CHAPTER 12

AUDIO FREQUENCY VOLTAGE AMPLIFIERS

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SECTION 1 : INTRODUCTION

(i) Voltage amplifiers

A voltage amplifier is one in which the voltage gain is the criterion of performance. To be strictly correct it is not possible to have voltage without power since infinite impedance does not exist in amplifiers, but for all ordinary purposes a "voltage amplifier" is one in which a "voltage" output is required. Voltage amplifiers generally work into high impedances of the order of 1 megohm, but in certain cases lower load impedances are used and there is no sharp demarcation between "voltage" and "power" amplifiers. In cases where transformer coupling is used between stages, the secondary of the transformer may be loaded only by the grid input impedance of the following stage and the numerical value of the impedance may not be known. In such cases the transformer is usually designed to operate into an infinite impedance, and the effect of normal grid input impedances on the transformer is very slight compared with the primary loading.

It is important to bear in mind the reversal in polarity which occurs in any valve when used as an amplifier with a load in the plate circuit. As a consequence, the a-f voltage from grid to plate is

$$E_{gp} = E_{gk} + E_{kp} = E_{gk}(1 + A) \tag{1}$$

where A is the voltage gain from grid to plate.

SECTION 2 : RESISTANCE-CAPACITANCE COUPLED TRIODES

(i) Choice of operating conditions (ii) Coupling condenser (iii) Cathode bias (iv) Fixed bias (v) Grid leak bias (vi) Plate voltage and current (vii) Gain and distortion at the mid-frequency (viii) Dynamic characteristics (ix) Maximum voltage output and distortion (x) Conversion factors with r.c.c. triodes (xi) Input impedance and Miller Effect (xii) Equivalent circuit of r.c.c. triode (xiii) Voltage gain and phase shift (xiv) Comments on tabulated characteristics of resistance-coupled triodes.

(i) Choice of operating conditions* (Fig. 12.1)

Any triode may be used as a r.c.c. amplifier, but for most purposes the valves specially suitable for this application may be grouped:

1. **General purpose triodes** with μ from 15 to 50, and plate resistance from 6000 to 10 000 ohms (with battery types somewhat inferior). These are also called "medium mu" triodes.

2. **High-mu triodes** with μ from 50 to 100, and plate resistances say from 50 000 to 100 000 ohms.

The **load resistance** (R_L) may be any value from a few ohms to many megohms, but for normal operation R_L should never be less than twice the plate resistance at the operating point, with a higher value preferred. The load resistance should never be greater than the following grid resistance (R_{g2}) and should preferably be not more than one quarter of R_{g2}. The following table is a good general guide, but capable of modification in special circumstances.

Valve type—	General purpose		High-mu		
Following grid resistor—	0.22 to 1	0.22	0.47	1	MΩ
Load resistance	0.1	0.22	0.22	0.22	MΩ
			or 0.47	or 0.47	MΩ

The optimum combination of R_L and R_{g2} for maximum output voltage is covered in Sect. 2(ix).

The **plate supply voltage** (E_{bb}) should generally be as high as practicable provided that the maximum ratings are not exceeded. Plate supply voltages up to 300 volts are safe for use with all types of indirectly-heated valves unless otherwise stated. Somewhat higher supply voltages may be used with triodes provided that the designer ensures that, other than momenttarily when switching on, the maximum plate voltage rating is not exceeded under any possible conditions.

FIG. 12.1

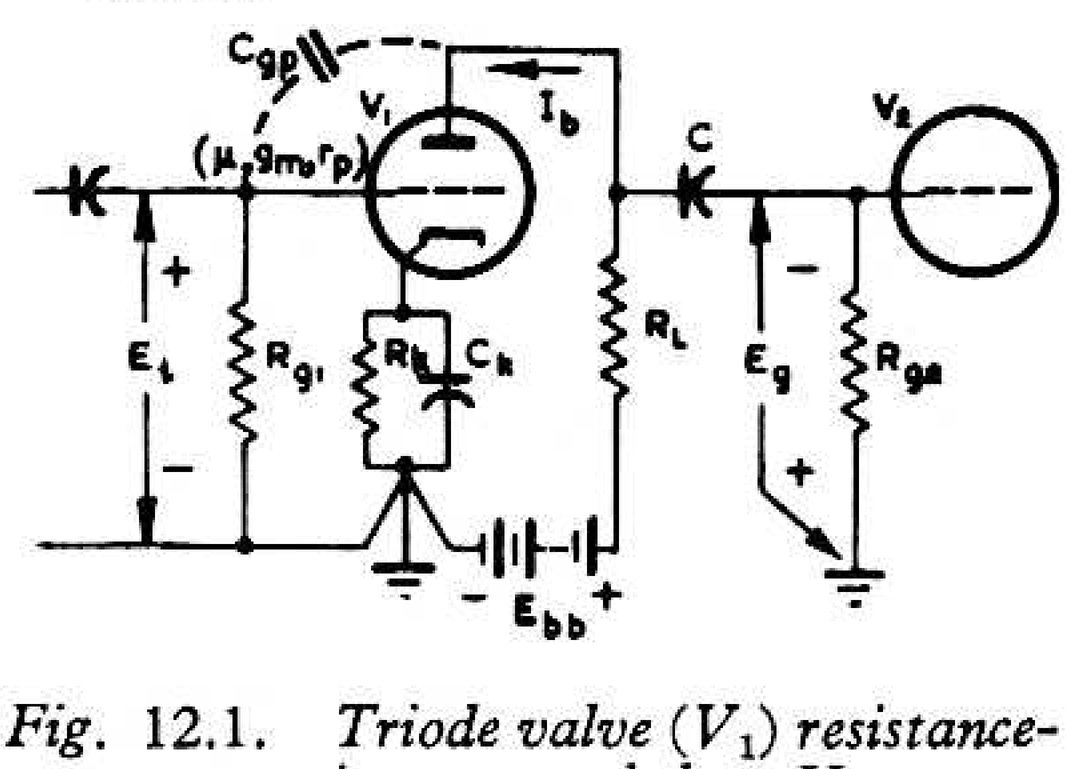


Fig. 12.1. *Triode valve (V_1) resistance-capacitance coupled to V_2.*

The **input grid resistor** (R_{g1}) should not normally exceed 1 megohm with indirectly-heated valves, although higher resistances are satisfactory with 1.4 volt battery valves. Higher grid resistances will not do damage to a r.c.c. triode, provided that the plate load resistance is not less than, say, 0.1 megohm with 300 volts plate supply, but the reverse grid current may be sufficient to shift the operating point into the region of distortion and lower gain. For this reason cathode bias is to be preferred to fixed bias and the input grid resistor should not be higher than necessary, particularly with high-mu triodes—see (iv) below. For example, little is gained by using a grid resistor having more than four times the resistance of the plate load resistor of the preceding stage.

Values of input grid resistor in excess of 1 megohm may only be used satisfactorily in low-level pre-amplifier operation where the valve is one specially manufactured or tested for this class of service, under a specification which ensures that the reverse grid current is very low. With a maximum reverse grid current of 0.2 μA and with

*See Chapter 3 Sect. 1(iv) for valve ratings and their limiting effect on operation ; also (v) Recommended practice and operation.

μ not greater than 40, a grid resistor up to 5 megohms may be used, provided that the total resistance of the plate load resistor and the plate decoupling resistor is not less than 0.1 megohm.

If grid-leak bias is used, R_g may be from 5 to 10 megohms.

The output grid resistor (R_{g2}) may be the **maximum recommended** for the following stage by the valve manufacturers—usually 0.5 megohm for power valves with cathode bias—or as determined by eqn. (6) or (7) in Chapter 3 Sect. 1(v)d. Lower values are desirable to reduce the effects of reverse grid current in the following stage, and there is no appreciable advantage in either gain or distortion through the use of a resistance greater than 4 R_L. Calculation of the maximum grid resistance for use with power valves is covered in Chapter 13 Sect. 10(i).

(ii) Coupling condenser

The coupling condenser (C) may be selected to give the desired low frequency response. The loss of voltage due to C may be calculated by the use of a vector diagram or by the following equation, which applies to the circuit of Fig. 12.1 provided that the input resistance of V_2 is very high.

$$E_g/E = R/|Z| \tag{1a}$$

where E_g = signal voltage on grid of V_2

E = signal voltage across R_L

$R = R_{g2} + r_p R_L/(r_p + R_L)$

or as an approximation, $R \approx R_{g2}$ if the plate resistance of V_1 is small compared with R_{g2},

$|Z|$ = magnitude of series impedance of R and C

$= \sqrt{R^2 + X_c^2}$

and $X_c = 1/\omega C = 1/2\pi f C$.

For example, if R = 1 megohm and f = 50 c/s, the following results will be obtained—

db loss	E_g/E	X_c/R	X_c	C
1	0.891	0.51	0.51 megohm	0.006 24 μF
2	0.794	0.76	0.76 megohm	0.004 19 μF
3	0.708	1.00	1.0 megohm	0.003 18 μF

The phase angle shift is given by

$$\phi = \tan^{-1} 1/\omega CR = \tan^{-1} 1/2\pi f CR \tag{1b}$$

In certain cases a low value of C is adopted intentionally to reduce the response to hum arising from preceding stages. However, the use of a low coupling capacitance, when the following grid resistor is 0.5 megohm or more, increases the hum contributed by the following valve through the a.c. operation of its heater. A low coupling capacitance should therefore not be used on a low-level stage. In high fidelity amplifiers a fairly large value of C is generally adopted, thus not only improving the low frequency response but also reducing phase shift and possibly also improving the response to transients. However, excessively large values of C are undesirable.

The following table gives the approximate values of C for certain selected conditions. Note that R must be as defined below eqn. (1a).

COUPLING CONDENSER

Attenuation 1 db at	12.5	25	50	100	200 c/s
2 db at	8.5	17	34	67	134 c/s
3 db at	6.5	13	26	51	102 c/s
R = 10 000 ohms	2.5	1.25	0.62	0.31	0.15 μF
50 000 ohms	0.5	0.25	0.12	0.06	0.03 μF
100 000 ohms	0.25	0.12	0.06	0.03	0.015 μF
0.25 megohm	0.1	0.05	0.025	0.012	0.006 μF
0.5 megohm	0.05	0.025	0.012	0.006	0.003 μF
1.0 megohm	0.025	0.012	0.006	0.003	0.001 5 μF
2 megohms	0.012	0.006	0.003	0.001 5	0.000 8 μF
5 megohms	0.005	0.002 5	0.001 2	0.000 6	0.000 3 μF

The effect of selected values of capacitance for $R = 1$ megohm is shown graphically in Fig. 4.36. These curves have an ultimate slope of 6 db/octave.

A general curve of attenuation and phase shift is given in sub-section (xiii) below and Fig. 12.9. A Nomogram of transmission factor and phase shift is given by Ref. A2.

(iii) Cathode bias

(A) Cathode bias (Fig. 12.1) is generally preferable to fixed bias as it is largely self-compensating. The plate current flowing through R_k produces a voltage drop which is smoothed out by C_k and applied through R_{g1} to the grid. The full voltage on the grid (in the absence of grid current) is

$$E_c = -I_b R_k \tag{1c}$$

where I_b = plate current in amperes
and R_k = resistance of cathode resistor in ohms.

If negative* grid current (e.g. gas or grid emission current) **is flowing,** the bias will be decreased by $I_g R_{g1}$
where I_g = direct grid current in microamperes (taken as positive)
and R_{g1} = resistance of grid resistor in megohms.

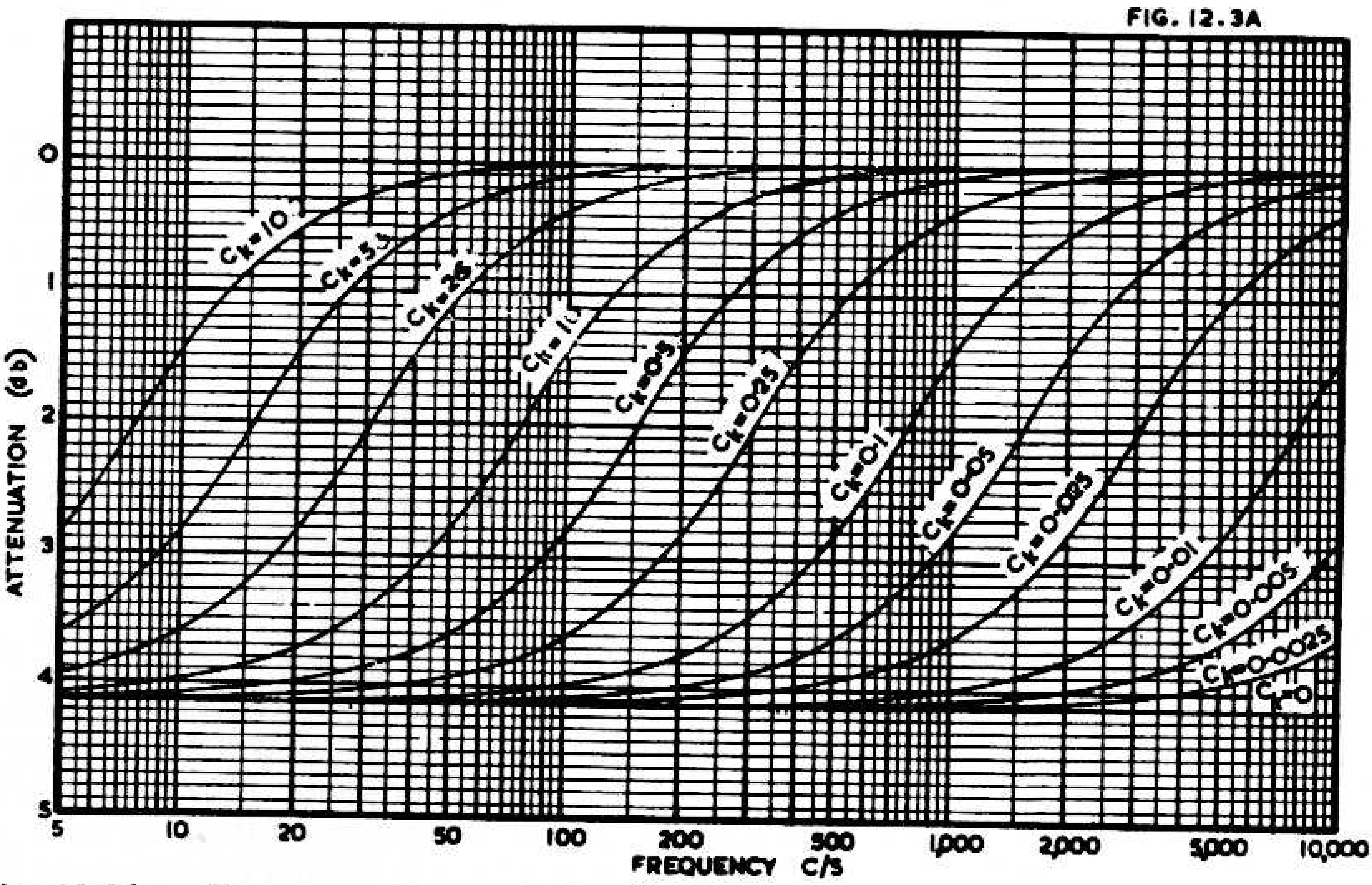


Fig. 12.3A. Frequency characteristics due to cathode by-pass condenser with general purpose triode having $\mu = 20$, $r_p = 10{,}000$ *ohms,* $R_L = 0.1$ *megohm,* $R_{g2} = 0.5$ *megohm,* $R_k = 2700$ *ohms.*

If positive grid current is flowing the bias will be increased by $I_g R_{g1}$.

The cathode by-pass condenser C_k is only fully effective at high frequencies, and it becomes increasingly ineffective as the frequency is decreased—under these conditions there is degeneration (negative current feedback—see Chapter 7). **The effect of R_k and C_k on the voltage gain** at any frequency is given by the equation

$$\left|\frac{A'}{A}\right| = \sqrt{\frac{1 + (\omega C_k R_k)^2}{\left[1 + \frac{(\mu + 1)R_k}{R' + r_p}\right]^2 + (\omega C_k R_k)^2}} \tag{2}$$

where A' = stage voltage gain at frequency f with self bias resistor R_k by-passed by condenser C_k
A = stage voltage gain with R_k completely by-passed
= mid-frequency voltage gain
$\omega = 2\pi f = 2\pi \times$ frequency of input signal

*Also known as reverse grid current.

C_k = capacitance of by-pass condenser in farads
R_k = resistance of self bias resistor in ohms
μ = valve amplification factor at the operating point
r_p = valve plate resistance in ohms at the operating point
R_L = resistance of plate load resistor in ohms
R_{g2} = resistance of following grid resistor in ohms
and $R' = R_L R_{g2}/(R_L + R_{g2})$.

The derivation of eqn. (2) is given in Ref. B11; see also Refs. A11, A13.

The **attenuation characteristics** of a typical general-purpose triode **with cathode bias** are given in graphical form in Fig. 12.3A. It will be seen that all curves have the same shape, but are shifted bodily sideways depending on the value of C_k. The maximum slope of the curves in Fig. 12.3A is 1.4 db/octave; the slope does not normally exceed 3 db/octave when R_k is the optimum value to provide bias.

The limiting loss of gain at zero frequency due to R_k is given by

$$\left|\frac{A'}{A}\right| = \frac{R' + r_p}{R' + r_p + (\mu + 1)R_k} \qquad (3)$$

Examples of limiting loss of gain at zero frequency (based on equation 3)

Type	R_L	R_{g2}	R'	R_k	μ*	r_p*	A'/A	loss
6J5	0.1MΩ	0.5MΩ	0.08MΩ	2700	18	17 000	0.65	3.7 db
6Q7	0.25MΩ	0.5MΩ	0.17MΩ	3000	68	70 000	0.53	5.5 db
6SQ7	0.25MΩ	0.5MΩ	0.17MΩ	3900	100	100 000	0.40	8.0 db

*At operating point.

