A.F. Amplification with the Cascode

AN OUTLINE OF THE ADVANTAGES OF THE TWIN-TRIODE OVER THE PENTODE FOR A.F.

By G. A. STEVENS

PRIOR to the development of Band III television tuners the cascode was very little used, its main use being in low noise r.f. stages, and it was used as a voltage amplifier in a stabilised power supply, where high gain was required.

The properties that enhance its use for r.f. amplification also indicate its suitability as an a.f. voltage amplifier. These advantages are as follows:—

1. High gain.
2. Low noise.
3. Low inter-electrode capacitances (particularly between input and output).
4. Capability of being designed with low phase shift (in feedback amplifiers).

The characteristics of the cascode that give rise to the above advantages are best illustrated by a simplified analysis of the circuit (a fuller analysis will be found in the " Radio Designers' Handbook " by Langford-Smith, Ch. 12, Sect. 9 XI).

In the basic cascode circuit as shown by Fig. 1, two halves of a double triode are connected in series, the bias for the lower valve, V1, being derived by grid current through the high value of grid resistor, the upper valves grid being held at a fixed potential V_B.

Since the anode potential of V1 is held constant by the cathode follower action of V2, the change in anode current of V1 caused by an alternating voltage V_{ip} on its grid is:—

$$\delta I_a = V_{ip} \cdot g_{m1}$$

where g_{m1} is the g_m of V1.

Providing V2 does not draw grid current, all current entering at its cathode must pass through the anode load R_L.

Hence,

$$V_{op} = \delta I_a \cdot R_L$$
$$= V_{ip} \cdot g_{m1} \cdot R_L$$
$$G = \frac{V_{op}}{V_{ip}} = g_{m1} R_L \quad .. \quad .. \quad .. \quad (1)$$

G. A. Stevens has for the past five years been a maintenance engineer with Rediffusion Television Ltd. From 1952 until being called up for National Service in the R.A.F. in 1959 he was in the Research Laboratory of E.M.I. which he joined on leaving school at the age of 19.

Hence the circuit has the same gain as a pentode with the same g_m as the lower valve.

Also, since for any valve that has an impedance R_K in its cathode, its anode impedance is increased by a factor:

$$(1 + g_m R_K).$$

We can write for the cascode:—

$$r_a = r_{a2} (1 + g_{m2} r_{a1}) \quad .. \quad .. \quad (2)$$

since the anode impedance of V1 is in series with the cathode of V2. Therefore, from equations (1) and (2) we can derive:—

$$\mu = r_a g_m = r_{a2} \cdot g_{m1} (1 + g_{m2} r_{a1}) \quad .. \quad (3)$$

From equation (2) it can be seen that for a twin triode

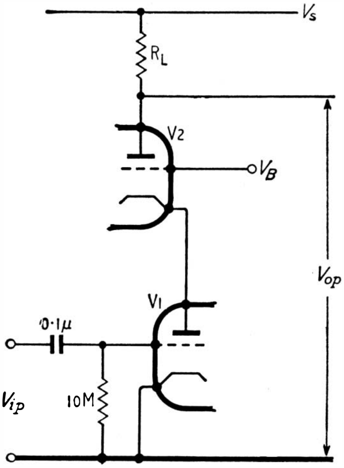


Fig. 1. Basic cascode circuit.

with $r_a = 100$ kΩ and $\mu = 70$ then its composite anode impedance is:—

$$r_a = 100 (1 + 70) \text{ k}\Omega \approx 7 \text{ M}\Omega$$

i.e., much more than a similar pentode.

The similarity of characteristics to a pentode is shown even more clearly by considering the knee voltage area. As stated above all the current flowing into the cathode of V2 must pass through the load R_L *providing no grid current is drawn.* However, once this point is reached, increasing the input drive only produces more grid current, so producing a very sharp pentode-like knee point as V2 bottoms. So far, it can be seen that the characteristics are those of a semi-idealized pentode, certainly better than the normal amplifying pentode. There are, however, two factors which make the cascode superior to the pentode for audio amplifier use.

Firstly, since there is only one grid in a triode, this taking no current, the partition noise generated by a pentode is absent, so reducing the noise level by a factor of three or more.

