuit with the transistor fundamental circuit we learn which electrodes correspond to each other. Thus the grid and the base, the cathode and the emitter, the anode and the collector all have the same task. All the following derivations are based in the first instance on this transistor fundamental circuit. By comparing it with the electron tube fundamental circuit we can discover the extensive formulary agreement between the two amplifying elements.

CHAPTER 4

THE FOUR QUADRANT CHARACTERISTIC FIELD

The measured transistor characteristics serve best at first to describe the technical behaviour of the transistor. We can obtain quantitative and qualitative statements concerning the amplification of a.c. voltages and currents from characteristics of this kind. For mathematical considerations characteristic values can be defined with the help of the characteristic curves. The measurement of characteristics for a.c. behaviour is not so simple for the transistor as for the electron tube because point by point plotting of "static characteristics" is not suitable for describing a.c. processes. Characteristics for dynamic transistor behaviour have to be plotted at constant crystal temperature (see temperature behaviour of the transistor). Each characteristic curve must therefore be dynamically measured from a working point of constant dissipation (oscillograph).

The a.c. behaviour of the transistor at low frequencies can be seen from the four quadrant characteristic field. This field is shown in Fig. 4.1. for the fundamental circuit, the emitter-base circuit. Why is a four quadrant characteristic field assigned to the transistor when two quadrants are obviously sufficient for the electron tube? To that we can reply that it is just as feasible for a.c. voltage control to give a two quadrant characteristic field for the transistor as it is for the electron tube. However, as the transistor in contrast to the electron tube, is not only a voltage amplifier but amplifies current as well, it is desirable that we should be able to investigate this property also from the characteristic curves. Apart from this, it is evident that the transistor has two amplification reciprocals, a voltage amplification reciprocal and a reciprocal of current amplification, i.e. compared with the electron tube which only shows a voltage reaction, the transistor has a current reaction as well. Nevertheless, both reactions in the transistor can be combined into a common amplification reciprocal as will be shown later.

In the characteristic field in Fig. 4.1. characteristic values for calculation can be obtained in the individual quadrants. If A is the working point in the quadrants concerned, a characteristic value for small signal amplification can be determined from the slope of the respective characteristics at this point. Such characteristic values are termed h-coefficients because they are used in the h-matrix in a four-pole representation of the transistor.

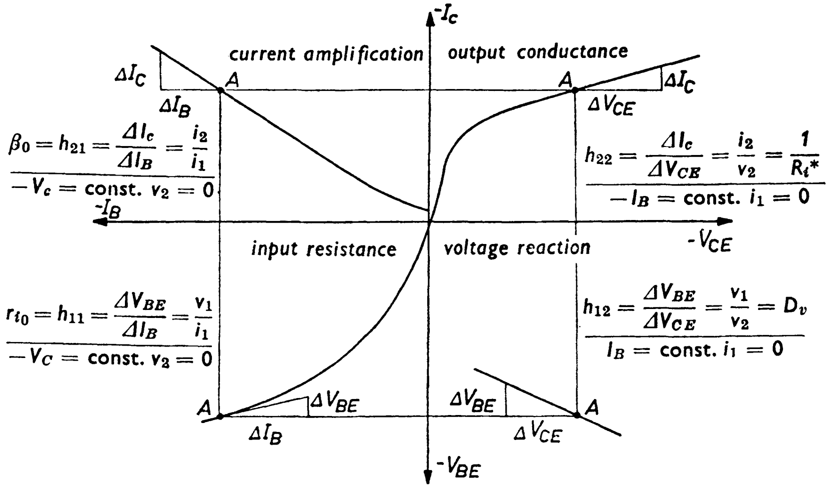


Fig. 4.1

In the first quadrant we obtain the
Output conductance:

$$h_{22} = \frac{i_2}{v_2} \qquad i_1 = 0. \tag{4.1}$$

In this case v_2 is the a.c. voltage applied at the output and i_2 the current which flows when the input is open for alternating current; thus $i_1 = 0$ (open circuit).

The second quadrant shows the
Current amplification:

$$h_{21} = \frac{i_2}{i_1} \qquad v_2 = 0. \tag{4.2}$$

Here i_1 is the a.c. input current when an a.c. input voltage has been applied and i_2 is the a.c. output current flowing when the output is short-circuited for alternating current; thus $v_2 = 0$ or V_{CE} is constant.

The third quadrant supplies the
Input short-circuit resistance:

$$h_{11} = \frac{v_1}{i_1} \qquad v_2 = 0. \tag{4.3}$$

This characteristic value results from the a.c. input current values for short-circuited output; thus $v_2 = 0$.

From the fourth quadrant comes the
Voltage reaction:

$$h_{12} = \frac{v_1}{v_2} \qquad i_1 = 0. \tag{4.4}$$

Here v_2 is the a.c. voltage supplied at the output and v_1 the resultant voltage measured at the input in the case when the input is open; thus $i_1 = 0$ (open circuit).

The connections between the characteristics in the individual quadrants and the validity of their information will be studied in Fig. 4.2.

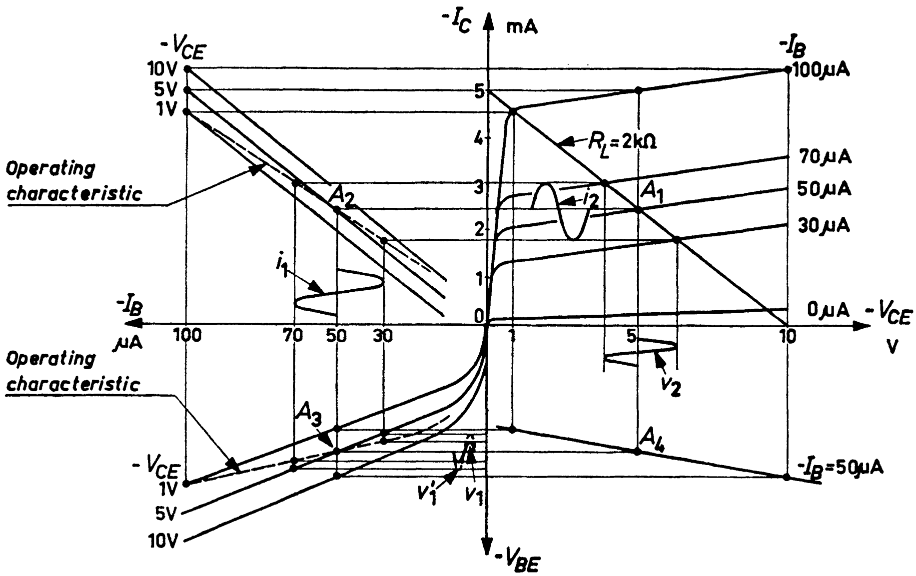


Fig. 4.2

The *first quadrant* shows the connection between collector voltage and collector current. Each characteristic is valid for constant base current $-I_B$. We notice that the general course of this characteristic is very similar to the I_a–V_a characteristic field of the pentode. The rise at the beginning of the curves is steeper than for the pentode; the so-called knee-voltage at the bend of the characteristic is therefore smaller in proportion. If the transistor is operated with a load resistance of $R_L = 2\ \text{k}\Omega$, a collector voltage $-V_{CE} = 5$ V appears, as Fig. 4.2 shows, with a working voltage of 10 V if the collector current $-I_C = 2.5$ mA, because the voltage drop at the load resistance is $V_L = 2\ \text{k}\Omega \cdot 2.5\ \text{mA} = 5\ \text{V}$. The working point A_1 is thus set with a base current $-I_B$ of 50 μA. For control with working point A_1, the slope of the characteristics, the output conductance h_{22} can now be stated. In the electron tube in this case the internal resistance R_i was read. The reciprocal value of h_{22}, the output conductance, is also an a.c. internal resistance here and will be denoted as $R_i{}^*$

$$R_i{}^* = \frac{1}{h_{22}} = \frac{v_2}{i_2} \qquad i_1 = 0. \tag{4.5}$$

This internal resistance is indicated as $R_i{}^*$ because it is not as in the electron tube, the actual effective source resistance. The difference arises from the fact that the characteristics apply for constant base current $-I_B$. The reaction of the collector voltage variation v_2 only extends to the current amplification in this characteristic field; we can therefore speak of a current reaction. The fact that there is also a voltage reaction on the input voltage alters this internal resistance $R_i{}^*$ and produces the actual transistor internal resistance R_i as will be described later.

We can learn the "current reaction" from the *second quadrant* which shows the connection between collector current and base current. With $-V_C = 5$ V constant, the "static characteristic" for current amplification has been constructed in Fig. 4.2 from the first quadrant. With a non-constant collector voltage, because of the load resistance $R_L = 2\ \text{k}\Omega$, the plotted points on the resistance straight line can be transferred to the second quadrant and we obtain the dotted operating characteristic. The slope of this dynamic characteristic for control in working point A_2 is less than that of the static characteristic whose slope is the current amplification.

$$\beta_0 = h_{21} = \frac{i_2}{i_1}\,. \tag{4.6}$$

The variation of current amplification through the output voltage v_2 can therefore actually be indicated as a reaction on the current, a "current reaction".

The *third quadrant* now gives the connection between input voltage and input current. The course of these characteristic curves depends, as Fig. 4.2. shows, on the collector voltage, i.e. each static characteristic is valid for a constant collector voltage, as in the second quadrant. The voltage reaction of the output voltage v_2 on the input voltage v_1 appears in this characteristic field. If there were no voltage reaction, there would have to be an input voltage v'_1 for current control, i.e. for the current i_1 corresponding, for instance, to the characteristic for $-V_{CE} = 5$ V. Nevertheless, because of the voltage reaction, the dotted characteristic applies if the collector voltage, in accordance with the resistance straight line for $R_L = 2\ \text{k}\Omega$, varies with the modulation. Only the smaller input voltage v_1 is now required for the same current control i_2. The slope of the static characteristic at working point A_3, the input short-circuit resistance h_{11}, is denoted as r_{i_0}

$$r_{i_0} = h_{11}. \tag{4.7}$$

The input resistance present during control with load resistance through voltage reaction, corresponding to the slope of the dotted curve at working point A_3, is greater than r_{i_0} and is always indicated as r_i.

The characteristics for the voltage reaction are illustrated in the *fourth quadrant*. This is the connection between the collector voltage at the input and the base voltage necessary for a certain base current. With the help of these characteristics, we can construct the characteristics in the fourth quadrant, as can be seen in Fig. 4.2. The slope of the characteristic at working point A_4 represents the voltage reaction h_{12}. As with the electron tube, this characteristic value can be termed the inverse voltage amplification $D_v(h_{re})$

$$D_v = h_{12}. \tag{4.8}$$

It will be shown later that we can also define an inverse current amplification D_i, corresponding to the current reaction in the first and second quadrants.

To *sum up*, the four quadrant characteristic field of the transistor gives information concerning the behaviour of the transistor during control. From the illustration of the transistor fundamental circuit, the emitter-base circuit, in Fig. 4.2, we can recognise in particular the *two* reactions of the transistor which must be regarded as a very important contrast to the electron tube with only one reaction. The current reaction of the collector voltage appears with reduced slope as dynamic characteristic for current. amplification. The voltage reaction is seen as dynamic characteristic for input current and input voltage, which flattens off here. The input resistance therefore becomes apparently less through the voltage reaction. Such characteristics can be used with particular advantage in large signal amplification. In this way, all interesting data such as current, voltage and output amplification, input resistance and so on, can be calculated. The probable non-linear distortions can also be estimated.

CHAPTER 5

TWO-POLE REPRESENTATION OF THE TRANSISTOR

We shall not deal with the four-pole representation of the transistor introduced earlier, because of its lack of clarity. On the other hand, reducing the tube to a two-pole representation has proved useful from the start. Why should a similar presentation not be equally applicable for the transistor? The essential difference compared with the tube, namely the two-fold reaction in the transistor, has already been discovered from the characteristics. It will now be shown how the transistor characteristic values S, R_i and D are obtained and what two-pole equations can be drawn up with them. Since the four-pole h-characteristics are usually given in data sheets for transistors, the two-pole characteristic values of the transistor will also be expressed through the h-characteristics.

The characteristic curves in the first, second and third quadrants are shown once more in Fig. 5.1, in simplified or idealised form. To begin with, only the current reaction will be considered. As far as the characteristics are concerned, the ratios are exactly as in the electron tube. We

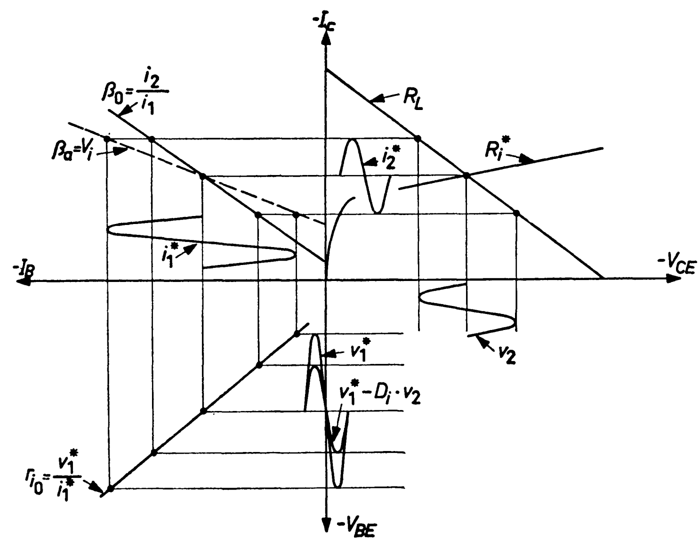


Fig. 5.1

learn the reaction from the dynamic characteristic for current amplification whose slope is less because of this reaction. The idea "current reaction" resulted from this fact. The second and third quadrants now obviously also supply information about the transistor slope

$$S^* = \frac{i_2}{v_1} = \frac{\beta}{r_{i_0}} = \frac{i_2}{i_1} \cdot \frac{i_1}{v_1} = \frac{h_{21}}{h_{11}}. \tag{5.1}$$

This slope S^* is also valid at first, therefore, without taking the voltage reaction into account, because the static characteristics for $-V_{CE}$ = constant were used in the definition. From the relevant internal resistance:

$$R_i^* = \frac{1}{h_{22}} \tag{5.2}$$

we should now obtain, according to Barkhausen, an inverse amplification, the "inverse current amplification":

$$D_i = D^* = \frac{1}{S^* \cdot R_i^*} = \frac{h_{11} \cdot h_{22}}{h_{21}}. \tag{5.3}$$

As in the electron tube, $v_2 \cdot D_i$ states the amount of output voltage which apparently operates in reverse at the input, and weakens the current amplification here. The characteristic diagram in Fig. 5.1 shows clearly that with the static characteristic β_0, a smaller input voltage is necessary than with the dynamic current characteristic β_a for modulating an equal current i_2^*. Since the dynamic current characteristic β_d always applies in the actual case, we can say that the input voltage v_1^* available is less effective in comparison with the static case with $v_2 \cdot D_i$. The actual control voltage is $v_1^* - v_2 \cdot D_i$, and the output current is therefore obtained from:

$$i_2^* = i_1^* \cdot \beta_d = \frac{v_1^* - v_2 \cdot D_i}{r_{i_0}} \cdot \beta_0.$$

Since β_0/r_{i_0} is now equal to S^*, the following equations are also valid:

$$i_2^* = S^*(v_1^* - D_i v_2) = S^*(v_1^* - D_i \cdot i_2^* \cdot R_L)$$

$$i_2^* = S^* v_1^* - S^* \cdot D_i \cdot i_2^* \cdot R_L = S^* \cdot v_1 - i_2^* \frac{R_L}{R_i^*}$$

$$i_2^* + i_2^* \frac{R_L}{R_i^*} = S^* \cdot v_1^*$$

$$i_2^* = v_1^* \cdot S^* \frac{1}{1 + \frac{R_L}{R_i^*}} = v_1^* \cdot S^* \frac{R_i^*}{R_i^* + R_L}. \tag{5.4}$$

The current i_2^* obtained without taking the voltage reaction into account thus results from the short circuit current $v_1^* \cdot S^*$ which appears at the parallel circuit of R_i^* and R_L as output current in the resistance R_L. The output current i_2^* must also have a dynamic slope S_d^*. Consequently $i_2^* = v_1^* \cdot S_d^*$ must be valid as well.

According to Eq. (5.4) the dynamic slope is:

$$S_d^* = S^* \frac{R_i^*}{R_i^* + R_L}. \tag{5.5}$$

This slope was the result of the "operating current amplification" β_a because the reaction affects the current amplification β_0, as we saw from Fig. 5.1. The "operating current amplification" can therefore be calculated with the same "reduction factor".

$$\beta_a = \beta_0 \cdot \frac{R_i^*}{R_i^* + R_L}. \tag{5.6}$$

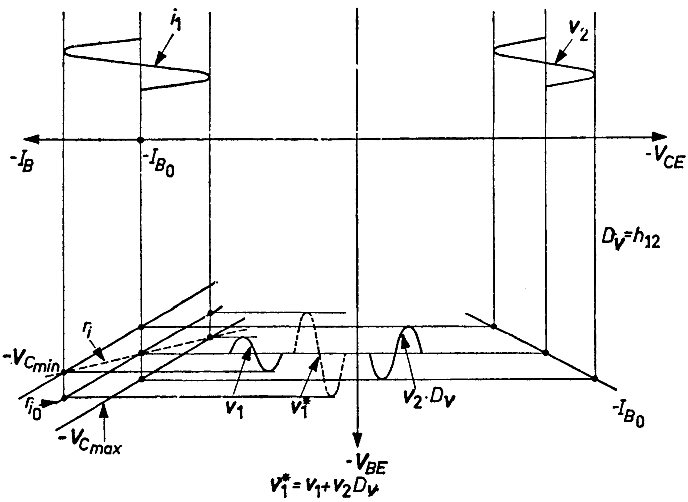


Fig. 5.2

To calculate the actual output current i_2 we must now also consider the voltage reaction. This reaction is shown again in simplified form in the third and fourth quadrants of Fig. 5.2. The static characteristics for r_{i_0} are given for variable collector voltage $-V_{CE}$, as the three parallel r_{i_0} characteristic lines show. If the two outer r_{i_0} characteristics apply for the largest and smallest collector voltages respectively with v_2, all other intermediate values thus lie on the dotted dynamic characteristic r_i, the effective input resistance. An input voltage v_1^* was required to control the input current i_1, ignoring the voltage reaction. If we now include the voltage reaction, only an input voltage v_1 is required, corresponding to the new input characteristic for r_t. This reaction voltage $v_2 \cdot D_v$ acts on the lines of a positive feedback and so reduces the necessary input voltage v_1.

We therefore obtain:

$$v_1 = v_1^* - v_2 \cdot D_v \quad \text{ou:} \quad v_1^* = v_1 + v_2 \cdot D_v\,. \tag{5.7}$$

If the voltage v_1^* from the preceding relation is now inserted into Equation (5.4) we obtain the actual current i_2, taking both transistor reactions into account.

We thus have:

$$i_2 = (v_1 + v_2 \cdot D_v) \cdot S^* \frac{R_t^*}{R_t^* + R_L}$$

$$i_2 = (v_1 + i_2 \cdot R_L \cdot D_v) \cdot S^* \frac{R_t^*}{R_t^* + R_L}$$

$$i_2 = v_1 \cdot S^* \frac{R_t^*}{R_t^* + R_L} + i_2 \cdot R_L \cdot D_v \cdot S^* \frac{R_t^*}{R_t^* + R_L}$$

$$i_2 \left(1 - R_L \cdot D_v \cdot S^* \frac{R_t^*}{R_t^* + R_L}\right) = v_1 \cdot S^* \frac{R_t^*}{R_t^* + R_L}$$

$$i_2 = \frac{v_1 \cdot S^* \dfrac{R_t^*}{R_t^* + R_L}}{1 - R_L \cdot D_v \cdot S^* \dfrac{R_t^*}{R_t^* + R_L}} = \frac{v_1}{\dfrac{R_t^* + R_L}{S^* \cdot R_t} - R_L \cdot D_v}$$

$$i_2 = \frac{v_1}{D_t \cdot R_t^* + D_t \cdot R_L - R_L \cdot D_v} = \frac{v_1}{D_t \cdot R_t^* + (D_t - D_v) \cdot R_L}$$

$$i_2 = \frac{v_1}{D_t - D_v} \frac{1}{\dfrac{R_t^*}{\left(\dfrac{D_t - D_v}{D_t}\right)} + R_L} = \frac{v_1}{D} \frac{1}{R_t + R_L}. \tag{5.8}$$

We now observe from the equation found through simple algebraic conversions, that the total reciprocal amplification in the transistor is obtained from the difference between the two reciprocal amplifications.

Reciprocal transistor amplification

$$D = D_t - D_v = \frac{h_{11} \cdot h_{22}}{h_{21}} - h_{12}. \tag{5.9}$$

The resulting internal resistance of the transistor is obtained from:

Transistor internal resistance

$$R_i = \frac{R_i^*}{1 - \dfrac{D_v}{D_i}} = \frac{1}{h_{22}\left(1 - \dfrac{h_{12} \cdot h_{21}}{h_{11} \cdot h_{22}}\right)}. \tag{5.10}$$

The slope of the transistor must as a result be derived from:

$$S = \frac{1}{D \cdot R_i} = \frac{D_i - D_v}{(D_i - D_v)R_i^* \cdot D_i} = \frac{1}{R_i^* \cdot D_i} = \frac{\hbar_{22} \cdot h_{21}}{h_{11} \cdot \hbar_{22}}.$$

We thus obtain the same slope as when the voltage reaction was not included because the static slope applies in principle for short-circuit on the output side, and so for constant output voltage, which is not the case for D and R_i.

Transistor slope

$$S = S^* = \frac{\beta_0}{r_{i_0}} = \frac{h_{21}}{h_{11}}. \tag{5.11}$$

With these "tube characteristics" for the transistor, it is now easy to calculate all the interesting working values for the loaded transistor.

For the a.c. output voltage v_2 in the fundamental circuit we have:

a.c. output voltage

$$v_2 = i_2 \cdot R_L = \frac{v_1}{D} \frac{R_L}{R_i + R_L} = v_1 \cdot S \frac{R_i \cdot R_L}{R_i + R_L} \tag{5.12}$$

according to Equation (5.8), when extended with R_i.

The voltage amplification A_v comes from:

$$A_v = \frac{v_2}{v_1} = \frac{1}{D} \cdot \frac{R_L}{R_i + R_L} \quad \text{or}$$

voltage amplification

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L}. \tag{5.13}$$

The input resistance r_{i_0} is reduced through the voltage reaction because the inverse voltage also drives the input current. Therefore,

$$i_1 = \frac{v_1 + v_2 \cdot D_v}{r_{i_0}}.$$

The input resistance is consequently obtained from:

$$r_i = \frac{v_1}{i_1} = \frac{v_1 \cdot r_{i_0}}{v_1 + v_2 \cdot D_v}.$$

If we abbreviate with v_1, we obtain:

Input resistance

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v} = \frac{h_{11}}{1 + A_v \cdot h_{12}}. \tag{5.14}$$

The current amplification A_i is found from the output current $i_2 = v_1 \cdot A_v \, 1/R_L$ and the input current $i_1 = v_1/r_i$, as:

Current amplification

$$A_i = \frac{i_2}{i_1} = A_v \frac{r_i}{R_L} = S \cdot r_i \frac{R_i}{R_i + R_L}. \tag{5.15}$$

By transistor capacitance amplification we understand the ratio of output capacity to input capacitance.

Capacitance amplification

$$A_p = \frac{v_2 \cdot i_2}{v_1 \cdot i_1} = A_v \cdot A_i. \tag{5.16}$$

Example 5.1

A transistor T_1 (for example, the OC 71) is operated at a working point $-V_{CE} = 3$ V and $I_E = 3$ mA. Its h-parameters have the following values:

$$h_{11} = 0.8 \text{ k}\Omega \qquad h_{12} = 5.4 \cdot 10^{-4}$$
$$h_{21} = 47 \qquad h_{22} = 80 \cdot 10^{-6} \text{ S}.$$

Transistor T_1 is loaded with $R_L = 3.3$ kΩ. The a.c. input voltage should amount to $v_1 = 10$ mV.

Required

The characteristic values, S, R_i and D and the complete working values.

Solution

$$S = \frac{h_{21}}{h_{11}} = \frac{47}{0.8 \cdot 10^3} = 58.8 \cdot 10^{-3} = 58.8 \text{ mA/V}$$

$$R_i = \frac{1}{h_{22} - \dfrac{h_{12} \cdot h_{21}}{h_{11}}} = \frac{1}{\dfrac{80}{10^4} - \dfrac{5.4 \cdot 47}{10^6 \cdot 0.8 \cdot 10^3}} = 20.8 \text{ k}\Omega$$

$$D = \frac{1}{S \cdot R_i} = \frac{10^3}{58.8 \cdot 20.8 \cdot 10^3} = 8.2 \cdot 10^{-4}$$

$$D_i = \frac{h_{11} \cdot h_{22}}{h_{21}} = \frac{0.8 \cdot 10^3 \cdot 80}{47 \cdot 10^6} = 13.6 \cdot 10^{-4}.$$

or

$$D_v = h_{12} = 5.4 \cdot 10^{-4}$$

$$D = D_i - D_v = (13.6 - 5.4) \cdot 10^{-4} = 8.2 \cdot 10^{-4}.$$

Voltage amplification

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L} = 58.8 \frac{20.8 \cdot 3.3}{20.8 + 3.3} = 168$$

$$A_{v_{\mathrm{dB}}} = 20 \lg 168 = 45 \text{ dB}.$$

Input resistance

$$r_i = \frac{h_{11}}{1 + A_v \cdot D_v} = \frac{0.8}{1 + 168 \cdot 5.4 \cdot 10^{-4}} = 0.735 \text{ k}\Omega.$$

Current amplification

$$A_i = A_v \frac{r_i}{R_L} = 168 \frac{0.735}{3.3} = 37.5.$$

Capacitance amplification

$$A_p = A_v \cdot A_i = 168 \cdot 37.5 = 6300$$

$$A_{p_{\mathrm{dB}}} = 10 \lg 6300 = 38 \text{ dB}.$$

Input current

$$i_1 = \frac{v_1}{r_i} = \frac{10 \cdot 10^{-3}}{0.735 \cdot 10^3} = 13.6 \,\mu\text{A}.$$

Output current

$$i_2 = i_1 \cdot A_i = 13.6 \cdot 37.5 = 510 \,\mu\text{A}.$$

Output voltage

$$v_2 = v_i \cdot A_v = 10 \cdot 168 = 1.68 \text{ V}.$$

Input capacitance

$$p_1 = v_1 \cdot i_1 = 10 \cdot 10^{-3} \cdot 13.6 \cdot 10^{-6} = 136 \cdot 10^{-9} \text{ W}.$$

Output capacitance

$$p_2 = v_2 \cdot i_2 = 1.68 \cdot 510 \cdot 10^{-6} = 858 \cdot 10^{-6} \text{ W}.$$

CHAPTER 6

TRANSISTOR EQUIVALENT CIRCUIT

6.1. Voltage equivalent circuit

The transistor equivalent circuit can now be evolved with the two-pole formulae for the transistor fundamental circuit in Fig. 3.1, the common emitter circuit. We learn from Equation (5.12) that the output voltage v_2 results from the initial voltage v_1/D which is split at the voltage divider R_i, R_L and appears at R_L as a voltage drop. We therefore obtain the simple equivalent circuit for the output parameters as for the electron tube.The input resistance is very high in the electron tube and is not usually included in the equivalent circuit diagram for this. In the transistor, however, the input parameters are dependent on this resistance. We thus have the transistor voltage equivalent circuit as shown in Fig. 6.1.

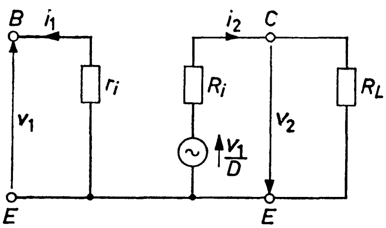


Fig. 6.1

6.2. Current equivalent circuit

According to Eq. (5.12) the output voltage v_2 also results from the current $v_1 \cdot S$ which can be regarded as supplied from a generator with an infinitely large internal resistance. v_2 is then the voltage drop produced by the current

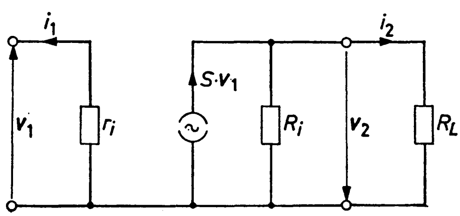


Fig. 6.2

$v_1 \cdot S$ at the parallel circuit of R_i and R_L. With this generator which supplies a constant current $v_1 \cdot S$ independent of the load resistance, we obtain the current equivalent circuit of the transistor as Fig. 6.2 shows.

Example 6.1

The transistor T_1 in Example 5.1. is to be represented in the voltage and current equivalent circuits.

Solution

See Fig. 6.3.

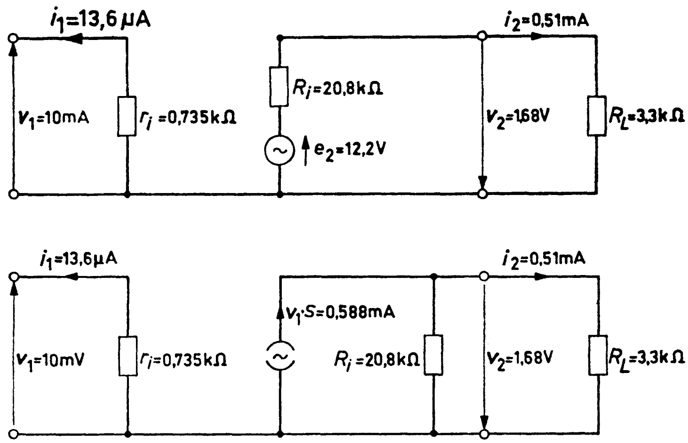


Fig. 6.3

CHAPTER 7

FIXING THE OPERATING POINT

The d.c. working voltages and currents must be fixed in accordance with the objective in hand. A negative bias is required at the base and the collector compared with the emitter. In the electron tube ,by contrast, a negative voltage is necessary at the grid and a positive one at the anode. With the electron tube, therefore, the bias can be produced through a cathode resistance. This bias production is not possible in the transistor by using a corresponding emitter resistance. Nevertheless, a resistance of this kind is present in most circuits; it produces no bias but serves for negative feedback.

The bias is fixed by means of a voltage divider connected to the working voltage. In practice, a voltage divider of this type may be high resistance or low resistance. A high resistance voltage divider has its lower part short-circuited by the low input resistance of the transistor in parallel with it and is thus transformed into a pure series resistance. The bias is therefore either produced through a high series resistance R_v or through a low resistance voltage divider as we see in Fig. 7.1.

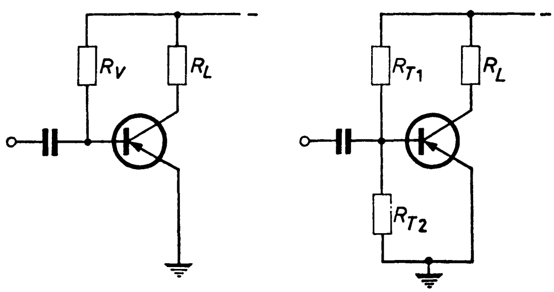


Fig. 7.1

7.1. Fixing with series resistance

The series resistance R_v represents the upper partial resistance of a voltage divider the lower part of which is formed through the low d.c. input resistance $-V_{CE}/-I_B$ ($\approx$ taken from the third quadrant) of the transistor. If the latter, for example, is 3 k,Ω the input voltage is $-V_{BE} = 0.3$ V, and $-I_B = 0.1$ mA, then with a working voltage of -12 V, R_v must amount to $12/0.1 - 3 =$ 117 kΩ. Even variations of the d.c. input resistance from transistor to transistor only cause slight alteration of the input current $-I_B$. The collector current and with it the collector working point can certainly deviate considerably through different current amplification and residual current and this can be very undesirable because it entails the possibility of exceeding the collector dissipation. For this reason the series resistance has to be adjusted for each transistor.

7.2. Adjustment with low resistance voltage divider

A low resistance voltage divider should set up a constant d.c. input voltage $-V_{BE}$ independent of variations in the transistor d.c. input resistance. This is only possible if the voltage divider transverse current is a multiple of the transistor input current. This has the drawback that the a.c. input resistance is reduced because for alternating current the two divider resistances lie parallel to the input resistance r_i. In calculating the values of resistance, we therefore always have to compromise. Moreover, the setting of the collector current working point is more difficult compared with the setting with series resistance because variations of the d.c. input resistance produce an additional displacement. Special adjustment of this type of voltage divider is thus more essential in transistor replacement than with the series resistance.

Why is the low resistance voltage divider used generally in spite of this? The reason is that we must have a low resistance voltage divider if we want to make use of working point stabilisation through series negative feedback by means of an emitter resistance (see Working Point Stabilisation 9.5.).

Thermistors are used as voltage divider resistances to compensate for the influence of temperature. The purpose of these voltage dividers will be described in more detail in the chapter concerning temperature influences.

Example 7.1

A transistor T_1 is loaded with $R_L = 2$ kΩ. The working voltage is $V =$ 9 V. The collector current working point should lie at $I_C = 2.25$ mA.

Required

The necessary series resistance R_V and the values R_{T_1} and R_{T_2} for a low resistance voltage divider. This voltage divider should only load the input voltage additionally with about 1 kΩ.

Solution

The resistance load line for $R_L = 2$ kΩ is plotted in the first quadrant of

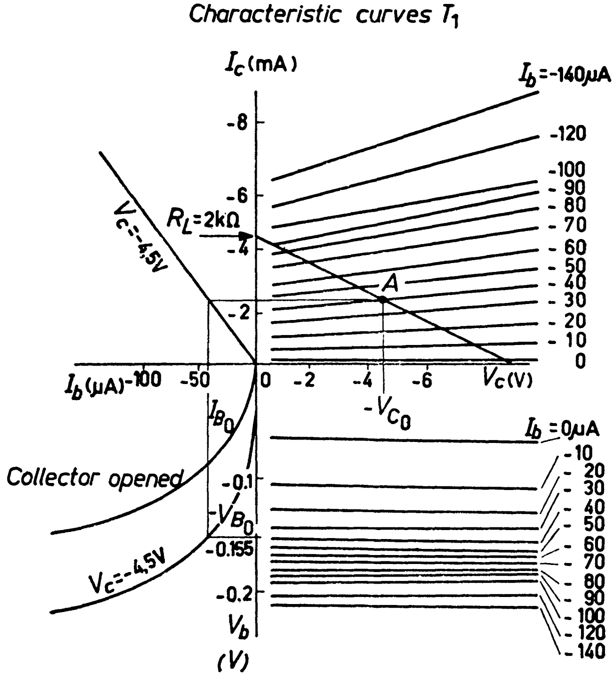


Fig. 7.2

the characteristic curves of transistor T_1 (Fig. 7.2.) This links $-V_{CE} = 9$ V with $-I_C = 4.5$ mA. The working point $-I_{C_0} = 2.25$ mA yields a static collector voltage $-V_{CE_0} = 9 - 2.25 \cdot 2 = 4.5$ V. This working point is obtained with a base current $-I_{B_0} = 40\ \mu$A. From the second and third quadrants we obtain approximately the base bias required $-V_{BE_0} = 0.155$ V. The d.c. input resistance is consequently:

$$R_{BE} = \frac{-V_{BE}}{-I_B} = \frac{0.155}{40 \cdot 10^{-6}} = 3.88\ \text{k}\Omega.$$

The series resistance required is then:

$$R_V = \frac{V}{I_B} - R_{BE} = \frac{9}{40 \cdot 10^{-6}} - 3.88 = 225 - 3{,}88 \approx 221\ \text{k}\Omega.$$

The low resistance voltage divider is obtained as follows:
An additional load which may amount to 1 kΩ is produced for the a.c. input voltage from the parallel circuit of R_{T_1} and R_{T_2}. Since with voltage division 9 : 0.155, the resistance R_{T_1} is about 60 times greater than R_{T_2}

its influence on the parallel circuit with R_{T_2} will not be substantial and R_{T_2} should therefore be chosen at 1 kΩ. The total d.c. input resistance is then:

$$R_{in} = R_{T_2} \parallel R_{BE} = \frac{1 \cdot 3.88}{1 + 3.88} = 0.8 \text{ k}\Omega.$$

Therefore,

$$R_{T_1} = R_{\text{tot}} - R_{in} = \frac{V}{I} - R_{in} = \frac{V}{V_{BE}/R_{in}} - R_{in} = \frac{9}{\frac{0.155}{0.8}} - 0.8 = 45.6 \text{ k}\Omega.$$

If the characteristic fields usually given are used, as this example shows, for calculating the d.c. values, for instance the d.c. input resistance R_{BE}, considerable errors can arise in certain circumstances. Strictly speaking, a four quadrant set of characteristic curves which has been statically recorded point by point should be used, because the normal published dynamic characteristics are only valid for a constant crystal temperature and therefore only apply for a particular working point of constant dissipation.

CHAPTER 8

APPARENT INTERNAL RESISTANCE WITH LOAD VARIATION

It must be noted that the voltage amplification changes with variation of the load resistance. Consequently the amount of the input resistance is also altered and according to Eq. (5.14) we can write:

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v} \approx \frac{r_{i_0}}{1 + D_v \cdot S \cdot R_L}.$$

In the equivalent circuit diagram, therefore, these two resistances can be represented as being coupled to each other (Fig. 8.1).

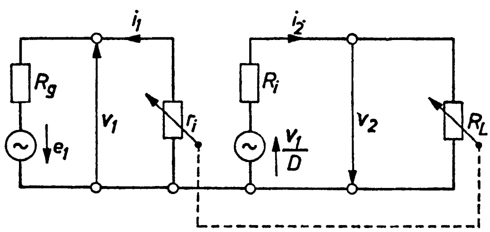


Fig. 8.1

If the source resistance R_g of the control generator e_1 is small compared with the transistor input resistance r_i, the voltage v_1 is practically constant if the load resistance and therefore the input resistance vary. However, if the source resistance R_g is not a negligible quantity, the amplification A_v is less, for example, with a smaller load resistance. This means, of course, that the input resistance r_i would become greater and the input voltage v_1 is increased as a result. Reduction of the amplification is equalised again to some extent through the rise in input voltage. This fact can also be interpreted as assigning a smaller output resistance R_{i_s} to the transistor because through the higher load the voltage has not decreased according to the actual R_i but to a lesser degree. In this consideration the input voltage is assumed as constant.

How do we obtain the altered output resistance R_{i_s} which is also called the "working output resistance" and which mainly acts as a dampling resistance in resonance amplifiers?

The equivalent circuit in Fig. 8.1 can naturally be calculated for two different load resistances, R_{L_1} and R_{L_2}. We then finally have the values $v_{2/1}$, $v_{2/2}$ and $i_{2/2}$. An output current change was produced by the load variation. As a result of this change in current, the voltage drop at the apparent internal resistance $R_{i_{eq}}$ is altered; this is the output voltage variation. The apparent internal resistance is then obtained from:

$$R_{i_{eq}} = \frac{v_{2/1} - v_{2/2}}{i_{2/2} - i_{2/1}}. \tag{8.1}$$

Example 8.1

A transistor T_1 with $S = 58.8$ mA/V; $R_i = 20.8$ kΩ;
$D = 8.15 \cdot 10^{-4}$; $r_{i_0} = 0.8$ kΩ;
$D_v = 5.4 \cdot 10^{-4}$
is loaded with $R_{L_1} = 2$ kΩ and $R_{L_2} = 0.5$ kΩ.
The control generator has an initial voltage $e_1 = 10$ mV and a source resistance $R_g = 1$ kΩ.
What apparent $R_{i_{eq}}$ is effective?

