Splitting the Load

OUTPUT STAGE USING TRANSISTOR WITH EMITTER AND COLLECTOR LOADS

By O. GREITER

THE natural approach of an engineer who is required to design an amplifier to deliver a fair amount of power is to begin by choosing a Class-B output stage. When the amplifier is to use transistors his move towards Class B is, if anything, even more pronounced. When he is called on to provide a very low output impedance his reaction is quite different. He may fly to positive feedback to help him out or he may adopt the cathode-follower or common-collector circuit but he will probably accept the inconvenience of Class-A working. In the more exacting cases this is not as bad as it seems, because frequently the requirement imposed on a transistor amplifier is that it should work satisfactorily up to a stated temperature which implies that the junction temperature will be very near its limit. Ingenious biasing circuits which use temperature-dependent biasing elements have been proposed for Class-B systems but these literally cannot be kept well enough in step with the junction to cope with practical operation even when, on steady-state tests, the tracking is good. The penalty is cross-over distortion, from which Class-A operation is the only escape.

The common-collector circuit would be ideal if it did not require so much drive: the base must be driven with a voltage which is roughly equal to the voltage across the load. Current is still required and it is found that the real problem has merely been moved back one stage. Even with valve stages it is often difficult to get sufficient swing at the grid to drive a cathode follower fully.

A common-emitter transistor can be driven very conveniently by a common-collector driver and this may in fact be achieved by forming the two into a compound pair. If this is done there is direct coupling between the two transistors: the leakage current of the first may upset the bias conditions of the second and, when this difficulty is avoided by a.c. coupling, we are faced by the problem of the very low impedance level.

The real trouble is that there is too much difference between the common-emitter connection and the common-collector connection. Each has its advantage, to excess, and each its disadvantage.

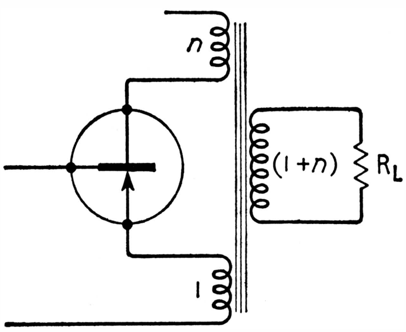


Fig. 1. Basic split-load stage.

We require a circuit in which we trade a loss in one characteristic against a gain in another. Fortunately such a circuit exists, although it seems to receive rather less than its due in the literature. This may be because it is possible to analyse it in a way which makes it almost incomprehensible: it can, however, be treated quite simply, so simply that one can only wonder why any other circuit is used.

Split Load Circuit

The skeleton of the circuit is shown in Fig. 1. It will be seen that instead of the primary of the output transformer being connected in the collector line of the common-emitter circuit, or in the emitter

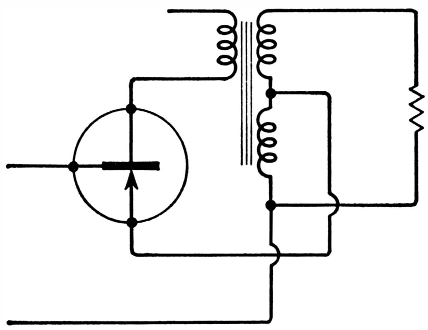
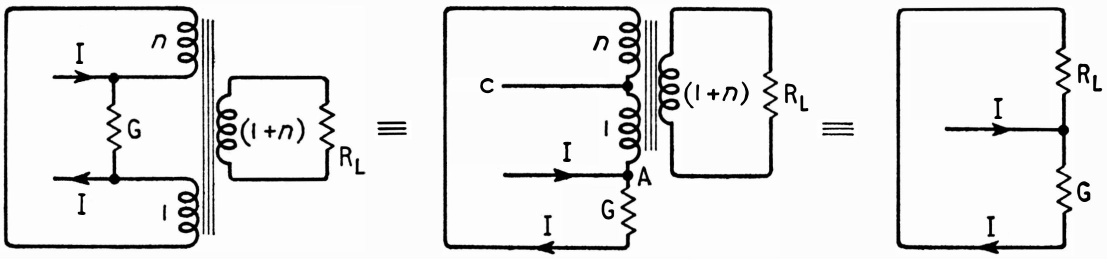


Fig. 2. Circuit of Fig. 1 re-drawn to show "application of feedback" principle.

line of the common-collector circuit, it is divided and the transistor is inserted at the division. By re-arranging the circuit in the form shown in Fig. 2 it becomes apparent that this is a way of providing negative voltage feedback round the output stage and it is well known that voltage feedback reduces the output impedance. It would appear to be a waste of time to go further, for the effect is known to involve a factor of $(1-\mu\beta)$. The difficulty is that μ is the unloaded gain and here there can be no doubt that even when the load resistor is removed the emitter path must constitute some sort of loading on the transistor. Referring, as one always does in these situations, to Bode* we see that we need to use the *fractionated gain* defined in Chapter V (5.11) and that this is the gain which would be obtained if it were possible to open the β circuit without affecting its impedance at either end. It may be that Mr. Roddam or Mr. Ray (Cathode variety) will care to show that this is easily done: it appears unlikely.