Secondly, in a pentode, the g_m is dependent on the anode current, which in turn depends on the load resistor

SYMBOLS	
$g_{m1}\ g_{m2}\ g_m$	Mutual conductance of V1, V2 and combination respectively.
$r_{a1}\ r_{a2}\ r_a$	Anode impedance of V1, V2 and combination respectively
$\mu_1\ \mu_2\ \mu$	Amplification factor of V1, V2 and combination respectively
$G_1\ G_2\ G$	Voltage gain of V1, V2 and combination respectively
G	Overall gain after application of negative feedback.
V_{ip}	Signal input voltage
V_{op}	Signal output voltage
R_L	Anode load
V_{k3}	The d.c. voltage at the cathode of V3
Z_o	The output impedance at the anode of V2
C_M	Effective Miller capacitance at the grid of V1
β	The fraction of the signal voltage at the cathode of V3, which is fed back to the grid of V2

and the supply voltage available. So that, for a given supply voltage, increasing the anode load decreases the g_m and a point is reached where increasing the load actually lowers the gain. In the cascode, however, the gain G is the product of the anode load of V2, and the g_m of V1.

Suppose now in Fig. 1 that a resistor was inserted between the anode of V1 (and cathode of V2) and the h.t. rail V_S. This would mean that V1 could be taking a reasonable current and so have a good g_m, whilst V2 could have a large value of anode load with its consequent low anode current.

This arrangement can, and does, give very high gain providing certain points are kept in mind.

1. Since V2 can now be cut-off without affecting the working conditions of V1, the bias of V2 (V_B) must be derived *via* potential divider of some sort from the anode of V2.
2. The anode load for V1 is the cathode of V2 and so is equal to $1/g_{m2}$ and when the mutual conductances of both valves are equal, the gain of V1 is equal to -1 and the Miller capacity at the grid of V1 is only about 3 pF (for an ECC83). However, with the top valve drawing only a small current its g_m is very low and presents a considerable load to V1. This means that the gain is shared between the two valves and the Miller capacity can be high, although the screening of input to output is still good, due to the earthed grid of V2.

Usually, a circuit with this gain would only be needed at the input of an amplifier, and should therefore be fed from a low-impedance source which swamps the effect of the Miller effect.

Since the grid of V2, and hence its cathode, is stabilised by the action of a potential divider from its anode, the effect of changes of bias on V1, due to signal level variation is, to a large part, compensated for, so allowing the use of grid current bias.

If, as is usual, the negative feedback for the whole audio amplifier in which the cascode is incorporated is applied to a small resistor in the cathode of V1 (about 100 Ω, which has little effect on the parameters of the circuit); then, the 0.1 μF capacitor and 10 MΩ resistor in the grid circuit are outside the feedback loop, and so do not affect the phase shift over the cascode. Similarly, since V2 is drawing very low anode current the voltage feed, V_B, to its grid can be supplied by another 10 MΩ resistor and 0.1 μF capacitor, so giving virtually no l.f. phase shift down to very low frequencies.

The bias voltage, V_B, can be supplied in one of two ways:—

1. By direct potential divider from the anode of V2. This method has two disadvantages, the maximum value of resistance usually made in standard ranges is 10 MΩ, and with this value as an anode load, the potential divider has a resistance of this order which shunts the signal, and so reduces the gain.
2. The anode load and potential divider can, if the valve of R_L is too high cut off V2 by reducing its anode voltage to the point where no current flows through the valve.

The above drawbacks limit the straight potential divider at high values of R_L, although when R_L is around 1 MΩ or less there is nothing to choose between this and the next bias method to be described.

If the potential divider, instead of being applied direct, is in the cathode load of a cathode follower, the grid of which is connected to the anode of V2 the above drawbacks disappear.

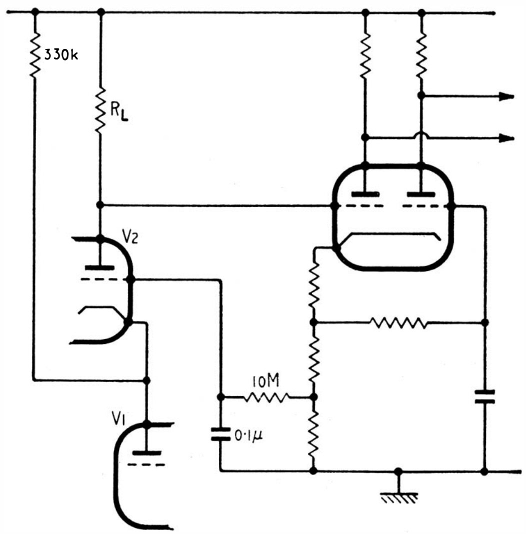
Usually in modern audio amplifiers the trend is to follow the input voltage amplifier with a directly coupled phase splitter, of the cathode-coupled variety, in this case the cathode resistor of the phase splitter may be replaced by two other resistors so forming the necessary bias network for V2, as illustrated by Fig. 2.