Solution

$R_{L_1} = 2$ kΩ

$$A_{v_1} = S\frac{R_i \cdot R_L}{R_i + R_L} = 58.8\,\frac{20.8 \cdot 2}{20.8 + 2} = 107$$

$$r_{i_1} = \frac{r_{i_0}}{1 + A_v \cdot D_v} = \frac{0.8}{1 + 107 \cdot 5.4 \cdot 10^{-4}} = 0.76\ \text{k}\Omega$$

$$v_{1/1} = e_1\frac{r_i}{r_i + R_g} = 10\,\frac{0.76}{0.76 + 1} = 4.32\ \text{mV}$$

$$v_{2/1} = v_1 \cdot A_{v_1} = 4.32 \cdot 107 = 462\ \text{mV}$$

$$i_{2/1} = \frac{v_{2/1}}{R_{L_1}} = \frac{462 \cdot 10^{-3}}{2 \cdot 10^3} = 231\ \mu\text{A}.$$

$R_{L_2} = 0.5$ kΩ

$$A_{v_2} = 58.8\,\frac{20.8 \cdot 0.5}{20.8 + 0.5} = 28.8$$

$$r_{i_2} = \frac{0.8}{1 + 28.8 \cdot 5.4 \cdot 10^{-4}} = 0.787\ \text{k}\Omega$$

$$v_{1/2} = 10 \frac{0.787}{0.787 + 1} = 4.4 \text{ mV}$$

$$v_{2/2} = 28.8 \cdot 4.4 = 126.5 \text{ mV}$$

$$i_{2/2} = \frac{126.5}{0.5} = 253 \ \mu\text{A}$$

$$R_{t_{\text{eq}}} = \frac{\Delta v_2}{\Delta i_2} = \frac{462 - 126.5}{253 - 231} = \frac{335.5}{22} = 15.2 \text{ k}\Omega.$$

A better method of accounting for the source resistance of the control generator and therefore for estimating the working output resistance $R_{t_{\text{eq}}}$ with load variations is given in Fig. 8.2.

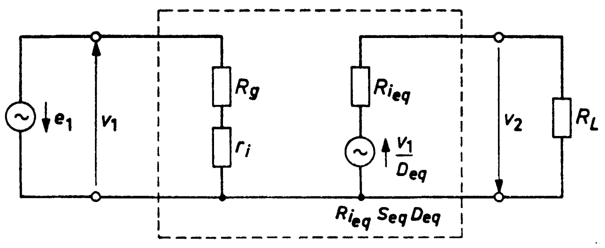


Fig. 8.2

If we assume the control generator to be without internal resistance, the input voltage is then $v_1 = e_1$. The source resistance R_g of the control generator is now connected in series with the input resistance r_t and thus added to the actual transistor. With this we obtain the equivalent transistor indicated in Fig. 8.2, whose characteristic values must be changed. The equivalent characteristic values S_{eq} $r_{t_{\text{eq}}}$ and D_{eq} are obtained as follows: The slope was $S = \beta/r_{t_0}$. If a resistance R_g is now connected in series with the input resistance, the input current i_1 and consequently the output current i_2 will be reduced for the same input voltage v_1. The slope S_{eq} is therefore similarly reduced and we obtain the new apparent slope from:

Working slope

$$S_{\text{eq}} = \frac{\beta_0}{r_{t_0} + R_g} = \frac{h_{21}}{h_{11} + R_g}. \tag{8.2}$$

Corresponding alterations have to be made in the expressions for R_t and D. The resistance R_g must be added to the input resistance r_{t_0} in these relations also. The following equations then apply for the

Working internal resistance:

$$R_{i_{eq}} = \frac{1}{\frac{1}{R_i^*} - \frac{D_v \cdot \beta_0}{r_{i_0} + R_g}} = \frac{1}{h_{22} - \frac{h_{12} \cdot h_{21}}{h_{11} + R_g}} \tag{8.3}$$

and for the
Reciprocal working amplification:

$$D_{eq} = \frac{1}{S_{eq} \cdot R_{i_{eq}}} = \frac{r_{i_0} + R_g}{R_i^* \cdot \beta_0} - D_v. \tag{8.4}$$

With these equivalent characteristic values, all the formulae are valid from now on, as for the transistor with its characteristic values S, R_i and D (see Chapter 4), except for the equation for the input resistance r_i which here runs:
Input resistance

$$r_{i_{eq}} = r_i^* + R_g = \frac{r_i}{1 + A_v \cdot D_v} + R_g. \tag{8.5}$$

To conclude the investigations into the apparent internal resistance $R_{i_{eq}}$ of the transistor, i.e. the "working output resistance" produced when a control generator is influenced by a source resistance which cannot be disregarded, it must again be pointed out that this need only be represented if the load resistance R_L is variable because only then does the reaction on the input resistance occur. The source resistance R_g of a control generator is again only effective as the result of a variable input voltage. For a specific load resistance a transistor circuit can always be calculated with the transistor characteristic values S, R_i and D even if this load resistance is afterwards altered. It is then sufficient, for instance in the case of the resonance circuit, to calculate the apparent resistance $R_{i_{eq}}$ which, as a damping resistance is in parallel to the resonance circuit. The numerical relationship can be found from the following example.

Example 8.2

We have a transistor T_1 with the known data:

$$r_{i_0} = h_{11} = 0{,}8\ \text{k}\Omega \qquad D_v = h_{12} = 5.4 \cdot 10^{-4}$$

$$\beta = h_{21} = 47 \qquad \frac{1}{R_i^*} = h_{22} = 80 \cdot 10^{-6}$$

$$S = 58.8\ \text{mA/V} \quad R_i = 20.8\ \text{k}\Omega \quad D = 8.15 \cdot 10^{-4}$$

This transistor is to be loaded with $R_L = 2$ kΩ. A control generator with $R_g = 2$ kΩ supplies an initial voltage $e_1 = 1$ mV.

Required

1. All the input and output data for constant R_L and the voltage equivalent circuit.
2. All the input and output data for variable R_L and the voltage equivalent circuit.

Solution to 1

$$A_v = S \cdot \frac{R_i \cdot R_L}{R_i + R_L} = 58.8 \frac{20.8 \cdot 2}{20.8 + 2} = 107$$

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v} = \frac{0.8}{1 + 107 \cdot 5.4 \cdot 10^{-4}} = 0.76 \text{ k}\Omega$$

$$v_1 = e_1 \frac{r_i}{r_i + R_g} = 1 \frac{0.76}{2 + 0.76} = 0.275 \text{ mV}$$

$$i_1 = \frac{v_1}{r_i} = \frac{0.275}{0.76} = 0.362 \ \mu\text{A}$$

$$v_2 = v_1 \cdot A_v = 0.275 \cdot 107 = 29.4 \text{ mV}$$

$$i_2 = \frac{v_2}{R_L} = \frac{29.4}{2} = 14.7 \ \mu\text{A}$$

$$A_i = A_v \frac{r_i}{R_L} = 107 \frac{0.76}{2} = 40.6$$

$$A_p = A_v \cdot A_i = 107 \cdot 40.6 = 4350$$

$$\frac{v_1}{D} = \frac{0.275}{8.15} \cdot 10^4 = 338 \text{ mV.}$$

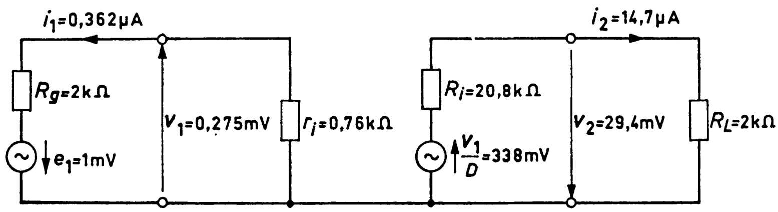


Fig. 8.3

Solution to 2

$$R_{i_{eq}} = \frac{1}{h_{22} - \dfrac{h_{12} \cdot h_{21}}{h_{11} + R_g}} = \frac{1}{\dfrac{80}{10^6} - \dfrac{5.4 \cdot 10^{-4} \cdot 47}{0.8 \cdot 10^3 + 2 \cdot 10^3}} = 14\ \text{k}\Omega$$

$$S_{eq} = \frac{\beta}{r_{i_0} + R_g} = \frac{47}{0.8 + 2} = 16.8\ \text{mA/V}$$

$$D_{eq} = \frac{1}{R_{i_{eq}} \cdot S_{eq}} = \frac{1}{14 \cdot 10^3 \times 16.8 \cdot 10^{-3}} = 42.5 \cdot 10^{-4}$$

$$A_{v_{eq}} = S_{eq} \frac{R_{i_{eq}} \cdot R_L}{R_{i_{eq}} + R_L} = 16.8 \frac{14 \cdot 2}{14 + 2} = 29.4$$

$$i_1 = \frac{e_1}{r_i + R_g} = \frac{e_1}{r'_i} = \frac{10^{-3}}{(0.76 + 2) \cdot 10^3} = 0.362\ \mu\text{A}$$

$$v_2 = e_1 \cdot A'_v = 29{,}4\ \text{mV}$$

$$i_2 = \frac{v_2}{R_L} = \frac{29.4}{2} = 14.7\ \mu\text{A}$$

$$A_{i_{eq}} = A'_v \frac{r'_i}{R_L} = 29.4 \frac{0.76 + 2}{2} = 40.6 = A_i$$

$$A_{p_{eq}} = A'_v \cdot A'_i = 29.4 \times 40.6 = 1190$$

$$\frac{v_{1_{eq}}}{D_{eq}} = \frac{e_1}{D_{eq}} = \frac{10^4}{42.5} = 235\ \text{mV}.$$

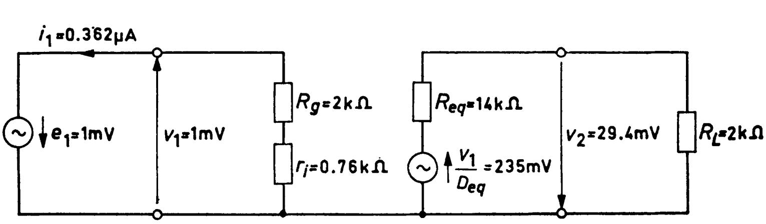


Fig. 8.4

CHAPTER 9

TRANSISTORS WITH NEGATIVE FEEDBACK

Negative feedback is a means much used with the tube to influence and stabilise the amplifying properties of an amplifier. Even greater use is made of these methods with transistors because the transistor is extremely subject to external influences; its dependence on temperature is an example. With negative feedback, as the name implies, part of the amplified a.c. voltage $\alpha \cdot v_2$ is allowed to operate against the applied input voltage v_1. In principle, therefore, we can distinguish between a series and a parallel coupling.

We have "series feedback" when the input voltage is in series with the feedback voltage (only applies with voltage control: see 9.11). In parallel feedback the input voltage has the feedback voltage at the input resistance superimposed on it; they operate in "parallel".

In the tube we usually refer to "current and voltage feedback". Current feedback occurs when a cathode resistance is connected in the circuit. Here the cathode current causes a voltage drop which as a feedback voltage is in series with the input voltage (series feedback). Voltage feedback takes place in the tube when a feedback voltage is fed back from the anode to the tube input.

Because of the high input resistance, it is possible in the tube to effect

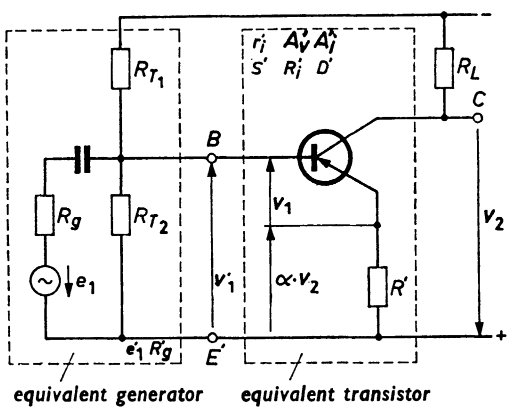


Fig. 9.1

the "voltage feedback" in series or parallel, though the parallel feedback predominates. As a result of its low input resistance, the transistor in the corresponding circuit, by returning the feedback voltage from the collector to the base, is more suitable for parallel feedback. Series feedback from collector to base is difficult to realise and will therefore not be further dealt with. We therefore only refer in the transistor to series and parallel feedback. The two forms of feedback will be treated in the following sections in such a way that an equivalent transistor with changed characteristic values is formed.

9.1. Series feedback

The series feedback is obtained if a resistance R' is inserted in the lead from the emitter to the "earth" line of the circuit. The base will have a positive bias because of the feedback resistance R'. However, as a negative bias is required, an over-compensation of this positive bias has to be produced through the voltage divider $R_{T_{1/2}}$, so that a negative bias results. The feedback voltage is $\alpha \cdot v_2$ which is connected in series with the transistor input voltage v_1. The two voltages supply the new input voltage v'_1 at the equivalent transistor between B and E'.

To calculate the equivalent characteristic values of the transistor with series feedback, we use the correct equivalent circuit for the control source and the transistor.

Control source equivalent circuit

We have a control source equivalent circuit based on the points B and E' which includes the voltage divider resistances $R_{T_{1/2}}$. The equivalent initial voltage is equal to the open circuit voltage between B and E' (transistor not yet connected) whereby $R_{T_1} \parallel R_2$ ($\parallel$ parallel) appears as a load and the following equation applies:

$$e'_1 = e_1 \frac{R_{T_1} \parallel R_{T_2}}{R_g + (R_{T_1} \parallel R_{T_2})}.$$

The new equivalent source resistance is equal to the resistance which lies between B and E' when e_1 is short-circuited. We then have:

$$R'_g = (R_{T_1} \parallel R_{T_2}) \parallel R_g = \frac{1}{\dfrac{1}{R_{T_1}} + \dfrac{1}{R_{T_2}} + \dfrac{1}{R_g}}$$

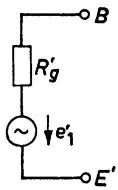


Fig. 9.2

The control source therefore has the following equivalent circuit (Fig. 9.2).

Transistor equivalent circuit

In the transistor fundamental equivalent circuit the feedback resistance R' is inserted on the output side and connected to the input circuit. The following equivalent circuit is produced as a result (Fig. 9.3).

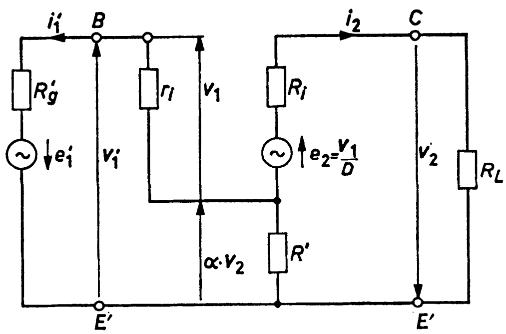


Fig. 9.3

Part of the output voltage $\alpha \cdot v_2$ acts as feedback voltage. This feedback is produced as the voltage drop of the current i_2 at the parallel circuit $R' \parallel (R'_g + r_i)$ if we disregard the very small base current, while v_2 decreases at the resistance R_L lying in series. Since the voltage in the series circuit behaves like the resistances, the feedback factor is found from:

$$\alpha = \frac{v_{fb}}{v_2} = \frac{R' \parallel (R'_g + r_i)}{R_L}.$$

As in most cases $R' < (r_i + R'_g)$ we obtain for the:

feedback factor

$$\alpha \approx \frac{R'}{R_L}. \tag{9.1}$$

For the correction at the input of the transistor equivalent circuit we get the equation:

$$v_1' - v_1 - \alpha \cdot v_2 = 0$$

or

$$v'_1 = v_1 + \alpha \cdot v_2.$$

With this we obtain the new input resistance r'_i as follows:

$$r'_i = \frac{v'_1}{i'_1} \quad \text{in which} \quad i'_1 = \frac{v_1}{r_i}$$

and consequently

$$r'_i = \frac{v'_1}{v_1} \cdot r_i = r_i \frac{v_1 + \alpha \cdot v_2}{v_1} = r_i \frac{v_1 + v_1 \cdot A_v \cdot \alpha}{v_1}.$$

Input resistance

$$r'_i = r_i(1 + \alpha \cdot A_v). \tag{9.2}$$

The voltage amplification A'_v for the equivalent transistor comes from:

$$A'_v = \frac{v_2}{v'_1} = \frac{v_2}{v_1 + \alpha \cdot v_2} = \frac{v_2}{v_1 + \alpha \cdot A_v \cdot v_1} = \frac{v_2/v_1}{1 + \alpha \cdot A_v}.$$

Voltage amplification

$$A'_v = \frac{A_v}{1 + \alpha \cdot A_v}. \tag{9.3}$$

If we investigate the new current amplification the following expression must apply:

$$A'_i = A'_v \frac{r'_i}{R_L} = \frac{A_v}{1 + \alpha \cdot A_v} \frac{r_i(1 + \alpha \cdot A_v)}{R_L} = A_v \frac{r_i}{R_L} = A_i.$$

Current amplification:

$$A'_i = A_i. \tag{9.4}$$

The current amplification is not altered with series feedback. The new equivalent characteristic values S', R'_i and D' for the equivalent transistor, as represented in Fig. 9.1, are obtained as follows: The new static equivalent slope S' must be given for the case of short-circuit on the output side, i.e. $R_L = 0$.

The following then applies:
$S' = i_2/v'_1$ in which $i_2 = v_1 \cdot S$ and therefore $S' = S(v_1/v'_1)$.

Now

$$v'_1 = v_1 + \alpha \cdot A_v \cdot v_1 = v_1 + v_{fb} = v_1 + i_2 \cdot R' = v_1 + v_1 \cdot S \cdot R'.$$

In consequence we now have

$$S' = S\frac{v_1}{v'_1} = S\frac{v_1}{v_1 + v_1 \cdot S \cdot R'} \qquad \text{or}$$

Equivalent slope

$$S' = \frac{S}{1 + S \cdot R'}. \tag{9.5}$$

The internal resistance R'_i is obtained according to the following considerations: The input voltage v'_i is set at zero, i.e. the input is short-circuited for alternating current. r_i then lies in parallel to R' and the only input voltage is now the feedback voltage. Instead of the load resistance a generator with negligible internal resistance is connected at the output at the same collector open-circuit current (the same working point), supplying an initial voltage e_2. As a result of this initial voltage, a current i_2 flows which produces at R' the input voltage $v_i = i_2 \cdot R'$, which wants to drive a current in the opposite direction to i_2 as feedback voltage. We therefore assume that R' is less than r_i. We can now apply the following equivalent circuit produced from the normal transistor equivalent circuit: (Fig. 9.4).

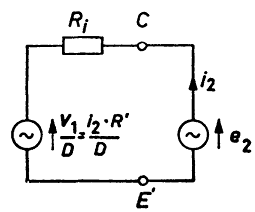


Fig. 9.4

From the equivalent circuit we obtain:

$$i_2 = \left(e_2 - \frac{i_2 \cdot R'}{D}\right)\frac{1}{R_i} = \frac{e_2}{R_i} - \frac{i_2 \cdot R'}{D \cdot R_i} = \frac{e_2}{R_i} - i_2 \cdot S \cdot R'$$

$$i_2(1 + SR') = \frac{e_2}{R_i}; \qquad i_2 = e_2\frac{1}{R_i(1 + S \cdot R')}.$$

From this now emerges the

Equivalent internal resistance

$$R'_i = \frac{e_2}{i_2} = R_i(1 + S \cdot R'). \tag{9.6}$$

The internal resistance increases with series feedback. The *equivalent amplifi-*

cation reciprocal is obtained, according to Barkhausen, from:

$$D' = \frac{1}{R'_i \cdot S'} = D. \tag{9.7}$$

The amplification reciprocal is not changed with series feedback. The following equivalent circuits can now be given for the equivalent transistor:

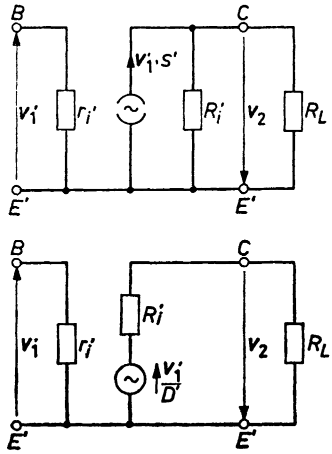


Fig. 9.5

Example 9.1

A transistor T_1 is operated with series feedback as the following circuit shows. The output voltage v_2 should amount to 1 V.

Required:
All transistor equivalent characteristic values and all a.c. input and output current values.

The transistor data are:

$$S = 58.8 \text{ mA/V} \quad R_i = 20.8 \text{ k}\Omega \quad D = 8.15 \cdot 10^{-4}$$
$$r_{t_0} = 0.8 \text{ k}\Omega \quad D_v = 5.4 \cdot 10^{-4}$$

The capacitances in Fig. 9.6 are so large that they represent short circuits for alternating current. The voltage divider resistances should be added to the control generator to produce base bias. The equivalent control generator then has an equivalent internal resistance R'_g equal to the resistance which can be measured against earth with short-circuited initial voltage e_1 at terminal *B*. R_g, R_{T_1} and R_{T_2} are then connected in parallel and

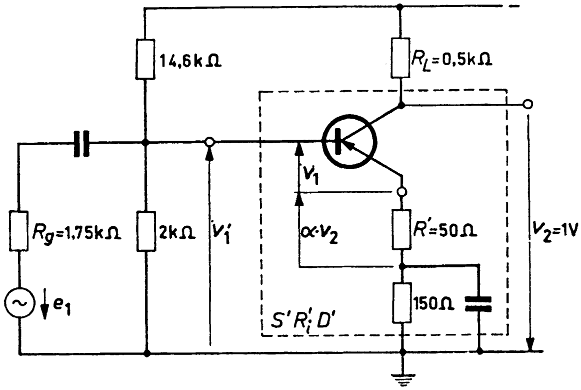


Fig. 9.6

$$R'_g = \frac{1}{\frac{1}{R_g} + \frac{1}{R_{T_1}} + \frac{1}{R_{T_2}}} = \frac{1}{\frac{1}{1.75} + \frac{1}{2} + \frac{1}{14.6}} = 0.875 \text{ k}\Omega.$$

The voltage amplification of the transistor is:

$$A_v = S\frac{R_i \cdot R_L}{R_i + R_L} = 58.8\frac{20.8 \times 0.5}{20.8 + 0.5} = 28.8.$$

The normal input resistance is:

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v} = \frac{0.8}{1 + 28.8 \times 5.4 \cdot 10^{-4}} = 0.79 \text{ k}\Omega.$$

The equivalent characteristic values for the transistor with negative feedback are given as follows:

$$\alpha = \frac{R' \parallel (r_i + R'_g)}{R_L} \approx \frac{50}{500} = 0.1$$

$$r'_i = r_i(1 + \alpha \cdot A_v) = 0.79(1 + 0.1 \times 28.8) = 3.06 \text{ k}\Omega$$

$$A'_v = \frac{A_v}{1 + \alpha \cdot A_v} = \frac{28}{1 + 0.1 \times 28.8} = 7.4$$

$$S' = \frac{S}{1 + S \cdot R'} = \frac{58.8}{1 + 58.8 \times 0.05} = 15 \text{ mA/V}$$

$$R'_i = R_i(1 + S \cdot R') = 20.8 \times 3.94 = 82 \text{ k}\Omega$$

$$D' = D = 8.15 \cdot 10^{-4}$$

$$A'_i = A_i = A_v\frac{r_i}{R_L} = 28.8\frac{0.79}{0.5} = 45.6.$$

With the equivalent characteristic values, the voltage amplification of the equivalent transistor is obtained from:

$$A'_v = S' \frac{R'_i \cdot R_L}{R'_i + R_L} = 15 \frac{82 \times 0.5}{82 + 0.5} = 7.4.$$

For the working values we get:

$$i_2 = \frac{v_2}{R_L} = \frac{1}{0.5} = 2 \text{ mA}$$

$$v'_1 = \frac{v_2}{A'_v} = \frac{1}{7.4} = 135 \text{ mV}$$

$$i_1 = \frac{i_2}{A_i} = \frac{2}{45.6} = 43.8 \ \mu\text{A}$$

$$= \frac{v'_1}{r'_i} = \frac{135}{3.06} = 43.8 \ \mu\text{A}$$

$$A'_p = A'_v \cdot A'_i = \frac{v_2 \cdot i_2}{v_i \cdot i_1} = \frac{1 \times 2 \times 10^{-3}}{0.135 \times 43.8 \cdot 10^{-6}} = 340.$$

The initial voltage e'_1 of the equivalent control generator is:

$$e'_1 = v'_1 \frac{R'_g + r'_i}{r'_i} = 135 \frac{3.06 + 0.875}{3.06} = 173 \text{ mV}.$$

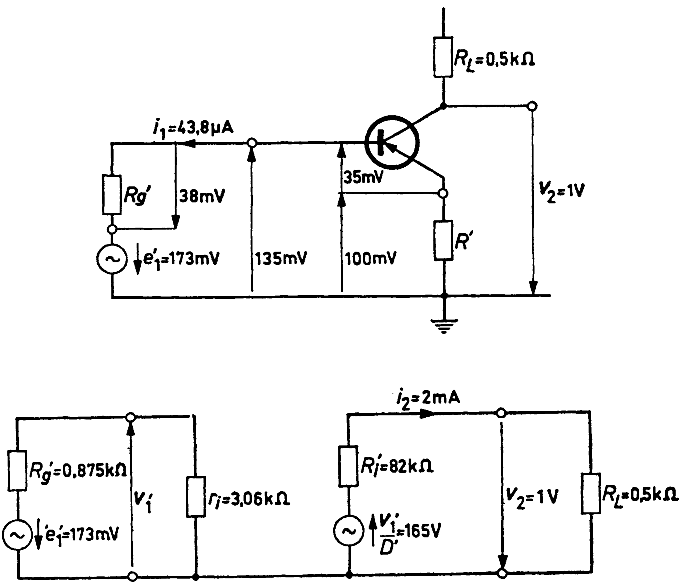


Fig. 9.7

As control voltage v_1 between base and emitter we have:

$$v_1 = \frac{v_2}{A_v} = \frac{1}{28.8} = 34.6 \text{ mV} \approx 35 \text{ mV}.$$

For the feedback voltage we obtain:

$$v_{\text{fb}}' = \alpha \cdot v_2 = 0.1 \times 1 = 0.1 \text{ V}.$$

The following equivalent circuits (Fig. 9.7) therefore apply. If we investigate the initial voltage e_1 in the actual circuit, we find that e'_1 is the open circuit voltage at the unloaded voltage divider. e'_1 thus lies at the parallel circuit of 2 || 14.6 kΩ; this resistance is

$$R_p = \frac{2 \times 14.6}{2 + 14.6} = 1.76 \text{ k}\Omega.$$

In this way,

$$e_1 = e'_1 \frac{R_p + R_g}{R_p} = 173 \frac{1.76 + 1.75}{1.76} = 346 \text{ mV}.$$

9.1.1. VOLTAGE CONTROL; INFLUENCE OF THE CONTROL GENERATOR

It is evident from Equation (9.3) for voltage amplification, that the effective amplification A'_v is only changed a little with a variation of the normal transistor amplification A_v caused by altered characteristic values. Voltage "stabilisation" occurs as a result of the series feedback. This stabilisation is only effective, however, when the transistor is voltage controlled. We have voltage control if the internal resistance R_g of the control generator is small in comparison with the effective input resistance r'_i because only then is the input voltage $v'_i \approx e_1$ independent of variations of the input resistance, according to Equation (9.2), with variable amplification A_v. We can also have voltage control when a naturally high resistance generator is artificially made low in value through a small load resistance, for instance, and through the voltage divider for the base voltage (see 9.1); an equivalent internal resistance R'_g is thus produced.

If there is no voltage control, that is if the control generator resistance R_g is no longer negligible, the voltage stabilisation deteriorates as can be understood from the following consideration. If we proceed from the constant initial voltage e_1 of the control generator, (Fig. 9.8), the transistor input voltage

$$v'_1 = e_1 \frac{r'_i}{r'_i + R_g}$$

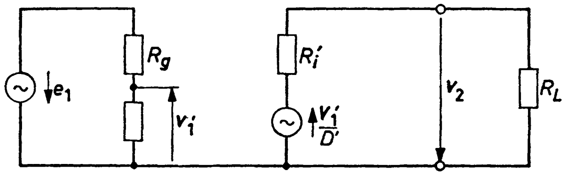


Fig. 9.8

corresponds to the voltage division between r'_i and R_g. The output voltage v_2 is therefore

$$v_2 = v'_1 \cdot A'_v = e_1 \frac{r_i(1 + \alpha \cdot A_v)}{r_i(1 + \alpha \cdot A_v) + R_g} \cdot \frac{A_v}{1 + \alpha \cdot A_v} \tag{9.8}$$

$$= e_1 \cdot S \cdot A_v'$$

and the total voltage amplification is based on the initial generator voltage e'_1

$$A_{v_{\text{tot}}} = \frac{v_2}{e_1} = S \cdot A'_v.$$

Therefore, if the amplification A_v varies with constant initial control voltage, v_2 will only vary with A'_v if R_g is $< r'_i$, because then the reduction factor

$$S = \frac{r_i(1 + \alpha \cdot A_v)}{r_i(1 + \alpha \cdot A_v) + R_g} \tag{9.9}$$

is constantly equal to 1. In the extreme case R_g could be $> r'_i$ (current control).

Then in Equation (9.8)

$$S \approx \frac{r_i(1 + \alpha \cdot A_v)}{R_g}$$

and the value $(1 + \alpha \cdot A_v)$ diminishes by A'_v, i.e. V_2 varies exactly as much as A_v and there is no longer stabilisation of the voltage amplification. In all other cases between the two extremes assumed we can calculate the amount of the stabilisation effect from Equation (9.8).

With a variation of the load resistance, the internal resistance R'_i operates according to Eq. (9.6) as long as there is voltage control ($R_g < r'_i$).

As we can see, the series feedback only reaches full effectiveness if the control generator is of low resistance compared with the input resistance r'_i. This fact can also be explained as follows:

The voltage v_1 between base and emitter is always decisive for the control of a transistor. If we want to alter anything through feedback, the voltage

v_1 between base and emitter also has to be changed through the feedback voltage. Fig. 9.3 shows that according to the principle of superposition, the voltage drop v_1 is produced from two voltage drops as the result of two initial voltages e' and e_2. One is produced through e'_1 and the other, in the opposite direction, through e_2. The feedback voltage effective at r_i is less than $\alpha \cdot v_2$. This is reduced through the voltage division between R_g and r_i according to the reduction factor

$$s' = \frac{r_i}{r_i + R_g}.$$

If therefore $R_g > r_i$, only a little of the feedback voltage $\alpha \cdot v_2$ becomes effective, i.e. with current control the "series feedback" is ineffective. It follows from this that in the transistor, the so-called "series feedback" is strictly speaking a parallel coupling because only the feedback voltage superimposed at r_i *) is decisive. In the electron tube R_g is usually $< r_i$ and therefore the feedback voltage at the cathode resistance is almost completely effective at the input of the electron tube, between grid and cathode. Consequently, we can consider this kind of feedback in the electron tube as a series feedback.

Example 9.2

In the feedback circuit in Example 9.1, the amplification A_v will be reduced by the factor $x = 0.7$, that is, to 70%.

How much percentage decrease would there be in amplification A'_v with pure voltage control and therefore complete stabilisation, and what is the actual value of this?

Given

$$A_v = 28.8; \qquad \alpha = 0.1; \qquad A'_v = \frac{A_v}{1 + \alpha \cdot A_v} = 7.4; \qquad r_i = 0.79 \text{ k}\Omega.$$

Solution

With pure voltage control R'_g would be $< r'_i$ and therefore $s' = 1$. In this case, the total amplification, based on constant generator initial voltage e'_1 is:

$$A_{v_{tot}} = s' \cdot A'_v = 1 \times 7.4 = 7.4.$$

If for any reason we now reduce the transistor amplification to $A_v \cdot x$, the total amplification becomes

$$A_{v_{tot\,x}} = s' \cdot A_{v\,x} = 1 \cdot \frac{A_v \cdot x}{1 + \alpha \cdot A_v \cdot x} = \frac{28.8 \times 0.7}{1 + 0.1 \times 28.8 \times 0.7} = 6.66.$$

*) Possibly weakened by s'

The percentage drop of variation of voltage amplification is thus:

$$p_{\%} = \left(1 - \frac{A_{v\,\mathrm{tot}_x}}{A_{v_{\mathrm{tot}}}}\right) \cdot 100 = \left(1 - \frac{6.66}{7.4}\right) \cdot 100 = 10\%.$$

Compared with a variation $A_v \cdot x$ of 30%, the amplification in this case with voltage control and series feedback is only reduced by 10%.

Actual case

The generator resistance R'_g of 0.875 kΩ is no longer negligible in comparison with the input resistance r_i of 0.79 kΩ. The total amplification based on e'_i is consequently

$$A_{v_{\mathrm{tot}}} = sA'_v = \frac{r_i(1 + \alpha \cdot A_v)}{r_i(1 + \alpha \cdot A_v) + R'_g} \cdot A'_v$$

$$= \frac{0.79(1 + 0.1 \times 28.8)}{0.79(1 + 0.1 \times 28.8) + 0.875} \cdot 7.4 = 5.8.$$

With a decrease in amplification to $A_v \cdot x$, we have:

$$s_x = \frac{r_i(1 + \alpha \cdot A_v \cdot x)}{r_i(1 + \alpha \cdot A_v \cdot x) + R'_g} = \frac{0.79(1 + 0.1 \times 28.8 \times 0.7)}{0.79(1 + 0.1 \times 28.8 \times 0.7) + 0.875}$$

$$= \frac{2.38}{2.38 + 0.875} = 0.72$$

and

$$A'_{v_x} = \frac{A_v \cdot x}{1 + \alpha \cdot A_v \cdot x} = \frac{28.8 \times 0.7}{1 + 0.1 \cdot 28.8 \cdot 0.7} = 6.66$$

The total amplification is thus only:

$$A'_{v_{\mathrm{tot}_x}} = s_x \cdot A'_{v_x} = 0.72 \times 6.66 = 4.8.$$

The total amplification has therefore dropped from 5.8 to 4.8. This corresponds to a percentage change of:

$$p\% = \left(1 - \frac{A_{v_{\mathrm{tot}_x}}}{A_{v_{\mathrm{tot}}}}\right) \cdot 100 = \left(1 - \frac{4.8}{5.8}\right) \cdot 100 = 17\%.$$

Compared with pure voltage control, the amplification loss is 7% greater and the stabilisation has deteriorated.

9.1.2. THE APPARENT TRANSISTOR CHARACTERISTIC VALUES WITH SERIES FEEDBACK

In the previous considerations the transistor has been mainly shown in series feedback, so that a specific input voltage was assumed. Through the derivation of the input resistance r'_i the influence of the control generator resistance could also be found.

The apparent internal resistance $R_{i_{eq}}$ or working output resistance, in the single transistor stage has already been mentioned. As a result of the feedback to the transistor input resistance with variable load resistance, this actual effective resistance $R_{i_{eq}}$ depends on the size of the control generator resistance. Even with series feedback we can add the control generator resistance R_g to the transistor. This equivalent transistor is then controlled from a generator whose internal resistance is nil. The input resistance from the control generator aspect no longer has any influence since the input voltage remains constant even when, for example, the load resistance is varied. With the resulting effective apparent transistor characteristic values $S_{v_{eq}}$, $R_{i_{eq}}$, D_{eq}, all working values for the equivalent transistor can then be calculated.

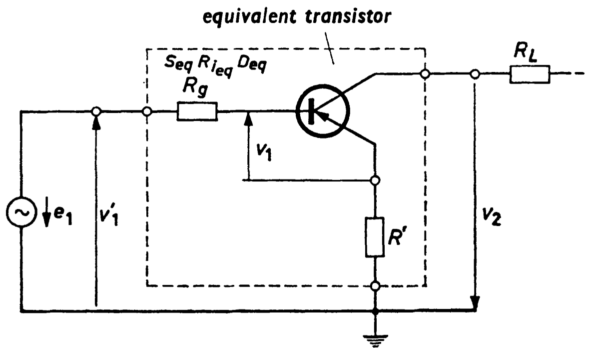


Fig. 9.9

The slope of the equivalent transistor is obviously less because only a small portion v_1 of the applied input voltage v'_1 is controlling. The input voltage lies at the series circuit of the resistances R_g, r_{i_0} and $R'(1 + \beta_0)$. The resistance R' appears increased by the factor $(1 + \beta_0)$ because the current in this resistance exceeds the input current by this factor. This control voltage v_1 appears as partial voltage at r_{i_0} and the equivalent slope S_{eq} is smaller corresponding to the voltage divider ratio. The following then applies:

$$S_{eq} = S \frac{r_{i_0}}{R_g + r_{i_0} + R'(1 + \beta_0)}. \tag{9.10}$$

Since $\beta_0 = S \cdot r_{i_0}$, and as rule $R_g + r_{i_0}$ is $< R'(1 + r_{i_0})$, we have:

$$S_{eq} \approx \frac{S \cdot r_{i_0}}{R' \cdot S \cdot r_{i_0}} \approx \frac{1}{R'} \tag{9.11}$$

i.e. in the extreme case, the slope of this stage is no longer dependent on the transistor characteristic values. The stabilisation can be found satisfactorily from the actual values which deviate from the extreme case.

The apparent internal resistance of the transistor R_{i_s} is obtained as follows:

To determine this resistance R_{i_s} a current i_2 is applied to the output with an initial voltage e_2. This current flows as $i_2 \cdot (1 + \beta_0)/\beta_0$, increased by the feedback resistance R'. The feedback voltage $V_{fb} = i_2(1 + \beta_0)/\beta_0 \cdot R'$ controls at the input resistance and is reduced according to the voltage division between R_g and r_{i_0}. In the corresponding equivalent circuit in Fig. 9.4, we have as a counteracting initial voltage

$$e_1 = \frac{v_{fb} \dfrac{r_{i_0}}{r_{i_0} + R_g}}{D} = \frac{i_2 \cdot R'}{D} \frac{r_{i_0}}{r_{i_0} + R_g}.$$

The current driven is now found from:

$$i_2 = \frac{e_2 - e_1}{R_i} = \frac{e_2}{R_i} - i_2 \frac{R'}{R_1 D} \frac{r_{i_0}}{r_{i_0} + R_g}$$

if we put $(1 + \beta_0)/\beta_0 \approx 1$.

By resolving according to i_2 we now have:

$$i_2 = \frac{e_2}{R_i\left(1 + S \cdot R' \dfrac{r_{i_0}}{r_{i_0} + R_g}\right)}.$$

Thus the apparent internal resistance $R_{i_{eq}}$ of the transistor is:

$$R_{i_{eq}} = \frac{e_2}{i_2} = R_i\left(1 + SR' \frac{r_{i_0}}{r_{i_0} + R_g}\right). \tag{9.12}$$

If $SR' > 1$ and $R_g < r_{i_0}$, then

$$R_{i_{eq}} \approx R_i \cdot S \cdot R'. \tag{9.13}$$

With a displacement of the working point, S', for example, is increased while R_i is decreased. It follows from this that even the effective internal resistance $R_{i_{eq}}$ of the equivalent transistor is stabilised.

The reciprocal amplification of the equivalent transistor is obtained from:

$$D_{eq} = \frac{1}{S_{eq} \cdot R_{t_{eq}}}. \tag{9.14}$$

Example 9.3

A resonance circuit is to be connected as load resistance R_L to a transistor T_1 with the known characteristic values $R_t = 20.8$ kΩ, $S = 58.8$ mA/V, $r_{t_0} = 0.8$ kΩ and a series feedback resistance $R' = 100$ Ω. The resonance resistance $R_0 = 100$ kΩ. The resonant circuit coil is tapped in the ratio 1 : 6 for matching to the transistor. The control generator has a resistance $R_g = 3$ kΩ. What is the effective R_D in parallel with R_0 as a damping resistance?

Solution

$$R_{t_{eq}} = R_t\left(1 + y_{te}R' \frac{r_{t_0}}{r_{t_0} + R_g}\right)$$

$$= 20.8\left(1 + 58.8 \times 10^{-3} \times 0.1 \times 10^3 \frac{0.8}{0.8 + 3}\right)$$

$$= 20.8 \times 2.24 = 46.5 \text{ k}\Omega$$

The effective damping resistance is therefore

$$R_D = R_{t_{eq}} v^2 = 46.5 \times 36 = 1.67 \text{ M}\Omega.$$

9.2. Parallel feedback

With parallel feedback the reverse coupling from the collector to the base is carried out through a resistance R'. The equivalent transistor with its terminals B, C, E, includes this resistance R' and is treated in effect like a normal transistor though naturally with different characteristic values. In the estimation the resistance in the emitter lead is only used for adjustment and stabilisation of the working point. It is short circuited for alternating current through the capacitance and therefore has no feedback (Fig. 9.10).

