Electronic Circuitry

Selections from a Designer's Notebook

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COUPLINGS IN D.C. AMPLIFIERS

DIRECT coupled amplifiers are used in a variety of applications ranging from amplifiers for oscilloscopes to industrial electronic equipment of various types. The design of such amplifiers is often quite specialized, and to those unfamiliar with these devices one of the commonest sources of difficulty is the question of the coupling between successive stages.

In ordinary amplifiers for audio signals, say, the coupling presents no difficulty and the ordinary RC coupling is commonly used. Such a coupling does not transmit zero frequency however, and as soon as extension to zero frequency is required, difficulties begin to arise. The simplest coupling in d.c. amplifiers is that shown in Fig. 1, which is really not a coupling at all inasmuch as the grid of the second stage is directly connected to the anode of the first. Unfortunately the anode of V_1 is positive to earth by the voltage V_a and this is equally true of the grid of V_2. For this reason the cathode of V_2 must be held at a positive voltage $V = (V_a + V_k)$, where V_k is the bias voltage for V_2.

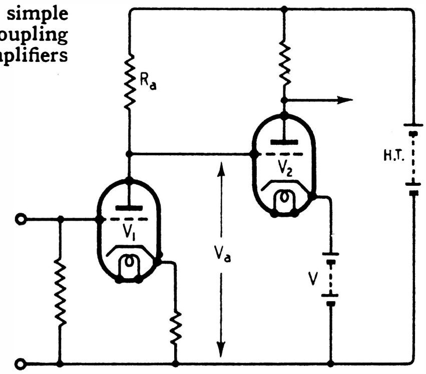


Fig. 1. A simple anode-grid coupling in d.c. amplifiers

In practice it is inconvenient to use a battery to supply V, so that the cathode of V_2 is either tapped into a bleeder network across the h.t. supply, or, better, held at the required voltage by a cathode follower as in Fig. 2. If V_2 and V_3 are similar valves, then if the grid of V_3 is held at a positive potential equal—or nearly so—to V_a, suitable working conditions will be obtained for V_2. The standing current in V_2 will be approximately $V_a/2R_k$, and the same will be true of V_3. It is obvious from the circuit that V_2 and V_3 form a cathode coupled pair with an anode load on one valve only (R_{a2} on V_2). If V_2 is a triode, then the usual Miller effect will throw a relatively large capacitance across R_{a1}. This may be avoided by short-circuiting R_{a2}, and placing an equal load (R_{a3}) in the anode circuit of V_3, as shown dotted. The gain of the V_2, V_3, circuit from the grid of V_2 to the anode of V_3 is given by:

$$A_{2,\,3} = \frac{\mu R_{a3}}{R_{a3} + 2r_a + \dfrac{r_a\,(R_{a3} + r_a)}{(\mu + 1)\,R_k}} \quad \cdots \quad \cdots \quad (1)$$

where $\left.\begin{array}{l}\mu = \text{amplification factor}\\ r_a = \text{anode resistance}\end{array}\right\}$ of V_2 and V_3

The disadvantage of this method of coupling is the high value of h.t. supply potential (V_b) required. Provided the grid-cathode bias of V_2 and V_3 is small compared with V_a, the effective supply potential for V_2 and V_3 is only ($V_b - V_a$), and this must be sufficient to supply the voltage drop across R_{a3} (or R_{a2}), and the anode potentials of V_2 and V_3. It would be more convenient if we could arrange to operate the cathode of V_2 at earth potential; this may be done with the circuit of Fig. 3.

In Fig. 3 a negative supply of V_c volts is used. If V_2 is to be operated at zero bias, it turns out that ($V_cR_1 = V_aR_2$) is the condition which must be satisfied. In fact the grid of V_2 will nearly always be

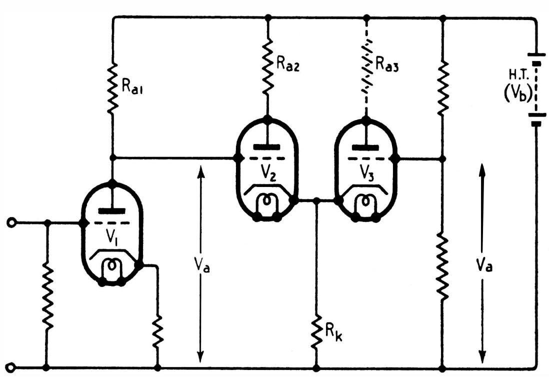


Fig. 2. Direct coupling on to cathode-coupled pair, with cathode of V_2 held at required voltage by V_3