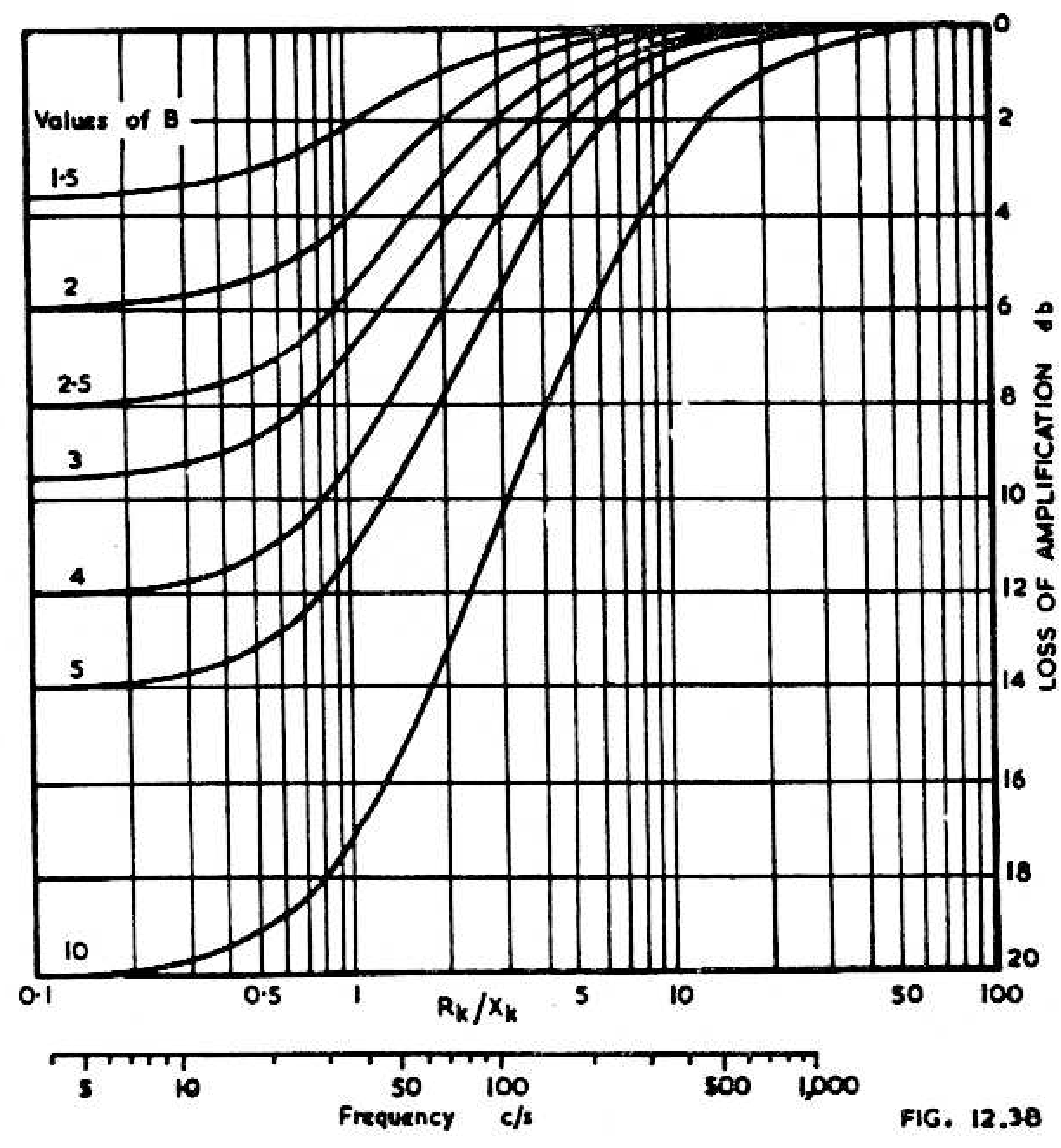


Fig. 12.3B. *Universal attenuation curves with cathode bias (Ref. A11 Part 4).*

Universal attenuation curves are given in Fig. 12.3B (Ref. A11 Part 4) in which

$$B = 1 + \frac{R_k(\mu + 1)}{r_p + R'}$$

$R' = R_L R_{g2}/(R_L + R_{g2})$
and $X_k = 1/(\omega C_k)$.

(B) The phase angle shift is a maximum at the frequency where the slope of the attenuation characteristic is a maximum, and drops towards zero at zero frequency and at higher audio frequencies. The maximum value for any normal r.c.c. amplifier does not exceed about 30°.

Universal curves showing phase angle shift are given in Fig. 12.3C in which the symbols have the same meaning as in Fig. 12.3B.

(C) Choice of cathode bias resistor

The best method is to determine the grid voltage as for fixed bias—see (iv) below—and then to calculate R_k. A fairly satisfactory approximation is to select R_k by the following table.

μ	R_L MΩ	Low level	Inter-mediate level	High level $R_g = R_L$	High level $R_g = 2R_L$
12 to 25	any	$R_k = 0.25R_L/\mu$	$0.5R_L/\mu$	$0.6R_L/\mu$	$0.8R_L/\mu$
30 to 50	any	$R_k = 0.5R_L/\mu$	$0.7R_L/\mu$	$0.7R_L/\mu$	$0.9R_L/\mu$
70 to 100	0.1	$R_k = 1.0R_L/\mu$	$1.2R_L/\mu$	$1.2R_L/\mu$	$1.6R_L/\mu$
	0.22	$R_k = 0.8R_L/\mu$	$1.0R_L/\mu$	$0.9R_L/\mu$	$1.2R_L/\mu$
	0.47	$R_k = 0.65R_L/\mu$	$0.8R_L/\mu$	$0.75R_L/\mu$	$1.0R_L/\mu$

Conditions : Plate supply voltage 200 to 300 volts. Values of μ are the published values.

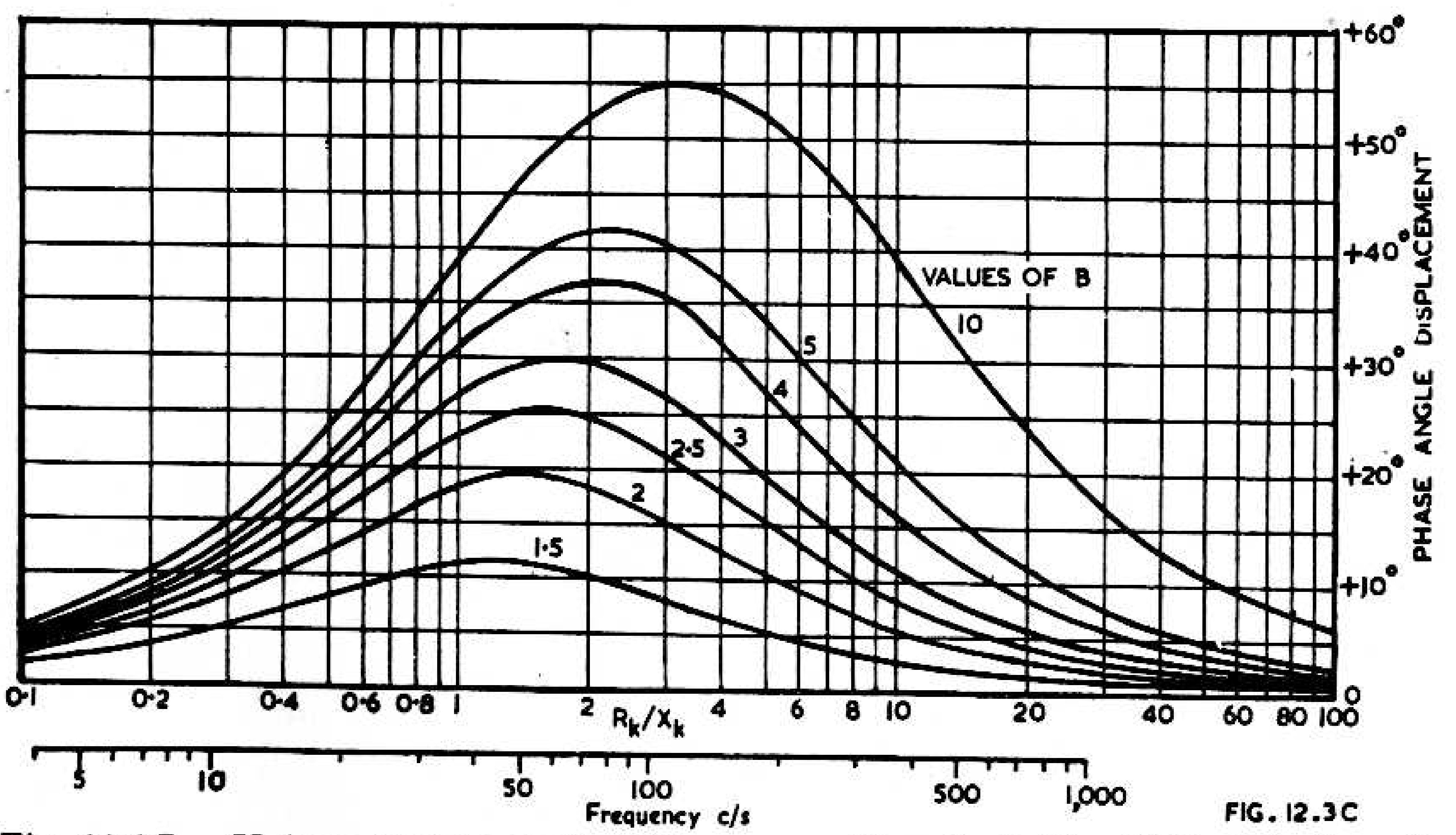


Fig. 12.3C. *Universal phase angle shift curves with cathode bias (Ref. A11 Part 4).*

(D) Cathode bias loadlines

Cathode bias loadlines may be drawn on the plate characteristics as on the mutual characteristics [Chapter 2 Sect. 4(v) and Fig. 2.27] but they will no longer be straight lines. Fig. 12.4 shows the loadline and plate characteristics for a high-mu triode. The cathode bias loadline may very easily be plotted for any selected value of R_k, e.g. 10 000 ohms. In Fig. 12.4 each curve applies to increments of 0.5 volt in E_c, so that (with R_k = 10 000 ohms) the successive increments of I_b are 0.05 mA—

$E_c = 0$	-0.5	-1	-1.5	-2	-2.5	-3 volts
$I_b = 0$	0.05	0.1	0.15	0.2	0.25	0.3 mA

The intersection of the selected R_k loadline with the R_L loadline gives the operating point (Fig. 12.4).

Alternatively, the intersection of the R_L loadline with each grid curve may be marked with the corresponding value of R_k; e.g. in Fig. 12.4 at point A, $E_c = -1.5$ V and I_b = 0.47 mA, therefore $R_k = (1.5/0.47) \times 1000 = 3200$ ohms. Similarly at point B, R_k = 7800 ohms. If greater accuracy is required in determining an individual point, the values of R_k may be plotted against E_c, and the desired value may be selected.

For greater accuracy, particularly with relatively high values of cathode resistance, the slope of the loadline in Fig. 12.4 should be that corresponding to $(R_L + R_k)$.

(E) Maximum grid resistance with cathode bias

It may be shown that the maximum permissible grid resistance with cathode bias is greater than that with fixed bias, as indicated by the approximation :

$$\frac{R_g \text{ with cathode bias}}{R_g \text{ with fixed bias}} \approx 1 + g_{md}R_k \tag{3a}$$

where g_{md} = slope of dynamic characteristic at operating point

$$\approx \frac{g_m r_p}{(r_p + R_L)} \approx \frac{\mu}{(r_p + R_L)}$$

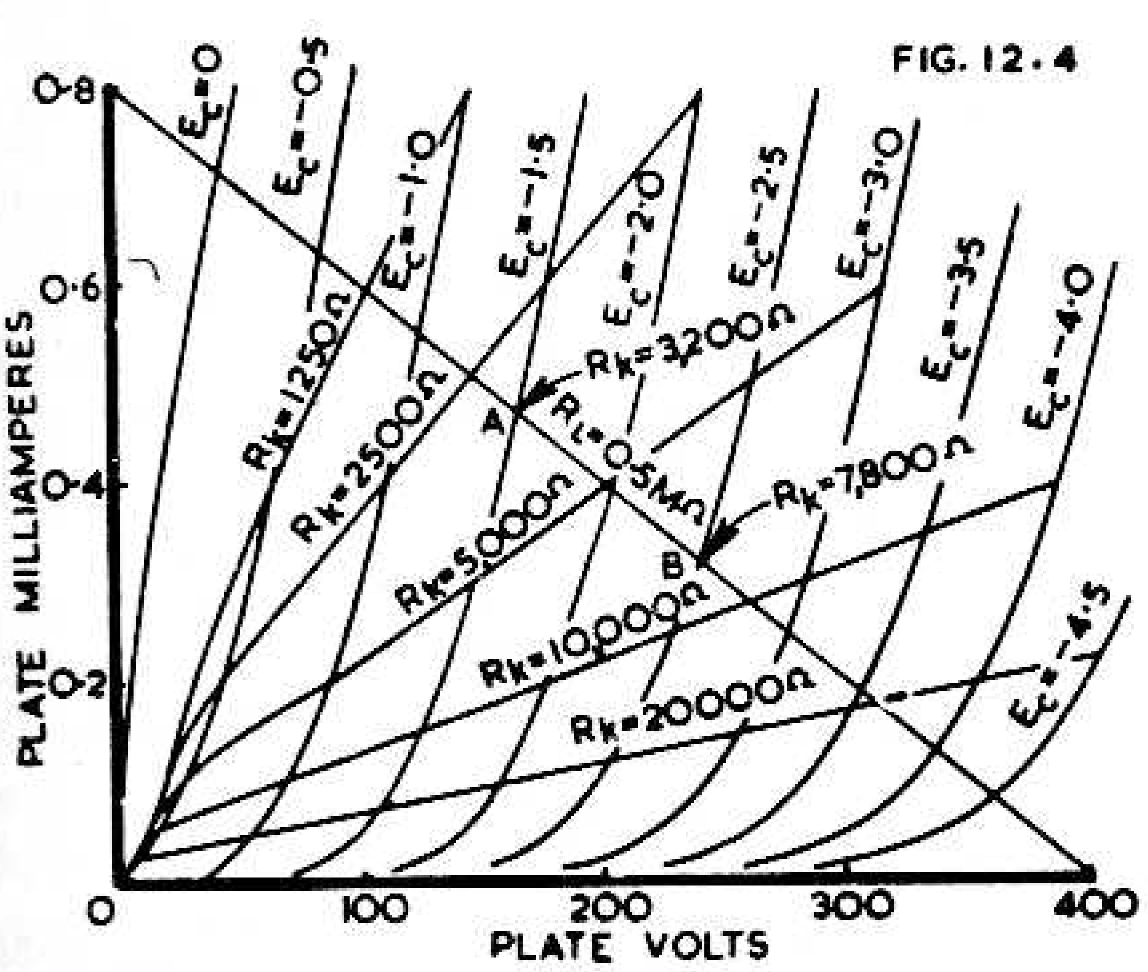
g_m = mutual conductance at operating point

μ = amplification factor at operating point

r_p = plate resistance at operating point

and R_k = cathode bias resistor.

This ratio (eqn. 3a) is usually between 1.2 and 2.



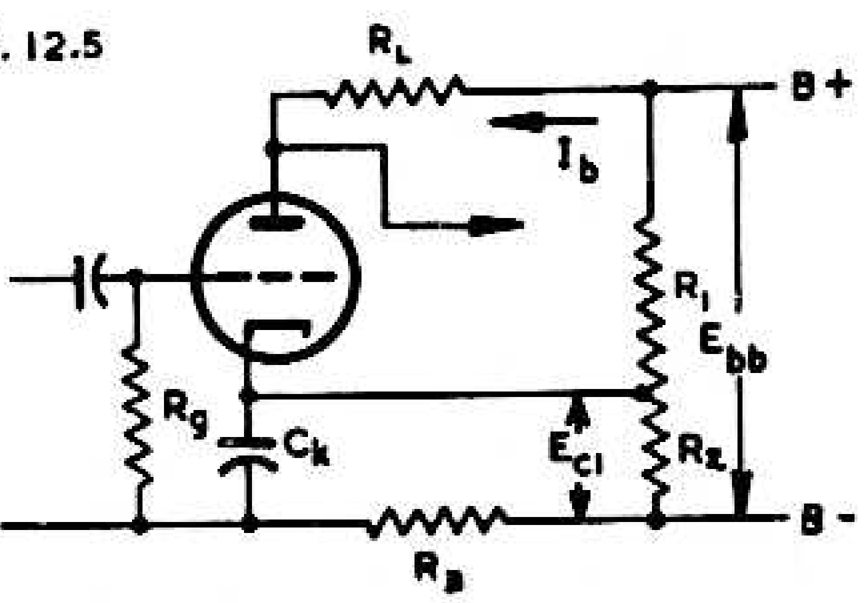


Fig. 12.4. *Plate characteristics of high-mu triode (6SQ7) with* E_{bb} = 400 *volts and* R_L = 0.5 *megohm. Cathode bias loadlines for selected values of cathode bias resistance* (R_k) *have been drawn in.*

Fig. 12.5. *Resistance-capacitance coupled triode with fixed bias from voltage divider.*

(iv) Fixed bias

(A) It is sometimes desired to operate the valve with fixed bias, either from a separate bias supply (battery or rectifier/filter combination) or from a voltage divider across the plate supply. Fig. 12.5 is an example of the latter, with R_1 and R_2 forming a voltage divider. Here

$$E_{c1} = \frac{R_2E_{bb} + R_1R_2I_b}{R_1 + R_2} \text{ and } R_2 = \frac{E_{c1}}{I_b + (E_{bb} - E_{c1})/R_1} \tag{4}$$

Condenser C_k is for by-passing R_2, but as R_2 may have a fairly low value, thus necessitating a large capacitance for C_k, it is frequently economical to add R_3 and so permit a smaller value of C_k. In addition, C_kR_3 forms a useful hum filter for the bias voltage.

(B) The optimum grid bias $(-E_c)$ is a function of the input voltage ; the bias should be the minimum (k_1) which can be used without running into damping due to positive grid current, with a margin for differences between valves, together with a further margin (k_2) to allow for the effects of reverse grid current in the grid resistor.

If the input voltage is known,

$$E_c = 1.41\, E_i + k_1 + k_2 \text{ volts} \tag{5a}$$

If the output voltage is known,

$$E_c = (1.41\, E_0/A) + k_1 + k_2 \text{ volts} \tag{5b}$$

where E_i = r.m.s. input voltage
E_0 = r.m.s. output voltage, from plate to cathode
A = voltage gain of stage
k_1 = 0.75 to 1.0 for high-mu indirectly heated valves*
= 0.6 (approx.) for general purpose indirectly heated valves*
= 0.25 (approx.) for battery valves*

$$k_2 \approx \frac{7.5\ R_{g1}}{100\ R_L} \text{ for high-mu valves}$$

$$\approx \frac{3\ R_{g1}}{100\ R_L} \text{ for general purpose valves.}$$

R_{g1} = resistance of grid resistor (upper limit)
and R_L = resistance of plate load resistor (lower limit).

These approximate values of k_2 are based on typical American manufacturing specifications for maximum grid current (limiting values), and on the assumption that the reverse grid current is proportional† to the plate current.