It is obvious from the circuit diagram that the input voltage v'_1 does not only produce the current i_1 which flows through the input resistance r_t. The feedback resistance R' carries an additional current i'_R and the total input current i'_1 is therefore $i_1 + i'_R$. The input resistance r'_t for the equivalent transistor has obviously decreased. The current i'_R in resistance R' is driven through a voltage which is obtained from:

$$v_{R'} = v'_1 + v_2 = v'_1(1 + A_v).$$

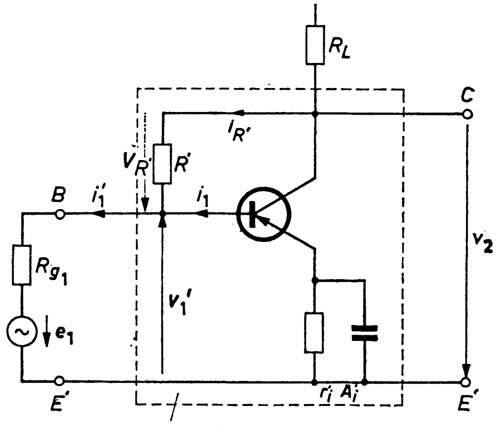


Fig. 9.10

If we consider the input voltage v'_1 as driving voltage for the current i'_R, we obtain the following relation:

$$i_{R'} = \frac{v'_1(1 + A_v)}{R'} = \frac{v'_1}{\dfrac{R'}{1 + A_v}}.$$

There is consequently at the voltage v'_1 a resistance $R'/(1 + A_v)$ connected in parallel with the transistor input resistance r_i. The new input resistance for the equivalent transistor is then found from:

$$r'_i = r_i \frac{\dfrac{R'}{1 + A_v}}{r_i + \dfrac{R'}{1 + A_v}}.$$

The following expression is also valid now:

$$r'_i = \frac{r_i \cdot R'}{(1 + A_v)\left(r_i + \dfrac{R'}{1 + A_v}\right)} = \frac{r_i \cdot R'}{r_i + A_v r_i + R'} = \frac{\dfrac{r_i \cdot R'}{r_i + R'}}{1 + \dfrac{r_i}{r_i + R'} \cdot A_v}.$$

If we now introduce the following value:

$$\alpha = \frac{v_{fb}}{v_2} = \frac{r_i}{r_i + R'} \qquad (9.15)$$

as feedback factor with open circuit on the input side (current control $R_g \gg r_i$) we then obtain the input resistance with parallel feedback from:

$$r'_i = \frac{r_i \parallel R'}{1 + \alpha \cdot A_v}. \qquad (9.16)$$

Since in most cases $R' > r_i$ we can set down as a close approximation:

$$r'_i \approx \frac{r_i}{1 + \alpha \cdot A_v}. \tag{9.17}$$

With this equivalent input resistance we find the input voltage from:

$$v'_1 = e_1 \frac{r'_i}{r'_1 + R_g}.$$

This input voltage is also at the transistor input resistance r_i and the output voltage v_2 is thus obtained with the old amplification factor A_v. We can therefore apply:

$$A'_v = A_v. \tag{9.18}$$

The voltage amplification does not vary with parallel feedback. Nevertheless, at a certain input voltage v'_1 which is required for a desired output voltage v_2, the input current i_1 increases by the factor $(1 + \alpha \cdot A_v)$ while the output current i_2 remains constant ($R_L \cdot i_2 = v_2$ has not changed). The current amplification $A'_i = i_2/i'_1$ has now become less and the following equation can be applied:

$$A'_i = A'_v \cdot \frac{r'_i}{R_L} = A_v \frac{1}{R_L} \frac{r_i \parallel R'}{1 + \alpha \cdot A_v};$$

$$A'_i = \frac{r_i \parallel R'}{R_L} \frac{A_v}{1 + \alpha \cdot A_v}, \tag{9.19}$$

so that

$$A'_i = \frac{r_i \cdot R'}{(r_i + R')R_L} \cdot \frac{A_v}{1 + \alpha \cdot A_v}.$$

If we put $A_v \cdot (r_i/R_L) = A_i$, we obtain:

$$A'_i = \frac{A_i}{1 + \alpha \cdot A_v} \frac{R_i}{r_i + R'} \approx \frac{A_i}{1 + \alpha \cdot A_v}. \tag{9.20}$$

Equation (9.19) shows that the current amplification is stabilised. Alteration of the characteristic values, which produces a change in voltage amplification and thus in current amplification, only causes very slight variation of the second term corresponding to the stabilisation factor $\alpha \cdot A_v$ (see 9.3). This stabilisation, however, only occurs with current amplification, i.e. with current control, as will be shown in the next section.

Equivalent Characteristic Values

The derivation of the working values for parallel feedback is based on a given input voltage v_1. We most therefore first assume a control generator with $R_g = 0$ because this given voltage v_1 is not influenced by the transistor. This applies to all the previous derivations (apart from the determination of the apparent characteristic values). The input resistance, for instance, was determined with these deductions and it was thus always possible to connect in and calculate a control generator in which R_g was not negligible.

The equivalent internal resistance R'_i now also holds good with constant input voltage v_1, that is with $R_g = 0$ and the input therefore short-circuited for alternating current. For an a.c. voltage v_2 supplied at the output, there is therefore a resistance available, the equivalent internal resistance of which is formed from the parallel connection of the normal R_i with the resistance R' earthed by the short circuit.

$$R'_i = R_i \parallel R'. \tag{9.21}$$

The given input voltage v_1 controls completely at the input resistance r_i and so the slope does not alter.

$$S' = S. \tag{9.22}$$

The equivalent reciprocal amplification is obtained from:

$$D = \frac{1}{S' \cdot R'_i} = D \frac{R_i + R'}{R'}. \tag{9.23}$$

It cannot be concluded from the expression $A_v/(1 + \alpha \cdot A_v)$ in Equation (9.19) that the voltage amplification is stabilised, for this is $A'_v = A_v$. If, however, we relate the voltage amplification to the generator voltage, the voltage amplification is also stabilised. Then e_1 yields the constant input current i_1 and the stabilised output current i_2 which once more results in a stabilised output voltage v_2. If, therefore, we consider the parallel reverse coupled transistor not at its input terminal but from the aspect of the initial control generator voltage e_1 a stabilisation of the voltage amplification corresponding to $A_v(1 + \alpha \cdot A_v)$ is also effective here (see also 9.21).

Example 9.4

A transistor T_1 is operated with parallel feedback (Fig. 9.11). The load resistance $R_L = 0.5$ kΩ is reduced through the feedback resistance $R' = 7.2$ kΩ because this appears as an extra load resistance in parallel with it.

All the operating values are to be calculated for an output voltage v_2 of 1 V. These values are already plotted in the circuit diagram in Fig. 9.11.

Given

$S = 58.8$ mA/V $R_i = 20.8$ kΩ $D = 8.15 \cdot 10^{-4}$

$r_{i_0} = 0.8$ kΩ $D_v = 5.4 \times 10^{-4}$.

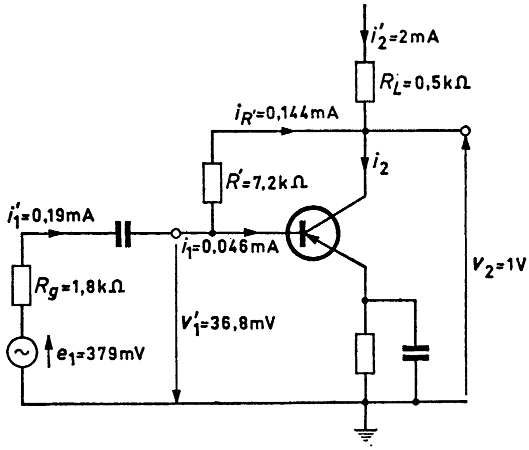


Fig. 9.11

Solution

For the transistor without parallel feedback we obtain:

Effective load resistance:

$$R'_L = \frac{R_L \cdot R'}{R_L + R'} = \frac{0.5 \times 7.2}{7.7} = 0.47 \text{ k}\Omega.$$

Voltage amplification:

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L} = 58.8 \frac{20.8 \times 0.47}{20.8 + 0.47} = 27.2.$$

Input resistance:

$$r_i = \frac{r_{i0}}{1 + A_v \cdot D_v} = \frac{0.8}{1 + 27.2 \times 5.4 \cdot 10^{-4}} = 0.79 \text{ k}\Omega$$

Current amplification

For the reverse coupled stage we find:

$$A_i = \frac{r_i}{R'_L} \cdot A_v = \frac{0.79}{0.47} \cdot 27.2 = 45.6.$$

Feedback factor:

$$\alpha = \frac{r_i}{R' + r_i} = \frac{0.79}{7.2 + 0.79} \approx 0.1.$$

Since R' is not negligible in comparison with r_i, the following values are obtained according to Eq. (9.16):

$$r'_i = \frac{r_i \parallel R'}{1 + \alpha \cdot A_v} = \frac{\frac{0.79 \times 7.2}{7.99}}{1 + 0.1 \times 27.2} = 0.19 \text{ k}\Omega$$

$$A'_i = \frac{r'_i}{R_L} \cdot A_{v'} = \frac{0.19}{0.47} \cdot 27.2 = 11$$

$$A'_v = A_v = 27.2$$

$$v'_1 = \frac{v_2}{A_v} = \frac{1}{27.2} = 36.8 \text{ mV} = v_1$$

$$i'_1 = \frac{v'_1}{r'_i} = \frac{36.8}{0.19} = 0.194 \text{ mA}$$

$$i'_2 = i'_1 \cdot A'_i = 0.194 \times 11 = 2.14 \text{ mA}$$

$$i_1 = \frac{v_1}{r_i} = \frac{36.8}{0.79} = 46.6\ \mu\text{A}$$

$$i_{R'} = \frac{v_1 + v_2}{R'} = \frac{1.0368}{7.2 \cdot 10^3} = 0.144 \text{ mA}$$

$$i'_1 = i_1 + i_{R'} = 0.0466 + 0.144 = 0.19 \text{ mA}$$

$$e_1 = v_1 + i'_1 \cdot R_{g_1} = 0.0368 + 0.19 \times 1.8 = 0.379 \text{ V}$$

$$A_p = \frac{v_2 \cdot i_2}{v'_1 \cdot i'_1} = \frac{1 \times 2 \cdot 10^{-3}}{36.8 \cdot 10^{-3} \times 0.19 \cdot 10^{-3}} = 286.$$

With the equivalent characteristic values we have:

$$R'_i = R_i \parallel R' = \frac{20.8 \times 7.2}{20.8 + 7.2} = 5.35 \text{ k}\Omega$$

$$S' = 58.8 \text{ mA/V}$$

$$D' = \frac{1}{R'_i \cdot S'} = \frac{1}{53.5 \text{ k}\Omega \times 58.8 \text{ mA/V}} = 3.18 \cdot 10^{-2}$$

$$A_v = S' \frac{R'_i \cdot R_L}{R'_i + R_L} = 58.8 \frac{5.35 \times 0.5}{5.35 + 0.5} = 27.$$

9.2.1. CURRENT CONTROL. INFLUENCE OF THE CONTROL GENERATOR

We can speak of current control when the current supplied from a control generator is only determined in volume by this generator. This control current is then independent of the instantaneous value of the transistor input resistance r'_i. This is obviously the case when the internal resistance of the control generator R_g is large in proportion to the input resistance r'_i. This condition can, of course, be produced artificially if a high resistance is connected in series with a low resistance generator.

Pure current control cannot always be realised. Stabilisation of current amplification as expressed by Eq. (9.19) then deteriorates. The amount of stabilisation in this case is found from the following:

With actual current control the input current is only decided by the initial voltage of the control generator, e_1, and through its internal resistance R_g.

$$i_{1_0} = \frac{e_1}{R_g}.$$

If the input resistance is no longer negligible, we obtain:

$$i_1 = \frac{e_1}{R_g + r'_i} = \frac{e_1}{R_g}\frac{R_g}{R_g + r'_i} = i_{1_0} \cdot s.$$

The input current is less than the desired control current i_{1_0}. The reduction of input current is the result of the weakening factor:

$$s = \frac{R_g}{R_g + \dfrac{r_i}{1 + \alpha \cdot A_v}}. \tag{9.24}$$

The weakened control current is amplified again, according to Eq. (9.20) by the reverse coupled transistor and we obtain:

$$A'_i = \frac{A_i}{1 + \alpha \cdot A_v} = \frac{r_i}{R_L}\frac{A_v}{1 + \alpha \cdot A_v}.$$

If the current amplification $A_{i_{tot}}$ is related to the control current i_{1_0} with $r'_i \ll R_g$, we get:

$$A_{i_{tot}} = s \cdot A'_i = \frac{i_1}{i_{1_0}} \cdot \frac{i_2}{i_1} = \frac{i_2}{i_{1_0}}.$$

By introducing the above expression we obtain:

$$A_{i_{tot}} = s \cdot A'_i = \frac{R_g}{R_g + \dfrac{r_i \parallel R'}{1 + \alpha \cdot A_v}} \cdot \frac{A_v}{1 + \alpha \cdot A_v}\frac{r_i}{R_L}. \tag{9.25}$$

The maximum control current i_{1_0} flows with pure current control if r'_C is $\ll R_g$. Then $s = 1$ and the current amplification $A_{i_{tot}} = A'_i$. Stabilisation of current amplification is then fully effective with $A_v/(1 + \alpha \cdot A_v)$ if there is a variation in the transistor characteristic values and consequently a change in A_v (see also 9.3.). In the extreme case R_g can become $< r'_i$ (voltage control). Eq. (9.25) then becomes

$$A_{i_{tot}} = \frac{R_g}{\dfrac{r_i}{1 + \alpha \cdot A_v}}\frac{A_v}{1 + \alpha \cdot A_v}\frac{r_i}{R_L} = \frac{R_g}{R_L} \cdot A_v.$$

There is no longer any stabilisation of current amplification since with a variation of A_v the current amplification also changes to the same degree. In all cases between pure current control and voltage control the stabilisation effect can be calculated from Eq. (9.25). If we investigate with current control the ratio v_2/e_1, i.e. the voltage amplification in relation to the initial control generator voltage, we find:

$$i_2 = i_{1_0} \cdot s \frac{A_v}{1 + \alpha \cdot A_v} \frac{r_i}{R_L}$$

in which

$$i_{1_0} = \frac{e_1}{R_g}$$

Now v_2 is also equal to $i_2 \cdot R_L$ and we obtain:

$$v_2 = \frac{e_1}{R_g} \cdot s \frac{A_v}{1 + \alpha \cdot A_v} \frac{r_i}{R_L}.$$

If the value from Equation (9.24) is inserted for s, we get:

$$v_2 = e_1 \frac{r_i}{R_g + \dfrac{r_i}{1 + \alpha \cdot A_v}} \frac{A_v}{1 + \alpha \cdot A_v}. \tag{9.26}$$

It is only with current control that $r_i/(1 + \alpha \cdot A_v)$ is $< R_g$ and can therefore be disregarded. The voltage amplification $A'_v = v_2/e_1$ is then stabilised according to the second fraction. With voltage control R_g would be $< r_i/(1 + \alpha \cdot A_v)$ and in this case R_g can be omitted in Equation (9.26). Then, however, $1 + \alpha \cdot A_v$ is also cut out of this expression and so there is no voltage stabilisation.

Example 9.5

In Example 9.4 we had $r'_i = 0.19$ kΩ and $R_g = 1{,}8$ kΩ. We now have to discover to what extent there is current control. We have to calculate the precentage drop in current amplification A'_i with a variation of amplification $A_v \cdot x$ when $x = 0.7$, for pure current control and for the actual case concerned with $R_g = 1.8$ kΩ.

Solution

With pure current control the following expression is valid according to Equation (9.19):

$$A'_i = \frac{r_i \parallel R'}{R'_L} \frac{A_v}{1 + \alpha \cdot A_v} = \frac{\dfrac{0.79 \times 7.2}{7.99}}{0.47} \cdot \frac{27.2}{1 + 0.1 \times 27.2} = 11.$$

With $x = 0.7$ we obtain:

$$A_{i'_x} = \frac{r_i \parallel R'}{R'_L} \frac{A_v \cdot x}{1 + \alpha \cdot A_v \cdot x} = 1.51 \frac{27.2 \times 0.7}{1 + 0.1 \times 27.2 \times 0.7} = 9.9.$$

The percentage decrease in current amplification is:

$$p\% = \left(1 - \frac{A'_{i_x}}{A_i}\right) \cdot 100 = \left(1 - \frac{9.9}{11}\right) \cdot 100 = 10\%.$$

The following calculation shows how little pure current control remains in the actual case.

The weakening of input current according to Equation (9.26) is:

$$s = \frac{R_g}{R_g + \dfrac{r_i \parallel R'}{1 + \alpha \cdot A_v}} = \frac{1.8}{1.8 + 0.192} = 0.9.$$

If the drop in amplification is $x = 0.7$, we get:

$$s_x = \frac{R_g}{R_g + \dfrac{r_i \parallel R'}{1 + \alpha \cdot A_v \cdot x}} = \frac{1.8}{1.8 + \dfrac{0.71}{2.9}} = 0.88.$$

The current amplifications based on the control current i_{1_0} are therefore:

$$A_{i_{\text{tot}}} = s \cdot A'_i = 0.9 \times 11 = 9.9$$

$$A_{i_{\text{tot}_x}} = s_x \cdot A'_{i_x} = 0.88 \times 9.9 = 8.7.$$

The actual percentage change in current amplification with 30% reduction of voltage amplification is:

$$p\% = \left(1 - \frac{A_{i_{\text{tot}_x}}}{A_{i_{\text{tot}}}}\right) \cdot 100 = \left(1 - \frac{8.7}{9.9}\right) \cdot 100 = 12\%.$$

Since the variation is only 2% greater than with pure current control, we can still speak of an almost pure current control.

9.2.2. THE APPARENT CHARACTERISTIC VALUES OF THE TRANSISTOR WITH PARALLEL FEEDBACK

The apparent transistor characteristic values are only of interest with variable load resistance. Here also, as in series feedback, an equivalent transistor can be visualised and added to the control generator resistance R_g. The transistor then has the apparent characteristic value $R_{i_{eq}}$, S_{eq}, D_{eq}; in the equivalent transistor there are the effective characteristic values.

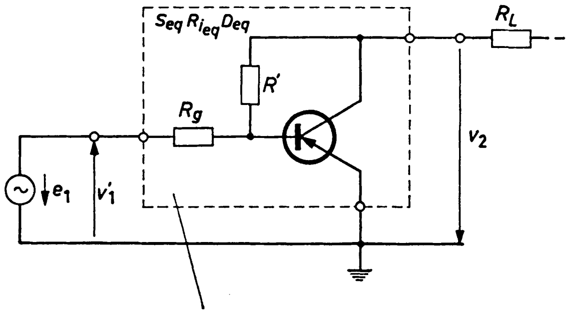


Fig. 9.12

The slope S_{eq} is obtained with the output short-circuited if R' is in parallel with the input resistance r_{i_0}. The apparent slope S_{eq} is less than the normal slope S because only a small part v_1 of the applied voltage v'_1 controls the transistor. Voltage division takes place in the series circuit of R_g with the parallel circuit of t_{i_0} and R'. We can therefore apply:

$$S_{eq} = S \frac{r_{i_0} \parallel R'}{(r_{i_0} \parallel R') + R_g}. \tag{9.27}$$

The apparent internal resistance $R_{i_{eq}}$ is also found here by taking the control generator resistance into account.
The voltage v_2 applied at the output drives the current i_2 and we have:

$$R_{i_{eq}} = \frac{v_2}{i_2}.$$

The current i_2 is mainly produced through the control voltage v_1 which drops through v_2 at the voltage divider R', $R_g \parallel r_{i_0}$. Here

$$v_1 = v_2 \frac{R_g \parallel r_{i_0}}{(R_g \parallel r_{i_0}) + R'}.$$

Therefore,

$$i_2 = v_1 \cdot S = v_2 \cdot S \frac{R_g \parallel r_{i_0}}{(R_g \parallel r_{i_0}) + R'}.$$

The remaining reaction of output voltage on the amount of output current, corresponding to R_i^* and D_v, can be disregarded as very small. We therefore obtain:

$$R_{i_{eq}} = \frac{v_2}{i_2} = \frac{1}{S} \frac{(R_g \parallel r_{i_0}) + R'}{R_g \parallel r_{i_0}}. \tag{9.28}$$

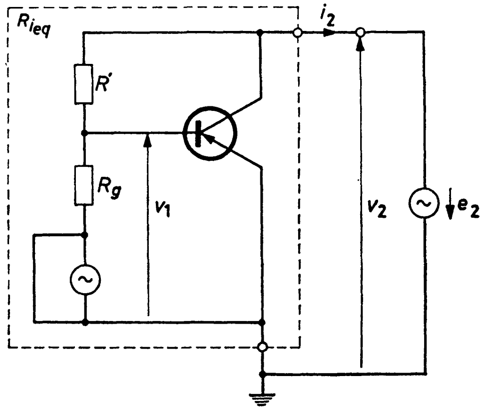


Fig. 9.13

The amplification reciprocal results from:

$$D_{\text{eq}} = \frac{1}{S_{\text{eq}} \cdot R_{t_{\text{eq}}}} = \frac{(r_{t_0} \parallel R') + R_g}{r_{t_0} \parallel R'} \frac{R_g \parallel r_{t_0}}{(R_g \parallel r_{t_0}) + R'}. \quad (9.29)$$

Since in general $R_g > r_{t_0} \parallel R'$, $p_t \parallel R' \approx r_{t_0}$, $R_g \parallel r_{t_0} \approx r_{t_0}$ we find that:

$$D_{\text{eq}} \approx \frac{R_g}{r_{t_0}} \frac{r_{t_0}}{r_{t_0} + R'},$$

$$D_{\text{eq}} \approx \frac{R_g}{r_{t_0} + R'}. \quad (9.30)$$

Example 9.6

The following apparent characteristic values are obtained for the stage with parallel feedback in Example 9.4.
We had:

$$R_g = 1.8\ \text{k}\Omega, \qquad R' = 7.2\ \text{k}\Omega,\ r_{t_0} = 0.8\ \text{k}\Omega,$$

$$S = 58.8\ \text{mA/V},\ e_1 = 379\ \text{mV},\ R_L = 0.5\ \text{k}\Omega.$$

Solution

$$S_{\text{eq}} = S \frac{r_{t_0} \parallel R'}{(r_{t_0} \parallel R') + R_g} = 58.8 \frac{0.72}{0.72 + 1.8} = 16.8\ \text{mA/V}$$

$$R_{t_{\text{eq}}} = \frac{1}{S} \frac{(R_g \parallel r_{t_0}) + R'}{R_g \parallel r_{t_0}} = \frac{10^3}{58.8} \frac{0.55 + 7.2}{0.55} = 240\ \Omega$$

$$D_{eq} = \frac{1}{S_{eq} \times R_{i_{eq}}} = \frac{10^3}{16.8 \times 0.24 \times 10^3} = 0.248$$

$$v_2 = e_1 \cdot S_{eq} \frac{R_{i_{eq}} \cdot R_L}{R_{i_{eq}} + R_L} = 0.379 \times 16.8 \times 10^{-3} \frac{0.24 \times 0.5}{0.24 + 0.5} 10^3$$

$$= 0{,}985 \approx 1 \text{ V.}$$

9.3. Stabilisation characteristic value

With series feedback we obtain stabilisation of voltage amplification when there is voltage control. Equation (9.3.) can be applied.

$$A'_v = \frac{A_v}{1 + \alpha \cdot A_v}.$$

Parallel feedback produces stabilisation of current amplification with current control, or, based on the initial control generator voltage, stabilisation of the voltage amplification also, according to the above Formula.

If the characteristic values are changed through an alteration or displacement of the working point, A_v, for instance, falls to $A_v \cdot x$ if x is < 1. The effectiveness of the stabilisation can now be learnt from the amount of variation in the effective amplification, $A'_v \cdot x'$, when the amplification alters by the factor x, x' being the reduction factor for the stabilised amplification.

There is thus an interesting connection between x' and x which can be represented graphically, and the parameters or values for a favourable shape for this curve, for it is naturally desirable that x' should be $> x$. The following formula applies for the stabilised amplification:

$$A'_{v_x} = \frac{A_v \cdot x}{1 + \alpha \cdot A_v \cdot x}. \tag{9.31}$$

Therefore

$$\frac{A'_{v_x}}{A'_v} = \frac{x(1 + \alpha \cdot A_v)}{1 + \alpha \cdot A_v \cdot x} = \frac{1 + \alpha \cdot A_v}{\frac{1}{x} + \alpha \cdot A_v}$$

or

$$A'_{v_x} = \frac{1 + \alpha \cdot A_v}{\frac{1}{x} + \alpha \cdot A_v} = A'_v \cdot x'. \tag{9.32}$$

The reduction factor x for the stabilised amplification A'_v with a reduction x of the normal amplification A_v, is therefore:

$$x' = \frac{1 + \alpha \cdot A_v}{\frac{1}{x} + \alpha \cdot A_v}. \tag{9.33}$$

It can be learnt from this formula that the stabilised amplification varies even less with x when:

$$\alpha \cdot A_v > \frac{1}{x}. \tag{9.34}$$

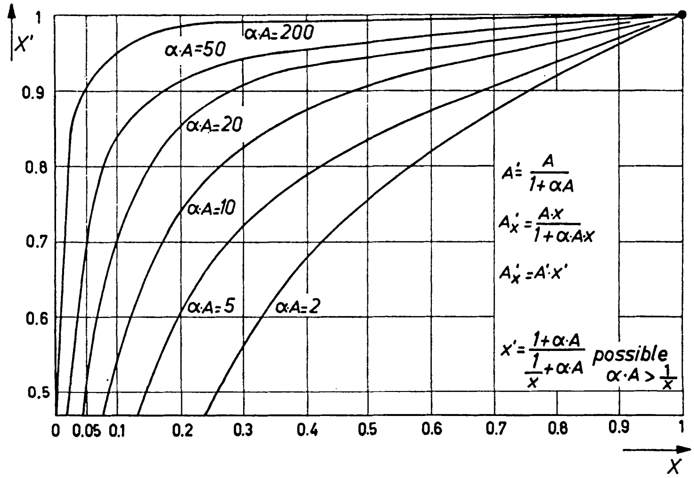


Fig. 9.14

We can therefore call the value $\alpha \cdot A_v$ the stabilisation characteristic value. The curves $x' = f(x)$ for different stabilisation characteristic values are plotted in Fig. 9.14. We see from these curves that with increasing stabilisation characteristic values the drop in stabilised amplification only occurs with greater reduction factors x.

Example 9.7

In Example 9.5, there is virtually current control. What is the stabilisation factor and how large is x' with $x_1 = 0.9$, $x_2 = 0.7$ and $x_3 = 0.5$?

Solution

$$\alpha \cdot A_v = 0.1 \cdot 27.2 = 2.72\ ;$$

$$x'_1 = \frac{1 + \alpha \cdot A_v}{\frac{1}{x_1} + \alpha \cdot A_v} = \frac{3.72}{\frac{1}{0.9} + 2.72} = 0.975$$

$$x'_2 = \frac{3.72}{\frac{1}{0.7} + 2.72} = 0.9$$

$$x'_3 = \frac{3.72}{\frac{1}{0.5} + 2.72} = 0.79$$

9.4. Non-linear distortions

The characteristics of a transistor are not straight lines. For this reason the characteristic values S, R_i and D are only valid for a specified working point with very little modulation. If a transistor is controlled with a larger sine voltage, the output current and the a.c. output voltage are no longer sinusoidal. In this case we speak of non-linear distortions because these arise at a non-linear characteristic. These distortions are smaller with current control than with voltage control since the characteristic for current amplification is only slightly curved. Output stages are amply modulated. Since the greatest distortions occur there, current control is mainly used for modulation (see 17.2.1.).

9.4.1. THE NON-LINEAR DISTORTION FACTOR

As we know, a non-sinusoidal voltage can be resolved into a fundamental wave of equal frequency (1st. harmonic) and a sum of sine voltages whose frequencies are multiples of the fundamental frequency (harmonic analysis). These multiples are also known as upper waves or higher harmonics. If the quadratic mean value is formed from the transient values of the fundamental wave (1st. harmonic) we obtain the effective value v_{FW} of the fundamental wave. If the quadratic mean value is formed from the transient value of the upper wave we obtain the effective value v_{UW} of the upper wave. The effective value of the non-sinusoidal voltage wave is found accordingly from the quadratic mean value of the transient values of this curve. However, since we already have the quadratic mean values for the fundamental wave v_{FW} and the upper waves v_{UW}, we obtain the effective value of the non-sinusoidal voltage from:

$$v_x = \sqrt{v_{FW}^2 + v_{UW}^2}. \tag{9.35}$$

To obtain an idea of the amount of non-linear distortions of this kind we form the ratio between the effective value of the upper wave and the effective value of the non-sinusoidal curve and call this the non-linear distortion factor.

$$d = \frac{v_{UW}}{v_x} = \frac{v_{UW}}{\sqrt{(v_{FW}^2 + v_{UW}^2)}}. \tag{9.36}$$

This factor can be calculated from the transmission characteristics or can be measured at the amplifier itself.

9.4.2. IMPROVING THE NON-LINEAR DISTORTION FACTOR

We are able to reduce the non-linear distortions of an amplifier by means of feedback. With series feedback the input voltage v'_1 is a sinusoidal voltage. The control voltage v_1 at the input resistance r_i is less than v'_i because of the feedback but it is sinusoidal all the same. As the amplification is now non-linear with large modulation, we obtain at the output a non-sinusoidal voltage:

$$v_{2_x} = v_1 \cdot A_v. \tag{9.37}$$

The fundamental wave v_{FW} of this output voltage operates with $\alpha \cdot v_{FW}$ against the input voltage v'_i and thus produces the control voltage v_1. In addition the upper wave voltage

$$v_{UW} = v_x \cdot d = v_1 \cdot A_v \cdot d \tag{9.38}$$

acts at the output as the impressed voltage.

This upper wave voltage, however, is no longer completely effective at the output load resistance because the feedback causes a weakening of the upper wave voltage through an opposing impressed voltage, the feedback voltage. Now only an upper wave voltage $v'_{UW} = v_1 \cdot A_v \cdot x$ is present at the load resistance. The voltage reacting inversely on the output circuit through the feedback is therefore:

$$(\alpha \cdot v_{UW}) \cdot A_v = (\alpha \cdot v_1 \cdot A_v \cdot x) \cdot A_v. \tag{9.39}$$

If we consider the fact that the sinusoidal fundamental wave at the output is obtained from Equation (9.35) as:

$$v_{FW} = \sqrt{(v_x^2 - v_{UW}^2)} = \sqrt{(v_x^2 - v_x^2 \cdot d^2)}$$

$$v_{FW} = \sqrt{[v_x^2(1 - d^2)]} = v_1 \cdot A_v \sqrt{(1 - d^2)} \tag{9.40}$$

we obtain the following equivalent circuit for the distortions in the output circuit (Fig. 9.15).

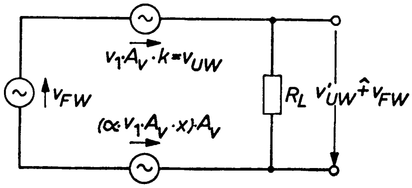


Fig. 9.15

The upper wave voltage resulting at the output is obtained from:

$$v_1 \cdot A_v \cdot x = v_1 \cdot A_v \cdot d - (\alpha \cdot v_1 \cdot A_v \cdot x)A_v$$

$$x = d - x \cdot \alpha \cdot A_v.$$

Therefore

$$x = \frac{d}{1 + \alpha \cdot A_v}. \qquad (9.41)$$

The new reduced non-linear distortion factor is consequently derived according to Equation (9.36) from:

$$d' = \frac{v'_{\mathrm{UW}}}{\sqrt{(v^2_{\mathrm{FW}} + v^2_{\mathrm{UW}})}}. \qquad (9.42)$$

If the values introduced in Fig. 9.15 are inserted, we thus get:

$$d' = \frac{v_1 \cdot A_v \cdot x}{\sqrt{[v_1^2 \cdot A_v^2(1 - d^2) + v_1^2 \cdot A_v^2 \cdot x^2]}} = \frac{x}{\sqrt{[(1 - d^2) + x^2]}} = \frac{1}{\sqrt{\left[\frac{1 - d^2}{x^2} + 1\right]}}.$$

If we insert the value found for x, this expression becomes:

$$d' = \frac{1}{\sqrt{\left[\frac{(1 + \alpha \cdot A_v)^2}{d^2}(1 - d^2) + 1\right]}} \approx \frac{d}{1 + \alpha \cdot A_v}. \qquad (9.43)$$

The non-linear distortion factor thus decreases in about the same measure as the amplification decreases through feedback. Therefore, this is only true of series feedback in the case of voltage control and of parallel feedback with pure current control.

If the non-linear distortion factor is improved in single stages through feedback, series connection of such stages will increase the factor for the whole amplifier because the percentage factor values of the individual stages are approximately added together. With a number of stages, therefore, it is better to undertake the feedback over the complete amplifier.

Example 9.8

It was established in Example 9.5 that there was virtually current control in the stage of Example 9.4. In the calculation of the non-linear distortion factor, the improvement through

$$1 + \alpha \cdot A_v = 1 + 0.1 \times 27.2 = 3.72$$

is consequently fully effective. If the non-linear distortion factor in this stage is $d = 5\%$ without feedback, there is a distortion factor of

$$d' \approx \frac{d}{1 + \alpha \cdot A_v} = \frac{5}{3.72} = 1.35\%$$

with feedback.

9.5. Working point stabilisation

The transistor characteristic values are strongly dependent on the working point, i.e. on the static voltages and currents, as will be described in Chapter 13. The working point selected and adjusted does not have to be absolutely stable, as we shall see in Chapter 10. Considerable working point displacements can occur through changes in the ambient temperature and through load variations. We now have a means of reducing displacements of the working point by making use of the feedback.

Adjustment of the working point is done by the base voltage divider (see Chapter 7). The criterion here is the static collector current which also determines the transistor dissipation. This collector current working point

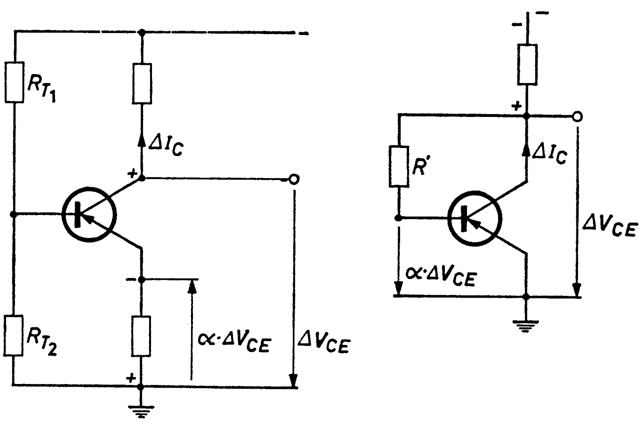


Fig. 9.16

above all should be kept constant. With a feedback which is not designed for a.c. voltage alone, as, for example, parallel feedback from the series connection of a capacitor with the resistance R', we obtain stabilisation of the d.c. voltage- or current working points as well as stabilisation of the amounts of alternating current.

The resulting voltage variations in series and parallel feedback are shown in Fig. 9.16 through an increase of ΔI_c in the collector current.
In both cases the increase of collector current causes a positive change ΔV_{CE} in the collector voltage and a similarly positive variation in the feedback voltage $\alpha \cdot \Delta V_{CE}$ at the base to the emitter. An alteration of the base voltage in the positive direction, however, means a reduction of the collector current, i.e. the actual increase of collector current will be less than it would be without feedback. If the variation of collector current or voltage without feedback is $\Delta I'_c$ or $\Delta V'_{CE}$ respectively, the variation reduced through feedback is obtained as follows:

The variation $\Delta V'_{CE}$ is decreased through the altered feedback voltage $\alpha \cdot \Delta V_{CE}$ which operates inversely with the amplification factor A_v. The following expression can therefore be applied for the reduced working point deviation:

$$\Delta V_{CE} = \Delta V'_{CE} - \alpha \, \Delta V_{CE} \cdot A_v.$$

If we resolve ΔV_{CE} we obtain:

$$\Delta V_{CE} = \frac{\Delta V'_{CE}}{1 + \alpha \cdot A_v}. \tag{9.44}$$

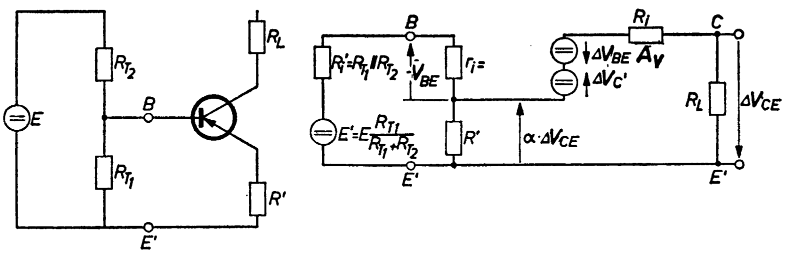


Fig. 9.17

This stabilisation, however, is only fully effective when the base voltage divider is made of low resistance with series feedback. As we also saw in Fig. 9.1 for a.c. voltages, the "generator resistance" for the base bias also counteracts the stabilisation with d.c. voltage feedback. This influence will be explained in an equivalent circuit (Fig. 9.17).

The feedback voltage $\alpha \cdot \Delta V_{CE}$ first appears at the feedback resistance and will operate as additional d.c. input voltage. However, only the voltage between base and emitter counts as d.c. input voltage. The feedback voltage reduced according to the voltage division acts at the input in exactly the same way as the bias e' (the open circuit voltage at the voltage divider) lies at the d.c. input resistance $r_{i=}$, reduced through the "internal resistance" of the voltage divider $R'_i = R_{T_1} \parallel R_{T_2}$. The reduction is:

$$s = \frac{r_{i=}}{r_{i=} + R_{T_1} \parallel R_{T_2}}. \tag{9.45}$$

The effect of the feedback voltage is reduced with this reduction factor. We can therefore imagine the feedback factor correspondingly reduced and with an insufficiently low resistance voltage divider we obtain the working point deviation from

$$\Delta V_c = \frac{\Delta V'_c}{1 + \alpha \cdot s \cdot A_v}. \tag{9.46}$$

Example 9.9

A transistor T_1 is connected as shown in Fig. 9.18. By what percentage will collector current variations be reduced?

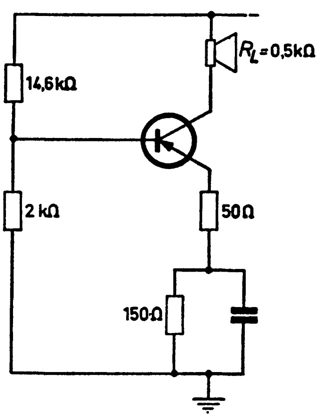


Fig. 9.18

$$S = 58.8 \text{ mA/V}; \quad r_{i=} = 2 \text{ k}\Omega;$$

$$A_v = S \cdot R_L = 29.9; \quad R' = 50 + 150 = 200 \ \Omega;$$

$$\alpha_{=} = \frac{R'}{R_L} = \frac{200}{500} = 0.4$$

$$R_{T_1} \parallel R_{T_2} = \frac{2 \times 14.6}{2 + 14.6} = 1.75\ \text{k}\Omega$$

$$s = \frac{r_{i=}}{r_{i=} + R_{T_1} \parallel R_{T_2}} = \frac{2}{2 + 1.75} = 0.534$$

$$p_{\%} = \frac{100}{1 + \alpha \cdot s \cdot A_v} = \frac{100}{1 + 0.4 \times 0.534 \times 29.9} = 13.5\%\,.$$

CHAPTER 10

THE INFLUENCE OF TEMPERATURE ON THE TRANSISTOR

In general the electron tube is insensitive to external temperature fluctuations. The reason for this is that the working temperatures of the electrodes, e.g. the cathode and the anode, lie far above the ambient temperature. The temperature gradient is therefore changed little relative to the variable external temperature. The tube characteristic values are thus not affected by temperature. Things are different with the transistor. The limit of permissible internal heating is 75 °C for germanium transistors and 130 °C for silicon transistors. If the limits are exceeded, the transistor becomes unserviceable The transistor internal temperature, the junction temperature T_j depends to a very great degree on the external or ambient temperature T_{amb} and on the conveyed dissipation which as current heat causes the transistor to heat up.

10.1. Permissible transistor dissipation

It is not possible to state a permissible dissipation for a particular transistor type as it is with the tube. This is because a maximum temperature difference $\Delta T_{max} = T_j - T_{amb_{max}}$ is required to yield a maximum dissipation. Since the highest junction temperature for germanium T_j is 75 °C, the influence of the ambient temperature on any possible temperature difference is thus very great. The maximum dissipation therefore depends on the maximum estimated ambient temperature $T_{amb_{max}}$.

In calculating the permissible transistor dissipation we must remember that the conveyed dissipation may only be equal to the dissipation carried off.

$$P_C = P_{off} \tag{10.1}$$

If the conveyed dissipation P_C were constantly greater than the energy carried off at the time, a continual temperature rise in the crystal would destroy it.