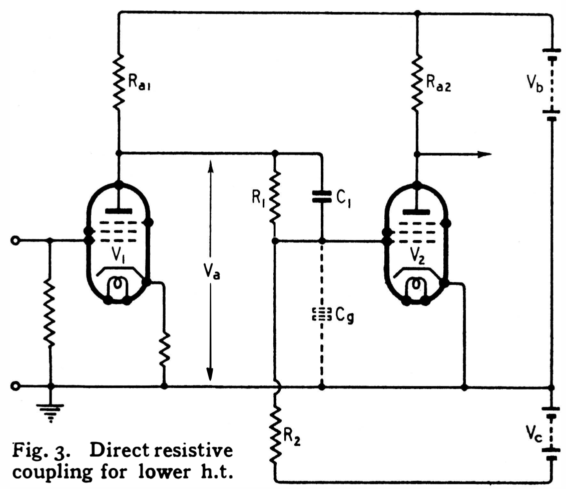
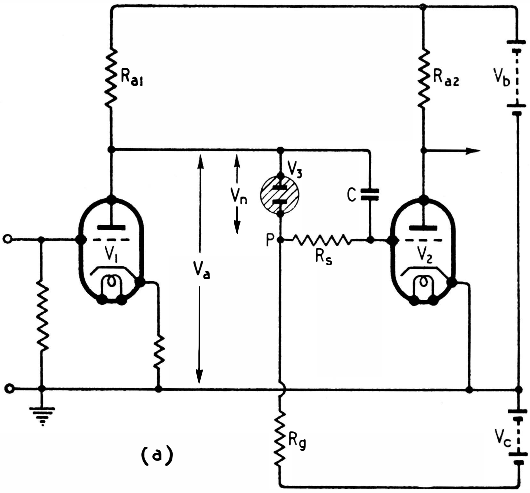


Fig. 3. Direct resistive coupling for lower h.t.



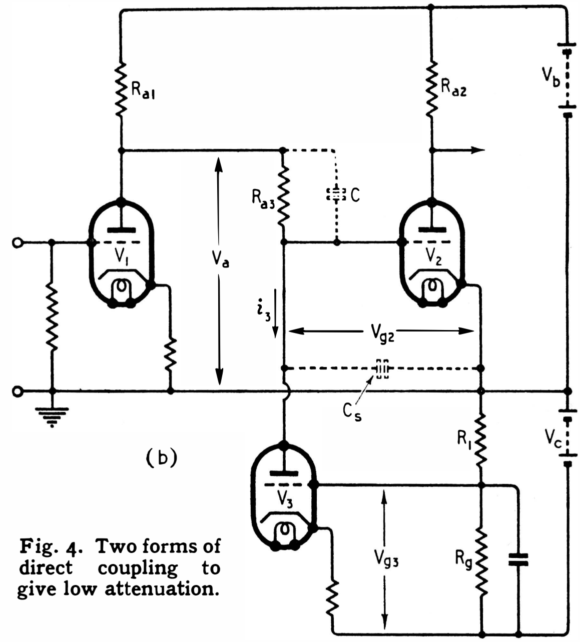


Fig. 4. Two forms of direct coupling to give low attenuation.

operated negative to earth, so that this simple expression is only an approximation. However if V_c is large compared with the grid base of V_2, the bias of V_2 may be neglected in the calculation of V_c. The disadvantage of this method is obviously the fact that an attenuation of $R_2/(R_1 + R_2)$ is inserted between the two stages, and if $V_a = V_c$ the loss will be one half. A virtue may be made of this disadvantage if high frequencies are important, as R_1 may be shunted by a condenser C_1 to make the time constant $R_1C_1 = R_2C_g$, where C_g is the input capacitance of V_2. The attenuation of the R_1, R_2 network is then independent of frequency, and the capacitance loading on the anode of V_1 by the grid circuit of V_2 is reduced from C_g to $C_g/(1 + C_g/C_1)$.

Low Attenuation Couplings

In purely low frequency applications, the loss of gain introduced by the coupling may be important and two means have been found for overcoming this disadvantage ; these are shown in Fig. 4(a)[1] and (b)[2]. At (a) a gas discharge stabilizer tube has been substituted for R_1 of Fig. 3. As such a tube has the property of maintaining a nearly constant potential across its terminals, it follows that any changes at the anode of V_1 will be transferred without alteration in amplitude to the point P. As the point P will have to be negative to earth by the bias of V_2, (V_g), a series resistor, R_g, to a negative supply, V_c, is needed to maintain sufficient current in the stabilizer to maintain the discharge. It is not difficult to see that the minimum safe value of V_c is given by :

$$V_{c\,min} = V_s + V_g$$

where V_s is the striking voltage of V_3.