Values of k_2 are tabulated below for three typical conditions for a plate supply voltage of 250 volts.

Case 1 : R_{g1} = 1 megohm

	General purpose triode	High-mu triode
R_L = 0.1 megohm	$k_2 = 0.31$	$k_2 = 0.75$ volt
R_L = 0.22 megohm	$k_2 = 0.14$	$k_2 = 0.34$ volt
R_L = 0.47 megohm	$k_2 = 0.07$	$k_2 = 0.16$ volt

The additional bias (k_2) provided to allow for the effect of reverse grid current is of no great consequence with general purpose triodes, but it is very serious with high-mu triodes

Case 2 : $R_{g1} = 2R_L$,
k_2 = 0.15 volt for high-mu triodes
k_2 = 0.06 volt for general purpose triodes.

Case 3 : $R_{g1} = 4R_L$
k_2 = 0.3 volt for high-mu triodes
k_2 = 0.12 volt for general purpose triodes.

Thus with **general purpose triodes operated as fixed bias r.c.c. amplifiers,** a grid resistor having a resistance of 1 megohm or even higher may be used without serious effects. However, it is always desirable to keep the resistance as low as practicable—not more than four times the resistance of the preceding plate load resistor, and not exceeding 1 megohm.

If the effect of reverse grid current on the operating point is to be kept small in **fixed-bias high-mu triodes,** it is essential to use a grid resistor R_{g1} (that is, its own grid resistor) having a resistance not more than twice that of its plate load resistor. This will restrict the maximum change of bias to 0.15 volt (i.e. k_2 = 0.15 volt). For example a load resistance of 0.47 megohm will permit the use of a grid resistor of 1 megohm, while a load resistance of 0.22 megohm will permit the use of a grid resistor of 0.47 megohm.

(C) Cathode bias is preferable to fixed bias in that the shift of operating point due to reverse grid current is minimized. With general purpose triodes employing cathode bias there is no advantage in making any adjustment to the bias voltage as calculated for fixed bias (eqns. 5a, 5b) but with high-mu triodes employing cathode bias the value of k_2 may be taken as approximately

$$k_2 \approx \frac{5\ R_{g1}}{100\ R_L} \text{ for high-mu valves with cathode bias.}$$

Consequently, with high-mu triodes, the resistance of the grid resistor R_{g1} may be made 1.5 times that with fixed bias for the same effect on the operating-point.

*These values of k_1 are only typical, and are likely to be exceeded by some valves. In each case, a value should be determined for the valves being used.

†This is approximately true when the reverse grid current is nearly all due to ionization—see Ref. G1, Fig. 1.

(D) Damping due to positive grid current

It is shown in Chapter 2 Sect. 2(iii) that damping on the positive peaks of grid input voltage may be quite serious, even when the peak grid voltage does not reach the grid current " cross-over point." This damping on positive peaks introduces an objecttionable form of distortion which is particularly important when the preceding stage has high effective impedance (looking backwards from the grid of the stage being considered) as, for example, with a high-mu triode or pentode. The damping at the positive peak of the input voltage is proportional to the grid conductance at this point, which value increases rapidly as soon as electrons commence to flow from cathode to grid, and even while the resultant grid current is still negative.

If grid current damping is to be avoided, the grid bias must be increased sufficiently to avoid this region.

(v) Grid leak bias

If a grid resistor of 5 or 10 megohms is used, it is possible to obtain the grid bias for a high-mu triode by means of the voltage drop in R_g [see Chapter 2 Sect 2(iii)]. This imposes damping on the input circuit due to the grid current, and the average input resistance is approximately $R_g/2$. The distortion is approximately the same as with the optimum fixed bias, but the great advantage is that it accommodates itself to variations from valve to valve, while fixed bias is critical.

It may also be used with general purpose triodes operating with input voltages not exceeding, say, 1 volt peak.

Grid leak bias is not very suitable for use with low-level (pre-amplifier) stages owing to hum.

(vi) Plate voltage and current

As discussed in Chapter 2 Sect. 3(i), the voltage on the plate is less than the supply voltage by the voltage drop in the load resistor. This is illustrated in Fig. 12.6 in which the loadline is fixed by the plate supply voltage E_{bb} and the load resistance R_L. The quiescent operating point Q is fixed by the intersection of the loadline and the grid curve for E_{c1}, the bias voltage having been previously determined. Alternatively, the cathode bias loadline may be drawn by the method described above, thus determining the operating point.

FIG. 12.6

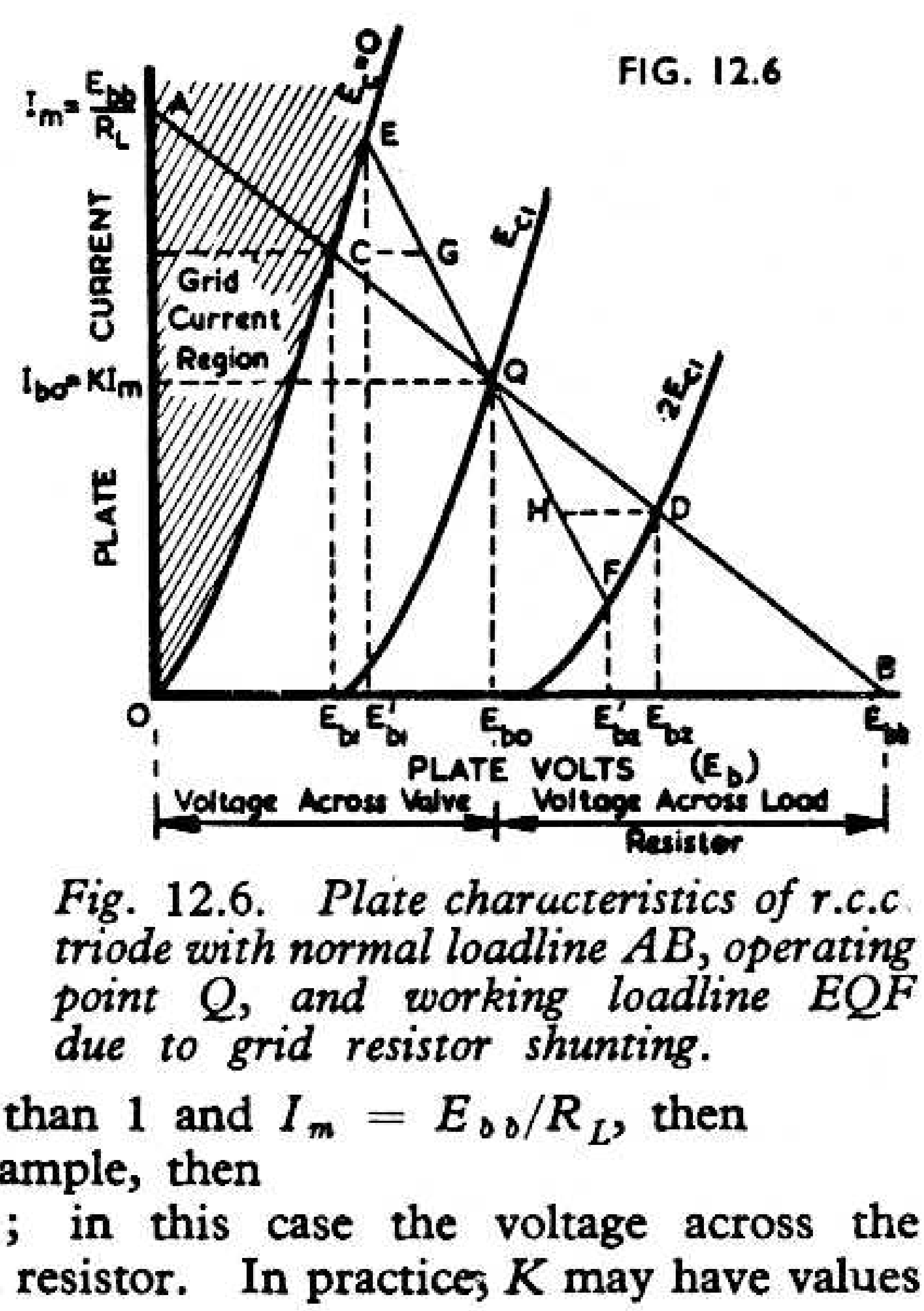


Fig. 12.6. Plate characteristics of r.c.c triode with normal loadline AB, operating point Q, and working loadline EQF due to grid resistor shunting.

Having fixed Q, the quiescent plate current I_{b0} and plate voltage E_{b0} will automatically be fixed. If

$I_{b0} = KI_m$, where K is a constant less than 1 and $I_m = E_{bb}/R_L$, then
$E_{b0} = (1 - K)E_{bb}$. If $K = 0.5$, for example, then
$I_{b0} = \frac{1}{2}I_m = \frac{1}{2}E_{bb}/R_L$ and $E_{b0} = \frac{1}{2}E_{bb}$; in this case the voltage across the valve is equal to the voltage across the load resistor. In practice, K may have values between about 0.25 and 0.85.

Typical values of K are tabulated below

	Low level	Intermediate level	High level*
General purpose triodes	0.75 – 0.85	0.6 – 0.75	0.5 – 0.6
High-mu triodes	0.6 – 0.65	0.5 – 0.6	0.45 – 0.55(R_L = 0.5MΩ)
			0.4 – 0.5 (R_L = 0.25MΩ)

*If R_{g2} is less than $2R_L$, then the values of K for high level operation should be decreased.

Relation between K and cathode bias resistor

Knowing K, it is possible to calculate R_k or vice-versa :

$$R_k \approx \frac{R_L - K(R_L + r_p)}{\mu K} \tag{6a}$$

$$K \approx \frac{R_L}{R_L + r_p + \mu R_k} \tag{6b}$$

Relation between K and grid voltage

$$E_c \approx [R_L - K(R_L + r_p)][(E_{bb}/\mu R_L). \tag{6c}$$

Eqns. (6a), (6b) and (6c) are exact for linear characteristics, but only approximate in practice.

Example : $\mu = 20$, $r_p = 10\,000$ ohms, $R_L = 0.1$ megohm, $E_{bb} = 250$ volts

$K =$	0	0.1	0.2	0.3	0.4	0.5	0.6	0.7	0.8	0.9	1.0
$E_c =$	−12.5	−11.1	−9.7	−8.4	−7.0	−5.6	−4.3	−2.9	−1.5	−0.1	+1.25

(vii) Gain and distortion at the mid-frequency

The voltage gain at the mid-frequency is

$$A_0 = \frac{\mu R_L}{R_L + r_p} = \frac{g_m}{(1/r_p) + (1/R_L)} \tag{6d}$$

where there is no a.c. shunt load, and where μ, g_m and r_p are the values at the operating point, that is at the operating plate current. Alternatively this may be calculated from the loadline (Fig. 12.6) :

$$A_0 = \frac{E_{b2} - E_{b1}}{2E_{c1}} \tag{6e}$$

where E_{b2} is plate voltage corresponding to negative peak grid signal voltage
E_{b1} is plate voltage corresponding to positive peak grid signal voltage
and E_{c1} is peak grid signal voltage.

When there is a following grid resistor R_{g2}, the loadline must be rotated about Q to a new position EQF with a slope of $-(1/R_L + 1/R_{g2})$. The voltage gain for the same input voltage is then

$$A' = \frac{E_{b2}' - E_{b1}'}{2E_{c1}} \tag{6f}$$

which may be shown by the equivalent circuit [see (xii) below] to be

$$A' = \frac{g_m}{1/r_p + 1/R_L + 1/R_{g2}} \tag{7}$$

also $A'/A = R_{g2}/(R + R_{g2})$ (8)
where $R = R_L r_p/(R_L + r_p)$
and g_m and r_p are the values at the operating point.

In practice, the new loadline EQF extends downwards with point F in the region of increasing distortion, and in the extreme case plate current cut-off may occur. This effect may be avoided by reducing the input voltage so that the output voltage will be

$R_{g2}/(R_L + R_{g2}) \times$ output voltage without shunting (9)

Gain in terms of g_m and g_p

Equation (6d) may be put into the alternative form

$$A_0 = \frac{g_m R_L}{1 + g_p R_L} \tag{9a}$$

where g_p = plate conductance at the operating point $= 1/r_p$.

When there is a following grid resistor R_{g2} the gain is given by

$$A_0 = \frac{g_m R}{1 + g_p R} \tag{9b}$$

where $R = R_L R_{g2}/(R_L + R_{g2})$.

The values of g_m and g_p may be derived from " G " curves (e.g. see Fig. 13.9B). By this means the actual values of g_m and g_p at the operating point may be determined with a good degree of accuracy without any assumptions or manipulations (Ref. A15).

Distortion

The percentage second harmonic distortion—see Chapter 13 Sect. 2(i)—is given by

$$H_2\% = \frac{EQ - QF}{2(EQ + QF)} \times 100 \qquad (10)$$

(viii) Dynamic characteristics

The dynamic characteristic of a resistance-loaded triode is described in Chapter 2 Sect. 3(i) and Fig. 2.19. In shape it closely resembles the dynamic characteristic of a resistance-loaded pentode, and a comparison between the two is made in Sect. 3(viii) below and Fig. 12.15. With the triode dynamic characteristic the greater part or the whole of the " top bend " is in the grid current region and therefore cannot be used. The most nearly straight portion is therefore that of minimum bias.

An " ideal " (linear) dynamic characteristic of a general purpose triode is shown in Fig. 12.7A. It is limited at the lower end by plate-current cut-off at A, while the other end is in the grid current region. The usable part extends from A to G, but in reality the lower part is curved, and is avoided as far as possible. The upper end (GB) is also curved, but as this is in the grid current region it cannot be used in any case. The curves are to scale for $\mu = 20$, $r_p = 10\,000$ ohms, $R_L = 0.1$ megohm and $E_{bb} = 250$ volts.

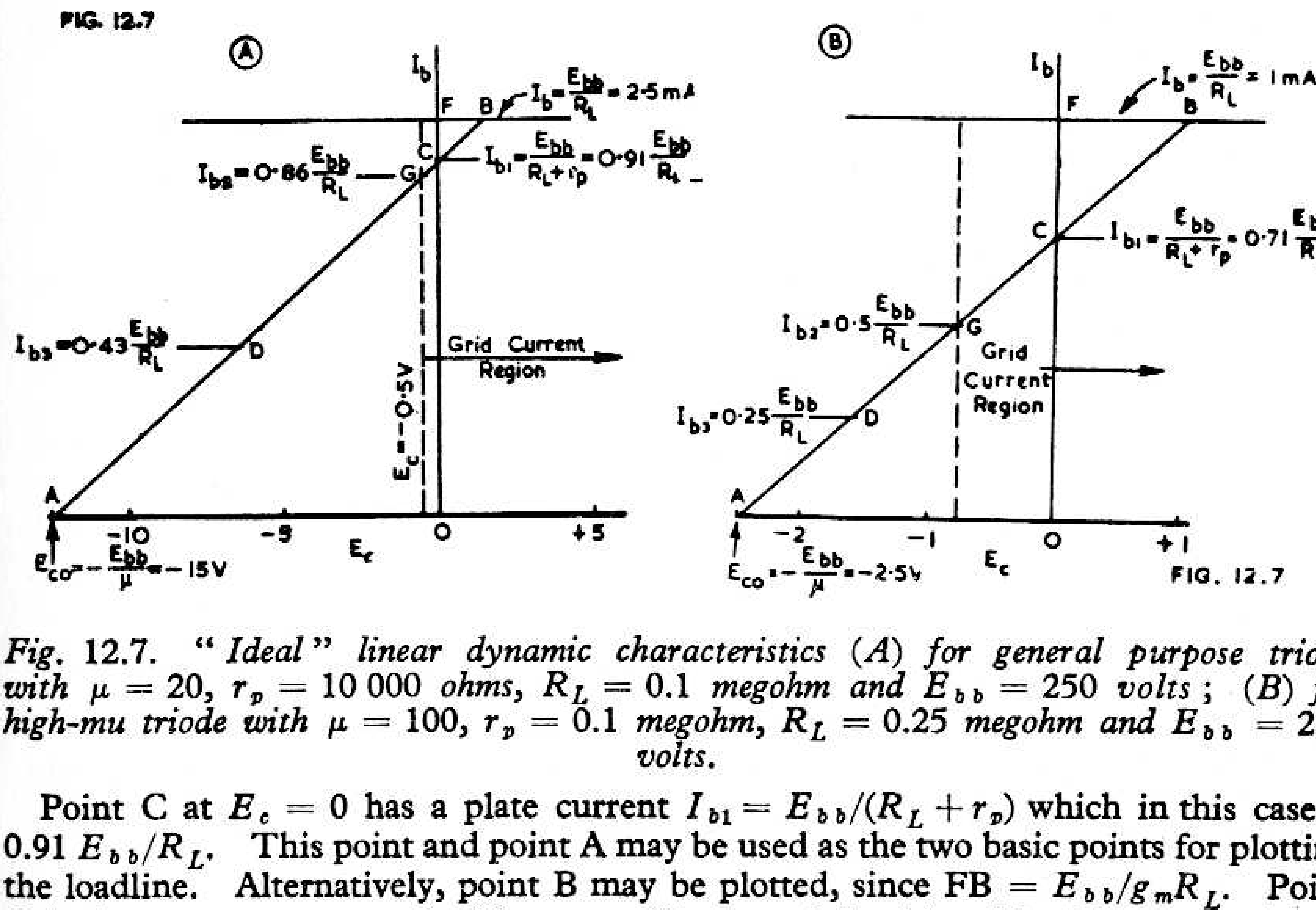


Fig. 12.7. " Ideal " linear dynamic characteristics (A) for general purpose triode with $\mu = 20$, $r_p = 10\,000$ ohms, $R_L = 0.1$ megohm and $E_{bb} = 250$ volts; (B) for high-mu triode with $\mu = 100$, $r_p = 0.1$ megohm, $R_L = 0.25$ megohm and $E_{bb} = 250$ volts.

Point C at $E_c = 0$ has a plate current $I_{b1} = E_{bb}/(R_L + r_p)$ which in this case is 0.91 E_{bb}/R_L. This point and point A may be used as the two basic points for plotting the loadline. Alternatively, point B may be plotted, since FB $= E_{bb}/g_m R_L$. Point G is the commencement of grid current ($E_c = -0.5$ volt) and in this case its $I_{b2} =$ 0.86 E_{bb}/R_L. The highest operating point cannot exceed 0.85 E_{bb}/R_L, and this is only possible for extremely small input voltages. The lowest useful operating point (for high-level operation) is D, which is the mid point of AG, with $I_{b3} = 0.43\, E_{bb}/R_L$. Thus the operating point must be within the limits 0.43 and 0.85 $\times\, E_{bb}/R_L$.