Practical cases

In order to investigate a practical case the two circuits shown in Figs. 3 and 4 (with and without the cathode follower bias network respectively) were constructed. The value of the 330 kΩ current bleed resistor was found experimentally for an anode load of 10 MΩ and on altering the load the value still seemed about optimum and so no experiments were conducted using values other than this.

S1 and R_2 were used to determine the output impedance of the circuit, R_2 being altered until closing S1 reduced the

Fig. 2. Bias arrangement of the cascode circuit to feed a cathode-coupled phase inverter.



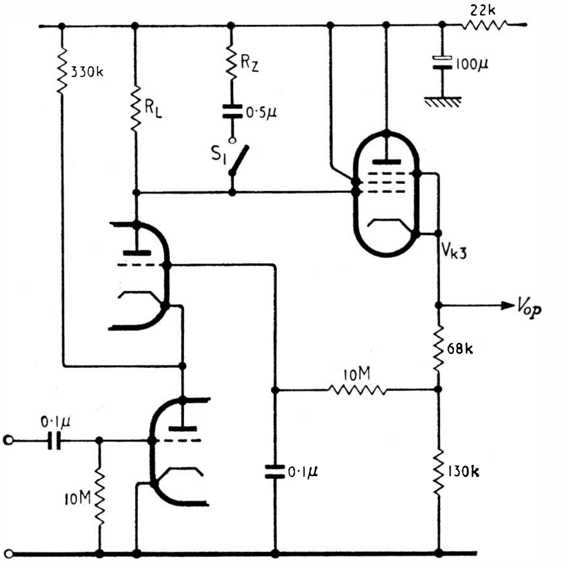


Fig. 3. Circuit used for measurements incorporating the cathode biasing method. Results are shown in Table I and Figs 6 and 8.

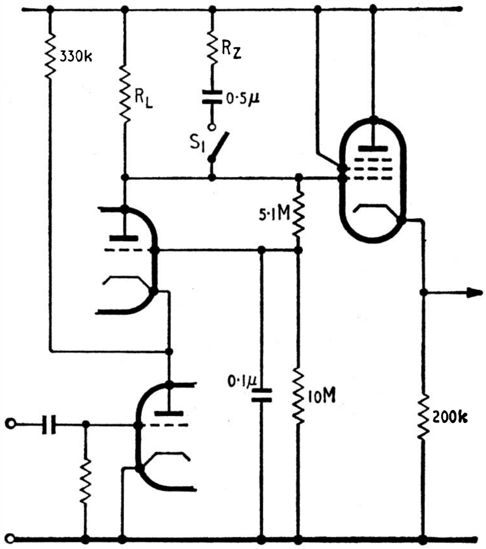


Fig. 4. Test circuit used for measurements. S_1 and R_2 were used to determine the output impedance. Results are shown in Table I.

output signal by 6 dB, R_2 then being equal to the output impedance (the series capacitor having negligible reactance).

In the experiments only the anode resistors were altered, together with slight adjustments to the stabilized power supply to give 300 V after the decoupling resistor.

The gain from the input to the anode of V1, was also measured for different loads, this giving the measure of Miller capacity on the input, and also the gain of the cascode if:—

1. The grid of V2 is not decoupled to earth; or
2. The effect of putting a frequency sensitive network between the cathode of V3 and the grid of V2 to act as an equalizer circuit.

The results are set out in Table I and in the graphs of individual parameters which follow later. The odd values of R_L being the measured values of standard resistors.

As can be seen, when the anode load is in the order of 10 MΩ the output impedance is 3.5 MΩ. In the test circuit 3 the stray capacity at the anode of V2 was 8 pF and so the response was 3 dB down at about 6 kc/s, the gain-bandwidth factor is given by the ratio G/Z_0 and as can be seen, increases with lower anode loads. The early roll-off associated with the highest gains can in fact be a desirable feature in feedback amplifiers, where it is necessary to reduce the loop gain below unity before the phase shift reaches 180° in order to avoid instability. When the cathode-coupled phase splitter uses a pentode in the first stage, so avoiding large values of Miller capacitance loading the anode of V2, the total stray capacitance should not exceed 16 pF and so the 3 dB upper