The conveyed dissipation depends on the adjustment of the working point. If we disregard the base-emitter dissipation which is always very small, the dissipation conveyed is practically equal to the collector-emitter dissipation.

The working point for an average collector current I_{C_0} and a voltage V_{C_0}

prevailing at the transistor, lies in the A-circuit. The maximum dissipation is therefore:

$$P_{C_{\max}} = V_{c_0} \cdot I_{c_0}.$$

The working point in the B-operation lies at current $I_{c_0} = 0$. If the load resistance is connected to the transformer as is the case in push-pull B-operation, the dissipation occurring depends on the maximum current modulation and we have

$$P_{C_{\max}} = V_{c_0} \frac{I_{c_{\max}}}{\pi}$$

if the modulation takes place sinusoidally.

Ohm's Law for the "flow of heat" can be applied for the thermal capacity which the transistor is able to yield.

$$c \frac{Q}{t} = P_{\text{off}}.$$

It reads:

$$P_{\text{off}} = \frac{\Delta T}{K} \tag{10.2}$$

Here P_{off} is the flow of energy, ΔT the thermal stress and K the heat resistance. The heat resistance K depends on the construction of the transistor. External cooling conditions also play a considerable part. The heat dissipation can be the result of conduction, radiation or convection. Heat conduction is promoted by good contact with the chassis when there is a large surface. Heat dissipation through convection is caused by motion of the air. The heat resistance is thus not an "apparatus constant" but can be approximately stated by observing certain prerequisites.

For transistors available up to the present, the heat resistance varies about

$$K = 0.8 \div 0.003°\text{C/mW}$$

The resistance concerned can possibly be calculated from a sum of heat resistances, e.g. $K = K_C + K_{\text{ch}}$, where K_C is the heat resistance between junction and case and K_{ch} is the resistance between the chassis and its surroundings.

Since the permissible dissipation can only be equal to the energy P_{off} which can be produced, it follows that with rising ambient temperature and therefore decreasing thermal stress ΔT, the permissible dissipation must also be reduced, or, conversely, must increase at constant ambient temperature with rising crystal temperature.

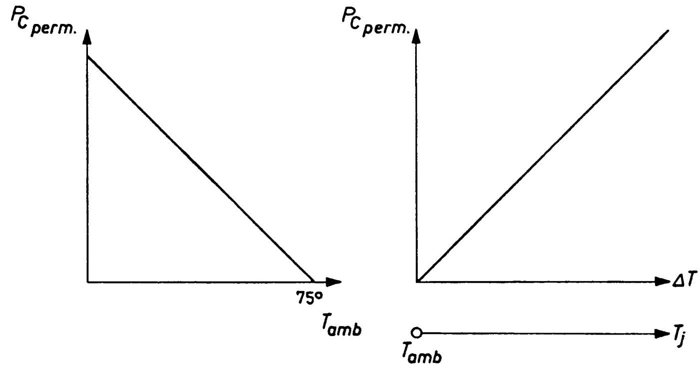


Fig. 10.1

If we assume a maximum ambient temperature of $T_{\mathrm{amb}_{\max}} = 45\,°\mathrm{C}$, $\Delta T_{\max}$ is then $75 - 45 = 30\,°\mathrm{C}$. With the heat resistance limits previously mentioned we obtain dissipations of:

$$P_{C_{\max}} = P_{\mathrm{ott}_{\max}} = \frac{\Delta T}{K} = \frac{30}{0.8} = 37.5\ \mathrm{mW}$$

or

$$P'_{C_{\max}} = P'_{\mathrm{ott}_{\max}} = \frac{\Delta T}{K'} = \frac{30}{0.003} = 10\ \mathrm{W}.$$

The maximum transistor dissipation up to now lies therefore at 10 W and is thus far below the maximum tube dissipation.

If the maximum dissipation in a particular case has been calculated according to the highest possible ambient temperature, the working point I_{C_0} can be set. If this adjustment is made with $T_{\mathrm{amb}} = 20\,°\mathrm{C}$, for instance, it is possible that the calculated maximum dissipation cannot be set at this temperature. With increasing ambient temperature the crystal temperature also rises and the result is an increase in the collector current. It is therefore possible that with $T_{\mathrm{amb}} = 45\,°\mathrm{C}$, the maximum dissipation adjusted with $T_{\mathrm{amb}} = 20\,°\mathrm{C}$ will be exceeded because of the larger collector current.

10.2. Temperature influence on the collector current

With increasing input voltage V_{BE} at initially constant crystal temperature, the input current I_B varies exponentially according to the equation:

$$I_B = C_1 \cdot e^{\frac{V_{BE}}{V_T}}\ . \tag{10.3}$$

C_1 is then a constant corresponding to the space charge constant pf the tube. The so-called temperature voltage has the value $V_T = 8.7 \cdot 10^{-5} T_{0_K}$. The collector current differs from the base current by the current amplification factor and with a new constant C_2 we can apply the expression:

$$I_c = C_2 \cdot e^{\frac{V_{BE}}{V_T}} . \tag{10.4}$$

With a base bias V_{BE_0} we obtain for the static collector current the relation:

$$I_{c_0} = C_2 \cdot e^{\frac{V_{BE_0}}{V_T}} . \tag{10.5}$$

If we now leave the bias $-V_{BE_0}$ constant we can establish that with rising temperature the collector current again increases according to an exponential principle. The relation

$$I_c = I_{c_0} \cdot e^{K \cdot \Delta T}$$

is then valid.

If we insert for the constant

$$K = \frac{D_T}{V_T}$$

we have:

$$I_c = I_{c_0} \cdot e^{\frac{D_T . \Delta T}{V_T}} . \tag{10.6}$$

By introducing (10.5) into (10.6) we then obtain:

$$I_c = C_2 \cdot e^{\frac{V_{BE_0} + D_T \cdot \Delta T}{V_T}} . \tag{10.7}$$

Here $D_T \cdot \Delta T$ is obviously a voltage which apparently also controls at the base and increases the collector current with growing temperature difference ΔT. We call D_T the reciprocal temperature difference.

The following values apply for germanium transistors.

I_cmA	0.1	0.2	0.5	1	2	5	10	20	50	100	200
$D_T \frac{\text{mV}}{°\text{C}}$	2.7	2.6	2.5	2.4	2.3	2.2	2.1	2	1.9	1.8	1.7

With mean values, acording to (10.6), we can approximately put:

$$I_c \approx I_{c_0} \cdot e^{\frac{\Delta T}{13}} , \tag{10.8}$$

i.e. for each 13 °C the current increases by about the factor e = 2.718 or with the factor 2 for every 9 °C temperature change.

This variation of collector current applies to the most unfavourable case with the ideal transistor. In practical operation the rise of collector current is mainly reduced through the internal base resistance r'_{bb} because the base current produces a voltage drop at this resistance which makes the base more positive. This extreme case will be applied in subsequent considerations just to be on the safe side.

After a transistor is connected in a circuit the junction temperature rises from the value of the ambient temperature. As a result the collector current is raised, the conveyed dissipation increases and the temperature can rise still further. If the conveyed dissipation at a certain temperature is not equal to the energy given off, thermal instability occurs, i.e. the temperature rises continuously until the transistor is destroyed. External switching measures can be used to prevent this kind of influence. Different cases will be studied in the following investigations.

10.2.1. CONSTANT D.C. BASE AND COLLECTOR VOLTAGE

This is the case, for instance, when a resonant circuit is present as load resistance, or a resistance is connected by means of a transformer. There is then no resistance voltage drop and the working voltage is equal to the d.c. collector voltage and therefore constant (Fig. 10.2).

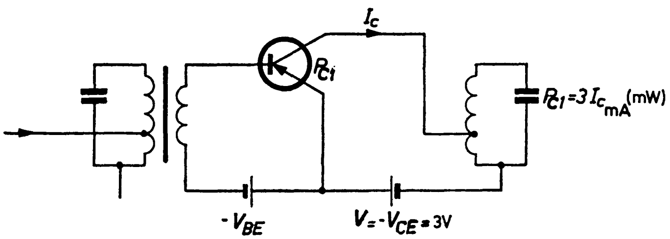


Fig. 10.2

The d.c. base voltage should also be produced from a low resistance source and because of the low resistance incoupling of the input bias, no additional ohmic resistance will be switched. in The base bias $-V_{BE_0}$ is thus constant and independent of the d.c. input resistance. The transistor dissipation in this case is now directly proportional to the collector current.

$$P_{C_1} = 3 \cdot I_{C_{mA}} [\text{mW}].$$

If we now plot in a diagram the current dependent on the temperature

T_j corresponding to a temperature difference $\Delta T = T_j - T_{amb}$ according to the Formula

$$I_C = I_{C_0} \cdot e^{\frac{\Delta T}{13}}$$

we find that the resulting curve can be used at the same time as energy curve P_C (Fig. 10.3).

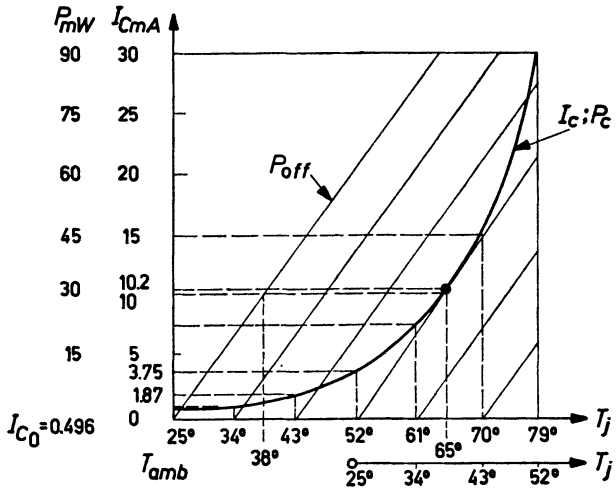


Fig. 10.3

If the transistor used has a heat resistance, for example, of $K = 0.432$°/mW, the energy which can be yielded would be

$$P_{off} = \frac{\Delta T}{0.432} = \frac{13}{0.432} = 30 \text{ mW}$$

that is to say, 30 mW for each temperature difference $T_j - T_{amb}$ of 13°.

If we assume a collector current $I_{C_0} = 0.496$ mA at an ambient temperature of $T_{amb} = 25$ °C, the current rises to twice its value for every 9 °C temperature increase in the crystal, as we see from the current curve in Fig. 10.3. By plotting the curve for dissipation energy as a straight line starting at the ambient temperature, we find that at 13 °C temperature increase this straight line goes through $P_{off} = 30$ mW. It can now be seen from the diagram that with a switch-on current $I_{C_0} = 0.496$ mA the curve of the dissipation P_C (current curve in the other scale) intersects the curve for dissipation with only a very slight temperature rise.

With an ambient temperature of $T_{amb} = 34$ °C, the P_{off} curve would begin at 34 °C because then the excess temperature or temperature difference $\Delta T = T_j - T_{amb}$ only starts at 34 °C. The point of intersection of the two energy curves already lies here with a somewhat greater temperature difference, as well as for an ambient temperature of 43 °C. The limiting case occurs with $T_{amb} = 52$ °C; there the two characteristics for P_C and P_{off} are tangential. The current at switch-on $I_{C_0} = 3.75$ mA now rises to $I_C = 10.2$ mA at 65 °C. The tangent to the P_C-curve is always given for a temperature difference $\Delta T = 13$ °C with a current rise by the factor $e = 2.718$. With a somewhat higher ambient temperature the two curves in the above case would no longer intersect. The conveyed dissipation then rises more than the energy given up. There is now thermal instability and the temperature on the crystal rises to the point of destruction. Whenever the collector current tends to rise above 2.5 times its value after switch-on, there is the danger of thermal instability and we have to switch off at once. We could now say that in Fig. 10.3 the ambient temperature of $T_{amb} = 52$ °C which brings about thermal instability is already very high and so the danger is not so great. This is true if the static current I_{C_0} with $T_{amb} = 25$ °C is only set at $I_{C_0} = 0.496$ mA.

If, however, I_{C_0} has already been adjusted at 3.75 mA with $T_{amb} = 25$ °C, the temperature scale T_j would apply in Fig. 10.3. The two energy curves would therefore already be tangential with $T_{amb} = 25$ °C. Even a slight increase in the ambient temperature can now lead to instability. This fact is important because the dissipation would only be $P_C = 3 \cdot 3.75 = 11.25$ mA with the switch-on current $I_{C_0} = 3.75$ mA. If we assume a maximum ambient temperature $T_{amb_{max}}$ of 45 °C, the maximum temperature difference ΔT could become $75 - 45 = 30$ °C. The maximum dissipation in this case with $K = 0.432$ is then:

$$P_{C_{max}} = \frac{\Delta T}{K} = \frac{30}{0.432} = 70 \text{ mW}.$$

Consequently, without a knowledge of the above considerations, the adjusted dissipation at switch-on of $P_C = 11.25$ mW would be justifiable. It can even rise to $P_C = 3 \cdot 10.2 = 30.6$ mW and be stable. Nevertheless, every minute rise in the external temperature would then lead to instability. We also have the case of constant base and emitter voltage in the measurement of transistor characteristics, possibily carried out point-by-point. If these characteristics have to be valid for a.c. behaviour, it is obvious that all characteristic curves must be taken up dynamically by working points of constant dissipation so that the crystal temperature remains constant, as is more or less the case in modulation with a.c. voltages.

10.2.2. LOAD RESISTANCE (ALSO PRINCIPLE OF SPLIT-LOAD OPERATION)

With a load resistance in the circuit the d.c. collector voltage $-V_{CE_0}$ is less than the d.c. operating voltage by the voltage drop $I_{C_0} \cdot R_L$ and is therefore no longer constant with variable static current. We can arrange for the dissipation to be at its greatest at the working point selected, decreasing with every change of the static current I_{C_0}. This is always the case when with static current half the operating voltage appears as collector voltage (Fig. 10.4).

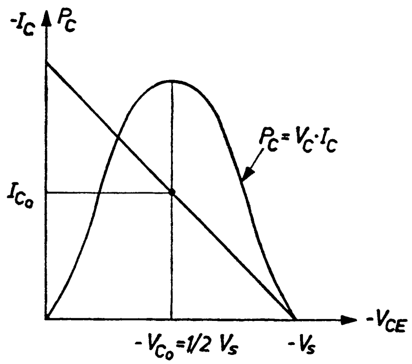


Fig. 10.4

With twice the static current the voltage at the transistor is theoretically zero, likewise the dissipation. Thermal instability can now be prevented with certainty through the increase of collector current with rising external temperature. This socalled "principle of split load operation" is based on the fact that an amplifier stage is operated for direct current with a resonance circuit as load resistance via a resistance with half the supply voltage. This

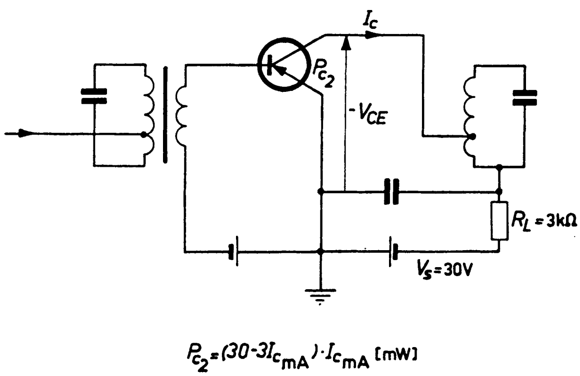


Fig. 10.5

resistance is capacitively short-circuited for a.c. voltage control (Fig. 10.5).

The dependence of dissipation and the working point on the ambient temperature will be seen in Fig. 10.6. The same transistor with a heat resistance $K = 0.432$ is again used. The following expression applies for the dissipation:

$$P_{C_2} = (30 - 3 \cdot I_{C_{\mathrm{mA}}}) \cdot I_{C_{\mathrm{mA}}} [\mathrm{mW}].$$

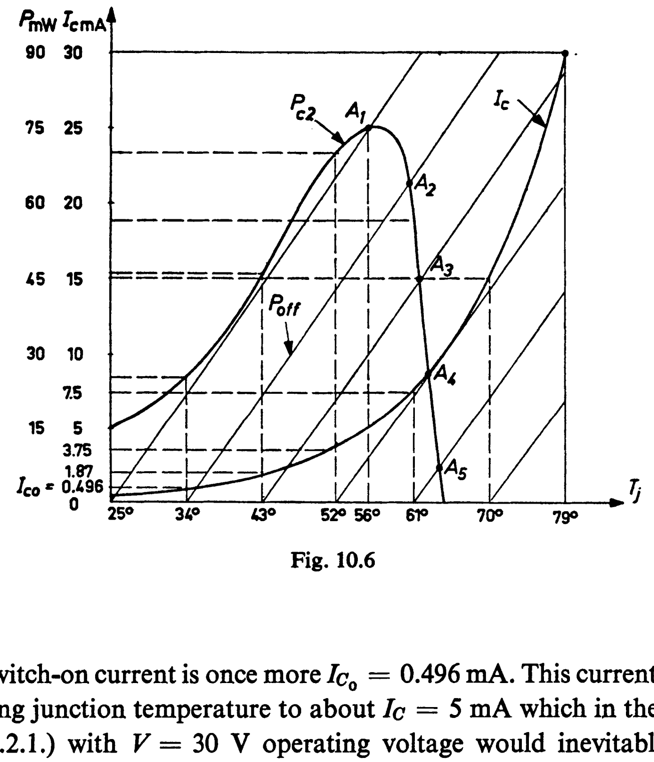


Fig. 10.6

The switch-on current is once more $I_{C_0} = 0.496$ mA. This current increases with rising junction temperature to about $I_C = 5$ mA which in the previous case (10.2.1.) with $V = 30$ V operating voltage would inevitably lead to destruction. Here, however, the two energy curves intersect at this current and give a stable working point in A_1. The transistor can therefore be used to much greater advantage, for the dissipation can now amount to $P_C = 75$ mW. In the case in (10.2.1.) the maximum dissipation in the stable limiting case was 30 mW, the normal dissipation was strongly dependent on temperature and at $T_{\mathrm{amb}} = 25$ °C was only about 1.5 mW. Here the dissipation at a maximum external temperature of $T_{\mathrm{amb}} = 45$ °C is now only reduced to 59% (A_3). Beyond that the ambient temperature can rise to 70 °C without endangering the transistor.

10.2.3. TEMPERATURE STABILISATION BY MEANS OF FEEDBACK

The chapter on feedback showed that each variation of the collector current is reduced through the feedback. For example, if it is 100% in the unstabilised case, the variation is reduced with feedback corresponding to the factor $1/(1 + \alpha \cdot A_v)$. If the current in the unstabilised case was

$$I_C = I_{C_0} \cdot e^{\Delta T/13}$$

or the current variation was

$$\Delta I_C = I_{C_0} \cdot e^{\Delta T/13} - I_{C_0}$$

with feedback it is

$$\Delta I'_C = \frac{I_{C_0}(e^{\Delta T/13} - 1)}{1 + \alpha \cdot A_v} \quad \text{or} \quad I_C = I_{C_0} + \Delta I'_C = I_{C_0}\left(1 + \frac{e^{\Delta T/13} - 1}{1 + \alpha \cdot A_v}\right).$$

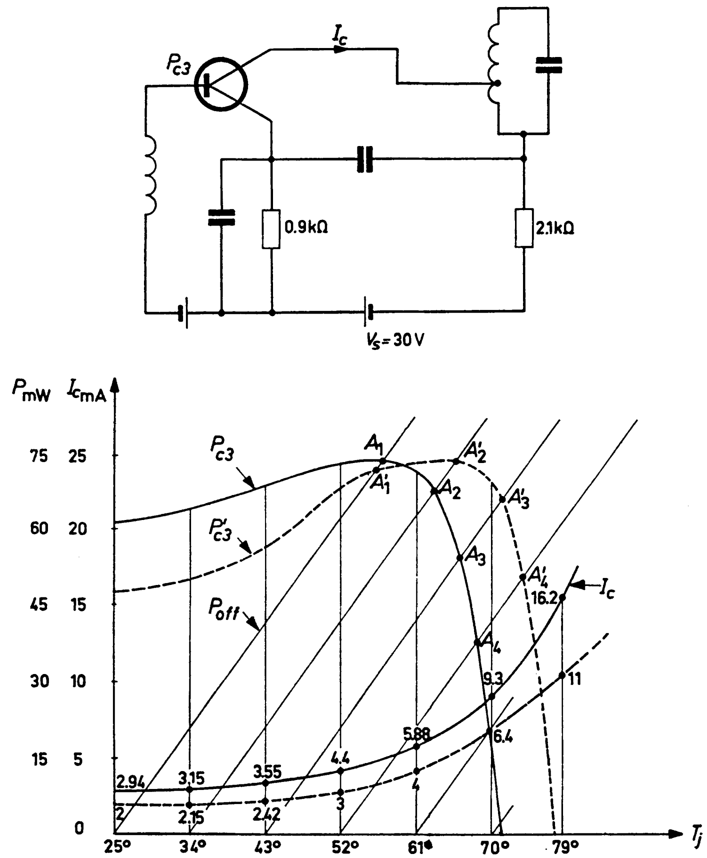


Fig. 10.7

If we now plot the current curve again as a function of the temperature T_j corresponding to a temperature difference $\Delta T = T_j - T_v$, according to the previous formula (Fig. 10.7), we obtain a much flatter curve. The switch-on current was adjusted to $I_{C_0} = 2.9$ mA so that the static current is again $I_{C_0} = 5$ mA. The circuit with series feedback is included in Fig. 10.7. Because of the equal total load resistance the dissipation is as in the case in 10.2.2.

$$P_{C_3} = (30 - 3I_C) \cdot I_C.$$

The feedback factor is $\alpha = 0.9/2.1 = 0.43$. All collector current variations are reduced with the factor

$$\frac{1}{1 + \alpha \cdot A_v} = \frac{1}{1 + 0.43 \times 30} = \frac{1}{14}$$

if we assume the voltage amplification to be about $A_v = 30$. It will be seen in Fig. 10.7 that the dissipation curve only drops at a higher temperature. The working points $A_1 \div A_3$ lie at higher collector currents than in case 10.2.2. (dotted curve). At $T_{amb} = 43$ °C the dissipation here only falls to 72%. The ratios become still more favourable if the switch-on current is set at $I_{C_0} = 2$ mA. Then the dissipation drop between $T_{amb} = 25$ °C $\div$ 43 °C only amounts to 12% (from A'_1 to A'_3).

10.2.4. TEMPERATURE COMPENSATION WITH THERMISTORS

Temperature stabilisation by means of a load resistance produces thermal stability by limiting the dissipation. The collector current working point is then displaced to higher values with rising temperature. Stabilisation with d.c. feedback gives a smaller displacement of the working point with consequent thermal stability. Constancy of the collector current working point, however, is not always desirable because even with constant working points the current amplification h_{21} and therefore the slope S can increase with rising temperature (direct temperature influence). In order that the transistor amplification does not increase with rising temperature, the collector current must not only not increase, but on the contrary it has to decrease. In this case we are forced to shift the base working point with increasing temperatures towards a smaller bias (see Chapter 13). This is easlily achieved by using a thermistor in the base bias voltage divider. A smaller base bias is obtained with increasing temperature because of the drop in resistance (Fig. 10.8).

In comparison with d.c. feedback this compensation circuit has the advantage that the collector voltage is not reduced and as a result the whole working voltage is available for modulation.

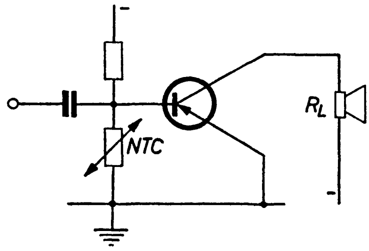


Fig. 10.8

This is particularly important in output stages. This compensation circuit has the drawback that it does not give absolute security against transistor overloading because the thermistor only responds to ambient temperature. The junction temperature of the transistor therefore does not directly enter into the compensation.

10.2.5. STABILISATION BY A BASE SERIES RESISTANCE

Stabilisation with a base series resistance would be the simplest method. That is to say, if the base bias is only produced with a series resistance this can be made relatively large by using the full operating voltage. The d.c. input resistance is small in comparison and as a result the inflowing base current is virtually determined only by the series resistor and the working voltage. The input current remains constant in spite of rising temperature but the collector current working point is very greatly displaced under certain conditions because the d.c. amplification is temperature dependent, and the base-collector/residual current I_{CB_0} shows particularly strong dependence on temperature. With small base currents this leads to large working point displacements (see Chapter 11).

CHAPTER 11

TRANSISTOR LEAKAGE CURRENTS

The current in the blocking direction in a diode is called the leakage current. There are no leakage currents in the transistor with its two diodes if one electrode is open at a time. The amounts of these leakage currents depend only a little upon the value of the voltage applied (saturation). With a germanium transistor the following leakage currents can be measured at 25 °C:

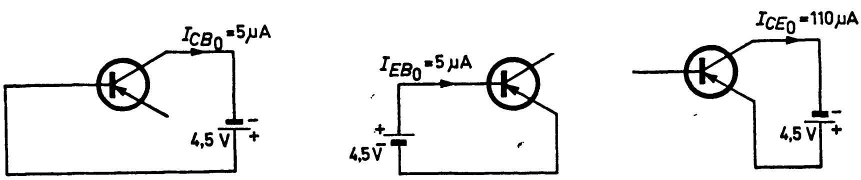


Fig. 11.1

These leakage currents are not interesting in themselves when the transistor is controlled with a.c. values because the leakage current does not directly contribute to the control. So why do they claim our attention? There are two reasons for this:

1. The leakage currents give information concerning the quality and the internal nature of a transistor. Leakage current measurements are therefore made by transistor testing.
2. The collector-base leakage current I_{CB_0} has to be accounted for in the practical circuit.

With reference to Point 2 we can say that in normal transistor operation in a circuit, only the collector-base leakage current I_{CB_0} is effective because only the collector-base section is operated in the blocking direction. This collector-base leakage current I_{CB_0}, therefore, always flows in the circuit, even during operation. It can influence the working point in the circuit, especially with temperature variation. The leakage current I_{CB_0} usually given for a junction temperature of 25 °C rises with each 9 °C increase in temperature to twice its value. With a temperature rise from 25 °C to 61 °C, i.e. by $4 \cdot 9 = 36$ °C, the leakage current increases by about the factor $2^4 = 16$.

A germanium transistor, for example, can have a leakage current of 12 μA at 25 °C. At 61 °C, however, this current amounts to $16 \cdot 12 = 192$ μA. The effect of this kind of current in the circuit with temperature variations will be investigated under different conditions.

11.1. Unstabilised circuit

A transistor would be operated entirely unstabilised if all the d.c. resistances supplying the d.c. working voltages were of very low value. Such would be the case, for instance, if only resonance circuits were connected into the base- and collector circuits.

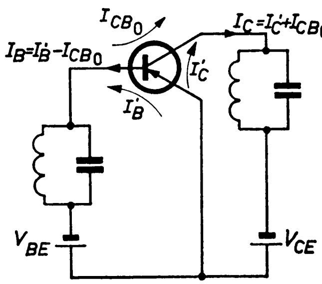


Fig. 11.2

The collector-base leakage current I_{CB_0} which is driven through $V_{CE} - V_{BE}$ is superimposed on the base current I'_B and the collector current I'_C. This superposition takes place without a reciprocal effect because the external voltages are practically constant as a result of the smaller resistances. With rising temperature the current I_{CB_0} increases to double its value for every 9 °C, while the actual collector current I'_C rises to twice its value for every 20 to 30 °C. This hardly ever happens, however, because stabilisation and therefore reduction of the current increase with rising temperature are ensured by connecting resistances in the circuit. It depends on the circuit whether this required stabilisation succeeds with the normal collector current and simultaneously with the residual current I_{CB_0}, as we shall see in the following section.

11.2. Leakage current influence with stabilised constant base current

If the base-emitter section is fed from a source with high voltage, a high resistance R_g can be used to divide the voltage. The second divider resistance is then the d.c. input resistance r_{BE}. Since $R_g > r_{BE}$, the base current is only determined through R_g, i.e. a constant base current is supplied. However, as the collector-base leakage current I_{CB_0} also has to be taken into account, the constancy of the actual base current I'_B is non-existent under certain circumstances. This only happens of the residual current is no longer small in comparison with the normal static base current.

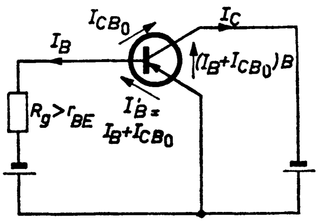


Fig. 11.3

It is clear from Fig. 11.3 that if I_B is the measured base current and the current I_{CB_0} flows from the base to the collector in the opposite direction to the base currents, the d.c. control current must have the value $I'_B = I_B + I_{CB_0}$ because I_B is equal to $I'_B - I_{CB_0} = (I_B + I_{CB_0}) - I_{CB_0}$. If B is the d.c. amplification the following expressions are valid for the collector current:

$$I_C = (I_B + I_{CB_0})B + I_{CB_0}$$
$$I_C = I_B B + I_{CB_0}(1 + B). \qquad (11.1)$$

If the base current is thus kept constant through a large generator resistance $R_g > r_{BE}$, the collector current can vary greatly with rising temperature, especially if I_B itself is small. The change in the collector current is then considerable:

$$\Delta I_C = \Delta I_{CB_0}(1 + B) \qquad (11.2)$$

The leakage current variation ΔI_{CB_0} which occurs with the factor 2 for each 9 °C, therefore takes place in the collector circuit amplified by the factor $(1 + B)$. We can also say that an apparent effective leakage current of the value

$$I_0 = I_{CB_0}(1 + B) \qquad (11.3)$$

always flows in the collector circuit.

Example 11.1

There are measuring instruments for determining the d.c. amplification of a transistor. Here, for instance, through a working voltage of 25 V a current of $I_B = 25\ \mu A$ is fed into the base via a resistance R_g of 1 MΩ. From this we assume that the current is constant independent of the d.c. input resistance r_{BE} which can amount to 8 kΩ. The instrument for measuring the collector current is therefore calibrated directly in d.c. amplification "B". This measurement, however, is very inaccurate as we can easily realise when we know, for example, that the leakage current can amount to 5 μA at 25 °C. In this case the actual controlling current I'_B is $25 + 5 = I_B = I_{CB_0} = 30\ \mu A$. With only 9 °C temperature rise, I_{CB_0} is then already 10 μA, which with 25 μA base current naturally means a large error because the actual control current is now $I'_B = 25 + 10 = 35\ \mu A$. Such measurements are thus very inaccurate because they show strong temperature dependence in estimating the current amplification factor.

11.3. The leakage current influence upon stabilisation of the emitter current

Stabilisation of the emitter current and consequently of the collector current can always be produced if there is a resistance R_E in the emitter lead. So that temperature variation only causes a slight change in the emitter current and therefore only a small displacement of the transistor working point, the emitter current is controlled through the proportional voltage drop at the emitter resistance. This voltage drop will react upon the transistor input as feedback voltage. The total variation $\Delta V_E = \Delta I_E \cdot R_E$ should as far as possible be effective at the input between base and emitter because then the counteraction and consequently the stabilisation is at its strongest. This is always the case when the d.c. voltage source for the base-emitter section is of low resistance. This condition is also suitable as a means of reducing the influence of the leakage current.

The following circuits belong in the same way to the group: "Stabilisation of emitter current".

These three circuits are the same in regard to the d.c. base supply. They only differ electrically, possibly through the size of the internal resistance of

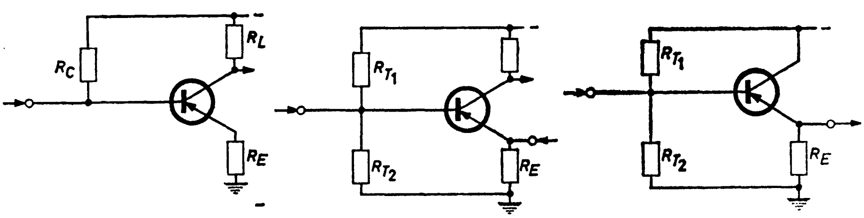


Fig. 11.4

the supply source. All three circuits can be replaced as far as direct current is concerned by the following circuits.

The supply generator resistance R_g is represented in circuit 1, Fig. 11.4, by $R_g = R_C$, in circuits 2 and 3 by $R_g = R_{T_1} \parallel R_{T_2}$. In Fig. 11.5 the leakage current I_{CB_0} which always flows from the base to the collector, is represented in a resistance R_{CB_0} which we can envisage in the transistor interior. This current is virtually driven through E_C because E_B is small.

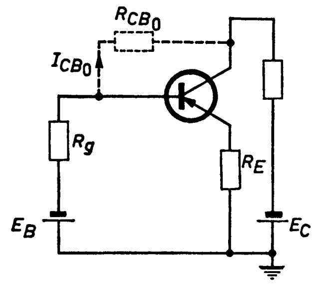


Fig. 11.5

For better understanding Fig. 11.5 can be somewhat elaborated. The base supply voltage lies at a voltage divider formed from the generator internal resistance R_g, the d.c. input resistance r_{BE} and a resistance $R_E \cdot (1 + B)$. This last resistance is produced because the base current flows through R_E increased by the current amplification factor B thus multiplied by $(1 + B)$. If we assume, however, that the base current alone flows via R_E, as is shown in the above circuit, the resistance R_E must then be increased by the factor $(1 + B)$ to produce the same voltage drop.

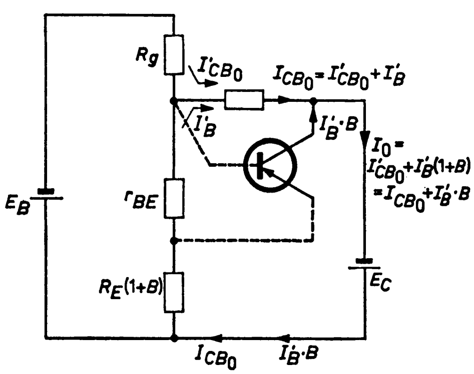


Fig. 11.6

The current I_{CB_0} driven through E_C comes from the voltage divider from point B. It is formed through two currents because the current I_{CB_0} coming from E_C branches at the resistances R_g and $r_{BE} + R_E\,(1 + B)$ which in this case lie parallel. The leakage current is now composed of the two partial currents I'_{CB_0} and I'_B. The current I'_B is chiefly interesting here because it flows through r_{BE} as control current and as such appears as collector current amplified by $B \cdot I'_B$. The effective leakage current I_0 in the collector circuit is therefore greater than the actual leakage current I_{CB_0}. The effective leakage current I_0 in the collector circuit obviously derives from:

$$I_0 = I'_{CB_0} + I'_B(1 + B) = I_{CB_0} + I'_B \cdot B. \tag{11.4}$$

It is thus composed of the actual leakage current and the amplified laekage current $I'_B \cdot B$.

The size of the controlling leakage current I'_B depends on the ratio of the parallel resistances R_g and $R_E\,(1 + B)$ if $r_{BE} < R_E \cdot (1 + B)$ is disregarded.

According to the current divider formula the following expression can be applied:

$$I'_B = I_{CB_0} \frac{R_E}{R_g + R_E(1 + B)}. \tag{11.5}$$

For the second partial current we obtain:

$$I'_{CB_0} = I_{CB_0} \frac{R_E(1 + B)}{R_g + R_E(1 + B)}. \tag{11.6}$$

By substitution in Equation (11.4) we get for the actual leakage current in the collector circuit:

$$I_0 = I_{CB_0}\left[\frac{R_E(1 + B)}{R_g + R_E(1 + B)} + \frac{R_g(1 + B)}{R_g + R_E(1 + B)}\right]$$

$$I_0 = I_{CB_0} \cdot (1 + B) \frac{R_E + R_g}{R_g + R_E(1 + B)}$$

$$I_0 = I_{CB_0} \frac{R_E + R_g}{R_E + \dfrac{R_g}{1 + B}}. \tag{11.7}$$

It can be seen from this equation that the leakage current I_0 acting in the collector circuit depends on the ratio of the two resistances R_E to R_g. If we assume, for example, that $R_g > R_E$, we then have:

$$I_0 = I_{CB_0}(1 + B).$$

This is practically the case in 11.2., "Leakage current influence with constant base current". The improvement through resistance R_E $(1 + B)$ then does not come into effect at all. On the other hand, if $R_g < R_E$, $I_0 = I_{CB_0}$. This is the case described under point 11.2, "Unstabilised circuit". The leakage current has no control effect; it simply superimposes the collector current.

What general influence has the effective leakage current I_0 on the stabilisation? The emitter resistance ensures almost constant emitter and collector current even with rising temperature. If the effective leakage current I_0 now increases with the temperature while the total collector current remains almost constant, the controllable dynamic current I_{C_F} must decrease.

$$I_C = \text{const.} = I_0 + I_{C_F}.$$

If the dynamic current I_{C_F} itself is small, it can easily happen that $I_C =$ const. $= I_0$, and the dynamic current will be completely squeezed out; the transistor is "starved". With the help of Equation (11.7) we can find out how high the effective leakage current is at a maximum temperature and whether there is still sufficient dynamic current $I_{C_F} = I_{C_0} - I_0$.

11.4. Leakage current influence on stabilisation of the collector current

The stabilisation of the collector current is called parallel feedback on account of the circuit arrangement.

We therefore obtain two circuits of equal value for the emitter and the base, as far as direct current is concerned.

Every variation of the collector current is controlled here through the voltage drop at the load resistance R_L. The larger R_L is, the stronger is a variation such as $\Delta V_L = \Delta I_C \cdot R_L$. This variation also occurs at the voltage divider R', r_i, so that the portion decreasing at r_i (corresponding to a feedback factor $\alpha = r_i(r_i + R'))$ counteracts the variation as feedback voltage at the

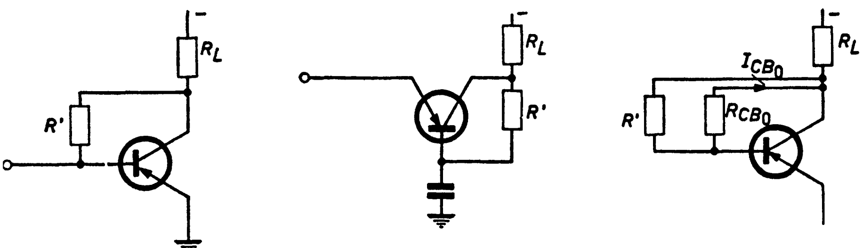


Fig. 11.7

transistor input. Therefore, the larger R_L can be made and the smaller R' is, the better is the stabilisation. According to the statements in 9.5. "Working point stabilisation", we could expect that a collector current variation caused by the leakage current I_{CB_0} would also be reduced by the factor $1/(1 + \alpha \cdot A_v)$. This, however, is not so because the leakage current not only produces a variation of the collector current but of the base current as well.

As Fig. 11.7 also shows, we can visualise a resistance R_{CB_0} between base and collector which allows the leakage current I_{CB_0} to flow. The question now is, with a specific leakage current I_{CB_0}, what proportion of this current flowing in the collector circuit is the effective leakage current I_0 in the collector circuit? This problem can best be solved by drawing the whole circuit; consider points C and B at which R_{CB_0} lies as generator terminals and calculate the current variation when R_{CB_0} is connected at these terminals and the generator is loaded with I_{CB_0}.

The voltage divider $(R' \parallel R_{CB_0}) + r_{BE}$ connected at the load resistance allows the d.c. base current I_B to flow via the d.c. input resistance r_{BE}. According to the current amplification factor B, however, a total current $I_B\ (1 + B)$ flows because the transistor itself allows a current $I_B \cdot B$ to flow. If we start with the input circuit concerned, we can then take into account the additional collector current I_C which is always linked with I_B, in such a way that, as Fig. 11.8 shows, all the resistances of the input circuit are reduced by the factor $1/(1 + B)$, or the currents are increased by $(1 + B)$; the total resistance connected at R_L then remains constant.