In order to maintain V_2 at the required bias (V_g) it is also not difficult to see that

$$V_n = V_a + V_g$$

where V_n is the running voltage of V_3.

The resistance R_s and the condenser C are included because stabilizer tubes often have an impedance varying with frequency.

If R_s is made large compared with the nominal internal resistance of V_3 (R_n = 500 Ω or so) and the time constant CR_s made sufficiently long (1/10 to 1/100 second), variations of R_n with frequency become unimportant at the higher frequencies, as C progressively approximates to a short circuit as the frequency increases.

Because of these difficulties with stabilizer tubes the alternative circuit of Fig. 4(b) is sometimes used. In this arrangement the valve V_3, and its anode and cathode loads R_{a3} and R_k, form the coupling network. As readers are well aware, the use of negative current feedback introduced by a cathode resistor leads to a very high effective anode resistance, r_{ao}. In fact

$$r_{ao} = (\mu_3 + 1)R_k + r_{a3} \quad \ldots \quad (2)$$

where r_{a3} = anode resistance of V_3. So that the transmission of the coupling is

$$T = \frac{(\mu_3 + 1)R_k + r_{a3}}{(\mu_3 + 1)R_k + r_{a3} + R_{a3}} \quad \ldots \quad (3)$$

If V_3 is a high-μ valve this transmission is generally not far short of unity.

The circuit is easy to design if V_3 is a high-μ valve, for then the current, i_3, in V_3 is nearly V_{g3}/R_k, and the voltage drop across R_{a3} is $i_3R_{a3} = \frac{R_{a3}}{R_k} V_{g3}$.

In practice low current high-μ valves like the 6Q7G, or 6F5G, are very suitable for the V_3 position, and the coupling transmission then approaches unity. Taking a practical example let us make V_a = 145 volts, V_{g3} = 50 volts, R_k = 100 kΩ, so that $i_3 = \frac{1}{2}$mA approx. If the required bias on V_2 (V_{g2}) is 5 volts, R_{a3} must drop 145 + 5 = 150 volts at $\frac{1}{2}$mA. Thus R_{a3} = 300k Ω. If μ_3 = 50 and r_{a3} = 70k Ω (for a 6Q7G) the transmission turns out to be

$$T = \frac{51 \times 100 + 70}{51 \times 100 + 370} = 0.945$$

If good high frequency performance is required—when V_1 and V_2 will usually be pentodes—R_{a3} may be shunted by a condenser C so that $CR_{a3} = C_s r_{ao}$ and

[1] Miller, S., *Electronics*, Nov. 1941.
Valley and Wallman, "Vacuum Tube Amplifiers," p. 486.

then the value of T will become independent of frequency. In our example above r_{ao} was $5.17\text{M}\Omega$, and if we assume $C_s = 20\text{pF}$, a likely value, $C = 5.17 \times 20/0.3 = 345\text{pF}$. In the circuit of Fig. 4(b), the bias of V_2 may be adjusted over a small range by variation of R_k or R_g with but little change in the transmission of the coupling.

One other[3] rather interesting coupling is very useful for wideband direct coupled amplifiers. This is a variant on the ordinary resistive coupling of Fig. 3, and is shown in Fig. 5. It is applicable only when V_1 is a pentode as shown. In wideband amplifiers it is essential to use a low value of anode load, in order that the inevitable stray shunting capacitance shall produce the usual droop in the high frequency response at some conveniently high frequency. If R_{a1} in any of the previous circuits is made small, then V_a would assume a value approximating to the full h.t. potential. This involves a greater loss in the coupling than is the case with Fig. 5.