An approach which does not involve too much mathematics, and which will at least provide an approximation, begins by writing the a.c. equivalent circuit shown in Fig. 3a and then rearranging it

* *Network Analysis and Feedback Amplifier Design,* by Hendrik W. Bode, Ph.D., Van Nostrand Company Inc., New York, U.S.A. (1945).



(a) (b) (c)

Fig. 3. (a) A.c. equivalent circuit of split-load stage (b) Rearrangement leading to (c) in which the term n has disappeared.

as in Fig. 3b. By taking the transformer secondary winding to have simply the ratio $(1+n)$ to the two windings n and 1 we get rid of n altogether in the final simplified form (Fig. 3c). Using the standard form for the transistor as a current generator we have the conductance G and we shall interchange between G and R $(=1/G)$, R_L and G_L $(=1/R_L)$ to suit the convenience of the analysis.

We see at once that the voltage which appears across the load is $I/(G+G_L)$. It will be convenient to work in terms of the transistor mutual conductance in producing a first approximation and we can write $I = g_m v_{be}$.

Across the section cA of the transformer we shall get a voltage of $I/(G+G_L)(1+n)$ and this must be added to v_{be} to find what the input voltage will be, thus

$$v_{in} = v_{be} + [I/(G+G_L)(1+n)]$$
$$= I[1/g_m + 1/(G+G_L)(1+n)]$$

This expression can be "processed" to obtain the gain but our first concern is with the output impedance. We remove R_L and from a source of infinite impedance we force in a current I_o, as shown in Fig. 4a. Turning this drawing round we find we can deal with the very simple form shown in Fig. 4b.

The voltage is $v = (I_o - I)R$ and the impedance must be v/I_o which we proceed to find. We have, as before, $I = g_m v/(1+n)$, for now G_L is zero and there is no external input supplied. Thus

$$v/I_o = R/[1+g_m R/(1+n)].$$

It is slightly more convenient to write this as an output conductance.

$$G_o = G + g_m/(1+n).$$

Typical Values

An indication of the order of magnitude may be useful at this stage. The Westinghouse WT10 series have an R of about 100 Ω and a g_m of about 2A/V, or $1/g_m \approx 0.5\,\Omega$. If we take as a dividing line the point at which $G = g_m/(1+n)$ we see that as n falls below about 50 the second term becomes the dominating one. A typical load for such a transistor might be of the order of 10 Ω and this could be matched by making $(1+n)/g_m = 10$ or, quite roughly, $n = 20$. By making $n = 4$ we should get an apparent source impedance of only 2.5 Ω. We can thus move the impedance to any point we choose in the range 0.5 to 100 Ω. The price we must pay is loss of gain. We can get an idea of the input impedance by writing $I = h_{fe} i_b$ and substituting this in a simplified form for the gain equation based on the fact that G is much smaller than G_L. Then

$$v_{in}/i_b = h_{fe}(1/g_m + 1/(n+1)G_L.)$$

Generally we shall find that $1/(n+1)G_L$ is sufficiently greater than $1/g_m$ for us to write $R_{in} = R_L h_{fe}/(n+1)$ so that, as we might expect, the reduction of n leads to an increase of input impedance. When we have $R_L = 10\,\Omega$, $n = 4$ and $h_{fe} = 5$ we get an input resistance of 8 Ω, for example.

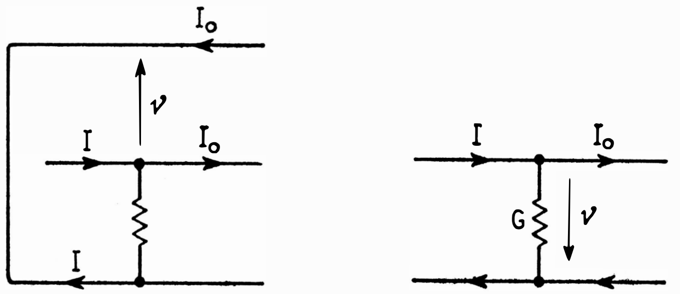
A very simple approximation shows that the power gain is about $(1+n)h_{fe}$, for the output voltage is multiplied by a factor of $(1+n)$ if the mutual conductance is high enough while the current is multiplied by a factor of h_{fe}. This approximation is very useful in the preliminary stages of design, when the transistor types to be used must be selected. Either a limit can be set on n or the size of the driver transistor estimated as soon as the output transistor is specified. Attention can thus be concentrated on the essential of the design.