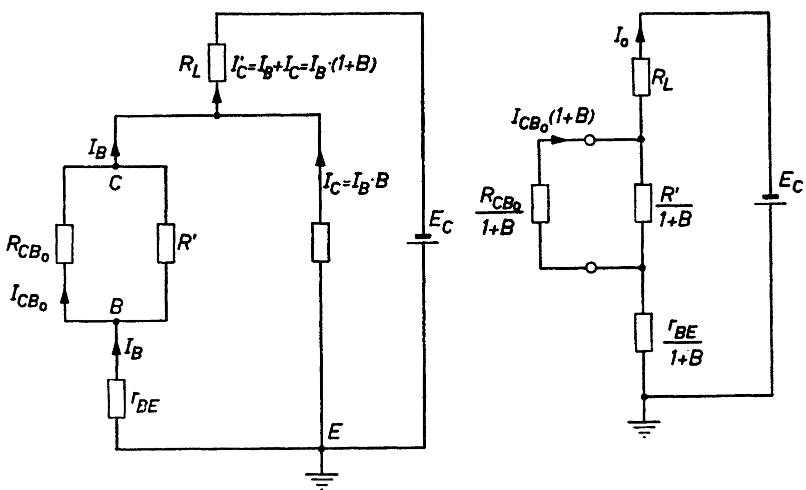


Fig. 11.8

From this it is easy to calculate the current I_0 produced through loading C and B with I_{CB_0}. If terminals C and B are open, the internal resistance of this generator is:

$$R_{t_{CB}} \approx \frac{R'}{1+B} \parallel R_L$$

if we disregard the resistance $r_{BE}/(1+B)$ which is small compared to R_L. If this generator is now loaded with I_{CB_0} $(1+B)$ we obtain a voltage variation $\Delta V = I_{CB_0} \cdot R_{tCB}$ $(1+B)$ at the terminals. The same voltage variation must take place at the load resistance because the total voltage E_C is constant and the small voltage drop at $r_{BE}/(1+B)$ can be ignored. In this way we now obtain the effective residual current I_0 in the load resistance. It is obtained from:

$$I_0 = \frac{\Delta V}{R_L} = I_{CB_0} \cdot (1+B) \frac{\frac{R'}{1+B} \cdot R_L}{\frac{R'}{1+B} + R_L} \cdot \frac{1}{R_L}$$

$$I_0 = I_{CB_0} \frac{R'}{\frac{R'}{1+B} + R_L} \approx I_{CB_0} \frac{R'}{R_L}. \tag{11.8}$$

It is evident from the above formula that with stabilisation of the collector current the effective leakage current I_0 in the collector circuit will always be greater than the actual leakage current I_{CB_0} because the ratio R'/R_L is always greater than one.

Here also, $I_C \approx$ const. $= I_{C_0} + I_0$ because of the collector current stabilisation. Therefore, if the effective leakage current I_0 increases, the dynamic current $I_{C_F} = I_{C_0}$ initially adjusted becomes gradually less. If I_{C_0} itself was small, it can even happen in this case that the dynamic current disappears with rising temperature. By calculating with Equation (11.8) we can check whether this condition is reached with higher temperatures. We can then carry out a corresponding adjustment such as providing for a larger static current I_{C_0}.

Example 11.2

We have a circuit as in Fig. 11.9. At 25 °C an emitter current $I_{E_0} = 0.18$ mA $\approx I_{C_0}$ flows, and $I_{CB_0} = 4.5$ μA when $1 + B \approx 100$. We wish to find the effective dynamic current at 43 °C.

Solution

$$R_g = 47\ \text{k}\Omega + 47\ \text{k}\Omega \parallel 22\ \text{k}\Omega = 62\ \text{k}\Omega$$

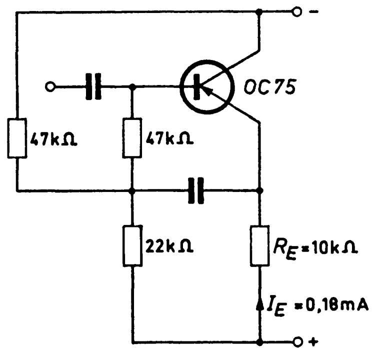


Fig. 11.9

at 25°C

$$I_0 = I_{CB_0} \frac{R_E + R_g}{R_E + \frac{R_g}{1+B}} = 4.5 \frac{10 + 62}{10 + \frac{62}{100}} = 30.4\,\mu\text{A}$$

àt 43°C

$$\Delta t = 43 - 25 = 18\text{ °C}$$

$$I_{0_{43°}} = 2^2 \cdot I_0 = 4 \cdot 30.4 = 121\,\mu\text{A}$$

$$I_{C_F} = I_{C_0} - I_{0_{43°}} = 180 - 121 = 59\,\mu\text{A}.$$

There is therefore only 59 μA collector current available for modulation With further heating there is thus the risk of "starvation".

This circuit is also interesting because the necessary low values of base voltage divider resistance are included for alternating current as the emitter resistance and there is therefore no great reduction of input resistance.

The emitter resistance has the value:

$$R_L = \frac{1}{\frac{1}{10} + \frac{1}{22} + \frac{1}{47}} = 6\text{ k}\Omega$$

for a.c. voltage.

The effective input resistance is:

$$r_{i_C} \approx R_L \cdot (1 + \beta_0) = 100 \cdot 6 = 600\text{ k}\Omega$$

if $1 + B_0 \approx 100$ according to Equation (14.10).

The low voltage divider resistance thus has only very little effect.

CHAPTER 12

DIRECT CURRENT AMPLIFICATION

When we are fully aware of the influence of the leakage current it will be obvious that it is by no means so simple to state the d.c. amplification. The following expressions,

$$B = \frac{I_C}{I_B} \tag{12.1}$$

as d.c. amplification for the emitter circuit, or

$$A = \frac{I_C}{I_E}. \tag{12.2}$$

for the base circuit, are only valid if the leakage currents are so small compared with the dynamic currents that they can be disregarded.

12.1. Emitter circuit

If I_B, I_E and I_C are measured in an unstabilised and therefore low resistance circuit, and the leakage current I_{CB_0} is no longer small in comparison with these currents, we then obtain the following formula for the d.c. amplification in the emitter circuit:

$$B = \frac{I_C - I_{CB_0}}{I_B + I_{CB_0}} = \frac{I'_C}{I'_B}. \tag{12.3}$$

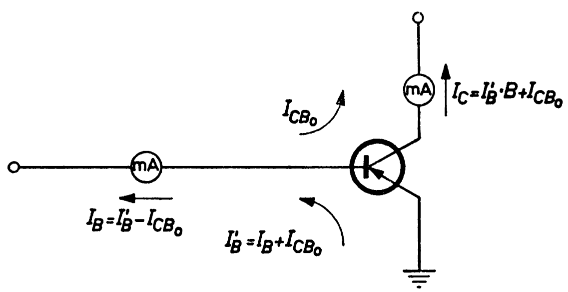


Fig. 12.1

The measured collector current must be decreased by the leakage current I_{CB_0} to obtain the controlled current I'_C. If the measured base current is

I_B but is flowing in the opposite direction to I_{CB_0}, the controlling base current must then have the value $I'_B = I_B + I_{BC_0}$. The d.c. amplification is thus produced from these two currents sharing in the control.

12.2. Base circuit

If the residual current is no longer small compared with the dynamic currents, the current amplification in the base circuit also has to be estimated by taking I_{CB_0} into consideration. In the unstabilised case the emitter current should be fed in as control current. It will be indicated by a measuring instrument and will not be affected by I_{CB_0}. The current I_C which can be measured at the collector again includes the leakage current I_{CB_0}, so that the actual controlled current reduced again by I_{CB_0}, appears as $I'_C = I_C - I_{CB_0}$.

The d.c. amplification is therefore obtained from:

$$A = \frac{I'_C}{I_E} = \frac{I_C - I_{CB_0}}{I_E}. \tag{12.4}$$

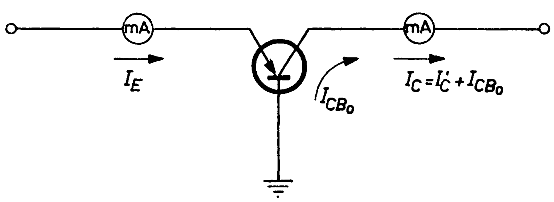


Fig. 12.2

If we wish to calculate the current amplification from any stabilised circuit, the influence of the prevailing leakage current must be taken into account as described in Chapter 11.

CHAPTER 13

CHARACTERISTIC VALUES AND THEIR DEPENDENCE UPON THE WORKING POINT

If we study the characteristics in the first to third quadrants in the four-quadrant set of characteristics we can discover the dependence of the transistor characteristic values upon the working point concerned. The different slope of the characteristics in various working points gives information concerning the change of the working points involved. An investigation of this kind naturally does not give the influence of the voltage feedback on these characteristic values, but this has practically no effect on the qualitative statements.

VARIATION OF THE INTERNAL RESISTANCE

If the working point in Fig. 13.1 is moved from B to A, i.e. from a smaller to a higher collector current, the slope of the characteristic increases and the internal resistance decreases.

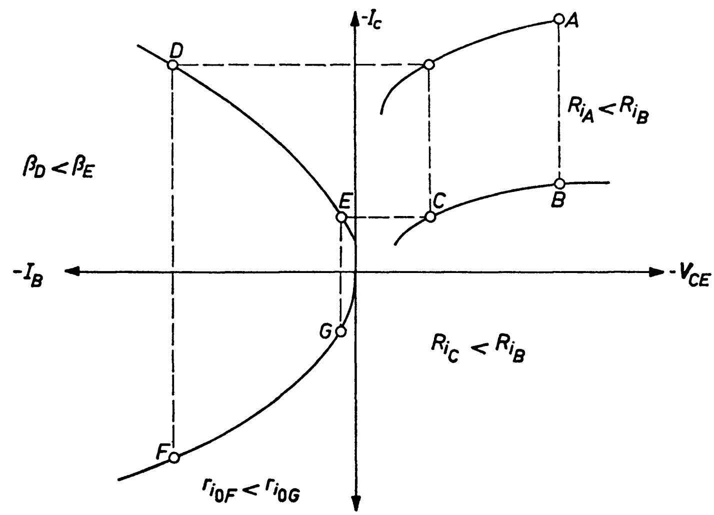


Fig. 13.1

The internal resistance acts more or less in inverse proportion to the collector current

$$R_i \sim \frac{1}{I_C}. \tag{13.1}$$

The collector voltage has less influence, as can be seen from a displacement of the working point from B to C. The internal resistance is somewhat less with smaller collector voltage.

VARIATION OF THE INPUT RESISTANCE

The working point F in Fig. 13.1 implies a larger collector current than in G. With the same a.c. input voltage the current variation is greater in working point F than with control in working point G. The input resistance therefore acts more or in less inverse proportion to the collector current

$$r_{i_0} \sim \frac{1}{I_C}. \tag{13.2}$$

VARIATION OF THE SLOPE

The slope is:

$$S = \frac{i_2}{v_1} = \frac{i_1 \cdot \beta_0}{i_1 \cdot r_{i_0}} = \frac{\beta_0}{r_{i_0}}.$$

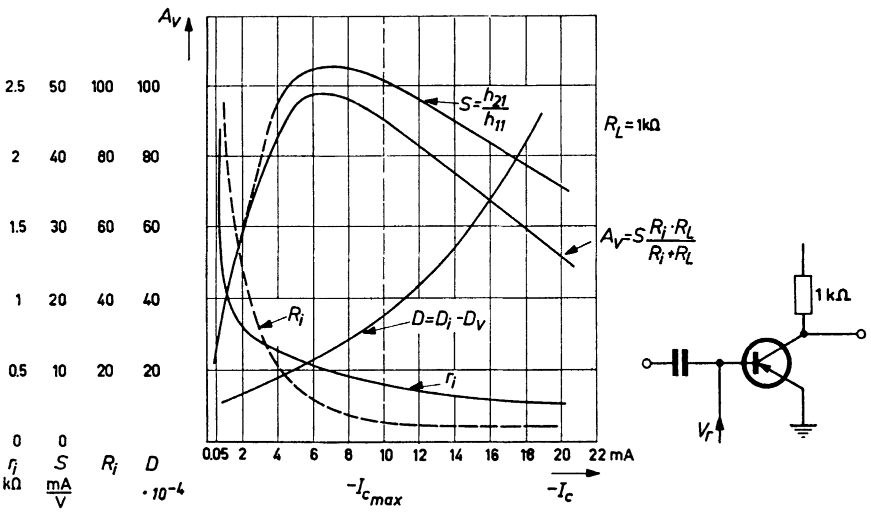


Fig. 13.2

Since B_0 only varies with higher collector currents, the slope at first increases in proportion to I_C.

$$S \sim I_C. \tag{13.3}$$

With larger collector currents β_0 can decrease and the slope can pass through a maximum.

By using the measured values for a transistor T_2 (corresponding to the OC 70) we can regard the typical curve of the characteristic values as a function of the collector current working point (Fig. 13.2).

13.1. The transistor as a regulating device

By means of an apparently power-less displacement of the grid working point, variable-mu valves permit a reaction-free slope variation with a ratio of more than 1 : 1000. This good regulating effect of the tube cannot be achieved with the transistor for three main reasons:

1. The transistor has to be able to displace the working point.
2. The displacement range of the working point $-V_{BE_0}$ only lies in the order of 100 mV.
3. The reaction upon the input resistance r_i can impair the regulating action.

The regulating voltage usually has to be amplified to control the regulating transistor. With small signal transistors in particular, and therefore with large modulation, the small displacement range is awkward and limits the maximum regulating action. With HF pre-amplifier transistors we reach regulating ratios in the dimension of 1 : 100. The harmful reaction is chiefly noticeable in AF transistors. The ratios for the most unfavourable case will now be shown in an example with the AF transistor T_2.

Example 13.1

Two transistors T_2 are connected together as illustrated in Fig. 13.3. The greatest input voltage should amount to $v_1 = 20$ mV in the small signal condition. We can see from the 3rd to the 1st quadrants of the characteristic curves for T_2 that with this limitation transistor T_2 can be stepped down to $-I_{C_{2/1}} = 0.5$ mA. It is obvious from Fig. 13.2 that regulation is possible in the other direction up to $-I_{C_{2/2}} = 5$ mA through regulating voltage V_r because the slope decreases beyond that. The regulating ratio for the total voltage amplification has to be calculated when the collector resistance is

a) $R_{C_{1/1}} = 5$ kΩ or only
b) $R_{C_{1/2}} = 0{,}5$ kΩ.

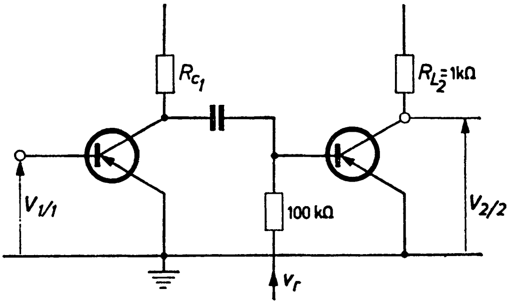


Fig. 13.3

Solution

From Fig. 13.3 we obtain:

$$S_2 = 13.6\ \text{mA/V} \qquad S'_2 = 49\ \text{mA/V}$$

$$A_{v_2} = 13.5 \qquad A'_{v_2} = 45$$

$$r_{i_2} = 2.2\ \text{k}\Omega \qquad r'_{i_2} = 0.55\ \text{k}\Omega$$

The load resistance for the first transistor is:

for a) $R_{L_1} = R_{C_{1/1}} \| r_{i_2}$ $\qquad R'_{L_1} = R_{C_{1/2}} \| r'_{i_2}$

$= 5\ \text{k}\Omega \| 2.2\ \text{k}\Omega$ $\qquad = 5\ \text{k}\Omega \| 0.55\ \text{k}\Omega$

$= 1.53\ \text{k}\Omega$ $\qquad = 0.496\ \text{k}\Omega$

for b) $R_{L_1} = 0.5\ \text{k}\Omega \| 2.2\ \text{k}\Omega$ $\qquad R'_{L_1} = 0.5\ \text{k}\Omega \| 0.55\ \text{k}\Omega$

$= 0.407\ \text{k}\Omega$ $\qquad = 0.272\ \text{k}\Omega$

The voltage amplification of the pre-amplifier transistor is:

for a) $A_{v_1} = S_1 \cdot R_{L_1} = 30 \times 1.53$ $\qquad A'_{v_1} = S_1 \cdot R'_{L_1} = 30 \times 0.496$

$= 46$ $\qquad = 14.9$

for b) $A_{v_1} = 30 \times 0.407$ $\qquad A'_{v_1} = 30 \times 0.272$

$= 12.2$ $\qquad = 8.15$

The total amplification is:

for a) $A_{v_{tot}} = A_{v_1} \cdot A_{v_2} = 46 \times 13.5$ $\qquad A'_{v_{tot}} = A'_{v_1} \cdot A'_{v_2} = 14.9 \times 45$

$= 620$ $\qquad = 670$

for b) $A_{v_{tot}} = 12.2 \times 13.5 = 165 \qquad A'_{v\,tot} = 8.15 \times 45 = 367.$

The regulating ratio is consequently:

for a) $$A = \frac{A_{v_{tot}}}{A'_{v_{tot}}} = \frac{620}{670} = 1/1.08$$

for b) $$A = \frac{165}{367} = 1/2.22.$$

The maximum regulating ratio of the second stage

$$A = \frac{A_v}{A'_v} = \frac{13.5}{45} = 1/3.33$$

is thus reduced in both cases. If we make use of a common collector stage as a pre-amplifier we can reach the maximum regulating ratio. The control generator resistance must therefore be as small as possible so that the reaction is made harmless.

CHAPTER 14

THE COMMON COLLECTOR CIRCUIT

In the common collector circuit (Fig. 14.1) the collector connection is the common base point for the input and output circuits because for a.c. voltage "minus" is equal to "earth".

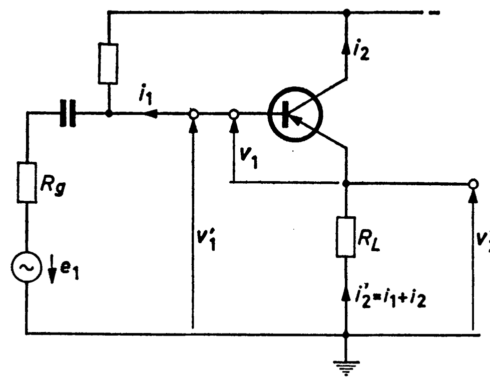


Fig. 14.1

The fundamental circuit for the transistor, the common emitter circuit, has been dealt with in detail. In contrast to this fundamental circuit, only the load resistance has been moved from the collector to the emitter. If we take into consideration this alteration in the fundamental equivalent circuit, the same must also apply here (Fig. 14.2).
For comparison the common emitter circuit has again been shown in Fig.

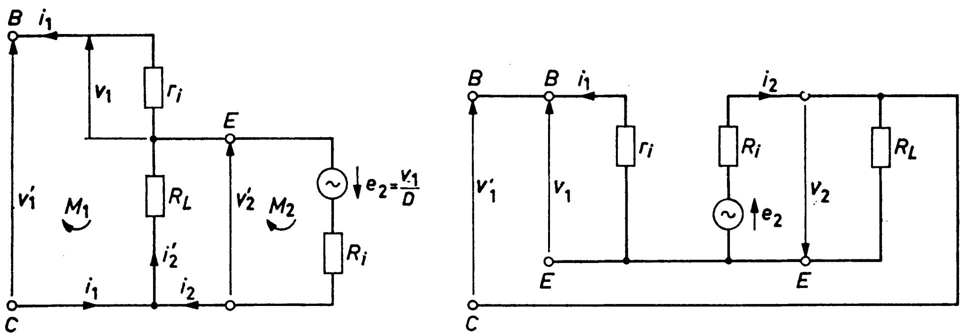


Fig. 14.2

14.2 alongside the common collector circuit which appears as an adaptation of the other .The only difference is that in the common collector circuit the input voltage v'_1 lies between base and collector. If we compare the plotted current and voltage directions these agree with Fig. 14.1. The next aim of the investigation is to find a simple equivalent circuit for the common collector circuit in the form of the fundamental equivalent circuit. If we look at the common collector circuit with its three connection terminals this represents an equivalent transistor, i.e. a new transistor with changed characteristic values can be imagined in relation to these three connection terminals. These equivalent characteristic values are calculated below (Fig. 14.3).

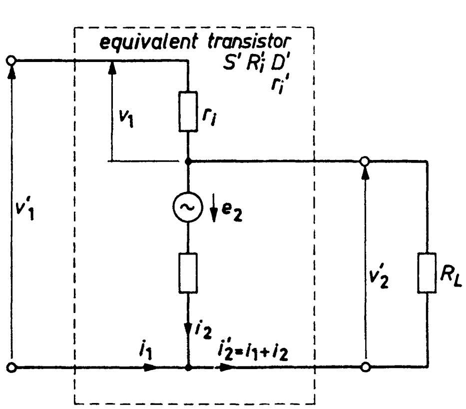


Fig. 14.3

Since these equivalent characteristic values correspond to the characteristic values in the fundamental circuit, it is clear that the fundamental equivalent is also valid with these equivalent characteristic values.

For the network M_1 drawn in Fig. 14.2 we have:

$$v'_1 - v_1 - v'_2 = 0$$

from which

$$v'_2 = v'_1 - v_1.$$

The output voltage is therefore always less than the input voltage v'_1. It is reduced through the control voltage v_1 appearing as voltage drop at r_i. The output voltage is equal to the voltage drop at R_L, therefore

$$v'_2 = (i_1 + i_2)\cdot R_L = i_2\left(\frac{1}{A_i} + 1\right)\cdot R_L = i'_2\cdot R_L. \tag{14.2}$$

The total current in R_L is thus:

$$i'_2 = i_2\left(\frac{1}{A_i} + 1\right).$$

The normal control voltage v_1 results from network M_2 in Fig. 14.2. There

$$e_2 = \frac{v_1}{D} = i_2 \cdot R_i + (i_1 + i_2) \cdot R_L = \frac{i'_2 \cdot R_i}{1 + 1/A_i} + i_2' \cdot R_L$$

$$= i'_2\left(\frac{R_i}{1 + 1/A_i} + R_L\right).$$

Resolved for v_1 this gives:

$$v_1 = i'_2\left(\frac{D \cdot R_i}{1 + 1/A_i} + D \cdot R_L\right). \tag{14.3}$$

If we insert (14.2) and (14.3) in (14.1) we then obtain:

$$v'_2 = i'_2 \cdot R_L = v'_1 - i'_2\left(\frac{D \cdot R_i}{1 + 1/A_i} + D \cdot R_L\right).$$

Resolved for i'_2 this gives

$$i'_2 = \frac{v'_1}{R_L + \dfrac{D \cdot R_i}{1 + 1/A_i} + D \cdot R_L} = \frac{v'_1}{\dfrac{D \cdot R_i}{1 + 1/A_i} + R_L(1 + D)}$$

$$i'_2 = \frac{\dfrac{v'_1}{1 + D}}{\dfrac{D \cdot R_i}{(1 + 1/A_i)(1 + D)} + R_L} = \frac{\dfrac{v'_1}{D'}}{R'_i + R_L}. \tag{14.4}$$

This relation for i'_2 represents a generator with the initial voltage v'_1/D' whose internal resistance is R'_i and which feeds a load resistance R_L. The equivalent characteristic values are consequently:

$$D' = 1 + D \approx 1$$

$$R'_i = \frac{D \cdot R_i}{(1 + D)(1 + 1/A_i)} \approx \frac{1}{S}$$

$$S' = \frac{1}{D' \cdot R'_i} = S\left(1 + \frac{1}{A_i}\right) \approx S.$$

VOLTAGE AMPLIFICATION A_{v_C}

If $A_v = v_2/v_1 = i_2 \cdot R_L/v_1$ is the voltage amplification in the common emitter circuit, the amplification A'_v in the common collector circuit based on the control voltage v_1 is somewhat greater because the input current i_1 also flows via R_L. We have:

$$A'_v = \frac{v'_2}{v_1} = \frac{(i_1 + i_2)\cdot R_L}{v_1} = \frac{i_2\left(\frac{1}{A_i} + 1\right)R_L}{v_1} = \frac{v_2\left(\frac{1}{A_i} + 1\right)}{v_1}$$

$$A'_v = A_v\left(1 + \frac{1}{A_i}\right). \tag{14.6}$$

Since the actual input voltage v'_1, however, is considerably greater than v_1 and is even greater than v_2, the voltage amplification in the common collector circuit is less than one. We have:

$$A_{v_C} = \frac{v'_2}{v'_1} = \frac{v'_2}{v_1 + v'_2} = \frac{\frac{v'_2}{v_1}}{1 + \frac{v'_2}{v_1}}$$

$$A_{v_C} = \frac{A_v\left(1 + \frac{1}{A_i}\right)}{1 + A_v\left(1 + \frac{1}{A_i}\right)} \approx \frac{A_v}{1 + A_v}. \tag{14.7}$$

CURRENT AMPLIFICATION A_{i_C}

If $A_i = A_v r_i / R_L = i_2/i_1$ is the current amplification in the common emitter circuit we have to take into account that the ouput current in the common collector circuit is greater by the input current i_1 than in the common emitter circuit. The following equation can therefore be applied:

$$A_{i_C} = \frac{i_1 + i_2}{i_1} = 1 + \frac{i_2}{i_1}$$

$$A_{i_C} = 1 + A_i. \tag{14.8}$$

INPUT RESISTANCE r_{i_C}

As the input voltage v'_1 does not lie at the resistance r_i and the output voltage e_2 acts against the input voltage v'_1, the input current i_1 is less than that in the common emitter circuit with the same input voltage. The effective input resistance r_{i_C} in the common collector circuit is therefore greater than r_i in the common emitter circuit and according to Fig. 14.3 is obtained as follows:

$$r_{i_C} = \frac{v'_1}{i_1} = \frac{v'_1}{\frac{v_1}{r_i}} = \frac{v'_1}{v_1}\cdot r_i\,.$$

If we replace v'_1 in this by Formula (14.1) and then replace v'_2/v_1 through Equation (14.6) we obtain:

$$r_{i_C} = \frac{v'_2 + v_1}{v_1} \cdot r_i = r_i\left(1 + \frac{v'_2}{v_1}\right)$$

$$r_{i_C} = r_i\left[1 + A_v\left(1 + \frac{1}{A_i}\right)\right] \approx r_i(1 + A_v), \tag{14.9}$$

since $A_v \approx S \cdot R_L$, $r_i \cdot A_v = r_i \cdot S \cdot R_L = (1 + \beta_0)R_L$ so also

$$r_{i_C} \approx r_i + (1 + \beta_0) \cdot R_L \tag{14.10}$$

EQUIVALENT CIRCUIT DIAGRAM

With the equivalent characteristic values found so far an equivalent circuit diagram can now be drawn which corresponds to the fundamental equivalent circuit diagram because the equivalent characteristic values for a transistor were derived from the common emitter circuit.

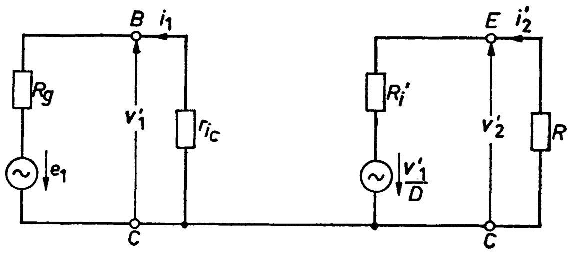


Fig. 14.4

Example 14.1

A transistor T_2 with the following data is to be operated in the common collector circuit. It is to be connected as an impedance transformer following a control generator with $e_1 = 10$ mV and $R_g = 61.2$ kΩ, and has to feed a load resistance $R_L = 2$ kΩ. All the equivalent and operating values have to be found as well as the equivalent circuit diagram with all the values. Given:

$$S = 13.6 \text{ mA/V} \quad R_i = 93.5 \text{ k}\Omega \quad D = 0.788 \cdot 10^{-3}$$

$$r_{i_0} = h_{11} = 2.2 \text{ k}\Omega \quad D_v = h_{12} = 9 \cdot 10^{-4} \quad R_L = 2 \text{ k}\Omega.$$

Solution

$$A_v = S\frac{R_i \cdot R_L}{R_i + R_L} = 13.6\frac{93.5 \cdot 2}{95.5} = 26.6$$

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v} = \frac{2.2}{1 + 26.6 \cdot 9 \cdot 10^{-4}} = 2.15 \text{ k}\Omega$$

$$A_i = A_v \frac{r_i}{R_L} = 26.6 \frac{2.2}{2} = 29.2$$

$$R'_i = \frac{1}{S} \frac{1}{1 + \frac{1}{A_i}} = \frac{10^3}{13.6\left(1 + \frac{1}{29.2}\right)} = 71\ \Omega$$

$$S' = S\left(1 + \frac{1}{A_i}\right) = 13.6\left(1 + \frac{1}{29.2}\right) = 14\ \text{mA/V}$$

$$D' \approx 1$$

$$r_{i_C} = r_i\left[1 + A_v\left(1 + \frac{1}{A_i}\right)\right] = 2.15\left[1 + 26.6\left(1 + \frac{1}{29.2}\right)\right] = 61.2\text{k}\Omega$$

$$A_{v_C} = \frac{A_v\left(1 + \frac{1}{A_i}\right)}{1 + A_v\left(1 + \frac{1}{A_i}\right)} = \frac{26.6 \cdot 1.034}{1 + 26.6 \cdot 1.034} = 0.965$$

$$A_{i_C} = 1 + A_i = 1 + 29.2 = 30.2$$

$$v'_1 = e_1 \frac{r_{i_C}}{R_g + r_{i_C}} = 10 \frac{61.2}{61.2 + 61.2} = 5\ \text{mV}$$

$$i_1 = \frac{v'_1}{r_{i_C}} = \frac{5 \cdot 10^{-3}}{61.2 \cdot 10^3} = 0.08\ \mu\text{A}$$

$$v'_2 = v'_1 \cdot A_{v_C} = 5 \cdot 0.965 = 4.8\ \text{mV}$$

$$i'_2 = i_1 \cdot A_{i_C} = 0.08 \cdot 30.2 = 2.4\ \mu\text{A}.$$

See also Example 11.1.

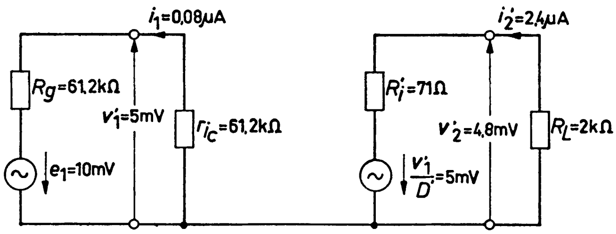


Fig. 14.5

14.1. The equivalent characteristic values of the common collector stage

Equivalent characteristic values for the transistor or effective characteristic values for the stage are also valid in the common collector stage if for example the influence of the control generator internal resistance has to be taken into account, with a variation of the load resistance of the stage. Here the control generator resistance is also added to the transistor. The equivalent internal resistance R_{i_S} is obtained as follows:

The current amplification in the collector base circuit is

$$A_{i_C} = 1 + A_i \approx 1 + \beta_0 \text{ et } r_i \approx r_{i_0}$$

because of the generally low resistance loading at the output. If we think of the current i_2 which flows in the load resistance R_L and is produced through the input current i_1, as flowing in full amount in the input circuit, all the resistances in this circuit must be reduced by the factor $1/(1 + \beta_0)$ so that all the voltage drops keep their original value.

For the drive source initial voltage all the resistances have been reduced by the factor $1/(1 + \beta_0)$. Even the load resistance which normally occurs as $R_L \cdot (1 + \beta_0)$ was multiplied by this factor because the current i_1 driven by e_1 appears amplified by i_1 $(1 + \beta_0)$ in resistance R_L, though this can be allowed for by a correspondingly larger resistance $R_L \cdot (1 + \beta_0)$ if we think of i_1 alone flowing via R_L.

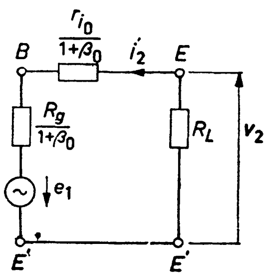


Fig. 14.6

From the equivalent circuit in Fig. 14.6 we see that the load resistance R_L lies at a generator whose internal resistance is equal to the sum of all the other resistances. We therefore have:

$$R_{i_{eq}} = \frac{R_g}{1 + \beta_0} + \frac{r_{i_C}}{1 + \beta_0}; \quad \frac{r_{i_C}}{1 + \beta_0} \approx \frac{1}{S}$$

$$R_{i_{eq}} = \frac{1}{S} + \frac{R_g}{1 + \beta_0}. \tag{14.11}$$

Since $\frac{1}{S}$ is usually $< \frac{R_g}{1+\beta_0}$ we have:

$$R_{i_{eq}} \approx \frac{R_g}{\beta}. \tag{14.12}$$

If the load resistance in Fig. 14.1 is considered as short-circuited to determine the slope S_{eq}, the input resistance to which the generator resistance R_g is connected in series, is then only r_{i_0}. The generator voltage e_1 therefore is effective as the input of the transistor reduced according to the voltage division. The slope S_{eq} is correspondingly less and we have:

$$S_{eq} = S \frac{r_{i_0}}{R_g + r_{i_0}}. \tag{14.13}$$

The reciprocal amplification is obtained from:

$$D_{eq} = \frac{1}{R_{i_{eq}} \cdot S_{eq}} = 1 \tag{14.14}$$

because

$$D_{eq} = \frac{R_g + r_{i_0}}{S \cdot r_{i_0}} \frac{1}{\frac{1}{S} + \frac{R_g}{\beta_0}} \qquad S \cdot r_{i_0} = \beta_0$$

$$D = \frac{R_g + r_{i_0}}{r_{i_0} + R_g} = 1.$$

Example 14.2

The equivalent transistor characteristic values are to be calculated for the common collector stage in Example 14.1. We had:

$$S = 13.6\ \text{mA/V},\ \beta_0 = S \cdot r_{i_0} = 30,\ R_g = 61.2\ \text{k}\Omega,\ e_1 = 10\ \text{mV}.$$

Solution

$$R_{i_{eq}} = \frac{1}{S} + \frac{R_g}{1+\beta_0} = \frac{10^3}{13.6} + \frac{61.2 \cdot 10^3}{1+30}$$
$$= 73.5 + 1970 \approx 2\ \text{k}\Omega$$

$$S_{eq} = S \frac{r_{i_0}}{R_g + r_{i_0}} = 13.6 \cdot 10^{-3} \frac{2.2}{61.2 + 2.2} = 0.472\ \text{mA/V}$$

$$D_{eq} = \frac{1}{R_{i_{eq}} \cdot S_d} = \frac{1}{2 \cdot 10^3 \times 0.472 \cdot 10^{-3}} = 1.06 \approx 1.$$

Check

$$v_2 = e_1 \cdot S_{eq} \frac{R_{i_{eq}} \cdot R_L}{R_{i_{eq}} + R_L} = 10 \cdot 10^{-3} \times 0.472 \cdot 10^{-3} \frac{2 \cdot 2}{2 + 2} 10^3 = 4.72\,\text{mV}.$$

This result practically agrees with that in Example 14.1.

CHAPTER 15

THE COMMON BASE CIRCUIT

If the base forms the common connecting point between input and output circuits we then have the common base circuit (Fig. 15.1).

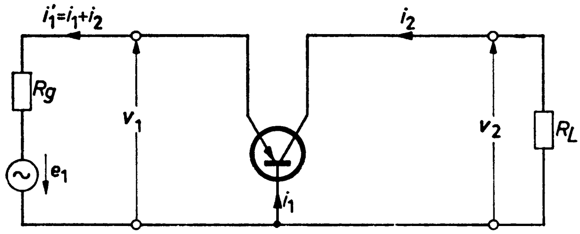


Fig. 15.1

The calculation of the common base circuit for alternating current will be derived from the common emitter circuit. Here too, the aim is to produce an equivalent circuit which corresponds to the common emitter equivalent circuit with its equivalent characteristic values. The circuit diagram in Fig. 15.1 can be represented with the equivalent circuit diagram found for the common emitter circuit, as Fig. 15.2 shows. The only difference is that the load resistance for alternating current does not lie at the emitter but at the base.

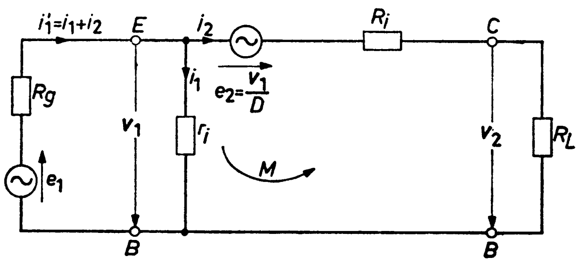


Fig. 15.2

All the characteristic values of the common emitter circuit are included in this circuit diagram. It only differs from this circuit in the fact that the load resistance for alternating current is connected to the base. The load current as well as the base current now flows via the control source and the following statement applies:

$$i'_1 = i_1 + i_2 = i_1(1 + A_t). \tag{15.1}$$

The input current i'_1 in the common base circuit exceeds that in the common emitter circuit by the factor $(1 + A_t)$.

The equivalent characteristic values are obtained from the equation for network M in Fig. 15.2.

$$v_1 - i_2 \cdot R_t - i_2 \cdot R_L = -\frac{v_1}{D}; \quad v_1\left(1 + \frac{1}{D}\right) = i_2(R_t + R_L).$$

If we resolve for i_2 we obtain:

$$i_2 = \frac{v_1}{\dfrac{1}{1 + 1/D}} \frac{1}{R_t + R_L} = \frac{\dfrac{v_1}{D'}}{R'_t + R_L}. \tag{15.2}$$

This equation again represents a generator whose initial voltage is v'_1/D' and whose internal resistance is not altered, in contrast to the common emitter circuit. The equivalent characteristic values for the common base circuit are therefore:

$$\begin{aligned} D' &= \frac{D}{1+D} \approx D \\ R'_t &= R_t \\ S' &= \frac{1}{D' \cdot R'_t} = S(1 + D) \approx S. \end{aligned} \tag{15.3}$$

The equivalent characteristic values for the common base circuit only deviate very slightly from those for the common emitter circuit.

INPUT RESISTANCE r_{t_B}

The input resistance r_{t_B} in the common base circuit differs very greatly from the common emitter circuit because the total output current also flows via the input. It follows from Equation (15.1) that the input resistance must decrease in inverse proportion to the current increase by the factor $(1 + A_t)$. We therefore have:

$$r_{t_B} = \frac{r_t}{1 + A_t} \quad . \tag{15.4}$$

The input resistance r_t, however, can also be represented in another way. We see from Fig. 15.2 that a branch which carries the current i_2 is connected in parallel to the resistance r_i. The resistance of this lead is obviously v_1/i_2, since i_2 flows as the result of the voltage v_1 across the parallel circuit. If we resolve Equation (15.2) for v_1/i_2 we obtain:

$$\frac{v_1}{i_2} = D'(R'_i + R_L).$$

This resistance in parallel to r_i is thus also the input resistance r_i.

$$r_{i_B} = r_i \| D'(R'_i + R_L). \qquad (15.5)$$

Since now R_L is mostly less than R'_i:

$$r_{i_B} \approx r_i \| \frac{1}{S}. \qquad (15.6)$$

Also, as r_i is generally greater than $1/S$ in the parallel circuit, we can almost always put:

$$r_{i_B} \approx \frac{1}{S}. \qquad (15.7)$$

CURRENT AMPLIFICATION A_{i_B}

As the control current plus the output current flow as the input current in the common base circuit, the current amplification must be less than one. According to Eq. (15.1) we have:

$$i'_1 = i_1(1 + A_i) = \frac{i_2}{A_i}(1 + A_i)$$

$$A_{i_B} = \frac{i_2}{i'_1} = \frac{A_i}{1 + A_i}. \qquad (15.8)$$

This formula naturally also applies for the short circuit current amplifications:

$$\alpha = \frac{\beta_0}{1 + \beta_0}. \qquad (15.9)$$

Since the current amplification in the common base circuit is less than one, this circuit is only important as a voltage amplifier.

VOLTAGE AMPLIFICATION A'_v

With the equivalent characteristic values we obtain the following for the voltage amplification:

$$A'_v = S' \frac{R'_i \cdot R_L}{R'_i + R_L} = S(1 + D) \frac{R_i \cdot R_L}{R_i + R_L}$$

$$A'_v = A_v(1 + D) \approx A_v. \qquad (15.10)$$

The voltage amplification is somewhat greater in the common base circuit than in the common emitter circuit.

EQUIVALENT CIRCUIT DIAGRAM

The equivalent circuit in Fig. 15.3 is based on the equivalent characteristic values derived. It is again a fundamental equivalent circuit with different characteristic values for the common base circuit.