The dynamic characteristic in Fig. 12.7B applies to a high mu triode with $\mu = 100$, $r_p = 0.1$ megohm, $R_L = 0.25$ megohm and $E_{bb} = 250$ volts. The grid current is taken as commencing at $E_c = -0.75$ volt. The usable part of the characteristic extends from A to G, and the operating point must be within the limits 0.25 and 0.5 multiplied by E_{bb}/R_L. The upper limit would be somewhat extended if the load resistance were increased to 0.5 megohm.

(ix) Maximum output voltage and distortion

It is difficult to lay down any limit to the maximum voltage output, since overloading occurs very gradually. It is assumed that in all cases the grid bias is sufficient to avoid positive grid current.

An approximate method for determining the conditions for maximum output voltage are given by Diamond (Ref. A16). In Fig. 12.7C, the tangent FCA at $E_c = 0$ cuts the voltage axis at F where $E_b = e_1$. The dynamic loadline AQD corresponds to R_L in parallel with the following grid resistor R_{g2}. Point B is predetermined by the minimum plate current permissible on account of distortion (I_{min}).

The optimum value of R_L is given by

$$R_L \approx \frac{r_p}{\delta + \sqrt{(\delta + \gamma/x)}}$$

and the maximum value of peak-to-peak voltage swing is then

$$E_0 \approx (E_{bb} - e_1)\frac{1 - \delta}{1 + (1/x) + 2(\delta + \sqrt{\delta + \gamma/x})}$$

where r_p = plate resistance at point C (as given by slope of FC)

$\delta = I_{min} r_p/(E_{bb} - e_1)$

$x = R_{g2}/r_p$

$$\gamma = \frac{\text{“ positive ” swing}}{\text{total swing}} = \frac{\text{GK}}{\text{GH}}$$

E_{bb} = plate supply voltage

E_0 = peak-to-peak total voltage swing = GH

and e_1 = plate voltage corresponding to intersection of tangent FC with voltage axis.

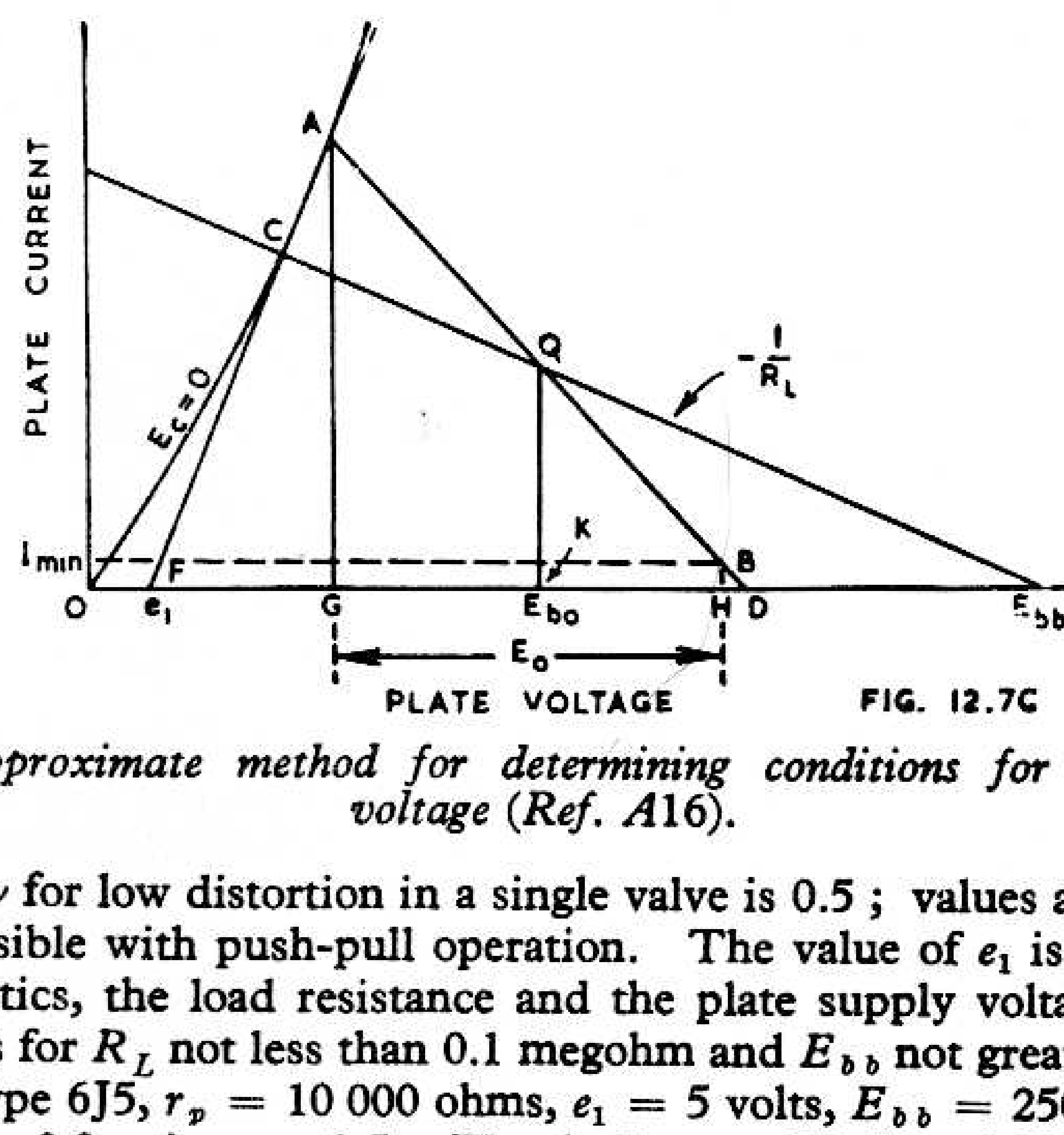


Fig. 12.7C. *Approximate method for determining conditions for maximum output voltage (Ref. A16).*

The value of γ for low distortion in a single valve is 0.5 ; values appreciably higher than 0.5 are possible with push-pull operation. The value of e_1 is a function of the valve characteristics, the load resistance and the plate supply voltage ; it is usually less than 10 volts for R_L not less than 0.1 megohm and E_{bb} not greater than 300 volts.

Example : Type 6J5, r_p = 10 000 ohms, e_1 = 5 volts, E_{bb} = 250 volts, R_{g2} = 0.5 megohm, I_{min} = 0.2 mA, γ = 0.5. We obtain x = 50, δ = 0.008 17, γ/x = 0.01, and from the equations, $R_L \approx$ 70 000 ohms optimum ; E_0 = 186 volts peak-to-peak.

Output voltage and distortion

The maximum output voltage for a general purpose triode (e.g. type 6J5) for 14% intermodulation distortion* and E_{bb} from 180 to 300 volts is given approximately by :

Load resistance	Following grid resistance	Maximum output voltage r.m.s.
0.1 megohm	0.1 megohm	$0.155 \times E_{bb}$
	0.2 megohm	$0.19 \times E_{bb}$
	0.4 megohm	$0.215 \times E_{bb}$
0.25 megohm	0.25 megohm	$0.20 \times E_{bb}$
	0.5 megohm	$0.24 \times E_{bb}$
	1.0 megohm	$0.27 \times E_{bb}$

*For details of intermodulation distortion see Chapter 14 Sect. 3.

For lower values of I.M. distortion, the preceding values of maximum output voltage should be multiplied by:

I.M.	$R_L = 0.1$ megohm	$R_L = 0.25$ megohm
10%	Factor = 0.8	Factor = 0.72
5%	0.48	0.36
2.5%	0.26	0.18

With a high-mu triode the voltage output for 14% intermodulation distortion is only about 0.17 E_{bb} to 0.21 E_{bb} for $R_L = 0.25$ and $R_{g2} = 0.5$ megohm.

Note that with intermodulation distortion the ouput voltage is taken as the arithmetical sum of the component voltages—see Chapter 14 Sect. 3(ii).

The intermodulation distortion of a triode and a pentode are compared in Fig. 12.16A [Sect. 3(ix)]. At low output voltages the pentode gives less distortion while at high output voltages the triode gives less distortion. The triode is, however, less critical than the pentode, and the distortion increases at a lower rate than with the pentode when the bias is made more negative than the optimum value.

(x) Conversion factors with r.c.c. triodes

Some valve manufacturers publish values of μ, g_m and r_p plotted against plate current. These may be used fairly accurately under any conditions of plate or grid voltage.

If these are not available:

1. The **amplification factor** is nearly constant. In most high-mu triodes it drops about 5% to 10% (15% for types 6SL7-GT and 5691) as I_b is reduced to 0.25 mA, while in general purpose triodes it drops about 15% as I_b is reduced to 1 mA.

2. The **mutual conductance** is largely a function of the plate current and is given by the approximate equation

$$g_m \approx F g_{m0} \tag{11a}$$

where g_m = desired mutual conductance at any operating point with plate current I_b

g_{m0} = published mutual conductance at plate current I_{b0}

and F is given by the following table based on the equation

$F = \sqrt[3]{F_i} = \sqrt[3]{I_b/I_{b0}}$.

I_b/I_{b0}	1.0	0.9	0.8	0.7	0.6	0.5	0.4	0.3	0.2	0.1
F	1.0	0.97	0.93	0.89	0.84	0.79	0.74	0.67	0.58	0.46

Alternatively, if the gain with resistance-coupled operation is known, the mutual conductance at the operating point may be calculated from

$$g_m = \frac{\mu \times \text{voltage gain}}{R(\mu - \text{voltage gain})} \tag{11b}$$

where $R = R_L R_g/(R_L + R_g)$,

and μ = amplification factor at the operating point.

3. The **plate resistance** may be calculated from $r_p = \mu/g_m$. It is approximately equal to $r_{p0}\sqrt[3]{I_{b0}/I_b}$.

(xi) Input impedance and Miller Effect

In the amplifier of Fig. 12.1 the grid to plate capacitance C_{gp} has impressed across it a voltage $(A + 1)E_i$ where A is the voltage gain of V_1. The current flowing through C_{gp} is therefore (A + 1) times the current which would flow through the same capacitance when connected from grid to cathode. This is one example of the Miller Effect which occurs in all amplifiers. For example take type 6SQ7 high-mu triode with $C_{gp} = 1.6$ $\mu\mu$F and $A = 62$. The effective input capacitance due to C_{gp} alone is therefore $63 \times 1.6 = 101$ $\mu\mu$F.

The total input capacitance also includes the capacitance from grid to cathode plus $(A + 1)$ times any stray capacitance from grid to plate.

If the plate load is partially reactive the input impedance is equivalent to a capacitance C' and a resistance R' in parallel from grid to cathode,

where

$$C' = C_{gk} + (1 + A\cos\phi)\,C_{gp} \tag{12}$$

$$R' = -1/(2\pi f C_{gp} A \sin\phi) \tag{13}$$

and ϕ = angle by which the voltage across the load impedance leads the equivalent voltage acting in the plate circuit (ϕ will be positive for an inductive load and negative for a capacitive load)

When the load is capacitive, as usually with a r.c.c. amplifier, R' is positive and there is some slight additional loading of the input circuit. When the load is inductive, R' is negative and self oscillation may occur in an extreme case.

(xii) Equivalent circuit of r.c.c. triode

The exact a.c. equivalent circuit of a r.c.c. triode is given in Fig. 12.8 where the valve V_1 is replaced by a generator of voltage μE_i in series with r_p. The plate load resistor R_L is shunted by C_0, which includes the valve output capacitance plus stray capacitance. The grid resistor of V_2 is shunted by C_i which includes the input capacitance of V_2 (including the Miller Effect capacitance from the plate) plus stray capacitance. Any additional condenser connected from plate or grid to earth should be added to C_0 or C_i respectively.

FIG. 12.8

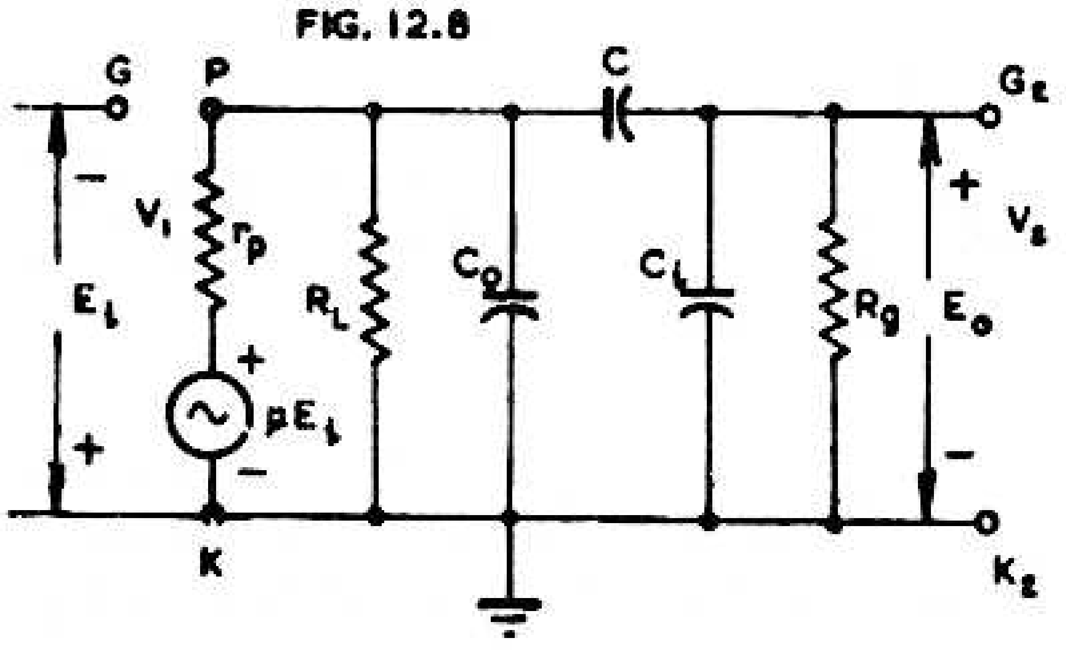


Fig. 12.8. *Exact equivalent circuit of r.c.c. triode.*

The cathode bias resistor is here assumed to be adequately by-passed at all signal frequencies.

This equivalent circuit is the basis of the calculations of gain and phase angle shift at all frequencies.

(xiii) Voltage gain and phase shift

It has been shown by Luck and others that the voltage gain/frequency characteristic of a r.c.c. amplifier has the same mathematical form as a tuned circuit except that the Q is very low (never greater than 0.5). There is a mid-frequency at which the gain is a maximum (A_0) and the phase shift zero. At both lower and higher frequencies the gain falls off, and when the gain is $A_0/\sqrt{2}$ the absolute values of the resistive and reactive components are equal. These two reference frequencies, which correspond approximately to 3 db attenuation, are the basic points on the attenuation characteristics, and, with the mid-frequency gain A_0, are sufficient to determine the whole frequency characteristic. The value of Q is given by

$$Q = f_0/(f_b - f_a) \tag{14}$$

where f_0 = mid-frequency (at which voltage gain is A_0)

f_b = (high) reference frequency at which voltage gain is 0.707 A_0

and f_a = (low) reference frequency at which voltage gain is 0.707 A_0.

The low reference frequency f_a is given by

$$f_a = 1/2\pi CR' \tag{15}$$

where $R' = R_g + r_pR_L/(r_p + R_L)$

and the constants are as in Fig. 12.8.

The high reference frequency (f_b) is given by

$$f_b = 1/2\pi(C_0 + C_i)R'' \tag{16}$$

where $R'' = r_pR_LR_g/(r_pR_L + r_pR_g + R_LR_g)$.

When f_a and f_b have been determined, the attenuation at any other frequency may be found by reference to Fig. 12.9A.

The phase shift at any frequency may be determined from Fig. 12.9B ; the phase shift at f_a and f_b is 45° leading and lagging respectively.

Alternatively, if the low reference frequency is known, the **gain at any frequency** f is given by (Ref. F1):

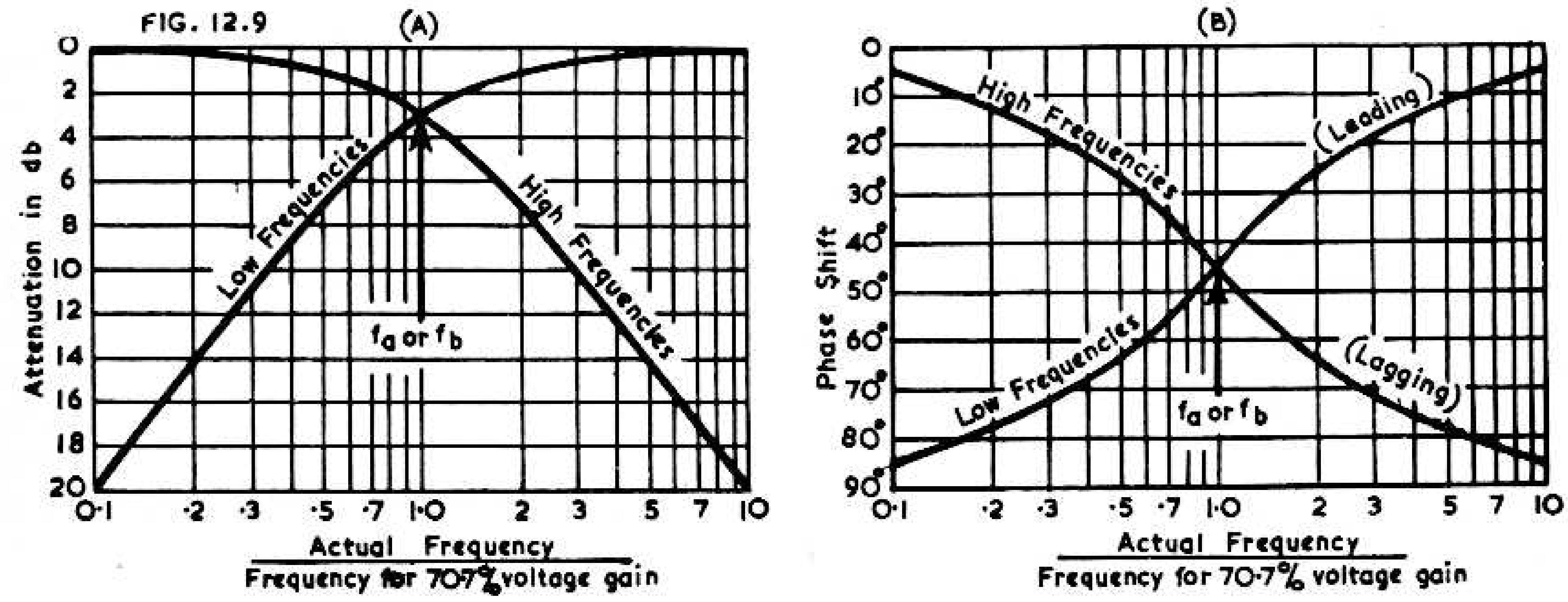


Fig. 12.9(A). *Attenuation in decibels for r.c.c. triode at low and high frequencies (B) Phase shift at low and high frequencies. Values of phase shift for frequencies beyond the limits of the curves may be estimated with reasonable accuracy by taking the angle as proportional or inversely proportional to the frequency.*

$$\frac{\text{gain at frequency } f}{\text{gain at mid-frequency}} = \cos \tan^{-1} \frac{f_a}{f} \tag{17}$$

while the **phase angle shift** is given by

$$\tan \phi = f_a/f \tag{18}$$

where f_a = low reference frequency (3 db attenuation).