Consider now the circuit of Fig. 5 at some high frequency, but not so high that the stray capacitance has produced noticeable loss of gain. At such a frequency C_d is a virtual short-circuit, and so is C_1. Thus the load on V_1 is R_a, and all the signal voltage across R_a is transmitted to the grid of V_2 through C_1. Now consider the circuit at zero frequency. The total load on V_1 is $(R_a + R_{ad})$ so that V_a may be made relatively low compared with V_b. At the same time any zero frequency signal is transmitted to V_2 with a loss of $R_2/(R_1 + R_2)$. If $(R_a + R_{ad})$ is small compared with r_a of V_1, the gain of V_1 will be proportional to the anode load. If the gain times the coupling loss can be made constant at all frequencies where loss of gain due to stray capacitance is negligible, a very advantageous result will have been secured. In the arrangement shown we may take the signal current (I_a) in V_1 to be constant with constant input (v), and in passing we see that

$$I_a = \frac{g_{m1}v}{1 + g_{m1}R_k} \quad \ldots \quad \ldots \quad \ldots \quad (4)$$

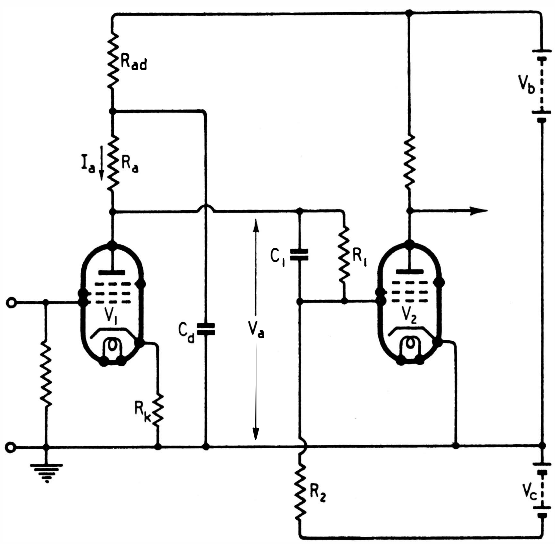


Fig. 5. A direct coupling used in wideband amplifiers.

[3] Edwards and Cherry, *J.I.E.E.*, Vol. 87, p. 178 (1940).

It turns out on analysis that if

$R_2 = nR_a$ and, $R_1 = nR_{ad}$ } Where n is any convenient numerical ratio, e.g. 20 or 50.

and $R_1C_1 = R_{ad}C_d$

the output signal voltage is simply

$$V_g = \frac{n}{n+1} \cdot I_aR_a$$

at all frequencies where stray capacitances are unimportant.

Consequently the gain of V_1 from the grid of V_1 to the grid of V_2 is

$$A = \frac{g_{m1} R_a}{1 + g_{m1} R_k} \cdot \frac{n}{n+1} \quad \ldots \quad \ldots \quad (5)$$

which is the usual expression for the gain of a pentode with an un-bypassed cathode bias resistor, except for the factor including n which may be made to approach unity by making n large.

The circuit of Fig. 5 is in fairly wide use now, and one of its main advantages is that R_{ad}, C_d, form a decoupling network which reduces the injection into the h.t. supply of the higher frequency components of the signal current in V_1. This is very useful in preventing instability and undesired feedback from one stage to an earlier one via a common impedance in the h.t. supply.

At low frequencies approaching zero, decoupling networks cease to be effective, so that in multi-stage direct coupled amplifiers a very low impedance h.t. supply—preferably stabilized—has to be used. The same is true of the negative supplies shown in various coupling circuits, although here it is usually stability rather than low impedance that is the prime consideration, since the currents taken from the negative supply are usually quite small.

No mention has been made of drift in d.c. amplifiers in the foregoing, but as any changes in the anode voltage of the first valve due to changes in the valve itself with changing temperature or other electrode potentials, are transmitted more or less completely to the next stage, a steady undesired drift in the anode potential of the output stage occurs only too frequently. There are various means of combating this, which can be found in the extensive literature of the subject. www.keith-snook.info

TELEVISION RECORDING

A SYSTEM of television recording has been developed by B.B.C. engineers. It is a combination of cinematographic and television apparatus and enables programmes to be "telefilmed"—as it is called—so that they can be re-transmitted at some future time with little loss of the original picture quality. The Service of Remembrance and the Lord Mayor's Show were among the first O.B.s to be telefilmed for a second transmission in the evening programmes.

The recording system uses a continuous-motion film camera in which the movement of the film is chased by an optical image of the television screen picture reflected from a rotating mirror drum. By this means all the 405 interlaced lines of the picture are recorded on the film and the difficulties of relating the television frame frequency to the picture repetition frequency on the film are overcome.

The method was proposed by H. W. Baker, Engineer-in-Charge at Alexandra Palace, and H. G. Whiting, now Engineer-in-Charge of Sutton Coldfield, in collaboration with D. R. Campbell, a senior engineer at Alexandra Palace, and was perfected by W. D. Kemp, of the B.B.C. Planning and Installation Department.