TABLE I

				FIG. 3 Cathode follower bias					FIG. 4 Potentiometer bias				
RL	V_{k3}	C_M	G_1	G	G(dB)	Z_0	G_2	G/Z_0	G	G(dB)	Z_0	G_2	G/Z_0
16 MΩ	153V	103pF	68	4,800	73.5	3.8 MΩ	70	1.26			Cut-off		
13 MΩ	153	103	68	4,800	73.5	3.67 MΩ	70	1.26	940	59.5	2.58 MΩ	13.8	0.365
10 MΩ	153	90	60	4,400	72.8	3.52 MΩ	73	1.25	2,350	67	2.2 MΩ	39	1.02
6 MΩ	154	75	50	3,750	71.3	2.48 MΩ	75	1.52	2,500	68	1.73 MΩ	50	1.44
3 MΩ	157	53	35.6	2,850	69.0	1.72 MΩ	80	1.64	2,100	66	1.35 MΩ	59	1.55
1.5 MΩ	161	34	22.5	1,850	65.2	1.13 MΩ	82	1.62	1,480	63.5	940 kΩ	66	1.58
1 MΩ	165	24.5	16.3	1,325	62.5	790 kΩ	81	1.67	1,130	61	693 kΩ	69	1.63
620 kΩ	173	16	10.8	900	59.0	540 kΩ	83	1.67	815	58	493 kΩ	77	1.66
400 kΩ	184	11.5	7.6	632	56.0	370 kΩ	83	1.71	610	55,5	334 kΩ	80	1.83
253 kΩ	195	8.5	5.6	423	52.5	235 kΩ	75	1.80	418	52	203 kΩ	75	2.03
137 kΩ	217	5.5	3.6	251	48.0	130 kΩ	70	1.93	250	48	125 kΩ	70	2.0
91 kΩ	232	4.4	2.9	176	45.0	87.5 kΩ	61	2.01	175	45	86 kΩ	61	2.03
52 kΩ	253	3.1	2.1	98	40.0	51.1 kΩ	47	1.92	98	40	50 kΩ	47	1.96
30 kΩ	260	2.5	1.7	56.5	35.0	28.8 kΩ	33	1.96	56.5	35	28 kΩ	33	2.02

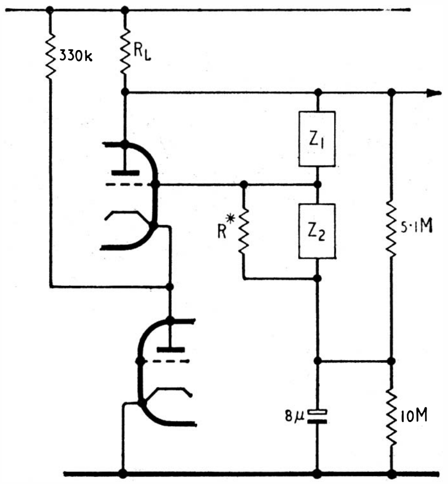


Fig. 5. Use of the cascode for equalization by frequency selective feedback. R^* is only required to provide bias if Z_2 is blocked to d.c.

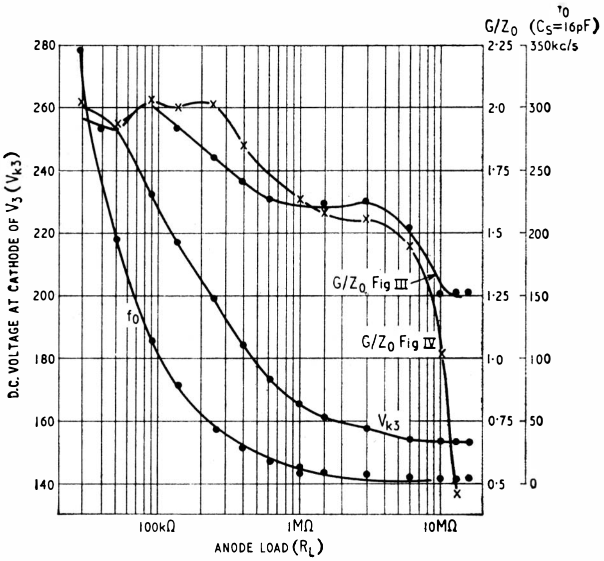


Fig. 6. Some of the results of Table I obtained from the circuits of Figs 3 and 4.

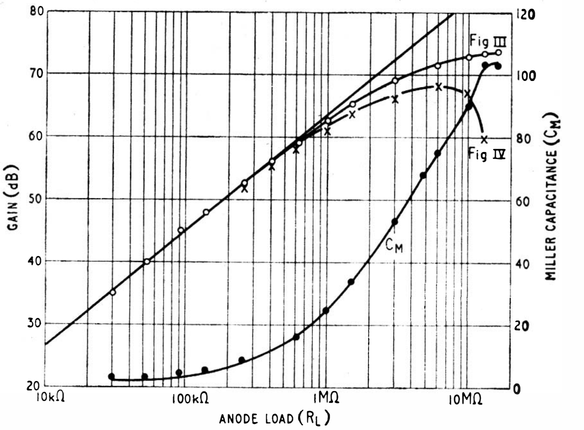


Fig. 8. Some of the results, listed in Table I, obtained from the circuits shown in Figs 3 and 4.