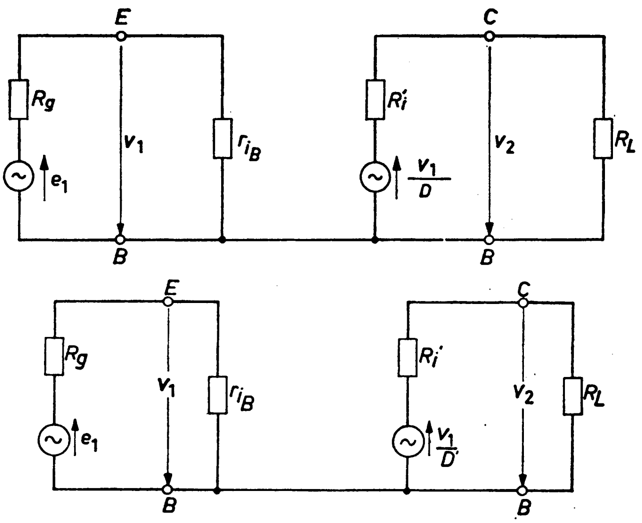


Fig. 15.3

Example 15.1

A transistor T_3 (e.g. the OC 70) is to be operated in the common base circuit. It has the characteristic values:

$S = 31$ mA/V $\quad R_i = 0.5$ MΩ

$r_{i_0} = 2.5$ kΩ $\quad D_v = h_{12} = 0.25 \times 10^{-3}$.

The control generator has the value $e_1 = 100\ \mu$V and $R_g = 300\ \Omega$. What working values arise with a load resistance $R_L = 1$ kΩ?

Solution

The reciprocal amplification is:

$$D = \frac{1}{S \cdot R_i} = \frac{10^3}{31 \times 0.5 \cdot 10^6} = 6.45 \cdot 10^{-5}.$$

The value $1 + D$ is obviously almost equal to one, i.e. the equivalent characteristic values can be equated with the characteristic values S, R_i

and D in the common emitter circuit and we then obtain the following values:

$$A_{v_B} \approx S' \cdot R_L = 31 \cdot 1 = 31$$

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v} = \frac{2.5}{1 + 31 \times 0.25 \cdot 10^{-3}} = 2.5\ \text{k}\Omega$$

$$A_i = A_v \frac{r_i}{R_L} = 31 \frac{2.5}{1} = 77.5$$

$$A_{i_B} = \frac{A_i}{1 + A_i} = \frac{77.5}{78.5} = 0.988$$

$$r_{i_B} = r_i \parallel \frac{1}{S} = 2.5\ \text{k}\Omega \parallel 32.2\ \Omega = 32\ \Omega$$

$$v_1 = e_1 \frac{r_{i_B}}{r_{i_B} + R_g} = 100 \frac{32}{332} = 9.64\ \mu\text{V}$$

$$v_2 = v_1 \cdot A_{v_B} = 9.64 \cdot 31 = 300\ \mu\text{V}$$

$$i_1 = \frac{v_1}{r_{i_B}} = \frac{9.64 \cdot 10^{-6}}{32} = 0.3\ \mu\text{A}$$

$$i_2 = i_1 \cdot A_i = 0.3 \times 0.988 = 0.296\ \mu\text{A}.$$

15.1. The equivalent characteristic values of the common base stage

The equivalent characteristic values $r_{i_{eq}}$, S_{eq} and D_{eq} can also be determined in the common base stage. Here the control generator resistance has to be taken into account with a change of the load resistance. These characteristic values, also shown as operating characteristic values, are obtained with the same formulae which were given in Chapter 8 for the common emitter circuit. Here the control generator resistance R_g is therefore also attached to the transistor, and the constant generator initial voltage is regarded as the input voltage. It is thus necessary to note that in calculating the equivalent characteristic values, the special characteristic values of the common base circuit have to be used because the amount of the voltage reaction is variable and in the reverse direction to that effective in the common emitter circuit. Consequently, in using Formula (8.3), for example, the sign in the second section of the denominator has to be changed. It therefore also happens that in the common base circuit with a large control generator resistance R_g (input side open circuited), the internal resistance R_i^* is greater than the normal internal resistance R_i for low resistance control. In the common emitter circuit it is known that the open circuited internal resistance R_i^* is smaller than the short circuit internal resistance R_i. As we have already seen, the amplification values can only be calculated with the assumption of low resistance control (input side short circuited) with the normal characteristic values because these are not dependent on the control generator resistance. The following characteristic amounts of the common base circuit are necessary to calculate the equivalent characteristic values:

1. The short-circuit input resistance $r_{i_{0b}} = h_{11_b}$
2. The open circuit internal resistance $R^*_{i_b} = 1/h_{22_b}$
3. The voltage reaction $D_{v_b} = h_{12_b}$

In this way we obtain the equivalent internal resistance according to Formula (8.3) from:

$$R_{i_{eq}} = \frac{1}{\dfrac{1}{R^*_{ib}} + \dfrac{D_{vb} \cdot \alpha_0}{r_{i0b} + R_g}}. \tag{15.11}$$

The equivalent slope is derived according to Formula (8.2) from:

$$S_{eq} = \frac{\alpha_0}{r_{i0b} + R_g}. \tag{15.12}$$

For the equivalent reciprocal amplification we obtain:

$$D_{eq} = \frac{1}{R_{i_{eq}} \cdot S_{eq}}. \tag{15.13}$$

Formula (8.5) can also be used for the new input resistance by taking into account the phase-rotated voltage reaction, and we can therefore apply:

$$r_{ib} = \frac{r_{i0b}}{1 - A_v \cdot D} + R_g. \tag{15.14}$$

CHAPTER 16

HIGH FREQUENCY BEHAVIOUR OF THE TRANSISTOR

In its behaviour at high frequencies, the transistor is much more difficult to represent than the tube. This is because the internal physical processes are more complicated, and compared with the ideal transistor, additional parasitic influences come into effect. So if we want to represent a transistor with absolute exactness an enormous mathematical process is necessary at high frequencies which in turn leads to complicated equivalent circuits.
In practice it is better to obtain an approximate solution through simple and clear representation and because of the spread in production of a particular type this does not usually need to be absolutely accurate. Apart from this, it is also much better for certain frequencies or frequency ranges to use the relevant characteristic values given by some manufacturers for certain transistor types. The most important relationships in the transistor for high frequency amplification will be set out, therefore, in the following section. These calculations are quite accurate enough for practical purposes within the range of the limiting frequencies described.

All the considerations are based on the fundamental circuit, the common emitter circuit.

16.1. Transistor capacitances

The a.c. behaviour of the transistor is determined at high frequencies by the transistor capacitances. In essence these are, as in the tube in the common emitter circuit, the input capacitance C_i, the reaction capacitance C_r and the output capacitance C_o. Compared with the tube capacitances, especially in the pentode, these capacitances are relatively larger and are also greatly dependent on the working point. Their influence is therefore greater than in the tube. In principle, we can think of two kinds of capacitance at the transistor barrier layer transitions, the barrier layer capacitance and the diffusion capacitance.

16.1.1. BARRIER LAYER CAPACITANCE C_r (REACTION CAPACITANCE)

The collector-base barrier layer is operated in the blocking direction. With this the capacitance is mainly decided by the barrier layer capacitance and the diffusion capacitance can virtually be disregarded. The barrier layer capacitance is formed by the charge-carrier deficient interface which acts as the dielectric of a capacitor. The two resulting zones at this boundary, the collector and the base thus form the foundation of this capacitor which takes up a charge according to the applied voltage and its capacitance C_r. As the blocking zone, with its charge-free zone, continually widens with increasing voltage, the sources of the capacitor also diverge further and further from each other and the capacitance decreases with increasing voltage.

The relation:

$$C_{cb} = C_r \approx \frac{K}{\sqrt{V_{cb}}}. \tag{16.1}$$

can be applied for the collector-base capacitance, the reaction capacitance C_r. This capacitance represents an analogue to the grid-anode capacitance of the tube.

16.1.2. DIFFUSION CAPACITANCE C_i (INPUT CAPACITANCE)

The base-emitter boundary layer is operated in the transmission direction. Here the diffusion capacitance chiefly takes effect. This capacitance results from the relatively great transit time of the charge carriers in the base area particularly when the charge carriers only move through the diffusion effect. The charge carriers arrive at the collector noticeably delayed with increasing frequency, i.e. the output current i_2 chases the input current i'_1 (Fig. 16.1). If the input current i'_1 is kept constant by means of a high resistance control generator, the phase angle between i_2 and i'_1 increases with rising frequency and the amplitude of i_2 decreases. This is the result of the carrier transit time in the base area where, with constant input current and increasing frequency, fewer charge carriers arrive at the collector (amplitude loss) or the charge carriers which do reach the colllector arrive delayed (phase rotation) (Fig. 16.1).

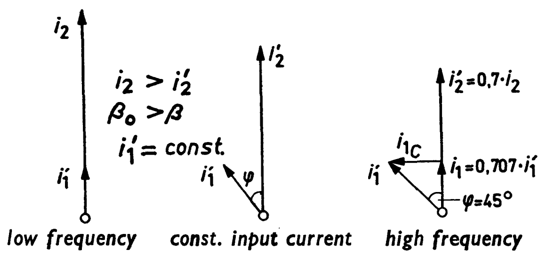


Fig. 16.1

The current amplification $\beta = i_2/i_1$ obviously decreases with increasing frequency.

If the control generator is of low resistance the input control voltage v_1 is constant. (The base area resistance at first plays no part at the lower end of the high frequency band). If we now allow the frequency to rise, starting with very low frequencies, we find that the input current increases and the output current remains constant. Naturally, the phase angle φ and the reduction of current amplification β are now the same, in spite of the fact that the output current remains constant. The input resistance is obviously less as a result of the diffusion process, i.e. the carrier transit time in the base area. The charge carriers which do not reach the collector now form a capacitive displacement current to the base electrode and so enlarge the input current or reduce the input resistance as the case may be. This fact can now be explained through the diffusion capacitance. Capacitance C_i is produced as result of the diffusion of the charge carriers in the base area and with increasing frequency this increases the input current from the base to the emitter; it is therefore connected as an input capacitance C_i in parallel with the input resistance r_i (which we have with very low frequencies) (Fig. 16.2).

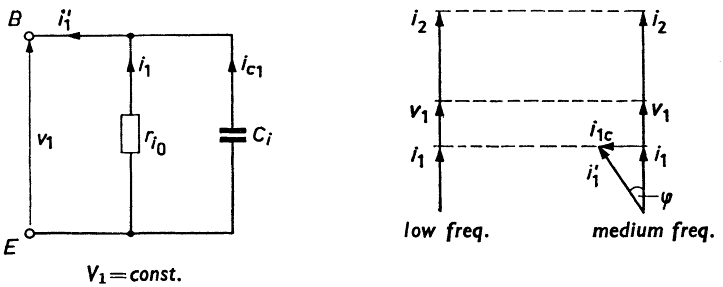


Fig. 16.2

Because of the diffusion capacitance, the input capacitance C_i, the input current i'_1 becomes larger with increasing frequency, since there is an increase in the capacitive current i_{C_1}. The effective component of the input current (the component in phase with the voltage v_1) i.e. the current i_1, can now be regarded as the actual control current since i_2 only depends on this. Therefore, if the voltage at r_i is constant, i_1 also remains constant with rising frequency and so i_2 is also constant because $i_2 = i_1 \cdot \beta_0$. The effective current amplification $\beta = i_2/i'_1$, however, decreases when i_{C_1} increases because i'_1 is correspondingly increased. The input capacitance depends on the working point I_C. We can calculate it from:

$$C_i \approx 6.2 \frac{I_C\,[\mathrm{A}]}{f_\alpha\,[\mathrm{c/s}]}\ [\mathrm{F}].$$

Here f_α is the α limiting frequency yet to be discussed.

16.1.3. THE OUTPUT CAPACITANCE C_o

If at first we regard the output capacitance as being only the capacitance of the collector electrode to the emitter electrode, it will be formed solely through the leads and the electrodes themselves. This capacitance C_o is in the order of 1 pF. The total output capacitance will nevertheless be increased through the reaction capacitance as can be seen from the equivalent circuit diagram below.

16.2. The high frequency equivalent circuit diagram

In the normal fundamental equivalent circuit diagram for low frequencies (Fig. 6.1) the input resistance

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v}$$

is connected at the input terminals. This input resistance r_i results from the short-circuit input resistance r_{i_0} with an additional resistance R_r' in parallel with it and produced by the voltage reaction. We can insert a reaction resistance in the fundamental equivalent circuit diagram and so obtain Fig. 16.3.

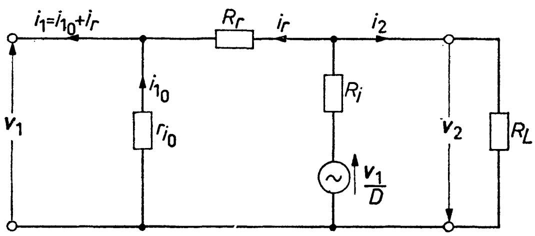


Fig. 16.3

The voltage reaction is represented here through the reaction resistance R_r. A greater input current flows because of the presence of R_r. We can therefore imagine a resistance R_r' directly in parallel with r_{i_0}. Since R_r lies at a voltage $v_1 + v_2 = v_1(1 + A_v)$ while R_r' is only at v_1, the resistance R_r' must be correspondingly less with the same current intake. We therefore obtain the old input resistance from:

$$r_i = \frac{r_{i_0} \cdot R'_r}{r_{i_0} + R'_r} = \frac{r_{i_0}}{1 + D_v \cdot A_v}.$$

If we resolve this equation for R_r' we have:

$$R'_r = \frac{r_{i_0}}{D_v \cdot A_v}.$$

We thus obtain the reaction resistance R_r from:

$$R_r = R'_r(1 + A_v) = \frac{r_{i_0}(1 + A_v)}{D_v A_v}$$

$$R_r \approx \frac{r_{i_0}}{D_v}. \qquad (16.3)$$

In Fig. 16.3 the voltage reaction is accounted for through the reaction resistance between collector and base. The reaction capacitance C_r already mentioned lies in parallel with this resistance and at high frequencies it therefore causes an additional voltage reaction. If we also take the input and output capacitances into consideration we obtain the high frequency equivalent circuit diagram (Fig. 16.4).

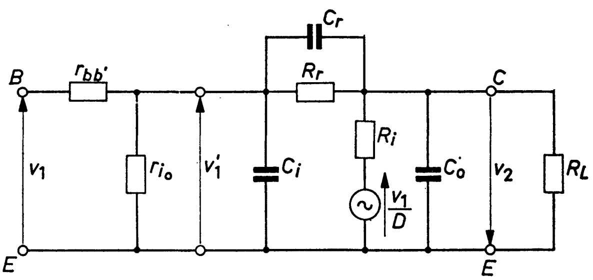


Fig. 16.4

It can be seen from this that the input capacitance C_i does not lie directly across the input terminals B/E. The reason is that in construction the material resistance, i.e. the "base area resistance" r_{bb}' lies between the actual base space B' which is only very small, and the base electrode.

This resistance is included in the input resistance at low frequencies and receives no further emphasis there. At very high frequencies, however, the series circuit of r_{bb}' and C_i is directly decisive for the frequency response because the actual control voltage v'_1 gradually decreases and the current in the input resistance, the control current, approaches zero. The dimension of the base area resistance is about 100 Ω and can be disregarded at low frequencies. With rising frequency the current amplification β will nevertheless decrease because of the growing capacitive input current. The voltage amplification, however, will at first still remain constant since the control voltage v'_1, through $r_{bb'}$, only decreases at very high frequencies.

The effective input resistance is reduced through the reaction capacitance. The resistance r_r lies parallel to this capacitance and both connect the collector with the base. As in parallel feedback we can regard this reaction resistance as operating at the input reduced by $1/(1 + A_v)$, or consider the reaction capacitance as increased by the factor $(1 + A_v)$. All the same, we have to take into account here that in the vicinity of the limiting frequency for the voltage amplification there is a phase rotation between the control voltages v'_1 and v_2 because of the capacitive loading of the output through C_i. The extra input current through $C_r(1 + A'_v)$ thus has not the same phase position as the capacitive input current through C_i. Its capacitive component therefore enlarges the capacitive input current to a lesser amount than would correspond to the capacitance $C_r(1 + A'_v)$. Because of the additional reaction capacitance the effective input capacitance C'_i depends on the amplification A_v and on the frequency.

$$C'_i = C_i + C_r(1 + A_v)K_f. \tag{16.4}$$

In this expression K_f is less than 1 depending on the frequency.

The effective output capacitance is given, as the circuit shows, with the capacitance C_o to which the series circuit of C_r and C_i is connected in parallel. In this series circuit the capacitance is virtually determined through C_r because C_i is less than C_r. As a result the following statement is valid for the effective output capacitance:

$$C'_o \approx C_o + C_r. \tag{16.5}$$

16.3. Limiting frequencies

"Limiting frequencies" are also given with tubes, e.g. an upper and a lower limiting frequency. There these critical frequencies are dependent on external circuit components if we ignore transit time effects which with the tube are only important in centimetric wavelengths. With the transistor, limiting frequencies can now be given as these already exist without taking external circuit components into account, a fundamental difference from the tube.

16.3.1. THE LIMITING FREQUENCY OF SHORT-CIRCUIT CURRENT AMPLIFICATION

Short-circuit current amplification at low frequencies is denoted by β_0 in the common emitter circuit or by α_0 in the common base circuit. Short-circuit current amplification means that the load resistance $R_L = 0$, the collector voltage is constant during modulation and there is no resulting reaction via the reaction capacitance C_r. If the short-circuit current amplification β is measured at higher frequency, the input capacitance C_i acts as a capacitive shunt to the input resistance r_{i_0}. The result is that with an input current i'_i which is kept constant (Fig. 16.1) the actual control current in resistance r_{i_0} becomes progressively less; the current amplification therefore decreases.

16.3.1.1. The Limiting Frequency f_β

If with constant input current supply i'_i, the output current drops at a frequency f_β to 70% of the value it had at low frequency, f_β is the β-limiting frequency. In this case, as Fig. 16.1 shows, the control current i_1 in resistance r_{i_0} has become equal to the capacitive current i_{1_C} in the input capacitance C_i. The control current then only amounts to $i_1 = i'_1 \cdot 0.7$ and the current amplification $\beta = \beta_0 \cdot 0.7$.

The resistances r_i and $1/C_i \cdot 2\pi f_\beta$ are equal at the limiting frequency f_β and from this condition we obtain the β-limiting frequency from:

$$f_\beta = \frac{1}{r_{i_0} \cdot C_i \cdot 2\pi}. \tag{16.6}$$

The input capacitance C_i can also be calculated from this relation:

$$C_i = \frac{1}{r_{i_0} \cdot 2\pi \cdot f_\beta}. \tag{16.7}$$

Here f_β is independent of the working point because this frequency is only determined by the diffusion effect, the transit time of the charge carriers in the base area. If, however, $r_{i_0} = h_{11}$ varies through a displacement of the working point, the effective input capacitance according to (16.7) or (16.4) will also be changed.

The control current from the input current i'_1 can be calculated as follows from the parallel circuit in Fig. 16.2, corresponding to the current distribution:

$$i_1 = i'_1 \frac{\dfrac{1}{\mathrm{j} C_i \omega}}{r_{i_0} + \dfrac{1}{\mathrm{j} C_i \omega}} = i'_1 \frac{1}{1 + \mathrm{j} r_{i_0} \cdot C_i \omega}$$

$$|i_1| = i'_1 \frac{1}{\sqrt{[1 + (r_{i_0} \cdot C_i \cdot 2\pi \cdot f)^2]}}. \tag{16.8}$$

It can be seen from the above relation how the actual control current i_1 decreases with rising frequency. The current amplification β is reduced in the same way and we have:

$$\beta = \beta_0 \frac{1}{\sqrt{[1 + (r_{i_0} \cdot C_i \cdot 2\pi \cdot f)^2]}}. \tag{16.9}$$

Since now $r_i \cdot C_i \cdot 2\pi = 1/f_\beta$, we can also set down:

$$\beta = \beta_0 \frac{1}{\sqrt{\left[1 + \left(\frac{f}{f_\beta}\right)\right]^2}}. \tag{16.10}$$

16.3.1.2. The Limiting Frequency f_α

In the base circuit the resistance $1/S$ lies in parallel with the input resistance r_i according to Eq. (15.5). Both resistances are again connected in parallel with the input capacitance C_i (Fig. 16.5) in which $r_i = r_{i_0}$ for the following considerations because f_α is also valid for $R_L = 0$.

The short-circuit input resistance in the base circuit $r_{i_B} = r_{i_0} \parallel 1/S$ at low frequency is less than the short-circuit input resistance r_{i_0} in the fundamental circuit. Since the same capacitance C_i is connected parallel to this resistance r_{i_B}, the limiting frequency f_α obviously has to be higher now than in the fundamental circuit because the capacitive resistance $1/C_i \cdot 2\pi \cdot f_\alpha$ must be smaller at the limiting frequency f_α. The input control current i_1 drops to 70% with constant input current i'_1 if

$$r_{i_0} \parallel \frac{1}{S} \text{ becomes equal to } \frac{1}{C_i \cdot 2\pi f_\alpha}.$$

Then the voltage drop at the parallel circuit is also only 70%. The α-limiting frequency is thus obtained from:

$$f_\alpha = \frac{S\left(r_{i_0} + \frac{1}{S}\right)}{r_{i_0} \cdot C_i \cdot 2\pi} = \frac{1 + S \cdot r_{i_0}}{r_{i_0} \cdot C_i \cdot 2\pi}$$

$$f_\alpha = \frac{\frac{1}{r_{i_0}} + S}{C_i \cdot 2\pi}. \tag{16.11}$$

The α-limiting frequency is thus always higher than the β-limiting frequency (16.6).

According to Equation (16.10) the current amplification can be derived in the common base circuit from the input parallel circuit if α_0 is known at low frequencies. We then have:

$$\alpha = \frac{\alpha_0}{\sqrt{\left[1 + \left(\frac{f}{f_\alpha}\right)^2\right]}}. \tag{16.12}$$

16.3.1.3. Converting f_α into f_β

We have:

$$f_\beta = \frac{1}{r_{t_0} \cdot C_t \cdot 2\pi} \qquad f_\alpha = \frac{1}{r_{t_B} \cdot C_t \cdot 2\pi}$$

$$\frac{f_\alpha}{f_\beta} = \frac{r_{t_0}}{r_{t_B}}. \tag{16.13}$$

The input short-circuit current i_{1_α} in the common base circuit is obtained from the input short-circuit current i_1 in the common emitter circuit when the output current $i_2 = i_1 \cdot \beta_0$ is added to it because in the common base circuit this flows with it via the input.

$$i_{1_\alpha} = i_1 + i_1 \cdot \beta_0 = i_1(1 + \beta_0). \tag{16.14}$$

As the input short-circuit current in the common base circuit is increased by the factor $(1 + \beta_0)$, the input short-circuit resistance must have the value:

$$r_{t_B} = \frac{r_{t_0}}{1 + \beta_0}. \tag{16.15}$$

If we insert (16.15) into (16.13) the following expression can be applied:

$$\frac{f_\alpha}{f_\beta} = 1 + \beta_0 = \frac{\beta_0}{\alpha_0}. \tag{16.16}$$

if we take into account that according to Equation (15.9):

$$\alpha_0 = \beta_0/(1 + \beta_0).$$

16.3.1.4. The frequency f_1 for $\beta = 1$

According to Eq. (15.10) the short-circuit current amplification decreases with rising frequency. At a certain frequency which we call f_1 the short-circuit current amplification eventually only has the value $\beta = 1$. If we ignore the figure 1 under the root sign in Formula (16.10) we have for $\beta = 1$:

$$\beta_0 \approx \frac{f_1}{f_\beta} \quad \text{or} \quad f_1 \approx f_\beta \cdot \beta_0. \tag{16.17}$$

Since $f_\alpha = f_\beta \frac{\beta_0}{\alpha_0} \approx f_\beta \cdot \beta_0$ according to Eq. (16.16) we can also state that the frequency f_1 is about equal to the α-limiting frequency.

$$f_1 \approx f_\alpha. \tag{16.18}$$

Practical values will be explained in an example.

Example 16.1

The following values are given for a transistor T_3:

$\beta_0 = 100$	$f_1 = 70$ Mc/s
$f = 450$ kc/s	$f = 10.7$ Mc/s
$g_{11} = 0.4$ mS	$g_{11} = 2.5$ mS
$b_{11} = 0.23$ mS	$b_{11} = 4.4$ mS
$C_{11} = 80$ pF	$C_{11} = 65$ pF
$g_{12} = 0.1\ \mu$S	$Y_{12} = 0.1$ mS
$b_{12} = 5.1\ \mu$S	
$C_{12} = 1.8$ pF	
$\lvert Y_{21}\rvert = 37$ mS	$\lvert Y_{21}\rvert = 32$ mS
$g_{22} = 1.1\ \mu$S	$g_{22} = 60\ \mu$S
$b_{22} = 14\ \mu$S	$b_{22} = 303\ \mu$S
$C_{22} = 5$ pF	$C_{22} = 4.5$ pF

The above values are given for calculation with conductance coefficients in the relevant four-pole equation. For example, the input short-circuit conductance would be given by $Y_{11} = g_{11} + jb_{11}$. All the values for the high frequency equivalent circuit can be estimated from the conductance values given, according to (16.18).

$$f_1 \approx f_\alpha = 70 \text{ Mc/s, according to (16.16) } f_\beta = \frac{f_\alpha}{\beta_0} = \frac{70}{100} = 0{,}7 \text{ Mc/s.}$$

Input capacitance and input resistance

An input capacitance $C_i'' = C_{11} = 80$ pF is given for $f = 450$ kc/s; with $f = 10.7$ Mc/s, $C_i''' = C_{11} = 65$ pF. The difference between the given capacitances is due to the base area resistance as can be observed from Fig. 16.4. With $f = 10.7$ Mc/s, the capacitive resistance of C_i''' comes into the same order as resistance $r_{bb'}$. The input capacitance C_i then no longer lies at the input voltage v_1 but at the smaller voltage v'_1. The capacitive loading of the input voltage v_1 at $f = 10.7$ Mc/s therefore becomes relatively less and it seems as though a reduced capacitance now lies at the input voltage v_1. With $f = 450$ kc/s the capacitive short-circuit input conductance b_{11} is 0.23 mS, or the capacitive input resistance is:

$$r_{i_b} \approx \frac{1}{b_{11}} = \frac{10^3}{0.23} = 4.35 \text{ k}\Omega.$$

This resistance lies in parallel with the short-circuit input resistance:

$$r_{i_0} = \frac{1}{g_{11}} = \frac{10^3}{0.4} = 2.5 \text{ k}\Omega.$$

Since the base area resistance is of the order of 100 Ω, we can say when $f =$ 450 kc/s that according to the size of the capacitive resistance $r_{i_b} = 4.35$ kΩ, there is no voltage division with r_{bb}' and the capacitance to be calculated from r_{i_b}

$$C_i'' = \frac{1}{r_{i_b} \cdot \omega} = C'_i$$

is practically equal to the input capacitance C_i' because both lie at about the same voltage (Fig. 16.6). As b_{11} applies for output-side short-circuit, the actual input capacitance according to (16.4) is obtained from:

$$C_i = C'_i - C_r = 80 - 1.8 = 78.2 \text{ pF}$$

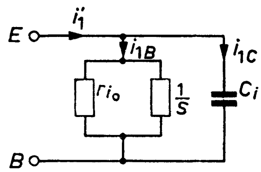


Fig. 16.5

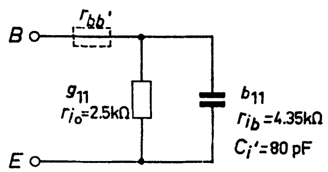


Fig. 16.6

Calculating the base area resistance r_{bb}'

The resistances r_{i_0} and r_{i_b} calculated from g_{11} and b_{11} already include the base area resistance. If we imagine r_{bb}' connected in series with about 100 Ω to the parallel circuit in Fig. 16.6, it is easy to understand that this resistance is negligible compared with the resistance of the parallel circuit.

With $f =$ 10.7 Mc/s it can no longer be ignored. The given values $g_{11} =$ 2.5 mS and $b_{11} =$ 4.4 mS are based on an equivalent parallel circuit. In reality, however, this comes from the parallel circuit of r_{i_0} and C_i which is connected in series with r_{bb}' (Fig. 16.7). With $f =$ 10.7 Mc/s the input capacitance represents a resistance of 190 Ω, and the parallel connected resistance r_{i_0} can therefore be disregarded. The parallel circuit in Fig. 16.7 is thus practically produced from the series circuit of r_{bb}' and $1/C_i\omega$. There-

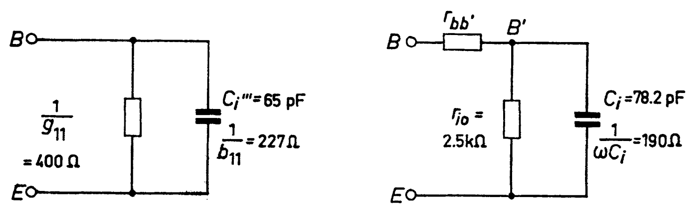


Fig. 16.7

fore, if the parallel circuit is converted back to a series circuit we can calculate the still unknown base area resistance from it.

The complex input resistance must be the same for both circuits in Fig. 16.7, so:

$$Z_{BE} = \frac{1}{g_{11} + \mathrm{j}b_{11}} = r_{bb}' - \mathrm{j}\frac{1}{C_i\omega}$$

$$= \frac{g_{11} - \mathrm{j}b_{11}}{g_{11}^2 + b_{11}^2} = r_{bb}' - \mathrm{j}\frac{1}{C_i\omega}$$

$$= \frac{g_{11}}{g_{11}^2 + b_{11}^2} - \frac{\mathrm{j}b_{11}}{g_{11}^2 + b_{11}^2} = r_{bb}' - \mathrm{j}\frac{1}{C_i\omega}.$$

Two complex numbers are equal when their real and imaginary portions are equal and so the following must be valid:

$$r_{bb}' = \frac{g_{11}}{g_{11}^2 + b_{11}^2} = \frac{1}{g_{11} + \dfrac{b_{11}^2}{g_{11}}}.$$

If we insert the given values we obtain:

$$r_{bb}' = \frac{1}{\dfrac{1}{400} + \dfrac{400}{227^2}} = \frac{1}{2.5 \cdot 10^{-3} + 7.75 \cdot 10^{-3}} = 97.5\ \Omega.$$

Reaction resistance R_r and reaction capacitance C_r

The conductances g_{12} and b_{12} are valid for input-side short-circuit, the current i_1 in the short-circuit being measured in $g_{12} = (v_2/i_1) \cos \varphi_{21}$. With the values given for $f = 450$ kc/s, C_r and R_r are obtained from this because the base area resistance does not yet play any part. (see also equivalent circuit diagram).

$$R_r = \frac{1}{g_{12}} = \frac{10^6}{0.1} = 10\ \mathrm{M}\Omega$$

$$X_{C_r} = \frac{1}{b_{12}} = \frac{10^6}{5.1} = 196\ \mathrm{k}\Omega.$$

These two resistances are therefore large in comparison with r_{bb}'. The reaction capacitance is obtained from:

$$C_r = \frac{b_{12}}{\omega} = \frac{5.1 \cdot 10^{-6}}{450 \cdot 2\pi \cdot 10^3} = 1.8\ \mathrm{pF}.$$

Output resistance R_i and output capacitance C_o

The output conductances $g_{22} = (i_2/v_2) \cos \varphi_2$ and $b_{22} = (i_2/v_2) \sin \varphi_2$ are also measured at input-side short-circuit. In this measurement C_r and R_r are connected in parallel with the output and have to be taken into account in the calculation of R_t and C_o. The effective output resistance is:

$$R'_t = \frac{1}{g_{22}} = \frac{10^6}{1.1} = 0.91 \text{ Mc/s.}$$

Now $R'_t = R_t \parallel R_r = 0.91\ \text{M}\Omega$ is valid, R_r being equal to 10 MΩ. Therefore $R_t = 1\ \text{M}\Omega$.

The effective output capacitance is:

$$C_o' = \frac{b_{22}}{\omega} = \frac{14 \cdot 10^{-6}}{450 \cdot 2\pi \cdot 10^3} = 5 \text{ pF.}$$

We now also have $C'_o = C_o + C_r$, so $C_o = C'_o - C_r = 5 - 1.8 = 3.2$ pF. The resistance of the output capacitance is:

$$X_{C_o} = \frac{1}{C_o\omega} = \frac{10^{12}}{3.2 \cdot 450 \cdot 2\pi \cdot 10^3} = 110\ \text{k}\Omega.$$

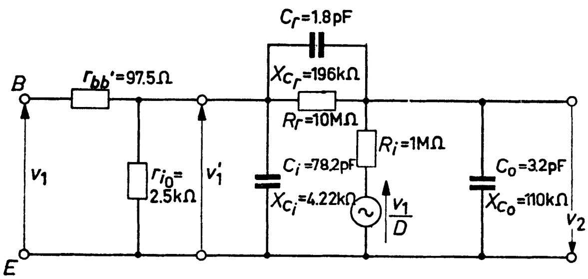


Fig. 16.8

The values calculated above will now be inserted into the equivalent circuit diagram (Fig. 16.8). From this it is easy to comprehend and estimate the resistance ratios for $f = 450$ kc/s. For instance, it can be established that v_1 is virtually equal to v'_1 because $r_{bb'}$ is still small in comparison with X_{C_o}. With voltage control the reaction via X_{C_r} as yet plays no part. With $f =$ 450 kc/s no appreciable drop is noticeable in the voltage amplification. The transistor slope is obtained from:

$$S = |Y_{21}| = \frac{i_2}{v_1} = 37 \text{ mS}$$

with short-circuited output ($v_2 = 0$).

The equivalent circuit diagram with resistance values for $f = 10.7$ Mc/s

The new reactances for $f = 10.7$ Mc/s (Fig. 16.9) are calculated with the capacitances from Fig. 16.8.

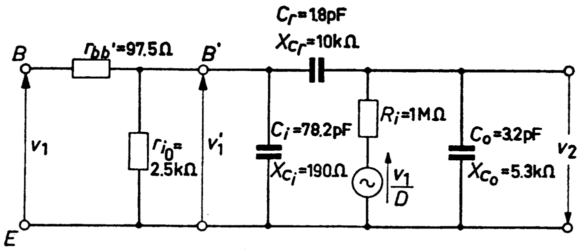


Fig. 16.9

It is easy to see from the resistances inserted in the equivalent circuit diagram that with $f = 10.7$ Mc/s, the control voltage v'_1 must now be smaller than the input voltage because voltage division takes place between $r_{bb}' = 97.5\ \Omega$ and $X_{C_i} = 187\ \Omega$. In this case the parallel resistance to X_{C_i} can be ignored. For the reduction of the control voltage we can set down:

$$s_i = \frac{v'_1}{v_1} = \frac{\dfrac{1}{\mathrm{j}C_i\omega}}{r_{bb}' + \dfrac{1}{\mathrm{j}C_i\omega}} = \frac{1}{1 + \mathrm{j}r_{bb}'\cdot C_i\omega}. \tag{16.19}$$

The amount is:

$$|s_i| = \frac{1}{\sqrt{[1 + (r_{bb}'\cdot C_i\cdot\omega)^2]}}. \tag{16.20}$$

If we consider that with output-side short-circuit to C_i, C_r is still connected in parallel, we find that

$$|s_i| = \frac{1}{\sqrt{[1 + (97.5 \times 80 \times 10^{-12} \times 10.7 \times 2\pi \times 10^6)^2]}} = 0.885.$$

With this reduction of the control voltage compared with low frequencies, we also obtain the new slope for 10.7 Mc/s

$$S' = S\cdot|s_i| = 37\cdot 0.885 = 32.7\ \mathrm{mS}$$

From the given values we derive:

$$Y_{21}| = 32\mathrm{mA/V} = S' = 32\mathrm{mS}$$

The agreement is thus very good.

Voltage amplification

The voltage amplification will now be calculated for both frequencies when the load resistance R_L is 1 kΩ.

f = 450 kc/s

We learn from the values inserted in the equivalent circuit in Fig. 16.8 that there is still no voltage division between r_{bb}' and C_i. Therefore the input voltage $v_1 \approx v'_1$ controls completely. A capacitive resistance of 110 kΩ ‖ 196 kΩ is connected in parallel with the load resistance $R_L =$ kΩ, at the output; this also has no influence. Through $R_L = 1$ kΩ the transistor with $R_i = 1$ MΩ practically operates as a short-circuit and we have:

$$A_v = S \cdot R_L = 37.$$

f = 10.7 Mc/s

As we see from the resistance values in Fig. 16.9, the capacitive resistances can no longer be disregarded with $f = 10.7$ Mc/s. The reduction in amplification is now obtained according to Formula (16.4) with the larger input capacitance C'_i because of reaction via C_r. The total input capacitance is assumed with $C'_i = 100$ pF. As a result,

$$s_i = \frac{1}{\sqrt{[1 + (r_{bb}' \cdot C'_i\omega)^2]}} = \frac{1}{\sqrt{[1 + (97.5 \cdot 10^{-10} \cdot 6.28 \cdot 10.7 \cdot 10^6)^2]}} = 0.82.$$

As the output generator current is constant because of the high internal resistance $r_i = 1$ MΩ, the current across the load resistance R_L, and with it the output voltage, is less corresponding to the current division through X_{C_o}. The drop in amplification at the output is then obtain from:

$$s_o = \frac{\dfrac{1}{jC_o\omega}}{R_L + \dfrac{1}{jC_o\omega}} = \frac{1}{1 + jC_o\omega \cdot R_L}. \tag{16.21}$$

The amount is:

$$|s_o| = \frac{1}{\sqrt{[1 + (R_L \cdot C_o \cdot \omega)^2]}} = \frac{1}{\sqrt{[1 + (10^3 \cdot 6.8 \cdot 10^{-12} \cdot 6.28 \cdot 10.7 \cdot 10^6)^2]}} = 0.9. \tag{16.22}$$

The amplification at $f = 10.7$ Mc/s is obtained with the two reduction factors from the amplification A_v at low frequencies as:

$$A'_v = s_i \cdot s_o \cdot A_v = 0.82 \cdot 0.9 \cdot 37 = 27.4.$$

Since $s_i \cdot s_o = 0.74$, the upper limiting frequency for voltage amplification is not yet reached at 10.7 Mc/s.

16.3.2. THE UPPER LIMITING FREQUENCY OF VOLTAGE AMPLIFICATION IN THE COMMON EMITTER CIRCUIT

By upper limiting frequency f_{0_v} we understand the frequency at which the output voltage v_2 is reduced to 70% when the input voltage remains constant.

It is evident from Example 15.1 and the Equations (16.20) and (16.21) that the reduction of amplification occurs at the input and output of the transistor. The reduction at the input has little influence; only the effect of the reaction capacitance is governed by the choice of R_L. At the output the reduction is strongly dependent on the load resistance, as is the case in the tube. The following statement must obviously apply for the limiting frequency f_{0_v}:

$$s_i \cdot s_o = \frac{1}{\sqrt{2}} = 0.7.$$

If we insert the relations (16.20) and (16.22) we get:

$$\frac{1}{\sqrt{[1 + (r_{bb'} \cdot C'_i \cdot \omega_{0_v})^2]}} \cdot \frac{1}{\sqrt{[1 + (R_L \cdot C_o \cdot \omega_{0_v})^2]}} = \frac{1}{\sqrt{2}}$$

$$[1 + (r_{bb'} \cdot C'_i \omega_{0_v})^2]\,[1 + (R_L \cdot C_o \omega_{0_v})^2] = 2$$

$$r^2_{bb'} \cdot C'^2_i \cdot \omega^4_{0_v} \cdot R_L^2 \cdot C_o^2 + (r^2_{bb'} \cdot C'^2_i + R_L^2 \cdot C_o^2)\,\omega^2_{0_v} - 1 = 0.$$

If we put $\omega^2_{0_v} = X$ we have:

$$r^2_{bb'} \cdot C_i'^2 \cdot R_L^2 \cdot C_o^2 \cdot X^2 + (r^2_{bb'} \cdot C'^2_i + R_L^2 \cdot C_o^2)\,X - 1 = 0.$$

If we also introduce:

$$a = r^2_{bb'} \cdot C'^2_i \cdot R_L^2 \cdot C'^2_o \tag{16.23}$$

$$b = r^2_{bb'} \cdot C'^2_i + R_L^2 \cdot C'^2_o \tag{16.24}$$

and

$$c = -1,$$

we obtain:

$$X_{1,2} = \frac{-b \pm \sqrt{(b^2 + 4ac)}}{2a},$$

or

$$f_{0_v} = \frac{\omega_{0_v}}{2\pi} = \frac{1}{2\pi}\sqrt{\frac{-b + \sqrt{(b^2 + 4a)}}{2a}}. \tag{16.25}$$

Example 16.2

How large is the upper limiting frequency f_{0_v} for the transistor T_3 in Example 15.1?