Approximations and Their Effects

A defect of the approximations made is that it has been assumed that the system is driven from a zero-impedance source. Although the transistor requires to have current fed into its base to provide the transistor action it is assumed that this current will be produced by the voltage indicated on the $v_b I_b$ characteristics. For the transistor we have selected the input resistance is only about 2 Ω: only when the source resistance is well below this will the approximation be valid. An examination of the gain will show us what happens: the circuit to be considered is just the base-emitter path shown in Fig. 5, with an input signal v_{in}, a voltage at the emitter of $v/(1+n)$, an external source resistance r_s, and a base current of i_b.

For this we have:—

$$i_b = [v_{in} - v/(1+n)]/(R_s + R_{in})$$
$$I = i_b h_{fe} \text{ and } v = R_L I$$
$$= R_L h_{fe}[v_{in} - v/(1+n)/(R_s + R_{in})] \text{ so that}$$
$$v[1 + R_L h_{fe}/(1+n)(R_s + R_{in})] = v_{in} R_L h_{fe}/(R_s + R_{in})$$



(a) (b)

Fig. 4. (a) Current I_o is supplied from infinite impedance to determine output impedance. Form of circuit shown in (b) is simplified from (a).

Let us write $h_{fe}/(R_s + R_{in}) = g_m'$ and we have

$v_{in}/v = 1/(1 + n) + 1/g_m' R_L$ or

$v_{in}/I = 1/g_m' + 1/(1 + n)G_L$

When this is compared with the result obtained earlier we see that it differs only in the dropping of G, as it is small, and the substitution of g_m' for g_m. For practical purposes we may now assume that this same substitution must be made in the expansion for the impedance, without further proof.

One more factor merits examination. The feedback winding must deliver power and thus imposes a load on the transistor. This affects the value of G which should be used in the expression above and may make G large enough for it to be taken into account. We have the voltage $v/(1 + n)$ operating at a current level of i_b, so that we are concerned with $vi_b/(1 + n)$ which is $vI/(1 + n)h_{fe}$. The load power, however, is vI, so that unless $(1 + n)h_{fe}$ is small we need not concern ourselves too closely with this effect: if $(1 + n)h_{fe}$ is small we have other worries, for this is the approximate power gain.

The purist may by now be in rebellion at this patchwork approach. In small but exquisite writing he is preparing his polemic. Lay down your pen, Sir, touch not a hair of that old gray head†. None of these approximations is nearly so coarse as the approximations involved once we begin to put numbers into the expressions. Only accountants are happy if the books balance to the last penny when the company has lost thousands of pounds. The exact solution of this problem can be left to wait for the exact transistor.

† "The Ballad of Barbara Allen", which explains the spelling, and is quoted from a very shaky memory.

Effect of Winding Resistance

It is not without interest to note the effect of the resistance of the emitter winding on the output impedance. A very simple analysis shows that it adds a term ($g_m rG$) to the output conductance, where r is the resistance in the emitter leg which is producing current feedback. In any practical case it is likely to be so small compared with G that this term will not lead us far from the very simple approximation $g_m/(1 + n)$. For example, a 1Ω resistance in the numerical case we have considered would represent a 20% correction when $n = 20$ and only 5% when $n = 4$.

The flexibility of this circuit, which makes it so attractive, makes it difficult to explore its possibilities adequately without considering a very wide range of examples. It will be clear that the freedom of choice of output impedance which the method offers provides the designer with a most valuable extra degree of freedom, and also leaves him with the painful task of deciding how best to use this freedom.

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Clive Loehnis, director, Government Communications HQ.

C.M.G.

J. B. Adams, lately director-general of the European Organization for Nuclear Research, who throughout the war was at T.R.E. working on the development of centimetric radar.

K.B.E.

A.V-M. Walter P. G. Pretty, Air Officer Commanding-in-Chief R.A.F. Signals Command.

C.B.E.

D. A. Barron, deputy engineer-in-chief at the Post Office.

Dr. E. G. Bowen, chief of the Division of Radiophysics in the Australian Commonwealth Scientific & Industrial Research Organization, who was in Watson-Watt's radar team.

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