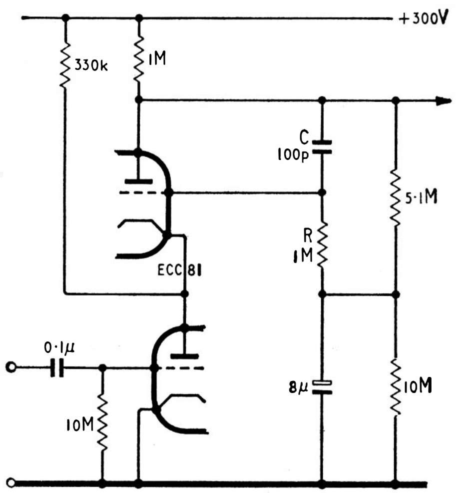


Fig. 7. Particular case of Fig. 5 for use as a tape playback equalizer.

frequency limit should be at about 15 kc/s with an anode load of 1 MΩ giving a gain of over 60 dB. The gain-bandwidth product may be obtained from the factor G/Z_o by multiplying it by 10^4 (assuming 16 pF strays).

An EF86 pentode with an anode load of 220 kΩ and a following grid resistor of 1 MΩ has a gain of about 200 (46 dB), the effective anode load due to the above two resistors is 180 kΩ, and the cascode gives a gain of 320 (50 dB) with a load of this value, while increasing the pentode load to 1 MΩ and running under starvation conditions gives a gain of 400 (52 dB), the cascode giving 1,200 (62 dB), an even more marked improvement over the standard pentode circuit.

The ECC83 has a normal heater, as opposed to the bifilar type of the EF86, and so the hum introduced by the circuit could be expected to be greater than that of the pentode. With the 10 MΩ load in circuit, (as with this value the hum and noise would be expected to be at their worst), and one side of the heaters earthed the total hum and noise at the output was 5 mV r.m.s., and when the heaters were on a d.c. supply the value fell to 1 mV r.m.s.—these correspond to an input level of 3.15 and 0.67 μV r.m.s. respectively over a frequency 25c/s—6 kc/s. The measurements were taken on Fig. 3 with the input shorted, and includes hum and noise introduced by the output cathode follower.

As can be seen in Fig. 4 the gain v load characteristic is linear up to 2 MΩ (for Fig. 3) and then falls off and becomes asymptotic to the 75 dB co-ordinate. This corresponds to the value of μ at low currents as given in the published curves of the ECC83.

These results show the superiority of the cascode circuit over the conventional pentode in audio applications, and whilst the investigations were concerned only with the ECC83, the newly-developed ECC807 should show an even more marked improvement in cases where better hum or gain figures might be needed, although the excellent results obtained with the older valve type would usually make higher gain unnecessary in all but a few cases.

Finally, the effects of valve changes on the gain and d.c. level at the cathode of V3 (V_{k3}) were investigated by putting seven different new valves in the V1 and V2 position. The total gain variation was 10%, and the output voltage variation was over an 18V range.

APPENDIX

Equalization with the cascode

If the grid of V2 is not decoupled to earth, but is included in a frequency sensitive network as shown in Fig. 5, then the overall gain is a function of frequency.

The circuit now behaves with V1 as a pre-amplifier with gain G_1, and V2 as a frequency-selective feedback amplifier. The gain of V2 as a feedback amplifier is given by:—

$$G'_2 = \frac{G_2}{1+G_2\beta} \text{ where } \beta = \frac{Z_2}{Z_1 + Z_2}$$

N.B. The application of feedback increases the input impedance of V2 and so also the gain of V1, this causes a reduction in the total lift supplied by the overall circuit, but for values of R_L greater than about 1 MΩ this effect is not serious, but in any case a little experimentation will soon determine the correct values.

Then the total gain of the circuit will be:—

$$\begin{aligned} G' &= G_1 \,.\, G'_2 \\ &= \frac{G_1\ G_2}{1+G_2\ \beta} \\ &= \frac{G}{1+G_2\ \beta} \quad = \frac{G\,(Z_1 + Z_2)}{Z_1 + Z_2\,(G_2 + 1)} \end{aligned}$$

If Z_1 is not blocked to d.c. a blocking capacitor would have to be inserted to maintain the correct bias level on V2.

An example of a tape equalizer circuit is given in Fig. 7, the values of C and R given provide a time constant of 100 μs with a total top cut of about 36 dB, the l.f. gain being 60 dB, dropping to 24 dB at h.f. www.keith-snook.info