In this example we have:

$$R_L = 1\ \text{k}\Omega, \quad r_{bb}' = 97.5\ \Omega, \quad C'_i = 100\ \text{pF}, \quad C'_o = 5\ \text{pF}.$$

We therefore obtain:

$$a = 0.975^2 \cdot 10^4 \cdot 10^{-20} \cdot 10^6 \cdot 6.8^2 \cdot 10^{-24} = 44 \cdot 10^{-34}$$

$$b = 0.975^2 \cdot 10^{-16} + 6.8^2 \cdot 10^{-18} = 1.41 \cdot 10^{-16}$$

$$f_{0_v} = \frac{1}{2\pi} \sqrt{\frac{-1.41 \cdot 10^{-16} + \sqrt{(1.41^2 \cdot 10^{-32} + 4 \cdot 0.44 \cdot 10^{-32})}}{2 \cdot 44 \cdot 10^{-34}}} = 12.3\ \text{Mc/s}.$$

16.3.3. THE UPPER LIMITING FREQUENCY OF VOLTAGE AMPLIFICATION IN THE COMMON BASE CIRCUIT

By re-drawing the common emitter equivalent circuit for high frequencies (Fig. 16.4) as a common base circuit, we obtain Fig. 16.10.

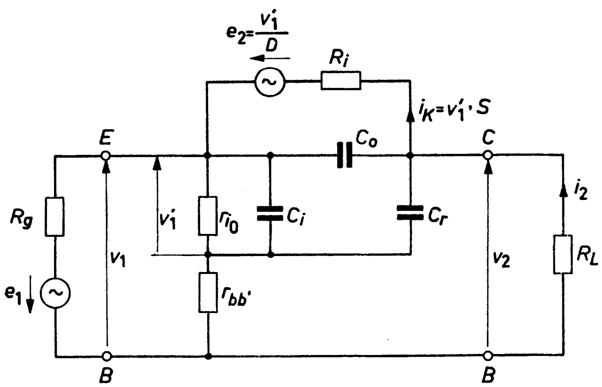


Fig. 16.10

If we first consider the frequency dependence of the output current and therefore of the output voltage, we find two influences which lead to a reduction of the output current.

1. As v_1 is less than v_2, the following equivalent circuit applies for the output:

The internal resistance of the transistor is always large by comparison with the load resistance and it therefore determines the current $i_K = v'_1 \cdot S_1$ supplied by the generator e_2. This current is only equal to the output current

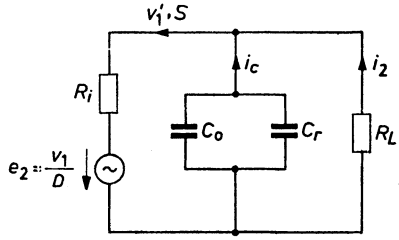


Fig. 16.11

i_2 at low frequencies. At high frequencies the output current becomes less because of current division via the parallel capacitances $C'_o = C_o = C_r$. According to Eq. (16.21) the "output reduction" is therefore:

$$s_o = \frac{1}{1 + \mathrm{j}\, C'_o \cdot \omega \cdot R_L}.$$

2. The second reduction of the output current i_2 results from the fact that the impressed voltage $e_2 = v'_1/D$ becomes less because the control voltage v'_i falls with rising frequency. This reduction s_i originating at the input results from the voltage division between $r_{bb'}$ and C_i according to Formula (16.19)

$$s_i = \frac{1}{1 + \mathrm{j}\, C_i \cdot \omega \cdot r_{bb'}}.$$

The entire frequency-dependent output voltage variation at constant input voltage v_1 would be accounted for with these two reduction factors $s_o \cdot s_i$. That is, with a very low resistance control generator (voltage control) the frequency dependence in the common base circuit is the same as in the common emitter circuit. Since there is no current amplification in the common base circuit, voltage control should always be aimed at. Nevertheless, a partial current control can, of course, always lead to adequate voltage amplification. Naturally the actual voltage amplification A_v of the transistor is then reduced corresponding to the input voltage division between R_g and r_{i_B}. If R_g is $4 \cdot r_{i_B}$, for example, the amplification will only amount to 1/5. Now, however, we have the benefit of the higher limiting frequency of the voltage amplification which gives the base circuit an advantage over the common emitter circuit. The reason for a higher limiting frequency with partial current control lies in the fact that the output current i_2 flows also as control current via the control generator. If the output current now tends to drop as a result of the two reduction factors s_o and s_i, the control generator is less loaded and the input voltage v_1 rises. The drop in output voltage is therefore reduced and the limiting frequency of voltage amplification is moved to higher values.

The rise of input voltage v_1 with increased frequency can be explained as follows:

According to Eq. (15.6) the input resistance of the common base circuit at very low frequencies is composed of $r_i \parallel 1/S$. Therefore r_i carries the input current in the emitter, and the resistance $1/S$ carries the output current i_2 which together give the input current in the common base circuit. If the current i_2 now decreases at high frequencies with the two reduction factors $s_o \cdot s_i$, the resistance $1/S$ which carries current i_2 at the input must be correspondingly larger. We therefore obtain the following equivalent circuit diagram:

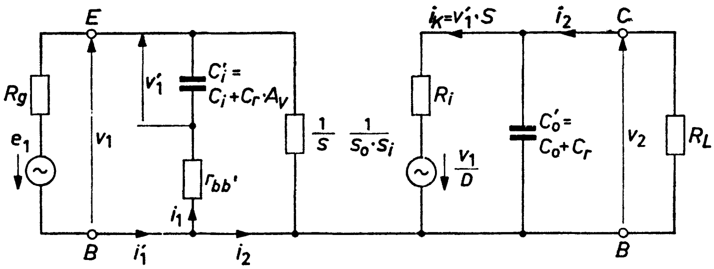


Fig. 16.12

The resistance $(1/S) \cdot (1/s_o \cdot s_i)$ now carries current i_2 at the input. The capacitance C_i is increased by $C_r \cdot A_v$ as we can see from Fig. 16.10, because C_r lies at voltage v_2. The input resistance of the common base circuit at high frequencies is thus obtained from:

$$r_{i_B} = \frac{1}{S \cdot s_o \cdot s_i + \dfrac{1}{r_{bb'} + 1/\mathrm{j}C'_i\omega}} =$$

$$= \frac{1}{\dfrac{S}{(1 + \mathrm{j}C'_i \cdot \omega \cdot r_{bb'})(1 + \mathrm{j}C'_o\omega \cdot R_L)} \quad \dfrac{1}{r_{bb'} + 1/\mathrm{j}C'_i\omega}} \tag{16.26}$$

This frequency-variable resistance governs the input voltage v_1 at high frequencies and partial current control. With very low frequencies the following equation can be applied:

$$v_1 = e_1 \frac{\dfrac{1}{S}}{\dfrac{1}{S} + R_g} = e_1 \cdot s_l.$$

The reduction at very low frequencies is consequently:

$$s_l = \frac{1}{1 + SR_g}. \tag{16.27}$$

At high frequencies we have:

$$v_1 = e_1 \frac{r_{i_B}}{r_{i_B} + R_g} = e_1 \cdot s_h.$$

The reduction at high frequencies is therefore:

$$s_h = \frac{1}{1 + \dfrac{R_g}{r_{i_B}}} =$$

$$= \frac{1}{1 + \dfrac{S \cdot R_g}{(1 + \mathrm{j}\, C'_i\, \omega r_{bb'})(1 + \mathrm{j}\, C'_o \omega R_L)} + \dfrac{R_g}{r_{bb'} + 1/\mathrm{j}\, C'_i \omega}}. \tag{16.28}$$

If the frequency rises from very low to high, the reduction then changes from s_l to s_h by an "operating reduction factor" $s_r = s_h/s_l > 1$. With this the change in input voltage is obtained from:

$$s_w = \frac{1 + S \cdot R_g}{1 + \dfrac{SR_g}{(1 + \mathrm{j}\, C'_i \omega r_{bb'})(1 + \mathrm{j}\, C'_o \omega R_L)} + \dfrac{R_g}{r_{bb'} + 1/\mathrm{j}\, C'_i \omega}}. \tag{16.29}$$

All the frequency-dependent variations are now covered with this. At constant input voltage v_1 the total reduction was $s_o \cdot s_i$. If there is partial current control the variation of input voltage is accounted for by s_w and the total frequency-dependent reduction is expressed by:

$$s = s_w \cdot s_o \cdot s_i \tag{16.30}$$

By inserting the values found we obtain:

$$s = \frac{1 + SR_g}{(1 + \mathrm{j}\, C'_o \omega R_L)(1 + \mathrm{j}\, C'_i \omega r_{bb'}) + S \cdot R_g + \dfrac{R_g(1 + \mathrm{j}\, C'_o \omega \cdot R_L)(1 + \mathrm{j}\, C'_i\, \omega r_{bb'})}{r_{bb'} + 1/\mathrm{j}\, C'_i \omega}}$$

$$s = \frac{1 + SR_g}{(1 + \mathrm{j}\, C'_o \omega R_L)(1 + \mathrm{j}\, C'_i \omega r_{bb'}) + SR_g + \mathrm{j}\, R_g (1 + \mathrm{j}\, C'_o \omega R_L)\, C'_i \omega}$$

$$s = \frac{1 + R_g \cdot S}{1 + \mathrm{j}\, C'_o \omega R_L + \mathrm{j}\, C'_i \omega r_{bb'} - C'_o C'_i \cdot r_{bb'} R_L \omega^2 + R_g S + \mathrm{j}\, R_g C'_i \omega - C'_o C'_i R_g R_L \omega^2)}$$

$$s = \frac{1 + R_g \cdot S}{(1 + R_g S) - (C'_o C'_i R_L)(R_g + r_{bb'})\omega^2 + \mathrm{j}\omega\, [C'_i (r_{bb'} + R_g) + C'_o R_L]}$$

At the upper limiting frequency ωo_v of voltage amplification, the amount of reduction is:

$$|s| = \frac{1}{\sqrt{2}} =$$

$$\frac{1 + R_g \cdot S}{\sqrt{\{(1 + R_g \cdot S) - (C'_o C'_i R_L)(r_{bb'} + R_g)\omega o_v^2]^2 + [C'_i(r_{bb'} + R_g) + C'_o \cdot R_L]^2 \omega o_v^2\}}} . \tag{16.31}$$

If we resolve this equation for the frequency, we find the limiting frequency fo_v of voltage amplification in the common base circuit:

$$2(1 + R_g S)^2 = (1 + R_g S) + (C'_o C'_i R_L)^2 (r_{bb'} + R_g)^2 \omega o_v^4 -$$

$$2(1 + R_g S)(C'_o C'_i R_L)(r_{bb'} + R_g)\omega o_v^2 + [C'_i(r_{bb'} + R_g) + C'_o \cdot R_L)^2 \omega o_v^2;$$

$$0 = \omega o_v^4 (C'_o C'_i R_L)^2 (r_{bb'} + R_g)^2 + \omega o_v^2 \{[C'_i(r_{bb'} + R_g) + C'_o R_L]^2 -$$

$$2(1 + R_g S)(C'_o C'_i R_L)(r_{bb'} + R_g)\} - (1 + R_g \cdot S)^2.$$

The solution is biquadratic. If we put in $\omega o_v{}^2 = X$ we have:

$$0 = X^2 \cdot a + X \cdot b - c; \qquad X = \frac{-b + \sqrt{(b^2 + 4ac)}}{2a}$$

in which

$$a = [(C'_o C'_i R_L)(r_{bb'} + R_g)]^2$$

$$b = [C'_i(r_{bb'} + R_g) + C'_o \cdot R_L]^2 - 2(1 + R_g S)\sqrt{a} \tag{16.32}$$

$$c = (1 + R_g S)^2.$$

The upper limiting frequency of voltage amplification in the common base circuit is consequently obtained from:

$$fo_v = \frac{1}{2\pi} \sqrt{\frac{-b + \sqrt{(b^2 + 4ac)}}{2a}}. \tag{16.33}$$

Example 16.3

The transistor T_3 is to be operated in the common base circuit with a load resistance $R_L = 1\ \text{k}\Omega$. A control generator with $R_g = 3.7 \cdot r_{i_B} = 3.7 \cdot 1/\text{S} = 3.7 \cdot 27 = 100\ \Omega$ is to be used for partial current control and therefore for reaching a higher voltage limiting frequency. What is the voltage limiting frequency obtained?

Given

$$R_g = 10^2\,\Omega, \quad r_{bb}' = 10^2\,\Omega, \quad S = 37\text{ mS}$$

$$R_L = 10^3\,\Omega, \quad C'_i \approx 10^{-10}\text{ F}, \quad C'_o \approx 5\cdot 10^{-12}\text{ F}$$

Solution

$$a = [(5\cdot 10^{-12}\cdot 10^{-10}\cdot 10^3)(10^2 + 10^2)]^2 = 10^{-32}$$

$$b = [10^{-10}(10^2 + 10^2) + 5\cdot 10^{-12}\cdot 10^3]^2 - 2(1 + 10^2\cdot 37\cdot 10^{-3})10^{-16}$$
$$= -\,3.15\cdot 10^{-16}$$

$$c = (1 + 3.7)^2 = 22$$

$$f_{0_v} = \frac{1}{2\pi}\sqrt{\frac{3.15\cdot 10^{-16} + \sqrt{(9.9\cdot 10^{-32} + 4\cdot 10^{-32}\cdot 22)}}{2\cdot 10^{-32}}}$$
$$= 40.6\text{ Mc/s}.$$

CHAPTER 17

MULTISTAGE LF AMPLIFIERS

RC sections or transformers are used to couple LF amplifier stages. The aim of connecting several transistors in series is to reach the greatest possible multiple of amplification at the same amplitude for all the frequencies to be transformed. All frequency-dependent sections must therefore be so designed that they play no part in the frequency range to be transformed. Apart from the coupling sections which are frequency dependent, the transistor is also a factor to be considered under certain circumstances, with its input capacitance which in transistor T_1, for example, is $\approx$ 5000 pF. The *RC* coupling is simple but has the drawback that the high source resistance R_t of the pre-amplifier transistor is loaded by the low resistance input capacitance of the following stage. The fact that transistor stages which are virtually operated in a short-circuit condition yield a voltage amplification is due entirely to the large slope of the transistors. Transformer coupling is more costly as regards material, but it has the advantage of a higher stage amplification because the input resistance r_t of the succeeding stage can be matched to the source resistance R_t*.

17.1. RC amplifier

If the a.c. output voltage of a transistor is applied through a large capacitance C_K to the input of a succeeding transistor, we then have an *RC* amplifier (Fig. 17.1).

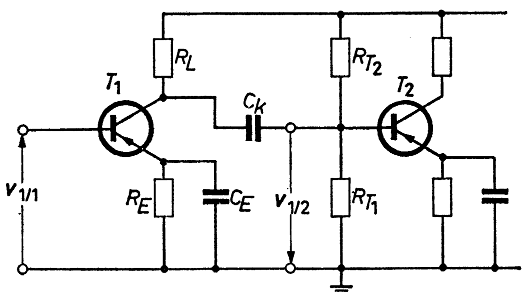


Fig. 17.1

We shall consider the ratios for one stage, for example between $v_{1/1}$ and $v_{1/2}$. There is a considerable difference here compared with a tube amplifier. The coupling capacitor is not followed by a high resistance input, but the resistances R_{T_1}, R_{T_2} and r_{i_2} lie in parallel to one another as the load. Because of these parallel connected resistances the load resistance R_L is reduced to R'_L even at medium frequencies. When the frequency response is studied, the input capacitance C_{i_2} also has to be taken into consideration if it assumes sizeable resistance values in the frequency band concerned.

All the interesting data will be calculated from the equivalent circuit diagram of two RC-coupled stages (Fig. 17.2).

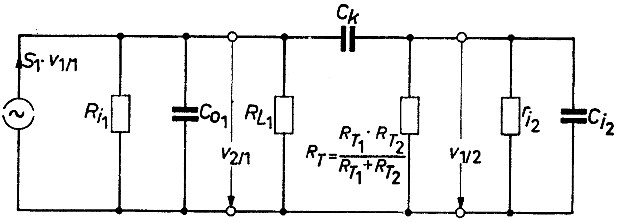


Fig. 17.2

Because of its extremely small value at AF, capacitance C_o will never have any effect. On the other hand, capacitance C_{i_2} which shows high values in certain transistors, already has some influence in the AF range. In medium audio frequencies the resistance of C_K is small and that of C_{i_2} is great; both can be disregarded. The effective load resistance for stage 1 is then:

$$R'_L = \frac{1}{\dfrac{1}{R_L} + \dfrac{1}{R_T} + \dfrac{1}{r_{i_2}}} \tag{17.1}$$

and the amplification $A_v \approx S_1 \cdot R'_L$.
If capacitance C_{i_2} is large enough, the load resistance R'_L can be reduced at high AF through the parallel-connected resistance $X_C = 1/(jC_{i_2}\omega)$. This results in a decrease of amplification.

17.1.1. UPPER LIMITING FREQUENCY

If the load resistance R'_L is equal to the capacitive resistance $1/(C_i\omega_0)$ of the large input capacitance C_i in the AF range, the new load resistance only amounts to $0.7 \cdot R_L$; the amplification is lowered by 30% or 3 dB. The frequency at which this drop occurs is termed the upper limiting frequency

ω_0 or f_0. We thus have:

$$\frac{1}{C_i\omega_0} = R'_L$$

$$\omega_0 = \frac{1}{C_i \cdot R'_L}; \quad f_0 = \frac{\omega_0}{2\pi}. \tag{17.2}$$

With a given C_i the upper limiting frequency can therefore be increased through a smaller R'_L though this is naturally at the cost of the amplification. By making use of HF transistors with small C_i, a drop in amplification at AF can nevertheless be completely eliminated at high frequencies (see also Chapter 16).

17.1.2. LOWER LIMITING FREQUENCY

If the frequency is allowed to become progressively less, the resistance of the coupling capacitor C_K increases. In contrast to the tube the load resistance also increases in this case, as also does the output voltage $v_{2/1}$. The drop in voltage $v_{1/2}$ across the voltage divider C_K and $R_T \parallel r_{i_2}$ will therefore not be so great. All the same, at a certain frequency, the lower limiting frequency f_v, ω_v the drop will eventually be 30% or 3 dB. We can calculate the lower limiting frequency as follows: If we take as input resistance

$$r'_i = r_i \| R_T = \frac{r_i \cdot R_T}{r_i + R_T} \tag{17.3}$$

the amplification at medium frequencies, with $R'_L < R_i^*$ is:

$$A_v = S\frac{r'_i \cdot R_L}{r'_i + R_L} = S \cdot R'_L. \tag{17.4}$$

At very low frequencies the effective stage amplification decreases because even with an increase in the load resistance which is now obtained from:

$$R''_L = \frac{R_L\left(r'_i + \dfrac{1}{\mathrm{j}C_K\omega}\right)}{R_L + r'_i + \dfrac{1}{\mathrm{j}\,C_K\omega}} \tag{17.5}$$

there is a drop in the output voltage through the voltage division at C_K and r'_i, resulting in a reduction

$$s' = \frac{r'_i}{r'_i + \dfrac{1}{\mathrm{j}C_K\omega}}. \tag{17.6}$$

The reduced stage amplification $A'_v = v_{1/2}/v_{1/2}$ is therefore:

$$A'_v = y_{\text{fe}} \cdot R''_L \cdot s' = S \frac{R_L \cdot r'_i}{R_L + r'_i + \dfrac{1}{\text{j} C_K \omega}}. \tag{17.7}$$

The reduction factor is found with (17.4) and (17.7) from:

$$s = \frac{A'_v}{A_v} = \frac{R_L \cdot r'_i}{R_L + r'_1 + \dfrac{1}{\text{j} C_K \omega}} \frac{R_L + r'_i}{R_L \cdot r'_i}$$

$$= \frac{1}{1 + \dfrac{1}{\text{j} C_K \omega (R_L + r'_i)}},$$

or the amount

$$|s| = \frac{1}{\sqrt{\left[1 + \dfrac{1}{C_K^2 \cdot \omega^2 (R_L + r'_i)^2}\right]}}. \tag{17.8}$$

Resolved for ω we find:

$$\frac{1}{|s|^2} = 1 + \frac{1}{C_K^2 \omega^2 (R_L + r'_i)^2}$$

$$\omega = \frac{1}{C_K (R_L + r'_i)} \sqrt{\frac{|s^2|}{1 - |s|^2}}. \tag{17.9}$$

The lower limiting frequency is obviously reached when the amplification recedes corresponding to a reduction factor $s = 1/\sqrt{2} = 0.707$. Then the irrational root in the above relation is equal to one and we obtain:

$$\omega_{v_g} = \frac{1}{C_K (R_L + r'_i)}. \tag{17.10}$$

This can also be explained by the fact that the control current which is supplied from the pre-stage transistor and at medium frequencies is determined by $R_L + r'_i$, now drops at very low frequencies to $i'_1 = s \cdot i_1 = 0.7 \cdot i_1$ because of resistance equality with the resistance $1/C_K\omega$ lying in series. If the fall of amplification in an amplifier with n stages amounts at a prescribed limiting frequency to $s_{\text{tot}} = 0.7$ over the whole amplifier, the reduction in one stage can only be

$$s = \sqrt[n]{s_{\text{tot}}}. \tag{17.11}$$

The necessary coupling capacitance C_K is then obtained from Eq. (17.9) and is:

$$C_K = \frac{1}{\omega_v (R_L + r'_i)} \sqrt{\frac{|s|^2}{1 - |s|^2}}. \tag{17.12}$$

Example 17.1

The stage amplification at $f_v = 15$ c/s should only drop by 5%.

$$R_L = 2\,\text{k}\Omega; r_i = 0.8\,\text{k}\Omega; R_{T_1} = 2\,\text{k}\Omega; R_{T_2} = 15\,\text{k}\Omega.$$

What coupling capacitance C_K is required?

Solution

$$|s| = 0.95 \qquad \omega_v = 2\pi\cdot 15 = 94.2$$

$$r_t = \frac{1}{\frac{1}{0.8}+\frac{1}{2}+\frac{1}{15}} = 0.55\,\text{k}\Omega$$

$$C_K = \frac{10^{-3}}{94.2\cdot 2.55}\sqrt{\frac{0.9}{0.1}} = 12.5\,\mu\text{F}.$$

17.1.3. THE EMITTER CAPACITANCE C_E

The emitter resistance R_E is required, as we saw in Chapter 9.5 for stabilisation of the working point. If there should be no feedback for a.c. voltages, R_E will be short-circuited with C_E. At very low frequencies, however, $1/C_E\omega$ also increases and can lead to a feedback. Then the lower limiting frequency caused by C_K which was calculated above, would be in question.

For the series feedback we have:

$$\alpha = \frac{R'}{R_L} = \frac{1}{C_E\omega\cdot R_L} \tag{17.13}$$

if R_E is short-circuited through $1/(C_E\omega)$ and so R_E is $> 1/(C_E\omega)$. Moreover we shall find: $A_v \approx y_{fe}\cdot R_L$ and the reduced amplification will be:

$$A'_v = \frac{A_v}{1+\alpha\cdot A_v} = A_v\cdot s.$$

The reduction of the amplification thus comes from:

$$s = \frac{1}{1+\alpha\cdot A_v}. \tag{17.14}$$

By inserting (17.13) into (17.14) we obtain:

$$s = \frac{1}{1+\frac{S}{\mathrm{j}\,C_E\omega}}.$$

The amount of s is consequently:

$$|s| = \frac{1}{\sqrt{\left[1+\left(\frac{S}{C_E\omega}\right)^2\right]}}. \tag{17.15}$$

If we resolve for C_E we then obtain:

$$\frac{1}{|s|^2} = 1 + \frac{S^2}{C_E{}^2\omega^2}$$

$$C_E = \frac{S}{\omega_v}\sqrt{\frac{|s|^2}{1-|s|^2}}. \qquad (17.16)$$

Example 17.2

The emitter short-circuit capacitor C_E is to be calculated for a transistor with $y_{fe} = 58.8 \cdot 10^{-3}$ A/V. With $f_v = 16$ c/s an $s = 0.95$ is required.

Solution

$$C_E = \frac{58.8 \cdot 10^{-3}}{2\pi\, 16}\sqrt{\frac{0.9}{0.1}} \approx 2000\,\mu\text{F}.$$

The emitter capacitance C_E calculated according to Equation (17.10) can be applied in the consideration of the single stage without taking the control generator into account, i.e. at constant input voltage v_1.

Nevertheless, if the control generator resistance R_g is no longer small in comparison with the input resistance, we have to take R_g into consideration. At very low frequencies the input resistance $r'_i = r_i\,(1 + \alpha\, A_v)$ becomes larger because of the increasing feedback. As a result the input voltage increases and the lower limiting frequency will be displaced to lower frequencies, or else with a limiting frequency that remains constant, the emitter capacitance can assume smaller values. The calculation is made as follows:

Since the current amplification $A'_i = A_i$ does not change with series feedback, the variation of the output voltage with variable feedback must be expressed through a corresponding output- and input current change. We can then derive the output voltage variation from the change in output current.

There is no feedback if the emitter connection is short-circuited to earth through C_E at medium frequencies.

The input current results from:

$$i_1 = \frac{e_1}{R_g + r'_i}.$$

With decreasing frequency the short-circuiting capacitive resistance $R_C = 1/C_E\omega$ increases and a feedback is produced. The input resistance becomes:

$$r'_i = r_i(1 + \alpha \cdot A_v) = r_i\left(1 + \frac{S}{\mathrm{j}C_E\omega}\right)$$

because

$$\alpha = \frac{R_C}{R_L} = \frac{1}{R_L \mathrm{j} C_E \omega} \quad \text{and} \quad A_v = S \cdot R_L.$$

For the input current at very low frequencies we therefore obtain:

$$i'_1 = \frac{e_1}{R_g + r'_i} = \frac{e_1}{R_g + r_i\left(1 + \dfrac{S}{\mathrm{j} C_E \omega}\right)}.$$

The weakening of the input current and consequently the output voltage is now found from the ratio of the two currents:

$$s = \frac{i'_1}{i_1} = \frac{R_g + r_i}{R_g + r_i\left(1 + \dfrac{S}{\mathrm{j} C_E \omega}\right)} = \frac{1}{\dfrac{R_g}{R_g + r_i} + \dfrac{r_i}{R_g + r_i} + \dfrac{r_i \cdot S}{(R_g + r_i)\mathrm{j} C_E \omega}}.$$

We now have:

$$\frac{R_g}{R_g + r_i} + \frac{r_i}{R_g + r_i} = 1 ,$$

and as a result

$$s = \frac{1}{1 + \dfrac{S \cdot r_i}{\mathrm{j}(R_g + r_i) C_E \omega}}. \tag{17.17}$$

If we compare this relation with Eq. (17.15) the necessary emitter capacitance is now obviously given by:

$$C_E = \frac{S \cdot r_i}{(R_g + r_i) \cdot \omega_1} \sqrt{\frac{|s|^2}{1 - |s|^2}}. \tag{17.18}$$

In the case where a decrease or weakening $s = 0.7$ is allowed, we call the frequency concerned the lower limiting frequency ω_{v_g} and we find:

$$C_E = \frac{S \cdot r_i}{(R_g + r_i) \cdot \omega_{v_g}} \tag{17.19}$$

or

$$\omega_{v_g} = \frac{S \cdot r_i}{C_E \cdot (R_g + r_i)}. \tag{17.20}$$

Example 17.3

The transistor in Example 17.2 is to be controlled with a generator whose internal resistance $R_g = 3.5$ kΩ. What capacitance C_E is now required if the input resistance $r_i = 0.7$ kΩ? How large would C_E have to be when $v_g = 16$ c/s?

Solution

$$C_E = \frac{S \cdot r_i}{(R_g + r_i)\omega_v} \sqrt{\frac{|s|^2}{1 - |s|^2}} = \frac{58.8 \cdot 10^{-3} \cdot 0.7 \cdot 10^3}{2\pi \cdot 16 \cdot (3.5 + 0.7) \cdot 10^3} \sqrt{\frac{0.9}{0.1}}$$
$$\approx 300\,\mu\text{F}$$

$$C'_E = \frac{S \cdot r_i}{(R_g + r_i) 2\pi \cdot f_{v_g}} \approx \frac{300}{3} = 100\,\mu\text{F}.$$

17.2. Transformer coupling

When the a.c. output voltage of a transistor is applied at the input of a following transistor through a transformer, we have the transformer coupling (Fig. 17.3).

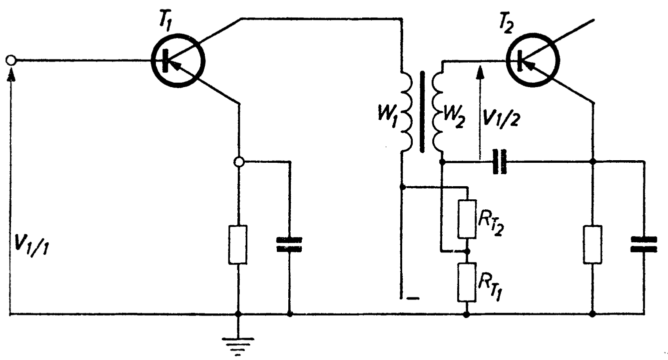


Fig. 17.3

With *RC* coupling the pre-amplifier transistor is loaded with the small input resistance and is therefore operated almost in short-circuit because the internal resistance is relatively high. Through the necessary low resistance base bias divider (see Chapter 7 and 9.5) a succeeding stage of this kind will still not be current controlled because through the voltage divider, the internal resistance of the equivalent control generator consists of the parallel circuit of R_i, R_L, R_{T_1} and R_{T_2} and is therefore of low resistance. The advantage gained through current control, for example in output stages, as regards non-linear distortions (see Chapter 9.4) cannot be utilised through *RC* coupling. On the other hand, it is possible to carry out current control by means of the transformer coupling in spite of low resistance base voltage division.

The second possibility of using transformer coupling to the pre-amplifier is in matching the output of the input resistance to the source resistance of the pre-amplifier transistor. This leads to a considerably higher stage amplification.

The frequency response with transformer coupling depends entirely on such data of the transformer as number of turns, spread, winding resistance, core material and construction.

17.2.1. CURRENT CONTROL

We see from Fig. 17.3 that the bias for the base is fed in at the "cold end" of W_2, that is, where W_2 is earthed to a.c. at the emitter, since C produces the "short-circuit". The a.c. equivalent circuit is obtained from Fig. 17.4.

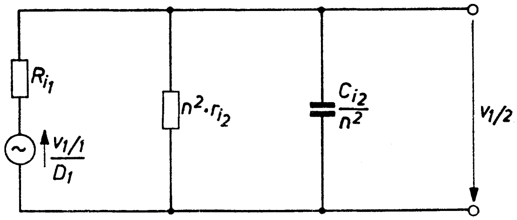


Fig. 17.4

The base voltage divider cannot load the transistor T_1 here. Even the load resistance R_L required with RC coupling for d.c. supply does not occur as bias. As a result the input resistance r_{i_2} with $n^2 \cdot r_{i_2}$ is connected as the only load at the pre-amplifier with its relatively high ohmic resistance R_{i_1}. If the transformation ratio is made so large that R_{i_1} is greater than $n^2 \cdot r_{i_2}$ the second transistor will then be current controlled because R_{i_1} virtually determines the control current. This coupling is used chiefly in front of output stages since there are few non-linear distortions here in spite of greater modulation through current control.

Current control is also used at the output of power transistors for the same reasons. The magnetisation current of the output transformer is then independent of the link in the magnetisation curve and therefore remains sinusoidal. The output voltage has a smaller distortion factor in this case.

17.2.2. OUTPUT MATCHING

A generator yields the greatest voltage when it is loaded with a resistance which is considerably larger than the internal resistance of the generator. In this case the generator operates in open circuit. With tubes we aim at this in order to achieve higher voltage amplification. This is possible at low and medium frequencies because of the high input resistance in a tube.

The transistor has an input resistance r_i which is always small in comparison with the internal resistance of the controlling pre-amplifier transistor. If the pre-amplifier transistor is directly loaded by r_i, it is operated almost in short-circuit.

The voltage- and output amplification are therefore small. If the voltage at the input resistance r_i of the following transistor should reach a maximum, the output yielded from the pre-amplifier transistor, i.e. from the control generator to the input resistance r_i, must also be maximum. As is generally known, the greatest output is drawn from a generator when its load resistance is equal to its internal resistance. As every transformer converts the resistances with n^2 we can make $n^2 \cdot r_{i_2} = R_{i_1}$ and so carry out output matching. In this case the input voltage and the input current have their highest value in the following transistor.

With output matching there is in effect still voltage control. Since the sharply-curved voltage-current-input characteristic is only slightly modulated in pre-amplifier transistors, the non-linear distortions are tolerable or can be kept small by means of feedback (see Chapter 9.4).

Example 17.4

A transistor with $R_i = 20.8$ kΩ is to be connected after a transistor with $r_i = 0.8$ kΩ. What is the necessary transformation ratio with current control if $R_i = 10 \cdot r'_i$? How large must the transformation ratio be made with output matching?

Solution

Current control

$$r'_i = \frac{R_i}{10} = \frac{20.8}{10} = 2.08 \text{ k}\Omega$$

$$r'_i = n^2 \cdot r_i$$

$$n^2 = \frac{r'_i}{r_i} = \frac{2.08}{0.8} = 2.6$$

$$n = 1:1.61.$$

Output matching

$$r'_i = R_i = n^2 \cdot r_i$$

$$n^2 = \frac{R_i}{r_i} = \frac{20.8}{0.8} = 26$$

$$n = 1:5.1.$$

CHAPTER 18

RESONANCE AMPLIFIERS

Selective amplifiers for transforming certain frequency bands are provided with resonance circuits, chiefly in radio and television technique. Transistor stages can naturally also have resonance circuits as load resistances. In tube amplifiers pentodes are used in conjunction with resonance circuits. The reason for this is that because of the high internal resistance R_i of the pentode the load resistance, that is the parallel resonance circuit, is fed with almost constant current and so the voltage wave form at the resonance circuit corresponds, as a function of the frequency, to the resistance resonance curve. We can also state that from the aspect of the current equivalent circuit diagram, the internal resistance R_i lying parallel to the resonance circuit has very little damping effect because of its very high resistance. The circuit quality ϱ is therefore only slighly impaired and the bandwidth of the circuit is scarcely enlarged. The succeeding tube does not load the resonance circuit resistivity either, as a result of its high resistance and so does not damp it. The capacitive load can be incorporated in the oscillation capacitance.

Transistor resonance amplifiers behave in a different way. The equivalent internal resistance $R_{i_{eq}}$ is often only of the same order as triode internal resistances and can therefore damp the resonance circuit quite considerably.

The output capacitance can also be included here in the oscillator circuit capacitance. Nevertheless, detuning can occur in various operating conditions as a result of the voltage dependence of the collector-base capacitance.

The loading of the following transistor is also important in the circuit arrangement of a resonance amplifier on account of the small a.c. input resistance r_i.

The transistor resistances are applied for matching at the tapping or special coupling windings of the resonance circuit inductance . As a result there is a transformation of the transistor resistances in parallel with the resonance circuit. (Fig. 18.1).

The damping resistances $n_1^2 \cdot R'_i$ and $n_2^2 \cdot r_i$ can be made large in accordance with the chosen tranformation ratio, and so keep the additional circuit damping small.

The output capacitance C_o which varies with the collector voltage, is

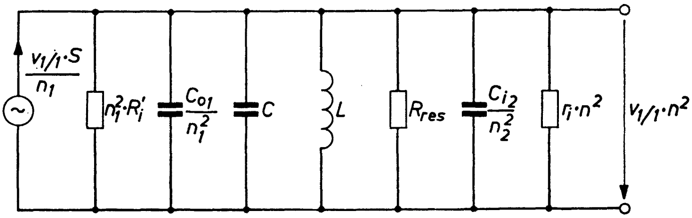


Fig. 18.1

also only effective with C_o/n_1^2 and its harmful influence on the circuit tuning can thus be reduced.

The voltage amplification A_v at resonance is obtained from Fig. 18.1 as follows:

$$R'_{res} = R_{res} \parallel n_1^2 \cdot R'_i \parallel n_2^2 \cdot r'$$

$$\frac{v_{1/1} \cdot S}{n_1} \cdot R'_{res} = v_{1/2} \cdot n_2$$

$$A_v = \frac{v_{1/2}}{v_{1/2}} = \frac{S \cdot R'_{res}}{n_1 \cdot n_2}. \tag{18.1}$$

Example 18.1

A resonance circuit stage is given with all the relevant values (Fig. 18.2). We wish to find:

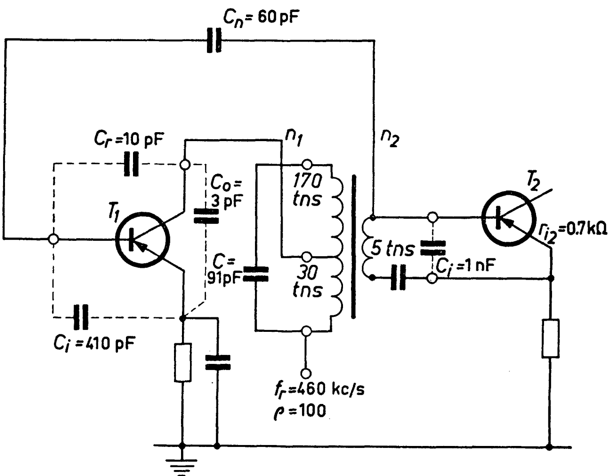


Fig. 18.2

1. The resonance resistance R'_{res}
2. The working bandwidth ϱ'
3. The voltage amplification with resonance A_v.

Solution

The values reduced to the resonance circuit are derived as follows:

$$n_1 = \frac{200}{30} = 66.6; \quad n_1^2 = 44{,}4$$

$$n_2 = \frac{200}{5} = 40; \quad n_2^2 = 1600$$

$$n_1^2 \cdot R'_i = 44.4 \cdot 20 = 888 \text{ k}\Omega$$

$$\frac{C_r + C_o}{n_1^2} = \frac{13}{44.4} \approx 0.3 \text{ pF}$$

$$n_1^2 \cdot r_i = 1600 \cdot 0.7 = 1120 \text{ k}\Omega$$

$$\frac{C_i}{n_2^2} = \frac{1000}{1600} \approx 0.6 \text{ pF.}$$

The resonance circuit alone has the following values:

$$X_C = \frac{1}{C \cdot \omega_r} = \frac{10^{12}}{91 \cdot 2\pi \cdot 0.46 \cdot 10^6} = 3.8 \text{ k}\Omega$$

$$R_{res} = \varrho \cdot X_C = 100 \cdot 3.8 = 380 \text{ k}\Omega$$

$$L = \frac{1}{\omega_r^2 \cdot C} = \frac{10^{12}}{0.46^2 \cdot 10^{12} \cdot 91 \cdot 6.28^2} = 1.32 \text{ mH.}$$

The parallel damping amounts to:

$$R_p = 888 \text{ k}\Omega \,\|\, 1120 \text{ k}\Omega \approx 500 \text{ k}\Omega.$$

As a result the resonance resistance is:

$$R'_{res} = R_{res} \,\|\, R_p = \frac{380 \cdot 500}{880} = 216 \text{ k}\Omega.$$

The circuit quality with the extra damping is consequently:

$$\varrho' = \frac{R'_{res}}{X_C} = \frac{216}{3.8} = 57.$$

The working bandwidth is:

$$B' = \frac{f_r}{\varrho'} = \frac{460}{57} \approx 8 \text{ kc/s.}$$

For the voltage amplification we obtain:

$$A_v = \frac{y_{fe} \cdot R'_{res}}{n_1 \cdot n_2} = \frac{39 \cdot 216}{6.66 \cdot 40} = 31.6 \cdot$$

18.1. Neutralisation

A portion of the a.c. output voltage is led back to the input through the collector-base capacitance C_r. In unfavourable cases this can lead to self-excitation. If, in addition, a voltage in phase opposition is also led back from the output to the input we are then "neutralising" (Fig. 18.3).

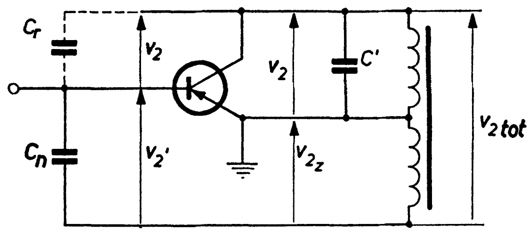


Fig. 18.3

It can be seen from Fig. 18.3 that the voltage led back from the output via C_r can exert no control action between base and emitter if the output voltage $v_{2_{tot}} = v_{2_z}$ at C_r and C_n is divided in equal ratio through suitable choice of value of the neutralising capacitance C_n. Then the diagonal voltage, i.e. the voltage between base and emitter in this bridge circuit is nil. In Example 18.1, Fig. 18.2, the voltage v_2 lies at the tapping with 30 windings. The opposite-phased extra voltage v_{2_z} lies at the output winding with 5 turns. The voltage ration is therefore 1 :6.