Similarly if the high reference frequency is known, the **gain at any frequency** f is given by

$$\frac{\text{gain at frequency } f}{\text{gain at mid-frequency}} = \cos \tan^{-1} \frac{f}{f_b} \tag{19}$$

while the **phase angle shift** is given by

$$\tan \phi = f/f_b \tag{20}$$

where f_b = high reference frequency (3 db attenuation).

(xiv) Comments on tabulated characteristics of resistance-coupled triodes

Published operating characteristics of resistance-coupled triodes (unless the test conditions are specified) appear to be taken under test conditions with a low source impedance, a low d.c. resistance between grid and cathode, and a small value of positive grid current. When this is so, they cannot be applied directly to the normal practical case where there is a high resistance d.c. path between grid and cathode, and where the source impedance may be high. The usual value of total harmonic distortion, unless otherwise specified, appears to be about 5% under the conditions of test (Ref. A17).

In a practical resistance-coupled amplifier the maximum output voltage for 5% total harmonic distortion will be less than the tabulated value, depending on the source impedance and the d.c. resistance from grid to cathode, in the presence of grid-circuit damping. This damping is caused, not only by positive grid current, but also by the grid input conductance which, on the extreme tip of the positive peaks, may be appreciable. See page 20 under Grid variational conductance and page 489 under Damping due to positive grid current.

Thus, for any practical amplifier, the tabulated maximum output voltage is misleading, and a lower value should be used for design purposes. In addition, there will be a slight effect on the gain.

SECTION 3 : RESISTANCE-CAPACITANCE COUPLED PENTODES

(i) Choice of operating conditions (ii) Coupling condenser (iii) Screen by-pass (iv) Cathode bias (v) Fixed bias (vi) Dynamic characteristics of pentodes (vii) Gain at the mid-frequency (viii) Dynamic characteristics of pentodes and comparison with triodes (ix) Maximum voltage output and distortion (x) Conversion factors with r.c.c. pentodes (xi) Equivalent circuit of r.c.c. pentode (xii) Voltage gain and phase shift (xiii) Screen loadlines (xiv) Combined screen and cathode load-lines and the effect of tolerances (xv) Remote cut-off pentodes as r.c.c. amplifiers (xvi) Multigrid valves as r.c.c. amplifiers (xvii) Special applications (xviii) Comments on tabulated characteristics of resistance-coupled pentodes.

(i) Choice of operating conditions

A r.c.c. pentode may be treated as a special case of a r.c.c. triode, and many features are common to both provided that the necessary adjustments are made for the differences in μ, g_m and r_p.

The **load resistance** (R_L) may be any value from a few ohms to many megohms, but for normal operation it is generally from 0.1 to 0.5 megohm. Lower values are used when a reduction in gain or an extended high frequency response is desirable, or when some form of tone correction is intended. Higher values are occasionally used for special applications, but the frequency response is seriously limited unless negative voltage feedback is used. The load resistance should never be greater than the following grid resistor (R_{g2}) and should preferably be not more than one quarter of R_{g2}. The choice of load resistance has only a small effect on the distortion, lower values of R_L (of the order of 0.1 megohm) being somewhat better in this respect.

Plate supply voltage : As for triodes.

Screen supply : A series resistor from the plate supply voltage is generally preferred. Valves in which the screen current has wide tolerances (e.g. some remote cut-off types and most tetrodes) should be supplied from a voltage divider.

Following grid resistor : As for triodes.

Input grid resistor : As for triodes.

(ii) Coupling condenser

The triode formulae and curves in Sect. 2(ii) may be used except that R may be taken as being approximately equal to $R_{g2} + R_L$. See Sect. 3(x) below for more accurate calculations involving r_p.

(iii) Screen by-pass

For normal performance it is necessary for the screen to be adequately by-passed to the cathode. In practice this is usually obtained by by-passing from screen to earth and also from earth to cathode, the value of C_k being normally much larger than C_s (Fig. 12.10).

FIG. 12.10

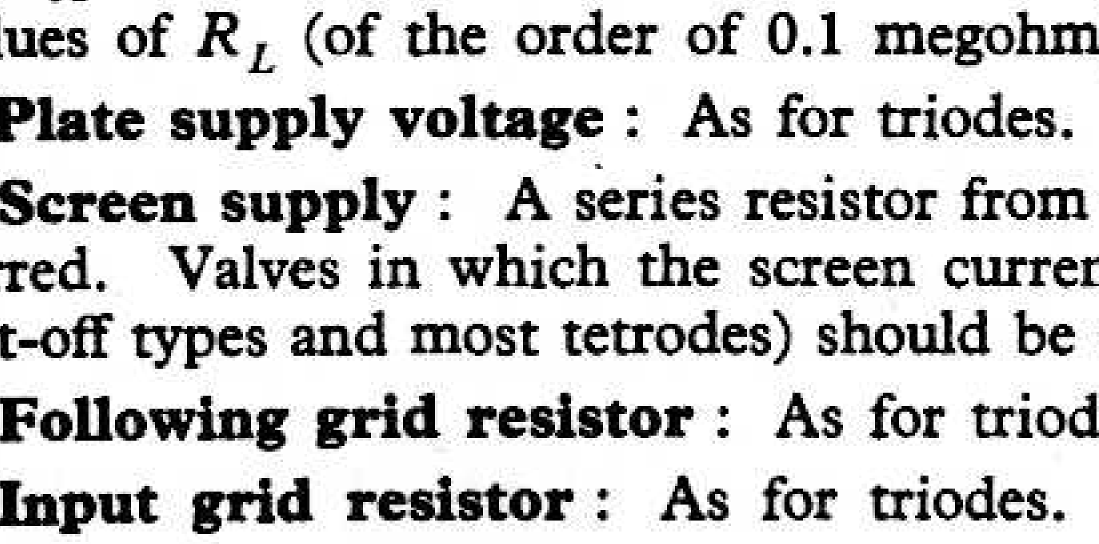


Fig. 12.10. *Circuit of resistance-capacitance coupled pentode with screen dropping resistor and cathode bias.*

The effect of incomplete screen by-passing on gain is given by eqn. (1) on the assumption of complete cathode by-passing (Fig. 12.10), the derivation being given in Ref. B3 (see also Refs. A11, A12, A13, B2).

$$\left|\frac{A'}{A}\right| \approx \sqrt{\frac{1 + R_s^2\omega^2C_s^2}{B^2 + R_s^2\omega^2C_s^2}} \qquad (1)$$

where A' = stage voltage gain at frequency f with screen resistor R_s by-passed by condenser C_s

A = stage voltage gain with R_s completely by-passed
= mid-frequency voltage gain

$\omega = 2\pi f$

$$B \approx 1 + \frac{R_s g_m}{m\mu_t(1 + R_L/r_p)} = 1 + \frac{R_s g_{md}}{m\mu_t} \qquad (1a)$$

$m = I_b/I_{c2}$ (assumed constant)

g_m = mutual conductance at operating plate current [see Sect. 3(vii)]

g_{md} = slope of dynamic characteristic at operating plate current

and μ_t = triode mu (with screen tied to plate).

If the screen voltage is obtained from a voltage divider (R_1, R_2) then R_s should be a ken as $R_1R_2/(R_1 + R_2)$.

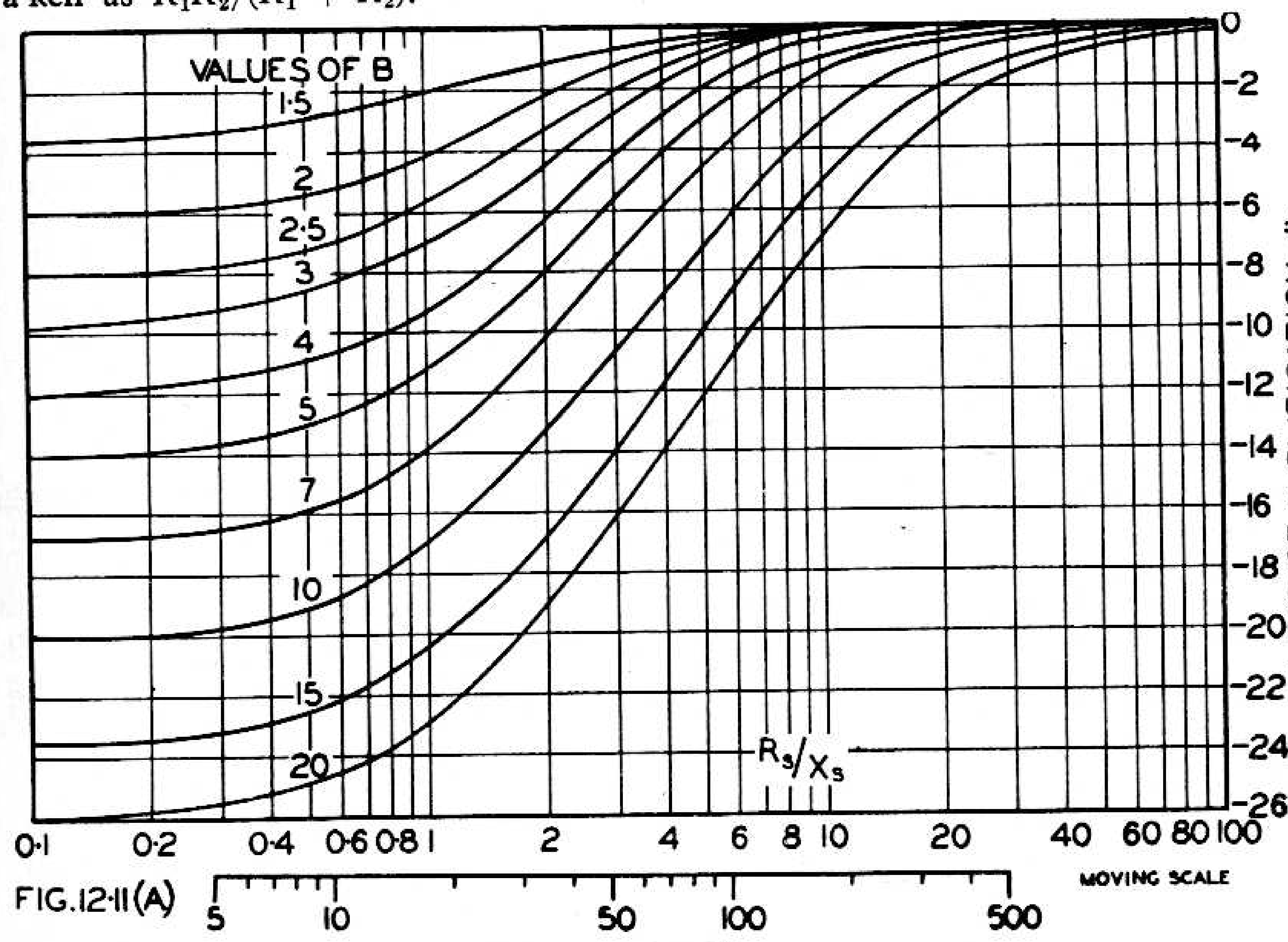


FIG. 12·11(A)

Fig. 12.11A. *Curves for the frequency response of a resistance-coupled pentode with screen dropping resistor R_s and screen by-pass reactance X_s ($= 1/\omega C_s$). The value of B is given by eqn. (1a). The frequency for $R_s/X_s = 1$ is given by $f = 1/2\pi R_sC_s$. The frequency scale should be traced and the tracing moved horizontally until f corresponds to $R_s/X_s = 1$ (method after Sturley).*

Eqn. (1) is plotted in Fig. 12.11A for selected values of B versus R_s/X_s which is equal to $2\pi fR_sC_s$ and therefore proportional to the frequency.

For most purposes it is sufficient to use the **approximation,** for a loss not exceeding 1 db at a frequency f:

$$C_s \approx B/\pi fR_s \qquad (2)$$

where $B \approx 1 + R_sg_m/m\mu_t$
and $m = I_b/I_{c2}$.

For example, type 6J7 with $R_L = 0.25$, $R_s = 1.5$ megohms, $g_m = 850$ µmhos, $\mu_t = 20$ and $m = 4$ has $B \approx 16.5$. For 1 db attenuation at $f = 50$ c/s, $C_s \approx 0.07$ µF.

The **limiting loss of voltage gain** at very low frequencies is $1/B$; in the example above this is 0.06 (i.e. 24.3 db). If the screen had been supplied from a voltage divider, the loss of gain at very low frequencies would have been much less.

An unbypassed screen resistor of $g_m r_s$ ohms gives the same degree of degeneration as a 1 ohm cathode resistor
where r_s = dynamic screen resistance $(\partial e_s/\partial i_s)$
$\approx$ triode plate resistance.
For example, when g_m = 2000 micromhos and r_s = 10 000 ohms, an unbypassed

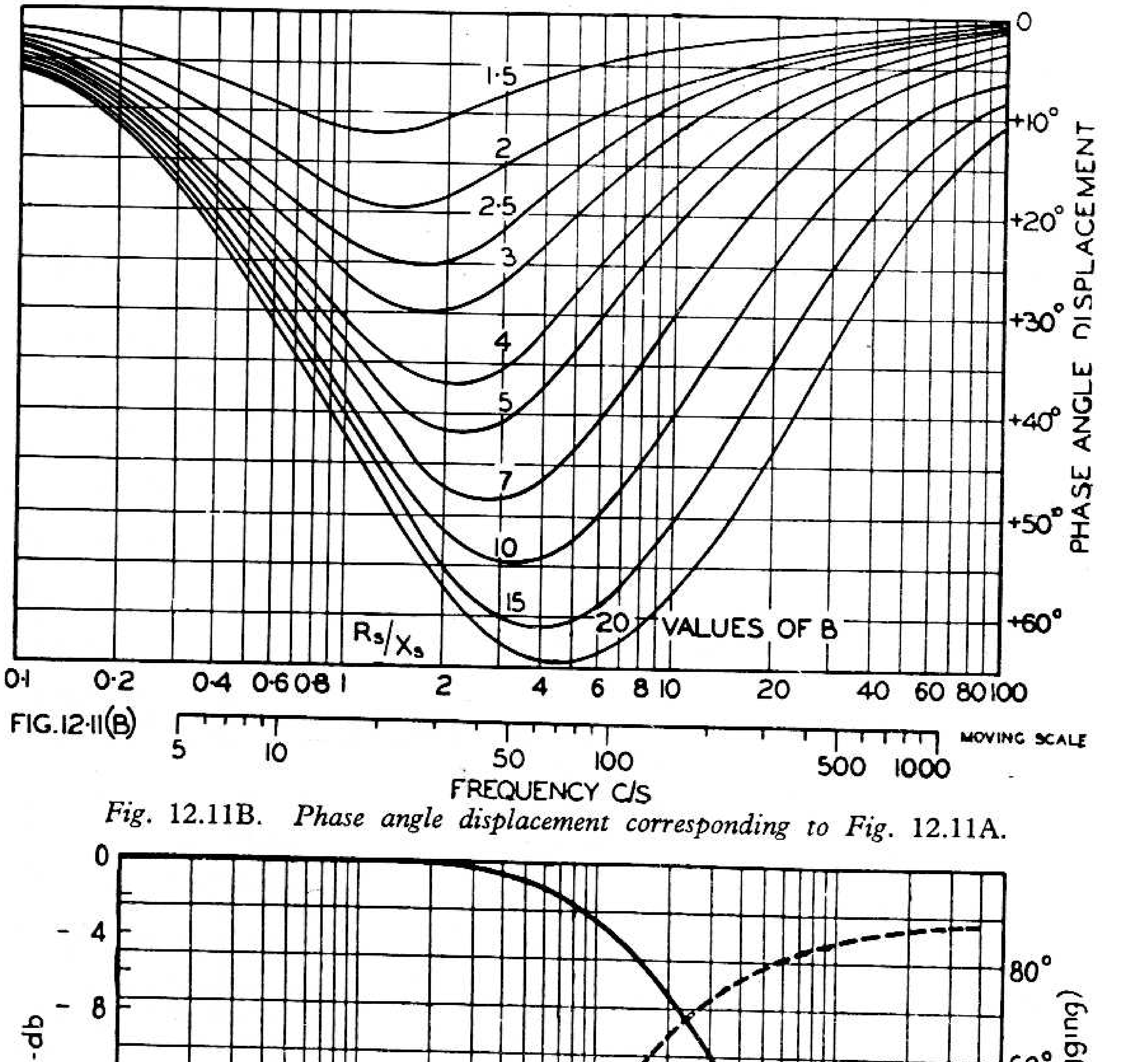


Fig. 12.11B. *Phase angle displacement corresponding to Fig.* 12.11A.

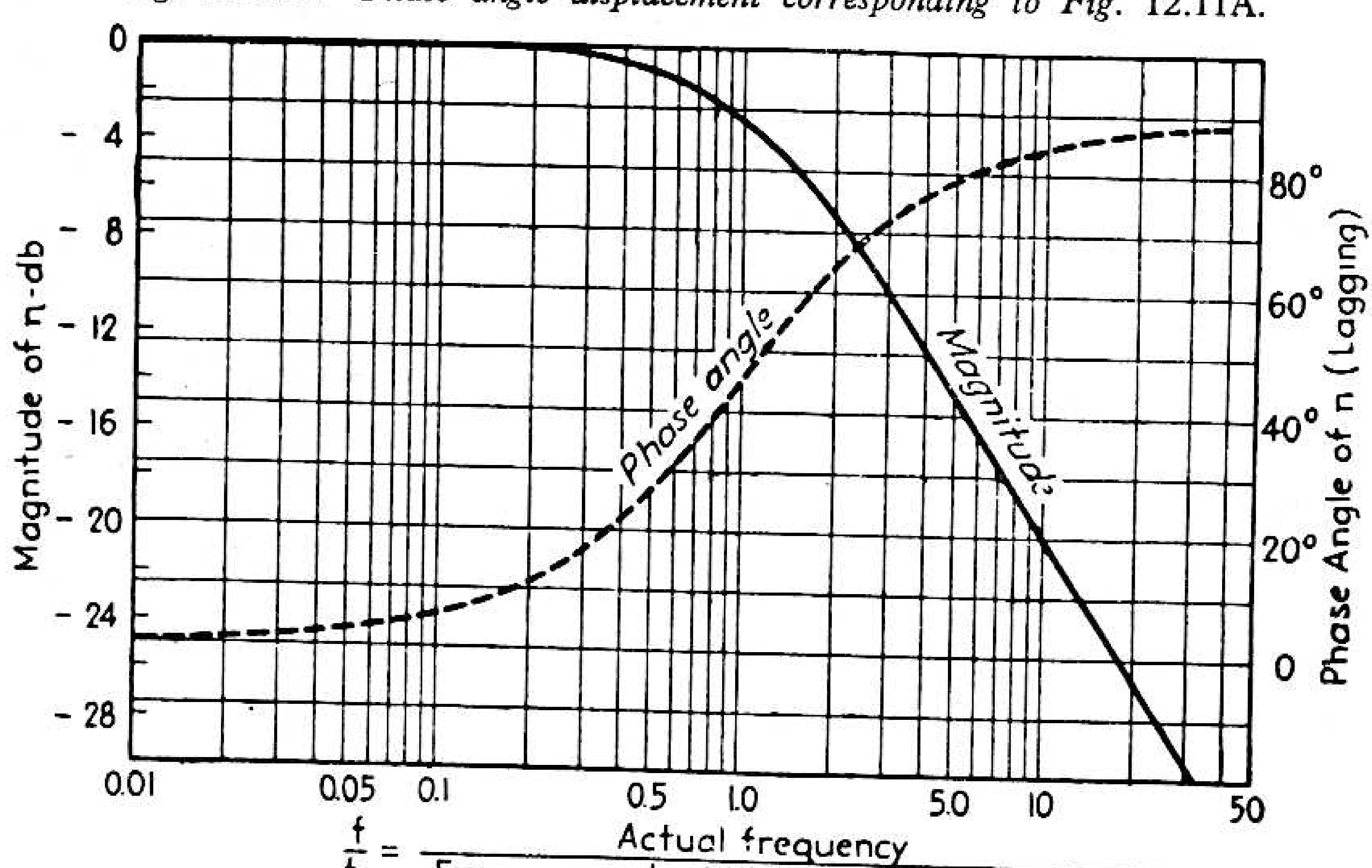


Fig. 12.11C. *Magnitude and phase of factor for use in Fig.* 12.11D (*Ref. A*12). *Reprinted by permission from RADIO ENGINEERS' HANDBOOK by F. E. Terman, copyrighted in* 1943, *McGraw-Hill Book Company Inc.*

screen resistor of 20 ohms will give the same degree of degeneration as a 1 ohm cathode resistor.

The effect of the screen by-pass capacitance on **phase shift** is indicated by Fig. 12.11B. Under normal conditions with a high resistance dropping resistor, the angle does not exceed 65°. This angle may be considerably reduced by reducing the effective value of R_s, as for example with a voltage divider. Such action is normally only necessary with negative feedback.

(iv) Cathode bias

(A) Effect of incomplete by-passing on gain

Provided that the screen is adequately by-passed at all frequencies of operation, the procedure is as for triodes (Sect. 2) except that the d.c. current flowing through

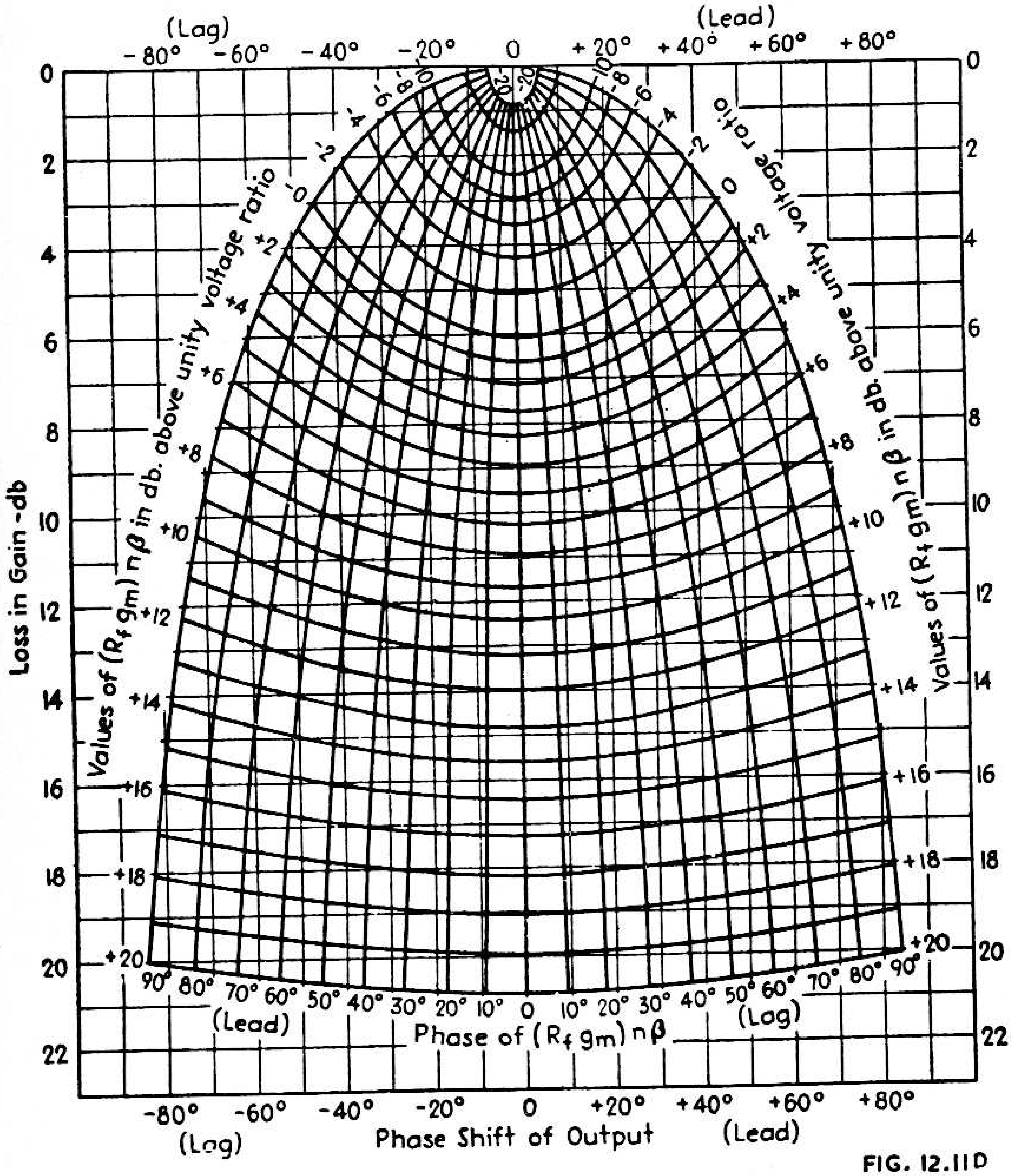


Fig. 12.11D. *Curves from which the loss in gain and phase shift of output at low frequencies resulting from resistance-condenser bias impedance R_fC_f can be calculated in a resistance-coupled amplifier for the general case where both screen and bias impedances are of importance (Ref. A12). Reprinted by permission from RADIO ENGINEERS' HANDBOOK by F. E. Terman, copyrighted in* 1943, *McGraw-Hill Book Company Inc.*

R_k is now $I_k = I_b + I_{c2}$. The effect of R_k and C_k on the voltage gain at any frequency is given by the approximation (derived from Sect. 2 Eqn. 2)

$$\left|\frac{A'}{A}\right| \approx \sqrt{\frac{1 + (\omega C_k R_k)^2}{(1 + g_m R_k)^2 + (\omega C_k R_k)^2}} \qquad (3a)$$

on the assumption that the screen is completely by-passed.

When the screen is not completely by-passed, so that both cathode and screen circuits are attenuating simultaneously, the cathode attenuation characteristic is affected by the screen circuit. Under these circumstances it is possible to use the method due to Terman (Refs. A12, B2) whose curves are reproduced in Figs. 12.11C and 12.11D.

Terman expresses the relationship in the form

$$\frac{\text{Actual output voltage}}{\text{Output voltage with zero bias impedance}} = \frac{1}{1 + R_f g_m \eta \beta} \qquad (3b)$$

where g_m = mutual conductance of valve in mhos, at operating point

R_f = cathode bias resistance in ohms

$\eta = \dfrac{1}{1 + j(f/f_4)}$

f = actual frequency in c/s

$f_4 = 1/2\pi C_f R_f$ = frequency at which the reactance of C_f equals bias resistance R_f

and $\beta = \left|\dfrac{A'}{A}\right|$ contributed by the screen circuit at frequency f as given by eqn. (1). (Note that $\beta = 1$ for complete by-passing).

The procedure in a practical case is

1. Determine $\beta = |A'/A|$ from equation (1).
2. Knowing f and f_4, calculate f/f_4.
3. Apply this value of f/f_4 to Fig. 12.11C, thus determining the magnitude of η in db, and also its phase angle.
4. Calculate the value of $R_f g_m \eta \beta$, in db above unity.
5. Apply the value of $R_f g_m \eta \beta$ and the phase angle of η to Fig. 12.11D to determine the loss in gain (in decibels).

It will be seen that incomplete screen by-passing results in a smaller value of attenuation by the cathode impedance than would occur with complete screen by-passing.

Attenuation curves for a typical r.c.c. pentode are given in Fig. 12.12 for complete screen by-passing.

As a useful **rule of thumb,** sufficient for most design purposes other than for amplifiers incorporating negative feedback over 2 or 3 stages,

$$fC_k \approx 0.55 g_m \text{ for 1 db attenuation} \qquad (4a)$$
$$fC_k \approx 0.35\, g_m \text{ for 2 db attenuation} \qquad (4b)$$

where C_k is in microfarads,

and g_m = mutual conductance in micromhos **at the operating plate current.**

The effect of incomplete cathode by-passing on phase angle displacement is approximately the same as for a triode, and the curves of Fig. 12.3C may be used, but the value of B may be taken as roughly

$$B \approx 1 + g_m R_k$$

on the assumption that μ is very large and that R' is small compared with r_p, where $R' = R_L R_{g2}/(R_L + R_{g2})$.

The combined effect of incomplete cathode and screen by-passing and of a grid coupling condenser on gain and on phase angle displacement may be determined by adding the attenuations in decibels and the phase angles in degrees, provided that eqn. (3b) is used for calculating the attenuation due to the cathode impedance.

(B) Cathode bias resistance

The cathode bias resistance should preferably be the smallest value which can be used without any danger of grid current on the maximum signal. Fortunately, however, the resistance is not critical and a higher value has very little deleterious

FIG. 12.12

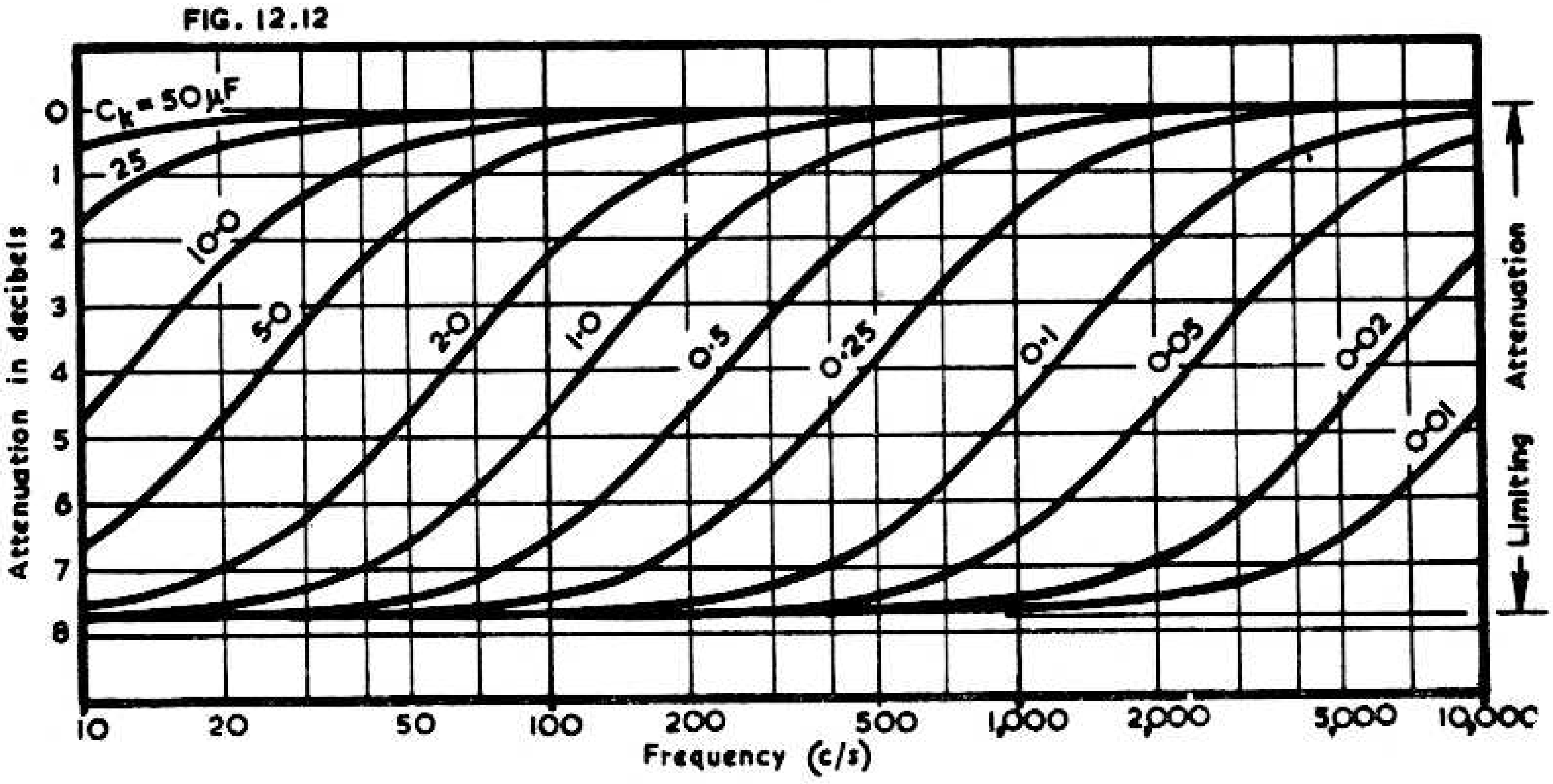


Fig. 12.12. *Attenuation characteristics for selected values of cathode by-pass condenser with valve type* 6J7 *and* $E_{bb} = 250$ *volts,* $R_L = 0.25$, $R_g = 0.5$, $R_s = 1.5$ *megohms,* $R_k = 2000$ *ohms : screen adequately by-passed.*

effect under normal operation, provided that the screen voltage is adjusted to give the correct operating plate current.

Optimum values may be calculated for any specific case by first finding the fixed bias and then calculating $R_k = E_{c1}/(I_b + I_{c2})$.

Alternatively, the value of R_k may be determined by eqn. (5f) below.

The procedure for the calculation of the cathode bias resistor when a series screen resistor is also used, is given in Sect. 3(vi)C, while the graphical method is described in Sect. 3(xiii) and (xiv).

(C) Effect of reverse grid current with cathode bias

When the screen voltage is maintained constant from a low impedance source, the following approximate relationship may be derived from Chapter 3 Sect. 1 eqn (6) :

$$\frac{R_{g1} \text{ for cathode bias}}{R_{g1} \text{ for fixed bias}} \approx 1 + R_k g_k \approx 1 + R_k g_m I_k / I_b \qquad (4c)$$

For a typical pentode, type 6J7, the ratio is approximately 1.7, 2.1 and 2.6 for $R_L = 0.1$, 0.22 and 0.47 megohm respectively. Now the value of R_{g1} for fixed bias is derived in (v)a below for a reverse grid current of 1 μA, so that for type 6J7 the maximum grid resistor under this condition with cathode bias and fixed screen voltage is approximately

0.56 megohm for $R_L = 0.1$ megohm,
0.91 megohm for $R_L = 0.22$ megohm,
1.5 megohms for $R_L = 0.47$ megohm.

With a series screen resistor the following approximate relationship may be derived from Chapter 3 Sect. 1 eqn. (6).

$$\frac{R_{g1} \text{ for cathode bias}}{R_{g1} \text{ for fixed bias}} \approx 1 + R_k g_k + \frac{P g_k R_{g2}}{\mu_{g1g2}} \qquad (4d)$$

The first two terms are the same as in eqn. (4c), while the third term is the same as the second term of eqn. (5b) below, so that the ratio in eqn. (4d) will normally exceed 7 : 1, for $R_L = 0.1$ megohm and will be considerably greater than this value for $R_L = 0.22$ megohm or higher.

However, it is generally advisable to limit the grid resistor to 2.2 megohms maximum for amplifiers having a reduced frequency range and 1 megohm or less for amplifiers having a maximum frequency of 10 000 c/s or more.

(v) Fixed bias

The optimum bias is the smallest which can be used without danger of positive grid current, provided that the screen voltage is adjusted to give the correct operating plate current.

If the input voltage is known,

$$E_{c1} = 1.41\,E_i + k_1 + k_2 \tag{4e}$$

If the output voltage is known,

$$E_{c1} = (1.41E_0/A) + k_1 + k_2 \tag{4f}$$

where E_i = r.m.s. input voltage
E_0 = r.m.s. output voltage from plate to cathode
A = voltage gain of stage
k_1 = bias voltage to avoid damping due to positive grid current—a value of 1.0 volt maximum would cover all normal indirectly heated valves*
and k_2 = increment of bias voltage to allow for the effects of reverse grid current in the grid resistor (see below).

If a fixed screen voltage is used, and the grid resistor has the maximum value determined in (a) below, then the value of k_2 may be taken as 0.1 volt for high slope valves (e.g. 6AU6) with R_L not less than 0.1 megohm, and the same value for low slope valves with R_L not less than 0.22 megohm. The value of k_2 may be taken as 0.2 volt for low slope valves having R_L less than 0.22 but not less than 0.1 megohm.

If a series screen resistor is used, and if the grid resistor does not exceed 1 megohm, k_2 may be taken as 0.1 volt maximum.