The neutralisation capacitance must act inversely to the reaction capacitance C_r, that is as 6 :1 ; C_n must therefore amount to 60 pF.

In neutralisation we also have to take into account that in certain circumstances an ohmic reaction resistance lies in parallel with the reaction capacitance. It may then be necessary to provide the neutralisation capacitance also with a parallel resistance to make phase-pure neutralisation possible.

18.2. Bandfilter coupling

As in the single circuit, there is also extra damping with the bandfilter when using transistors. Tappings are therefore provided at the primary or secondary circuit in this case also. As a rule this is done so that the working damping for both circuits is of equal value (equal ϱ').

CRITICAL COUPLING

The ratios have to be studied for critical coupling. If there is to be critical coupling in operation, $X' = \varrho' \cdot K$ must be equal to one. Here K is the coupling factor which with inductive coupling is obtained by measurement from the inductivity L_{14} of the two coils in series circuit (terminals 2 and 3 connected) and the inductivity L_{24} in opposite circuit (terminals 1 and 3 connected), as well as the inductance L of one of the two equal coils. We then obtain:

$$K = \frac{L_{24} - L_{14}}{4 \cdot L}. \tag{18.2}$$

The working quality ϱ' is again obtained from the working damping which for the primary circuit is derived here from: $n_1{}^2 \cdot R'_{i_{eq}}$. As a result,

$$R'_{\text{res}_1} = R_{\text{res}_1} \parallel n_1{}^2 \cdot R_{i_{eq}}.$$

The following expression applies for the secondary circuit:

$$R'_{\text{res}_2} = R'_{\text{res}_1} = R_{\text{res}_2} \parallel n_2{}^2 \cdot r_i.$$

Both operating resonance resistances give the working damping from:

$$\varrho' = \frac{R'_{\text{res}}}{L \cdot \omega_r} = R'_{\text{res}} \cdot C \cdot \omega_r = \frac{R'_{\text{res}}}{\sqrt{\frac{L}{C}}}. \tag{18.3}$$

The bandwidth comes from:

$$B = 1.41 \frac{f_r}{\varrho'}. \tag{18.4}$$

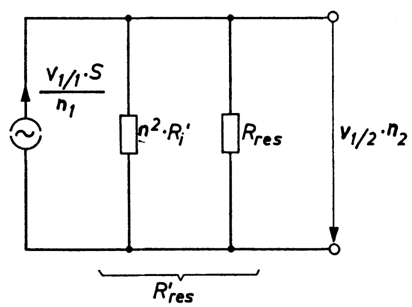


Fig. 18.4

With critical coupling the voltage at the primary circuit is equal to the voltage at the secondary circuit and at resonance the following equivalent circuit is valid (Fig. 18.4):

$$R'_{res} = R_{res_1} \parallel R_{res_2} = \tfrac{1}{2} R_{res_1}$$

The voltage amplification is consequently obtained as follows:
We have:

$$\frac{v_1 \cdot S_1}{n_1} \cdot R'_{res} = v_{1/2} \cdot n_2$$

therefore

$$A_v = \frac{v_{1/2}}{v_{1/1}} = \frac{S_1 \cdot R'_{res}}{n_1 \cdot n_2}. \tag{18.5}$$

Example 18.2

The following values are given for a band-filter:

$$f_r = 460\ \text{kc/s};\quad L = 520\ \mu\text{H};\quad C = 195\ \text{pF};$$

$$\varrho = 100;\quad K = 0.023;\quad R_{res} = \varrho\sqrt{\frac{L}{C}} = 162\ \text{k}\Omega.$$

The necessary transistor data can be taken from the circuit fiagram (Fig. 18.5).

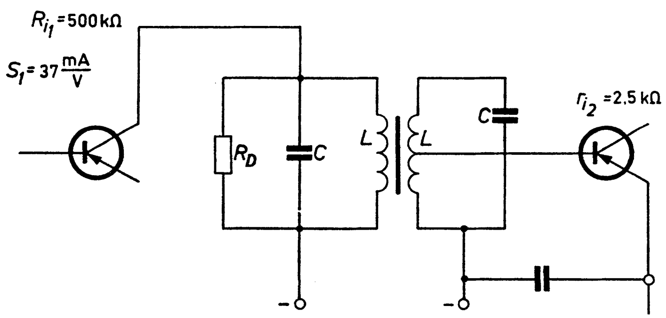


Fig. 18.5

1. Transformation ratio n_2 for critical coupling.
2. The damping resistance R_D required.
3. The working bandwidth.
4. The voltage amplification.

Solution

The working quality must be:

$$\varrho' = \frac{1}{K} = \frac{1}{0.023} = 43.5.$$

The operating resonance resistance is therefore:

$$R'_{\text{res}} = \varrho' \sqrt{\frac{L}{C}} = 43.5 \cdot 1.62 = 70 \text{ k}\Omega.$$

The parallel damping to $R_{\text{res}} = 162$ kΩ is obtained from:

$$\frac{n_2^2 \cdot r_i \cdot R_{\text{res}}}{n_2^2 \cdot r_i + R_{\text{res}}} = R'_{\text{res}}$$

$$\frac{n_2^2 \cdot r_i \cdot 162}{n_2^2 \cdot r_i + 162} = 70; \qquad 70 \cdot n_2^2 \cdot r_i + 11340 = 162 \cdot n_2^2 \cdot r_i$$

$$n_2^2 \cdot r_i = \frac{113 \cdot 40}{92} = 123 \text{ k}\Omega.$$

1. The transformation ratio is given by:

$$n_2^2 = \frac{123}{r_i} = \frac{123}{2.5} = 49.3$$

$$n_2 = 7.$$

2. The damping resistance R_D must be connected additionally because the internal resistance only damps with $R_{i_{eq}} = 500$ kΩ while the total damping needed is 123 kΩ. The following values must therefore apply:

$$\frac{500 \cdot R_D}{500 + R_D} = 123 \text{ k}\Omega$$

$$R_D = \frac{123 \cdot 500}{500 - 123} = 163 \text{ k}\Omega$$

3. $$B' = 1.41 \frac{f_r}{\varrho'} = 1.41 \frac{460}{43.5} = 15 \text{ kc/s}$$

4. $$A_v = \frac{S \cdot R'_{\text{res}}}{n_1 \cdot n_2} = \frac{37 \cdot 35}{1 \cdot 7} = 185.$$

CHAPTER 19

AMPLIFIER NOISE

In amplifiers with high amplification factors the so-called "noise voltage" takes effect at the output; this can actually be noticed in a loudspeaker as an irregular noise. It is caused by resistance noises and tube or transistor noises.

19.1. Resistance noise

Slight displacements of charge carriers are produced in every resistance through thermal agitation. These irregular electron movements, in a conductor, for example, are noticed as noise voltage. The frequencies of these erratic a.c. voltages originating in statistical distribution, can range from the lowest to the highest technical frequencies. With a wide frequency spectrum of this kind we speak, as with light, of "white" noise because in white light all the frequencies or colours are also present. The noise of a resistance after a selective amplifier often has only a limited frequency band and can therefore be called coloured noise. The noise power increased at the output by an amplifier will rise in proportion to the widening of the amplified noise spectrum.

We can also imagine that in the "noise resistance" concerned, the frequency band later narrowed by the following amplifier is already present. Then, according to Nyquist, the noise power generated in each resistance per cycle bandwidth is:

NOISE POWER

$$p = 4 \cdot k \cdot T_0. \qquad (19.1)$$

In this expression the Bolzmann constant

$$k = 1.37 \cdot 10^{-23} \qquad [Ws]$$

and T_0 is the absolute temperature in degrees Kelvin. For a medium room temperature of 20 °C the noise power produced in a resistance is therefore

$$p = 16 \cdot 10^{-21} \qquad [Ws]. \qquad (19.2)$$

If the width of the transformed frequency band is $B_{c/s}$ we are then interested in an apparent generated noise power

$$P = p \cdot B \tag{19.3}$$

i.e. the noise power per cycle bandwidth adds up at the larger bandwidth B to the total power P.

NOISE VOLTAGE

The effective noise power P now results at the terminals of the resistance concerned, of value R, in an impressed noise voltage v_r which is obtained from:

$$v_r = \sqrt{P \cdot R} = \sqrt{4 \cdot k \cdot T_0 \cdot B \cdot R}. \tag{19.4}$$

This voltage, generally called the noise voltage, is by nature an impressed voltage, that is, a generated voltage, and we have the following equivalent circuit (Fig. 19.1).

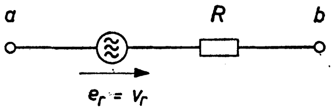


Fig. 19.1

The noise voltage v_r is therefore present at terminals a, b of the resistance in open circuits, i.e. when these terminals are not loaded.

Example 19.1

The noise voltage is to be calculated for a resistance $R = 1\ \text{k}\Omega$ at $t = 20\ °\text{C}$ and $t' = 100\ °\text{C}$ for a bandwidth $B = 10$ kc/s.

Solution

$$v_r = \sqrt{4 \cdot k \cdot T_0 \cdot B \cdot R} = \sqrt{16 \cdot 10^{-21} \cdot 10^4 \cdot 10^3} = 4 \cdot 10^{-7} = 0.4\ \mu\text{V}$$

$$v'_r = \sqrt{4 \cdot 1.37 \cdot 10^{-23} \cdot 373 \cdot 10^7} = 0.45\ \mu\text{V}.$$

SERIES CIRCUIT

For the noise voltage of a resistance we have:

$$v_r^2 = 4 \cdot k \cdot T_0 \cdot B \cdot R\,.$$

In the series circuit of two resistances, $R = R_1 + R_2$, and we obtain:

$$v_r^2 = 4kT_0 \cdot B(R_1 + R_2) = 4kT_0BR_1 + 4kT_0BR_2 = v_{r_1}^2 + v_{r_2}^2.$$

If two noise generators are thus connected in series as in Fig. 19.1, the noise voltage produced must be added geometrically.

$$v_r = \sqrt{v_{r_1}^2 + v_{r_2}^2}. \tag{19.5}$$

This relation must always be used when two series-connected resistances have different noise temperatures. If we insert the total resistance $R = R_1 + R_2$ in Equation (19.4) the noise resistances must have the same noise temperature.

Example 19.2

a. Two resistances $R_1 = R_2 = 1\ \text{k}\Omega$ are connected in series at the same noise temperature $t = 20\ °\text{C}$. How big is the total noise voltage at a bandwidth $B = 10$ kc/s.
b. One of the two noise resistances has reached a temperature $t = 100\ °\text{C}$. What total noise voltage do we have now?

Solution

With the values from Example 19.1 we find:

a. $$v_r = \sqrt{4 \cdot kT_0 \cdot B(R_1 + R_2)} = 1.41 \cdot 0.4 = 0.564\ \mu\text{V};$$

b. $$v_r = \sqrt{(v_r{}^2 + v'_r{}^2)} = \sqrt{(0.4^2 + 0.45^2)} = 0.6\ \mu\text{V}.$$

PARALLEL CIRCUIT

If two noise resistances R_1 and R_2 are connected in parallel, we can regard the resistance of the parallel circuit $R = R_1 \cdot R_2/(R_1 + R_2)$ as the noise resistance if both resistances have the same noise temperature, and the noise voltage can be calculated from Equation (19.4).

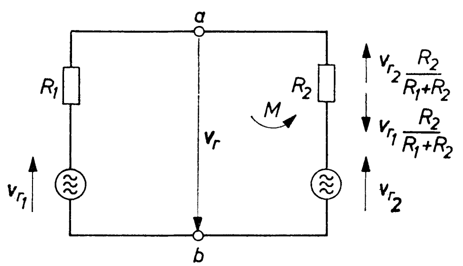


Fig. 19.2

The noise resistance R_1 loaded in the parallel circuit through a second noise resistance R_2, is viewed in a different way and the following equivalent circuit is applied (Fig. 19.2) which is derived from Fig. 19.1.

The two noise generators each drive a current according to the principle of superimposition and the voltage reductions quoted are produced, for example, at resistance R_2. The voltage drops originating from different noise voltages must again be totalled geometrically and we therefore have for network M

$$v_r + v_{r_2}\frac{R_2}{R_1 + R_2} - v_{r_1}\frac{R_2}{R_1 + R_2} = v_{r_2}$$

or

$$v_r = v_{r_2}\left(1 - \frac{R_2}{R_1 + R_2}\right) + v_{r_1}\frac{R_2}{R_1 + R_2}.$$

The noise voltage at the parallel circuit is then:

$$v_r = \sqrt{v_{r_1}^2\left(\frac{R_2}{R_1 + R_2}\right)^2 + v_{r_2}^2\left(\frac{R_1}{R_1 + R_2}\right)^2}\ . \tag{19.6}$$

This formula must always be used when two parallel noise resistances are at different noise temperatures.

Example 19.3

a. Two resistances $R_1 = R_2 = 1$ kΩ with equal noise temperature $T_0 = 293$ °K are connected in parallel. What noise voltage lies at the parallel circuit when $B = 10$ kc/s?
b. What noise voltage appears when one resistance has the noise temperature $T_0 = 373$ °K?

Solution

With the values already calculated in Example 19.1 we obtain:

a.
$$R = \frac{R_1 \cdot R_2}{R_1 + R_2} = 0.5\ \text{k}\Omega = \tfrac{1}{2}\cdot R_1$$

$$v_r = \sqrt{(4\cdot k\cdot T_0\cdot B\tfrac{1}{2}R_1)} = \frac{v_{r_1}}{\sqrt{2}} = \frac{0.4}{\sqrt{2}} = 0.284\ \mu\text{V};$$

b.
$$v_r = \sqrt{v^2_{r_1}\left(\frac{R_2}{R_1 + R_2}\right)^2 + v_{r_2}^2\left(\frac{R_1}{R_1 + R_2}\right)^2}$$

$$= \sqrt{[0.4^2(\tfrac{1}{2})^2 + 0{,}45^2(\tfrac{1}{2})^2]}$$

$$= \tfrac{1}{2}\sqrt{[0.4^2 + 0.45^2]} = 0.3\ \mu\text{V}.$$

19.2. Transistor noise

The causes of noise in transistors are not yet all known. Some sources of noise which are present in the tube can also be identified in the transistor.

These are the resistance noise, the Schott effect (shot effect), the current distribution noise, and at very low frequencies, a noise process similar to the Schott effect. This last is based on spontaneous resistance change in the barrier layer. Apart from frequency dependence we can also notice here a dependence on the collector current in which a small collector current gives smaller noise.

Moreover the special semiconductor noise is also dependent on the collector voltage. The collector voltage has to be kept small with specially low noise stages. The amount of the transistor noise is thus not a constant but depends on the operational working point.

As well as the internal noise of the transistor, there is always in operation the extra noise of the control source resistance connected at the input. We have found it useful to make statements concerning the amount of transistor noise in conjunction with the control source resistance. Since the values of such influences as frequency range, collector voltage, collector current and source resistance strongly affect the transistor noise, the advance calculation of transistor noise based on the general characteristic values can only be done with difficulty. It is better to work with measured characteristic values which are also given by the transistor manufacturers for the purposes concerned and are called noise factors. It is possible to compare different transistor types with the aid of noise figures of this kind. In addition, the apparent noise voltage lying at the transistor input can be calculated with them, and so, with known voltage amplification, the noise voltage at the output of an amplifier can eventually be estimated. By comparing the signal voltage of the control generator with the total impressed voltage which can be determined with the noise factor, we find the noise distance, a measure of the quality of amplification in relation to the noise.

NOISE FACTOR

The noise factor F is stated for a transistor in connection with a particular control generator whose internal resistance R_g is given. At the transistor output we obtain a noise voltage or noise power at the load resistance which is governed by the noise of the control generator source resistance R_g and by the noise of the transistor itself. The source resistance R_g has a noise power $p = 4kT_0$ per cycle bandwidth.

We now think of the noise of the transistor as produced in the control generator source resistance R_g which means that the noise power in R_g must be greater than the power for the pure resistance noise. We therefore write:

$$p = 4kT_0 \cdot F. \tag{19.7}$$

The noise factor F thus indicates how much greater the noise power must be in the source resistance of the control generator when the transistor noise is based on this resistance. We therefore regard the transistor itself as noise-free.

The following value applies for the noise voltage produced in the source resistance at a bandwidth B:

$$v'_r = \sqrt{4kT_0 \cdot F \cdot B \cdot R_g}. \qquad (19.8)$$

The above ratios can be expressed by means of the following equivalent circuit (Fig. 19.3).

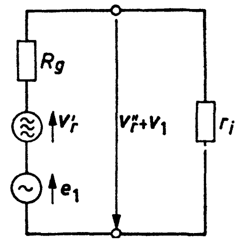


Fig. 19.3

The control generator included in the equivalent circuit can be, for instance, an aerial with voltage e_1 and resistance R_g. With the calculated noise voltage v'_r we immediately obtain the noise lead for the effective signal

$$Q = \frac{e_1}{v'_r}. \qquad (19.9)$$

Both voltages e_1 and v'_r are effective at the transistor input as v''_r and v_1 corresponding to the voltage divider ratio

$$s = \frac{r_i}{r_i + R_g}.$$

The controlling noise voltage at the transistor input is obtained from:

$$v''_r = v'_r \frac{r_i}{r_i + R_g} = \sqrt{[4 \cdot kT_0 \cdot F \cdot B \cdot R_g]} \frac{r_i}{r_i + R_g}. \qquad (19.10)$$

For the special case of matching when $R_g = r_i$ we use the following expression with $B = 1$ c/s,

$$v''_r = \sqrt{[4 \cdot kT_0 \cdot F \cdot 1 \cdot R_g]} \ \tfrac{1}{2}$$
$$v''_r = \sqrt{F \cdot kT_0 \cdot r_i} = \sqrt{p_{r_i} \cdot r_i}. \qquad (19.11)$$

This relation states that for matching the noise power per cycle at the transistor input resistance is found from:

$$p_{r_i} = F \cdot kT_0. \tag{19.12}$$

The noise factor F in this special case has therefore also been called the kT_0 number (noise figure). $F \cdot kT_0$ is thus the noise power which we can visualise at an input resistance r_i considered as noise-free in itself, with output matching of R_g. In an ideal transistor without noise, F would be equal to one and the noise power at the input resistance would equal kT_0. This representation of the "kT_0 number" which only applies for output matching, is discarded in favour of the noise factor because there is not always output matching and as a result the noise power at the input resistance assumes a different value to $F \cdot kT_0$.

There is consequently a considerable difference between the noise factor F and the kT_0 number F. The noise factor F is always valid for a transistor in conjunction with a particular control generator whose source resistance R_g is given. We get different noise factors F for the same transistor according to the size of the source resistance R_g. The kT_0 number, on the other hand, always applies for output matching $R_g = r_i$ and therefore has only one value for each transistor. The kT_0 number is definite and is therefore better for comparing various transistors.

Example 19.4

We are given a VHF input circuit (Fig. 19.4) with a transistor which has a noise factor $F = 7$ for a working point $I_{C_0} = 2$ mA and a control generator resistance $R_g = 60\ \Omega$. The transforming bandwidth will amount to $B = 200$ Mc/s. The control generator is a 60 Ω dipole which is connected to the transistor input by a transformer with a transformation ratio of $n = 1$.

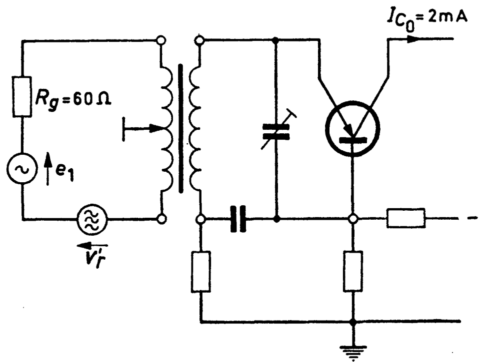


Fig. 19.4

We want to find the respective impressed aerial voltage e_1 for the noise leads 20 dB, 40 dB and 60 dB.

Solution

$$v'_r = \sqrt{4kT_0 \cdot B \cdot F \cdot R_g} = \sqrt{16 \cdot 10^{-21} \cdot 2 \cdot 10^5 \cdot 7 \cdot 60} = 1.16\ \mu\text{V}.$$

For 20 dB noise lead with tolerable reproduction we obtain:

$$Q = 10 = \frac{e_1}{v'_r}; \quad e_1 = 10 \cdot 1.16 = 11.6\ \mu\text{V}.$$

At 40 dB with good reproduction

$$Q = 100; \qquad e_1 = 100 \cdot 1.16 = 116\ \mu\text{V}$$

and at 60 dB with high quality reproduction the following values must be obtained:

$$Q = 1000; \qquad e_1 = 1000 \cdot 1.16 = 1.16\ \text{mV}.$$

CHAPTER 20

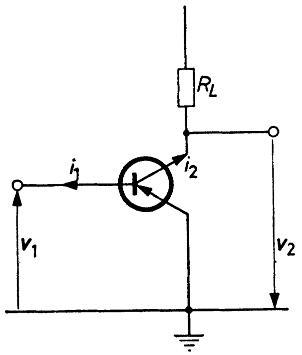
SUMMARY OF COMPARISONS BETWEEN THE TRANSISTOR AND THE ELECTRON TUBE

The differences between the two amplifying devices will have become obvious to all those who are familiar with the tube in the two-pole presentation. The transistor is often only distinguished in application as an amplifying device by differences of component values. For example, the electron tube also has an "input resistance", namely the grid leak resistance, determined by the circuit. Even though this resistance is several orders higher than the transistor input resistance, it often has to be taken into account in electron tube circuits .If this necessary resistance is thus added to the electron tube it often occurs that an electron tube amplifies without being powerless. This is only an example to draw attention to the fact that in this case the two-pole theoretical representation was also adequate for the electron tube. In spite of this the idea has become widespread that the transistor can only be correctly treated as a quadripole. The consistent use of two-pole theoretical principles shows that the transistor representation is considerably clearer than in a four-pole theoretical investigation. As well as these advantages in the qualitative study of the transistor, the quantitative results are equally accurate and easy to comprehend. For instance, if we calculate the electron tube amplification from $A_v = S \cdot R_L$ and we can use the same simple relation for the transistor, it is at least an advantage for the circuit designer. It is amazing how far there are agreements in formulae between the transistor and the electron tube. In the following section we give a comparison which describes the tube alongside the transistor and indirectly with its help.

20.1. Formula and circuit comparison

TRANSISTOR

Common emitter circuit
(fundamental circuit)



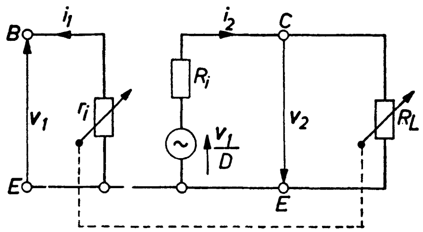
Characteristic values

$$S = \frac{h_{21}}{h_{11}} - R_i = \frac{1}{h_{22}\left(1 - \frac{h_{12} \cdot h_{21}}{h_{11} \cdot h_{22}}\right)}$$

$$D = \frac{1}{S \cdot R_i} = D_i - D_v$$

$$D_v = h_{12} \qquad r_{i_0} = h_{11}$$

Fundamental equivalent circuit



(very low frequencies)

Voltage amplification

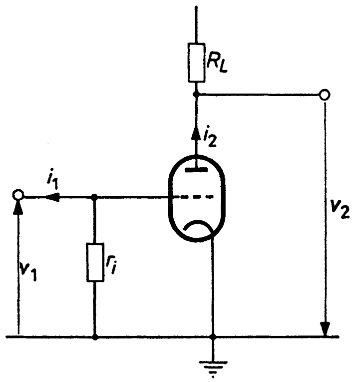
$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L}$$

Input resistance

$$r_i = \frac{r_{i_0}}{1 + A_v \cdot D_v}$$

ELECTRON TUBE

Cathode base circuit
(fundamental circuit)

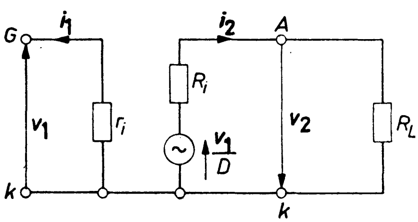


Characteristic values

$$S \quad R_i \quad D$$

$$r_i = R_{\text{grid}}$$

Fundamental equivalent circuit



(very low frequencies)

Voltage amplification

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L}$$

Input resistance

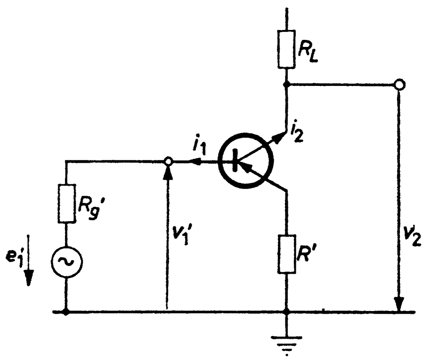
$$r_i = R_{\text{grid}}$$

TRANSISTOR	ELECTRON TUBE
Current amplification	*Current amplification*
$A_i = A_v \frac{r_i}{R_L}$	$A_i = A_v \frac{r_i}{R_L}$
Output amplification	*Output amplification*
$A_p = A_v \cdot A_i$	$A_p = A_v \cdot A_i$
equivalent internal resistance with load variation (resonance circuit)	. . . . .

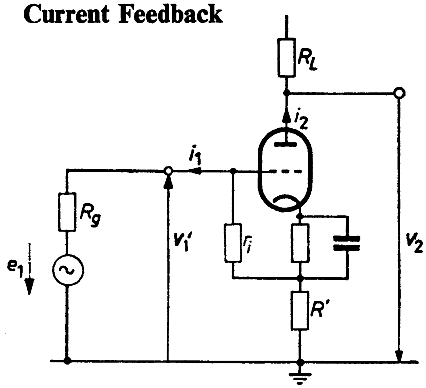
$$R_{i_{eq}} = \frac{1}{h_{22}\left(1 - \frac{h_{12} \cdot h_{21}}{h_{22}(h_{11} + R_g)}\right)}$$

R_g = control generator resistance

Series feedback



Current Feedback



TRANSISTOR	ELECTRON TUBE
Provided	*Provided*
$R'_g < r_i$	$R_g < r_i$
Feedback factor	*Feedback factor*
$\alpha = \frac{R'}{R_L}$	$\alpha = \frac{R'}{R_L}$
Voltage amplification	*Voltage amplification*
$A'_v = \frac{v_2}{v'_1} = \frac{A_v}{1 + \alpha \cdot A_v}$	$A'_v = \frac{v_2}{v'_1} = \frac{A_v}{1 + \alpha \cdot A_v}$
Input resistance	*Input resistance*
$r'_i = \frac{v'_1}{i_1} = r'(1 + \alpha \cdot A_v)$	$r'_i = \frac{v'_1}{i_1} = r_i(1 + \alpha \cdot A_v)$
Current amplification	*Current amplification*
$A'_i = A_i$	$A'_i = A_i$

TRANSISTOR

Equivalent slope

$$S' = \frac{S}{1 + S \cdot R'}$$

Equivalent internal resistance

$$R'_i = R_i(1 + S \cdot R')$$

Equivalent reciprocal amplification

$$D' = D$$

Apparent internal resistance

$$R_{i_{eq}} = R_i \left(1 + SR' \frac{r_{i_0}}{r_{i_0} + R_g}\right)$$

ELECTRON TUBE

Equivalent slope

$$S' = \frac{S}{1 + S \cdot R'}$$

Equivalent internal resistance

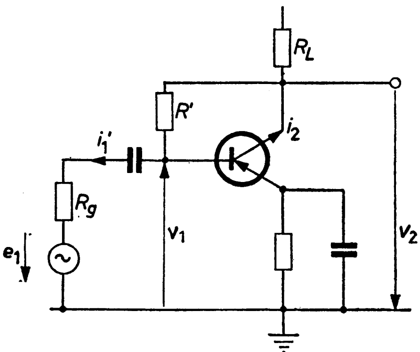
$$R'_i = R_i(1 + S \cdot R')$$

Equivalent reciprocal amplification

$$D' = D$$

. . . . .

Parallel feedback



Provided

$$R_g > r'_i$$

Feedback factor

$$\alpha = \frac{r_i}{r_i + R'}$$

Voltage amplification

$$A'_v = \frac{v_2}{v_1} = A_v$$

Input resistance

$$r'_i = \frac{r_i \| R'}{1 + \alpha \cdot A_v}$$

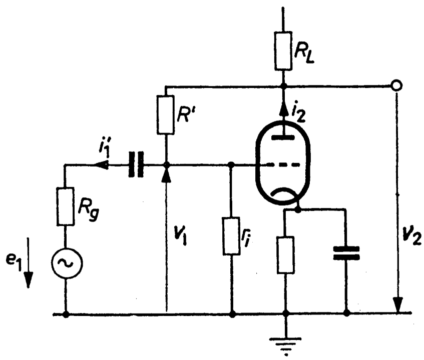
Current amplification

$$A'_i \approx \frac{A_i}{1 + \alpha \cdot A_v}$$

Equivalent characteristic values

$$S' = S, \quad R'_i = R_i, \quad D' = D$$

Voltage feedback



Provided

$$R_g > r'_i$$

Feedback factor

$$\alpha = \frac{r_i}{r_i + R'}$$

Voltage amplification

$$A'_v = \frac{v_2}{v_1} = A_v$$

Input resistance

$$r'_i = \frac{r_i \| R'}{1 + \alpha \cdot A_v}$$

Current amplification

$$A'_i = \frac{A_i}{1 + \alpha \cdot A_v}$$

Equivalent characteristic values

$$S' = S, \quad R'_i = R_i, \quad D' = D$$

TRANSISTOR	ELECTRON TUBE

TRANSISTOR

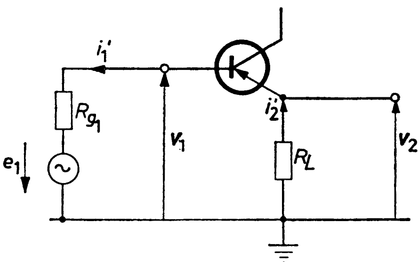
Total voltage amplification

$$A_{v_{\text{tot}}} = \frac{v_2}{e_1} = \frac{r_i}{R_g + \dfrac{r_i \parallel R'}{1 + \alpha \cdot A_v}} \frac{A_v}{1 + \alpha \cdot A_v}$$

Apparent internal resistance

$$R_{i_{\text{eq}}} = \frac{1}{S} \frac{(R_g \parallel r_{i_0}) + R'}{R_g \parallel r_{i_0}}$$

Common collector circuit



Voltage amplification

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L}$$

$$A_{v_C} = \frac{A_v}{1 + A_v}$$

Current amplification

$$A_{i_C} = 1 + A_i$$

Input resistance

$$r_{i_C} = r_i(1 + A_v)$$

Equivalent characteristic values

$$S' \approx S$$

$$R'_i \approx \frac{1}{S}$$

$$D' \approx 1$$

ELECTRON TUBE

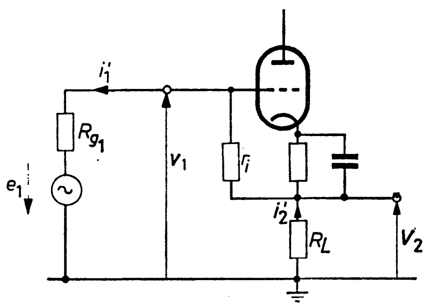
Total voltage amplification

$$A_{v_{\text{tot}}} = \frac{v_2}{e_1} = \frac{r_i}{R_g + \dfrac{r_i \parallel R'}{1 + \alpha \cdot A_v}} \frac{A_v}{1 + \alpha \cdot A_v}$$

Apparent internal resistance

$$R_{i_{\text{eq}}} = \frac{1}{S} \frac{(R_g \parallel r_i) + R'}{R_g \parallel r_{i_0}}$$

Earthed anode circuit



Voltage amplification

$$A_v = S \frac{R_i \cdot R_L}{R_i + R_L}$$

$$A_{v_{AB}} = \frac{A_v}{1 + A_v}$$

Current amplification

$$A_{i_{AB}} = 1 + A_i$$

Input resistance

$$r_{i_{AB}} = r_i(1 + A_v)$$

Equivalent characteristic values

$$S' \approx S$$

$$R'_i \approx \frac{1}{S}$$

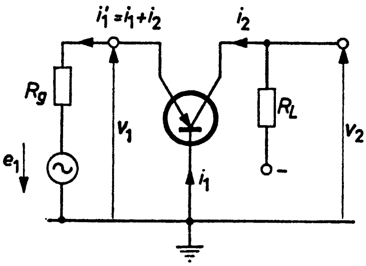
$$D' \approx 1$$

TRANSISTOR

Apparent internal resistance

$$R_{i_{eq}} = \frac{1}{S} + \frac{R_g}{1+\beta_0} \approx \frac{R_g}{\beta_0}$$

Common base circuit



Voltage amplification

$$A_v = S\,\frac{R_i \cdot R_L}{R_i + R_L}$$

$$A_{v_B} = A_v(1 + D)$$

Current amplification

$$A_i = A_v\,\frac{r_i}{R_L}$$

$$A_{i_B} = \frac{A_i}{1 + A_i}$$

Input resistance

$$r_{i_B} = r_i \parallel \frac{1}{S}$$

Apparent internal resistance

$$R_{i_{eq}} = \frac{1}{\dfrac{1}{R^*_{i_b}} + \dfrac{D_{v_b} \cdot \alpha_0}{r_{i_{0b}} + R_g}}$$

Equivalent characteristic values

$$S' = S(1 + D)$$

$$R_i' = R_i$$

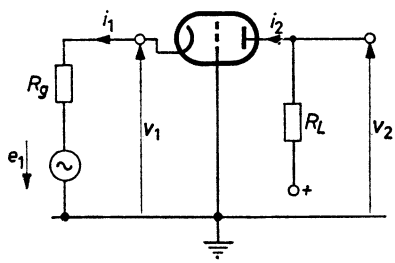
$$D' = \frac{D}{1 + D}$$

ELECTRON TUBE

Apparent internal resistance

$$R_{i_{eq}} = \frac{1}{S} \quad \frac{r_i + R_g}{r_i}$$

Earthed grid circuit



Voltage amplification

$$A_v = S \cdot \frac{R_i \cdot R_L}{R_i + R_L}$$

$$A_{v_{GB}} = A_v(1 + D)$$

Current amplification

$$A_i = 1$$

Input resistance

$$r_{i_{GB}} = \frac{1}{S}$$

. . . . .

Equivalent characteristic values

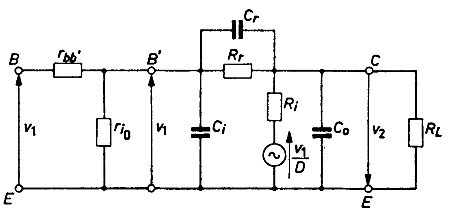
$$S' = S(1 + D)$$

$$R'_i = R_i$$

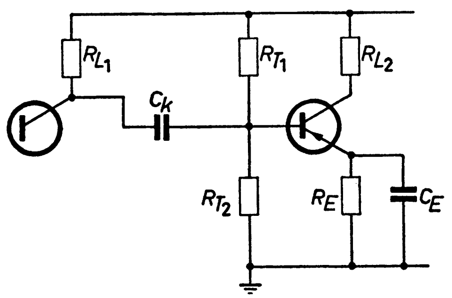
$$D' = \frac{D}{1 + D}$$

TRANSISTOR

High frequency equivalent circuit



Coupling capacitance C_K



Permissible amplification reduction

$$s = \frac{A'_v}{A_v}$$

At a lower frequency

$$\omega_v = 2\pi \cdot f_v$$

$$C_K = \frac{1}{\omega_v(r'_i + R_{L_1})} \sqrt{\frac{s^2}{1 - s^2}}$$

Lower limiting frequency ($s = 0.7$)

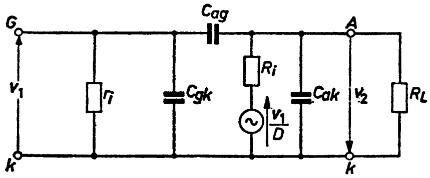
$$\omega_{v_g} = \frac{1}{C_K(R_{L_1} + r'_i)}$$

Emitter capacitance C_E

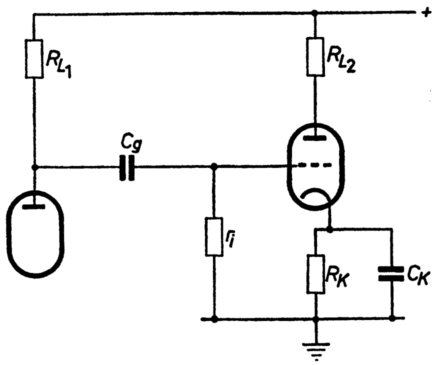
$$C_E = \frac{S \cdot r_i}{(R_g + r_i)\omega_v} \sqrt{\frac{s^2}{1 - s^2}}$$

ELECTRON TUBE

High frequency equivalent circuit



Coupling capacitance C_g



Permissible amplification reduction

$$s = \frac{A'_v}{A_v}$$

At a lower frequency

$$\omega_v = 2\pi \cdot f_v$$

$$C_g = \frac{1}{\omega_v \cdot r_i} \sqrt{\frac{s^2}{1 - s^2}}$$

Lower limiting frequency ($s' = 0.7$)

$$\omega_{v_g} = \frac{1}{C_g \cdot r_i}$$

Cathode capacitance C_K

$$C_K = \frac{s}{\omega_v} \sqrt{\frac{\left(S + \frac{1}{R_K}\right)^2 - \left(\frac{1}{R_K \cdot s}\right)^2}{1 - s^2}}$$

SYMBOLS

V	d.c. voltage
v	a.c. voltage
I	direct current
i	alternating current
S	slope
R_i	internal resistance
D_i	inverse current amplification
D_v	inverse voltage amplification
r_i	a.c. input resistance
r_{BE}	d.c. input resistance
R_g	source resistance, control generator
R_L	load resistance
R'	feedback resistance
α	feedback factor
r_{bb}'	base area resistance
s	reduction factor
A_v	voltage amplification
A_i	current amplification
β_0	short-circuit amplification
A	d.c. amplification for the emitter circuit
B	d.c. amplification for the base circuit
A_p	power amplification
P_c	dissipation
T	temperature (°C)
C_i	input capacitance
C_o	output capacitance
C_r	reaction capacitance
C_k	coupling capacitance
C_E	emitter capacitance

BIBLIOGRAPHY

P. A. NEETESON, *Junction Transistors in Pulse Circuits*, Philips Technical Library.

D. J. W. SJOBBEMA, *Using Transistors* (4th enl. ed.), Philips Technical Library.

G. FONTAINE, *Diodes and Transistors*, Philips Technical Library.

R. K. RICHARDS, PH. D., *Arithmetic Operations in Digital Computers*, D. van Nostrand Company, Inc., Princeton, New Jersey.

I. V. IDELSON, *The Introduction of Digital Modules in Industry*, Transactions of the Society of Instrument Technology, Volume 14, number 4, 1962.

S. HUMPHREY WATTS, JR., *Switching Circuits with Computer Applications*, McGraw-Hill Book Company, Inc., 1958.

RICHARD F. SHEA, *Transistor Circuit Engineering*, John Wiley & Sons, Inc., New York.

C. LE CAN, K. HART, C. DE RUIJTER, *The Junction Transistor as a Switching Device*, Philips Technical Library

W. W. GÄRTNER, *Transistors, Principles, Design and Applications*, D. van Nostrand Company, Inc., New York, 1960. Chapter 7.

R. A. GREINER, *Semiconductor Devices and Applications*, McGraw-Hill Book Company, Inc., New York, 1961, Chapter 12.

R. D. MIDDLEBROOK, *An Introduction to Junction Transistor Theory*, New York: J. Wiley & Sons, Inc., London: Chapman & Hall Ltd.

L. P. HUNTER, *Handbook of Semiconductor Electronics*.

DOSSE, *Der Transistor*, Verlag Oldenbourg, München.

RUSCHE/WAGNER/WEITZSCH, *Flächentransistoren*, Springer Verlag.

SCHLEGEL, *Der Transistor*, Verlag Schütz, Hannover.

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