With some valves, as the screen voltage is decreased, the grid current commencement (or cross-over) point may tend to move to a more negative value in indirectly-heated types, and to a less positive value (possibly even negative) in directly-heated battery pentodes. **Low screen voltages should therefore be avoided with zero-bias operation of battery pentodes, particularly when a high plate load resistance is used.**

The effect of reverse grid current with fixed grid bias

(a) With fixed screen voltage

The position of the operating point in pentodes, for minimum distortion, is fairly critical. Thus when the correct screen and grid bias voltages have been applied, any reverse grid current that may flow through the grid resistor R_{g1} will cause a change in bias and hence a change in plate current to a value less than the optimum. The writer considers a change of 0.1 E_{bb}/R_L as being the maximum permissible change in plate current due to the flow of grid current. On this basis, the maximum permissible grid resistance with fixed bias and fixed screen voltage can be shown to be given approximately by

$$R_{g1} \approx K_p\left(\frac{R_L}{1000}\right)^{0.3p} \tag{5a}$$

where R_{g1} = maximum grid resistance in ohms,
R_L = load resistance in ohms,
$K_p = \dfrac{8I_{b0}}{g_{m0}E_{bb}I_{c1}}$
I_{b0} = plate current in amperes at which both g_{m0} and I_{c1} are measured,
g_{m0} = mutual conductance in mhos at plate current I_{b0},
E_{bb} = plate supply voltage,
and I_{c1} = maximum rated reverse grid current in amperes at plate current I_{b0}.

In this calculation the assumption was made that the ionization current is proportional to the cathode current, which relationship only holds approximately and then only when the grid leakage current is small compared with the ionization current and provided that the latter does not increase during operation.

Values of K_p have been derived for three valve types—

$$\left.\begin{array}{ll}\text{Type 6J7} & K_p = 5.2 \times 10^4\\ \text{Type 6SJ7} & K_p = 5.8 \times 10^4\\ \text{Type 6AU6} & K_p = 5.7 \times 10^4\end{array}\right\}\begin{array}{l}\text{for } E_{bb} = 250 \text{ volts}\\ \text{and } I_{c1} = 1\ \mu\text{A}.\end{array}$$

Values of the function of R_L are given below—

*Higher values may occur in a few cases, but are not typical. See also comments on triodes in Sect. 2(iv)B. For grid damping see Sect. 2(iv)D.

R_L	$\left(\frac{R_L}{1000}\right)^{0.38}$
100 000	5.75
220 000	7.8
470 000	10.4

From which the following values may be calculated :

Type			
	R_L = 0.1	0.22	0.47 megohm
6J7	R_{g1} = 0.3 max.	0.4 max.	0.54 max. megohm
6SJ7	R_{g1} = 0.33 max.	0.45 max.	0.6 max. megohm
6AU6	R_{g1} = 0.33 max.	0.44 max.	0.59 max. megohm

We may therefore conclude that, for a maximum reverse grid current of 1 μA, the value of R_{g1} with fixed bias and fixed screen voltage should not exceed

0.33 megohm for R_L = 0.1 megohm,
0.43 megohm for R_L = 0.22 megohm,
0.56 megohm for R_L = 0.47 megohm.

Where the variations of reverse grid currents are such that the large majority of valves would have values below half the maximum value—i.e. in this case below 0.5 μA—double these values of R_{g1} would be satisfactory. This would also apply in any cases where the maximum value of reverse grid current is specified as 0.5 μA.

(b) **With series screen resistor**

When fixed bias is used in conjunction with a series screen resistor supplied from the plate voltage source, it may be shown from eqn. (9) of Chapter 3 Sect. 1 that

$$\frac{R_{g1}\text{ for series screen resistor}}{R_{g1}\text{ for fixed screen voltage}} \approx 1 + \frac{Pg_kR_{g2}}{\mu_{g1g2}} \qquad (5b)$$

where $P = I_{c2}/I_k$

$g_k = g_m(I_k/I_b)$ at the operating point,

R_{g2} = resistance of series screen resistor in ohms,

and μ_{g1g2} = "triode" amplification factor.

Ratios calculated from eqn. (4c) for typical valves, with a load resistance of 0.1 megohm or more, exceed 6 times, so that **when a high-resistance series screen resistor is used, the grid resistor may be at least 6 times the value quoted above for fixed screen voltage.**

However, it is generally advisable to limit the grid resistor to 2.2 megohms maximum for amplifiers having a reduced frequency range and 1 megohm or less for amplifiers having a maximum frequency of 10 000 c/s or more.

FIG. 12.13

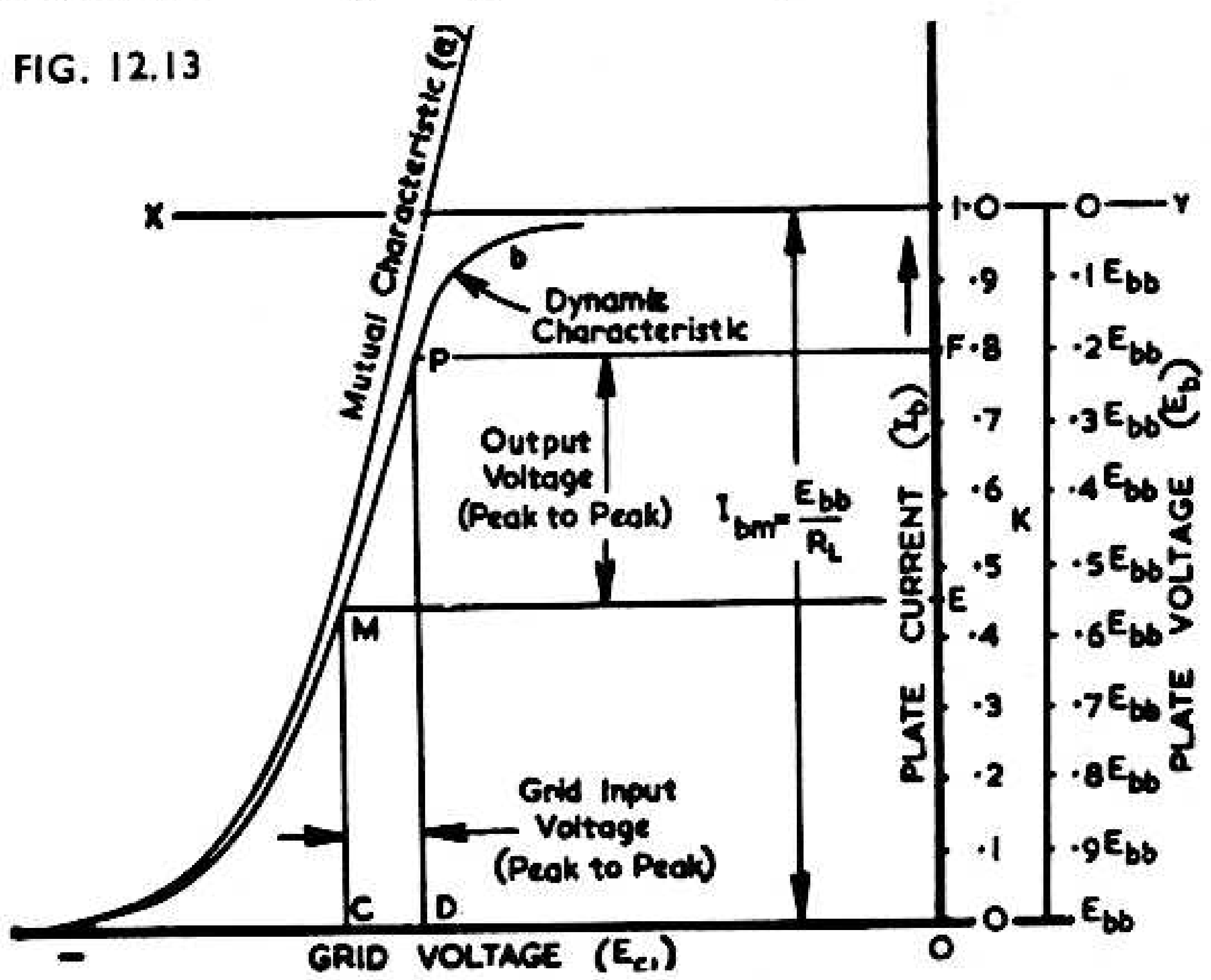


Fig. 12.13. *Dynamic characteristic of r.c.c. pentode for fixed screen voltage (b) compared with mutual characteristic (a).*

(vi) Dynamic characteristics of pentodes

(A) General description

A single dynamic characteristic of a pentode, together with the equivalent mutual characteristic, are shown in Fig. 12.13. The slope of the dynamic characteristic at any point is given by

$$g_{md} = g_m \frac{r_p}{r_p + R_L} = \frac{g_m}{1 + (R_L/r_p)} \tag{5c}$$

The values of g_m and g_{md} are normally within 10% provided that th^ operating plate current does not exceed 0.8 E_{bb}/R_L. Values of g_m and g_{md} for valve types 6J7 and 6SJ7 are given in (vii) below.

Pentodes differ from triodes in that there is an unlimited number of dynamic characteristics for any selected plate voltage and load resistance, there being one characteristic for each value of screen voltage. Some typical dynamic characteristics are shown in Fig. 12.14A from which it will be seen that these have much the same general form, but that they are moved bodily sideways. Careful examination will show that the curves for lower screen voltages are less curved than those for higher screen voltages. As a general rule, the screen voltage should therefore be as low as possible, provided that the correct operating point can be maintained and that positive grid current does not flow.

FIG. 12.14A

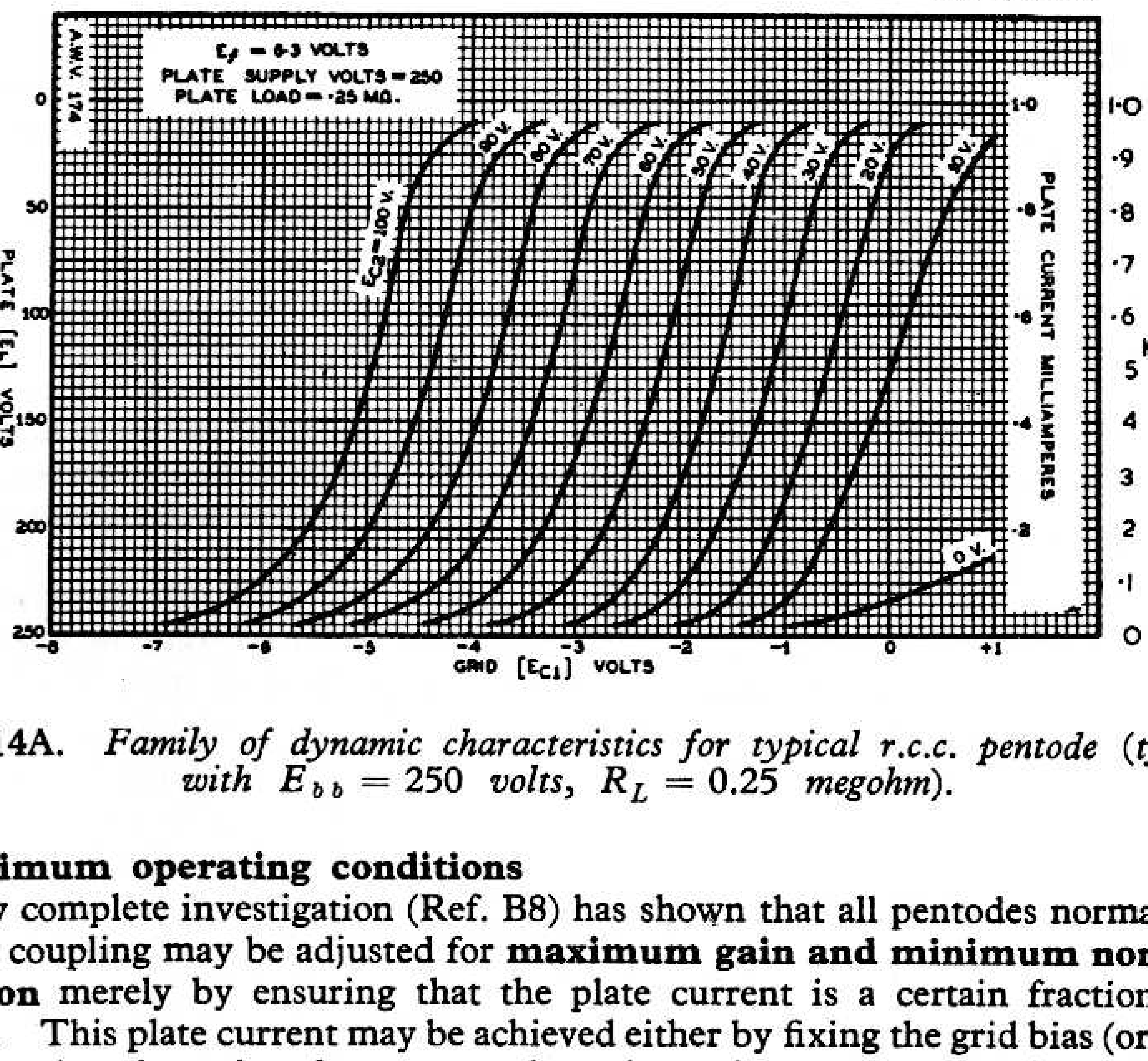


Fig. 12.14A. *Family of dynamic characteristics for typical r.c.c. pentode (type* 6J7 *with* $E_{bb} = 250$ *volts,* $R_L = 0.25$ *megohm).*

(B) Optimum operating conditions

A very complete investigation (Ref. B8) has shown that all pentodes normally used with r.c. coupling may be adjusted for **maximum gain and minimum non-linear distortion** merely by ensuring that the plate current is a certain fraction (K) of E_{bb}/R_L. This plate current may be achieved either by fixing the grid bias (or cathode bias resistor) and varying the screen voltage (or series screen resistor) or vice versa. The effect of reverse grid current in the grid resistor is to decrease the effective grid bias by an amount which may be appreciable with fixed screen voltage, but may usually be neglected when a high-resistance series screen resistor is used. In the former case, with fixed screen voltage, it is suggested that the design be based on a value of half the maximum specified reverse grid current flowing through the grid resistor.

The optimum value of K is not appreciably affected by either the following grid resistor or the electrode voltages, provided only that the screen voltage is kept reasonably low. The optimum value of K for minimum distortion is, however, a function of the output voltage as indicated below.

Output voltage (r.m.s.) =	$0.04E_{bb}$	$0.08E_{bb}$	$0.16E_{bb}$	$0.24E_{bb}$
Output voltage (r.m.s.)* =	10	20	40	60
Optimum K =	0.78	0.76	0.70	0.62

*For $E_{bb} = 250$ volts.

where $K = I_b/I_{bm}$ and $I_{bm} = E_{bb}/R_L$, and $R_g = 2R_L$.

For example, with $E_{bb} = 250$ volts and $R_L = 0.25$ megohm, $I_{bm} = 1$ mA.

The **voltage between plate and cathode** is a function of K: $E_b = (1 - K)E_{bb}$.

When K =	0.78	0.76	0.70	0.62
Then E_b =	$0.22E_{bb}$	$0.24E_{bb}$	$0.3E_{bb}$	$0.38E_{bb}$

(C) **Determination of series screen and cathode bias resistors when plate and screen current curves are available**

If curves of both plate and screen currents versus grid voltage under resistance loaded conditions (e.g. Figs. 12.14A and 12.18) are available, the procedure is—

1. Determine E_{c1}—see (v) above.
2. Determine the optimum value of K—see (vi)B above.
3. From the I_b curves, determine E_{c2} to give the required values of E_{c1} and K; also determine I_b at the operating point.
4. From the I_{c2} curves, determine I_{c2} at the operating point.
5. Then $R_k = E_{c1}/(I_b + I_{c2})$ (5d)

and $R_s = (E_{cc2} - E_{c2})/I_{c2}$ (5e)

(D) **Calculation of series screen and cathode bias resistors**

The procedure for calculating R_s and R_k is—

1. Determine E_{c1} as for fixed bias—see (v) above.
2. Determine the optimum value of K—see (vi)B above.
3. Then $R_k \approx \dfrac{E_{c1}R_L m}{KE_{bb}(m+1)}$ (5f)

where E_{c1} = control grid voltage (taken as positive)

and $m = I_b/I_{c2}$.

Typical values of m at normal voltages are:

Type	6J7	6SJ7	6SF7	6SH7	6AC7	6AU6	1S5	1U5
m	4.0	3.75	3.75	2.63	4.0	2.54	4.0	4.0

Equation (5f) is perfectly general, and may be used with any screen voltage source. The only approximation is the assumption that m is constant.

4. Then $E_{c2} \approx E_{c1}\mu_t + E_2 - \dfrac{KE_{bb}\mu_t}{R_{L1}g_{md}}\left(1 - \dfrac{R_{L1}}{R_L}\right)$ (5g)

where E_{c2} = screen voltage
μ_t = triode mu
E_2 = screen voltage at which $I_b = 2$ mA for $E_{c1} = 0$
$\approx$ 22 (for 6AU6), 25 (for 6SJ7), 39 (for 6J7)
$R_{L1} = 0.1\ M\Omega$

and g_{md} = max. slope of dynamic characteristic with $R_L = 0.1\ M\Omega$.

Equation (5g) may also be used with fixed bias E_{c1}.

5. Then $R_s \approx \dfrac{mR_L}{KE_{bb}}(E_{bb} - E_{c2})$ (5h)

Note: This procedure is based on several approximations (e.g. that m, g_{md} and E_2 are constant) but is sufficiently accurate for all practical purposes for any pentode.

It will be seen that a single operating condition is incapable of giving optimum performance for both low and high levels. For low level operation, E_{c1} may be taken as -1.3 volts, while for high level operation E_0 may arbitrarily be taken as $0.24E_{bb}$ volts r.m.s. unless it is desired to determine some intermediate condition.

It will be seen from eqn. (5g) that the optimum screen voltage is a function, not only of E_{c1}, but also of R_L—a higher value of load resistance permits a lower screen voltage.

(E) Screen supply from voltage divider

Although a series screen resistor is normally preferable, the screen may be supplied in certain applications from a voltage divider. The procedure is straight forward, but the voltage should be adjusted manually or else the two resistors should have $\pm 5\%$ tolerances.

When a screen voltage divider is used, the equivalent series screen resistance is given by the resultant of the two sections of the voltage divider in parallel. This value may be used as R_{g2} in eqn. (6) of Chapter 3 Sect. 1, although if R_{g2} is less than 50 000 ohms, its effect is small.

If the equivalent series screen resistance is less than 50 000 ohms, the screen voltage may be considered as fixed, and it is then necessary to increase the grid bias to allow for the **effect of reverse grid current.**

The increase in grid bias is

$$\Delta E_{c1} = R_{g1} I_{c1} I_b / I_{b0} \tag{6a}$$

where I_{b0} = plate current at which the reverse grid current is measured, and I_{c1} may be taken as half the maximum specified reverse grid current (say 1 μA); i.e. (say) $I_{c1} = 0.5\ \mu$A.

The increased value of cathode bias resistor is given by

$$R_k' = R_k + \Delta R_k \tag{6b}$$

where $\Delta R_k = \dfrac{\Delta E_{c1}}{I_b + I_{c2}} = \dfrac{\Delta E_{c1}}{I_b} \cdot \dfrac{m}{m+1}$

and E_{c1} is given by equation (6a).

Using the suggested value of I_{c1}, we have

$$R_k' = R_k + \frac{R_{g1} \times 0.5 \times 10^{-6}}{I_{b0}} \cdot \frac{m}{m+1} \tag{6c}$$

For example, if R_{g1} = 1 megohm, I_{b0} = 2 mA, and $m = 4$, then $R_k' = R_k + 200$ ohms.

(vii) Gain at the mid-frequency

(A) The voltage gain at the mid frequency (for small input voltages) is given by

$$A_0 = \frac{g_m}{(1/r_p) + (1/R_L) + (1/R_g)} \tag{7}$$

where g_m = mutual conductance at operating plate current,
r_p = plate resistance at operating point,
and R_g = resistance of following grid resistor.

The mutual conductance at the working plate current is often an unknown factor, and methods for determining it are given below (E).

(B) Alternatively, the gain may be calculated from the slope of the dynamic characteristic:

$$A_0 = g_{md} R\left(\frac{r_p + R_L}{r_p + R}\right) \approx g_{md} R \tag{8}$$

where g_{md} = slope of dynamic characteristic at the operating point—see (E) below,
and $R = R_L R_g/(R_L + R_g)$.

(C) **The gain may also be calculated graphically from the dynamic characteristic** (Fig. 12.13). If the peak-to-peak grid input voltage is CD, then the instantaneous plate current will swing between the extreme limits M and P, giving a peak-to-peak output voltage EF measured on the plate voltage scale. To allow for the effect of R_g, the output voltage must be multiplied by the factor $R_g/(R_g + R')$ where $R' = r_p R_L/(r_p + R_L) \approx R_L$.

Typical voltage gains are:

R_L	R_g	6J7	6SJ7	6AU6
0.1 megohm	0.5 megohm	94	104	168
0.25 megohm	0.5 megohm	140	167	230*
0.5 megohm	2.0 megohms	230	263	371**

*R_L = 0.22 megohm. **R_L = 0.47 megohm.

If the resistance of the following grid resistor R_g is limited, the **maximum gain on low input voltages** is obtained when the load resistance is approximately equal to the following grid resistance, i.e. $R_L \approx R_g$.

(D) Gain in terms of g_m and g_p

Eqn. (7) may be put into the alternative form

$$A_0 = \frac{g_m R}{1 + g_p R} \tag{9a}$$

where $R = R_L R_g/(R_L + R_g)$
and g_p = plate conductance $= 1/r_p$.

Unfortunately " G," curves for pentodes hold only for a fixed screen voltage and are not flexible (g_m curves for type 7E7 are shown in Ref. A15).

(E) The determination of g_m, g_{md}, and r_p

The values of g_m, g_{md} and r_p for use in equations (7), (8) and (9) are difficult to calculate with any precision, but may be measured on an average valve.

Curves of constant g_{md} could, with advantage, be plotted on the family of dynamic characteristics such as Fig. 12.14A, although this has not yet been done.

Mutual conductance

For the general case see (x)A below. Fig. 12.14B gives an approximation to the maximum value of g_m with any load resistance for two typical pentodes. This value is believed to be accurate within $\pm$ 10% for an average valve.

Slope of dynamic characteristic

The value of the slope of the dynamic characteristic at any point is given by

$$g_{md} = g_m/(1 + R_L/r_p) \tag{9b}$$

where g_{md} = slope of dynamic characteristic at operating point
g_m = mutual conductance at operating point
and r_p = plate resistance at operating point.

FIG. 12.14B

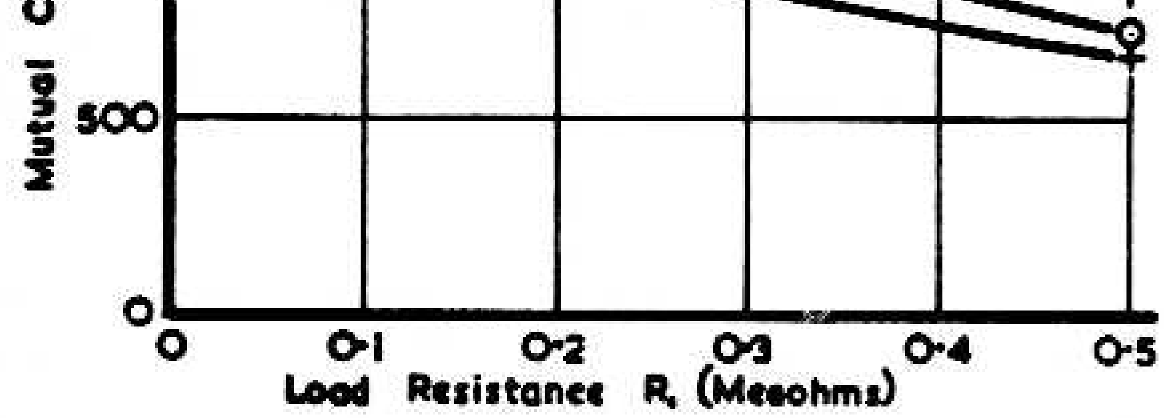


Fig. 12.14B. *Mutual conductance of resistance-capacitance coupled pentodes types* 6J7 *and* 6SJ7 *plotted against load resistance. In each case the value of K is taken as being adjusted to give maximum mutual conductance.*

Plate resistance

The value of plate resistance r_p at the operating point may be estimated from Sect. 3(x) below.

Data for types 6J7 and 6SJ7

Detailed values of mutual conductance (g_m), slope of the dynamic characteristic (g_{md}) and plate resistance (r_p) for types 6J7 and 6SJ7 are tabulated below. It is emphasized that there are considerable variations between valves, particularly with the plate resistance.

Valve type		6J7			6SJ7			
Plate supply voltage		250			250			volts
Published g_m*		1225			1650			μmhos
R_L MΩ	K	g_m μmhos	g_{md} μmhos	r_p MΩ	g_m μmhos	g_{md} μmhos	r_p MΩ	
0.1	0.78	1065	980	0.6	1390	1280	0.7	
	0.76	1080	1010	0.67	1380	1290	0.9	
	0.70	1130	1080	1.1	1370	1310	1.3	
	0.62	1120	1080	1.5	1340	1290	1.4	
	0.55	1100	1060	2.1	1335	1280	1.7	
0.25	0.78	896	834	1.6	1050	990	1.7	
	0.76	880	836	1.9	1040	985	1.8	
	0.70	860	830	3.0	1000	960	2.4	
	0.62	820	790	3.3	940	900	3.3	
	0.55	790	765	3.7	920	845	3.4	
0.5	0.78	670	640	3.0	720	660	3.0	
	0.76	660	630	3.6	710	660	3.0	
	0.70	647	620	4.5	650	648	3.7	
	0.62	595	575	5.2	590	580	5.0	
	0.55	550	540	6.0	519	495	5.4	

*With E_b = 250 volts, E_{c2} = 100 volts, E_{c1} = − 3 volts.

(viii) Dynamic characteristics of pentodes and comparison with triodes

A family of dynamic characteristics for a typical pentode is shown in Fig. 12.14A. There is one curve for every possible value of screen voltage, but it is obvious that those for lower screen voltages (e.g. 30 volts) are less curved than those for higher voltages. In actual operation the quiescent operating point is fixed, and the signal voltage on the grid causes the plate current to swing along the dynamic curve. This only holds strictly when there is no following grid resistor or other shunt load, but it is sufficient for design purposes since allowance can readily be made for the effect of R_g in reducing both the voltage gain (eqn. 8) and the output voltage.

The voltage drop in R_L is proportional to the plate current, and therefore a plate voltage scale may be added to Fig. 12.14A. The general treatment and the calculation of gain are identical with those for ideal linear dynamic characteristics as applied to triodes in Sect. 2(viii). In the case of pentodes the operation can be kept outside the grid current region, without any other loss, by selecting a suitable screen voltage.

A single pentode dynamic characteristic (that is one for a fixed screen voltage) is drawn as curve *b* in Fig. 12.13. For comparison, curve *a* has also been included, this being the ordinary mutual characteristic for the same screen voltage. At low plate currents the two are practically identical but they diverge steadily up to the point of inflexion* P, beyond which the dynamic characteristic forms the " top bend."

The line XY is at a plate current $I_{bm} = E_{bb}/R_L$ at which the factor $K = 1$. This curve is typical of the shape of all pentode dynamic characteristics which differ mainly in the slope and the horizontal displacement of the curve. The effects of changes in R_L are largely overcome by the use of the factor K in place of the actual plate current.

Comparison between triode and pentode dynamic characteristics

Both triode and pentode dynamic characteristics are shown in Fig. 12.15 in such a way as to enable a comparison to be made between them. Over the region from $K = 0.4$ to $K = 0.6$ they appear to be very similar, but in the region from $K = 0.15$ to 0.4 the pentode characteristic appears less curved than the triode, while in the region from $K = 0.6$ to 0.8 the triode characteristic appears less curved than the pentode.

*The point of inflexion is the point at which the curvature changes from one direction to the other, and is the point of greatest slope.

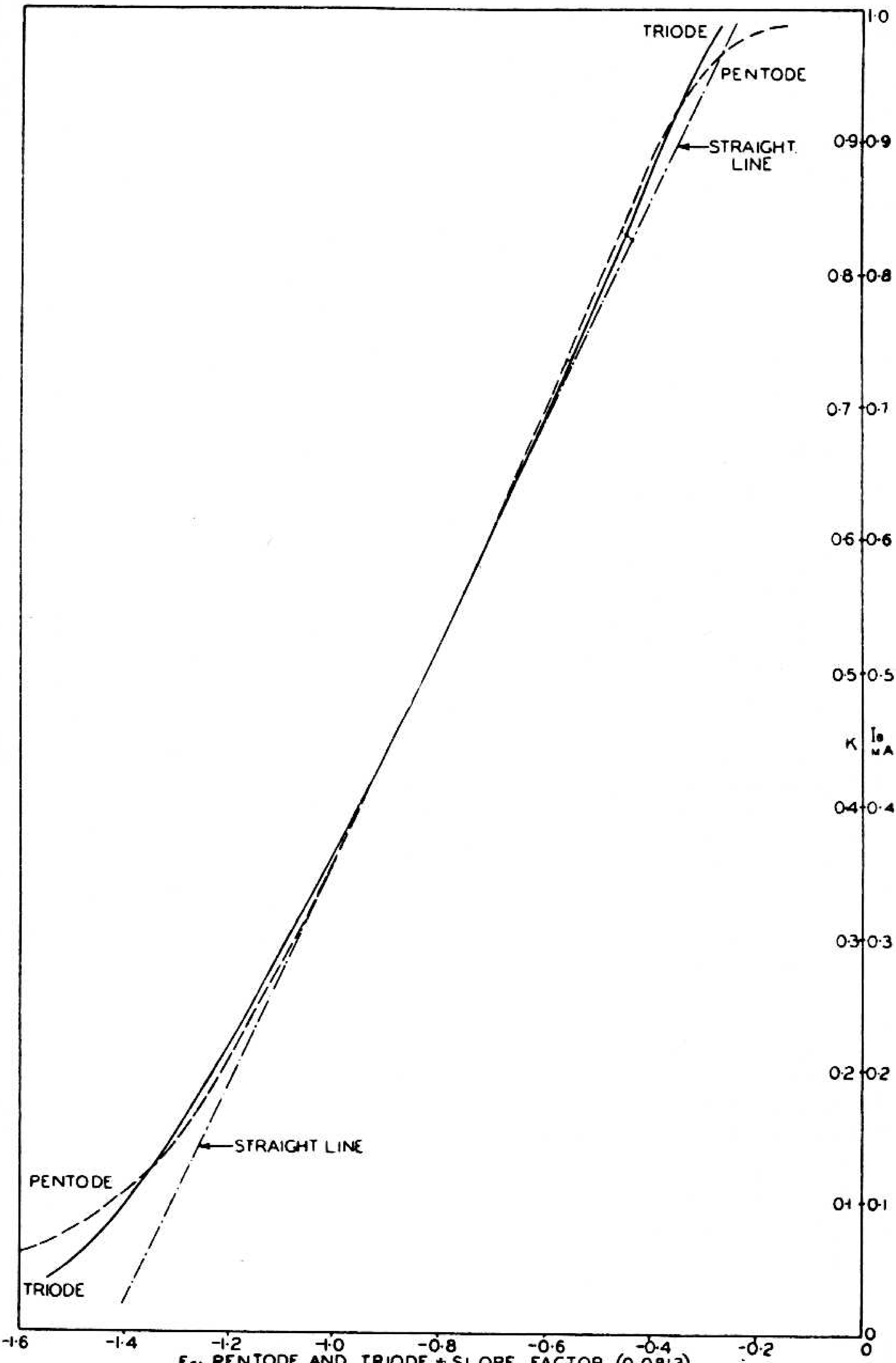


FIG. 12·15

Fig. 12.15. *Comparison between pentode and triode dynamic characteristics of the same valve* (6SJ7 *with* E_{bb} = 250 *volts*, R_L = 0.25 *megohm*, E_{c2} = 22.5 *volts for pentode*, E_{c2} *connected to plate for triode characteristics; triode characteristics divided by slope factor* 0.0813 *and superimposed so that points for* I_b = 0.5 *mA coincide*).

(ix) Maximum voltage output and distortion

The maximum voltage output, with any specified limit of distortion, is obtained when R_g is much greater than R_L. If R_g is limited, the maximum voltage output is obtained when R_L is much less than R_g. The following table is typical of r.c.c. pentodes such as types 6J7 and 6SJ7, and also enables a comparison to be made with triodes. The optimum value of K has been used throughout.

Intermodulation distortion* with E_{bb} = 250 volts

Output voltage†		10	19	37	63
Valve type 6J7		Intermodulation distortion %			
$R_L = 0.1$ MΩ	$R_g = 0.1$ MΩ	0.35	2.2	8.6	—
	0.2	0.3	1.5	4.6	14.0
	0.4	0.26	0.85	3.9	12.5
$R_L = 0.25$ MΩ	$R_g = 0.25$ MΩ	0.65	1.9	7.4	37
	0.5	0.5	1.6	4.4	17
	1	0.35	1.4	3.5	15
$R_L = 0.5$ MΩ	$R_g = 0.5$ MΩ	1.0	3.7	14.0	54
	1	0.8	2.6	9.5	32
	2	0.5	2.0	6.5	19
Valve type 6SJ7					
$R_L = 0.1$ MΩ	$R_g = 0.1$ MΩ	1.0	2.9	12.0	—
	0.2	0.9	2.2	8.3	24
	0.4	0.8	1.5	5.9	20
$R_L = 0.25$ MΩ	$R_g = 0.25$ MΩ	2.1	6.0	16.0	67
	0.5	1.1	3.8	13.0	38
	1	1.0	3.0	11.0	23
$R_L = 0.5$ MΩ	$R_g = 0.5$ MΩ	3.0	8.3	18.0	76
	1	2.1	6.6	15.0	42
	2	1.5	4.4	12.5	28

The maximum output voltage for limited intermodulation distortion (r.m.s. sum) is indicated below for type 6SJ7 pentode with $R_g = 4R_L$.

R_L	$E_{0(r.m.s.)}/E_{bb}$			
	I.M. = 2.5%	5%	10%	20%
0.1 MΩ	0.09	0.13	0.18	0.25
0.25	0.08	0.11	0.16	0.22
0.5	0.07	0.09	0.13	0.20

When $R_g = 2R_L$, the factors above should be multiplied by 0.87. When $R_g = R_L$, the factors above should be multiplied by 0.7. The values are for E_{bb} = 250 volts, but hold closely over the range from 200 to 300 volts. Optimum operating conditions are assumed.

Harmonic distortion

Type 6SJ7 with E_{bb} = 250 volts, R_L = 0.25 MΩ, R_g = 0.5 MΩ

K =	0.8	0.76	0.72	0.65	0.6	0.56
$E_{0(r.m.s.)}$ volts	8	16	30	52	62	78
$E_{0(r.m.s.)}/E_{bb}$	0.03	0.06	0.12	0.21	0.25	0.31
H_2 %	0.3	0.9	2.6	3.5	5.4	4.4
H_3 %	0.24	0.19	1.25	4.2	5.0	12.0
H_4 %	0.14	0.11	0.19	1.9	2.2	2.85
H_5 %	0.01	0.02	0.18	0.67	0.28	0.17

*Modulation method—r.m.s. sum. For details and for relation between intermodulation and harmonic distortion see Chapter 14 Sect. 3.

†The arithmetical sum of the r.m.s. values of the two component waves.

Comments

1. Type 6J7 (with published $g_m = 1225$ μmhos) has less distortion than type 6SJ7 (with published $g_m = 1650$ μmhos) under the same conditions.

The distortion for a given output voltage increases when a valve is replaced by another having higher mutual conductance, although there are also differences between valve types having approximately the same mutual conductance.

2. Load resistance $R_L = 0.1$ megohm provides lower distortion than higher values of load resistance.

3. The distortion under any given conditions decreases when the resistance of the following grid resistor is increased.

Comparison between triode and pentode

[Refer Sect. 2(ix)]

The comparison is based on intermodulation distortion with type 6SJ7 as both triode and pentode, having $R_L = 0.25$ and $R_g = 1.0$ megohm (Fig. 12.16A). Generally similar results are obtained with type 6J7 and with other load resistances (Ref. B8).

1. At the level used in the first a-f stage in a typical receiver ($E_0 = 10$ volts r.m.s.) the pentode gives only about one eighth of the intermodulation distortion given by a triode, when both are adjusted for minimum distortion.

2. The two curves in Fig. 12.16A cross, and the intermodulation distortion is therefore the same for both triode and pentode, at about 31 r.m.s. volts output.

3. At higher output voltages the pentode gives the greater intermodulation distortion, the ratio being 2.3 : 1 at 63 r.m.s. volts output.

4. The pentode, to give the minimum value of distortion, requires fairly critical adjustment. Under the working conditions recommended in this section, however, the distortion with a pentode is likely to be less than with a general-purpose triode at output voltages up to about 20 volts r.m.s.

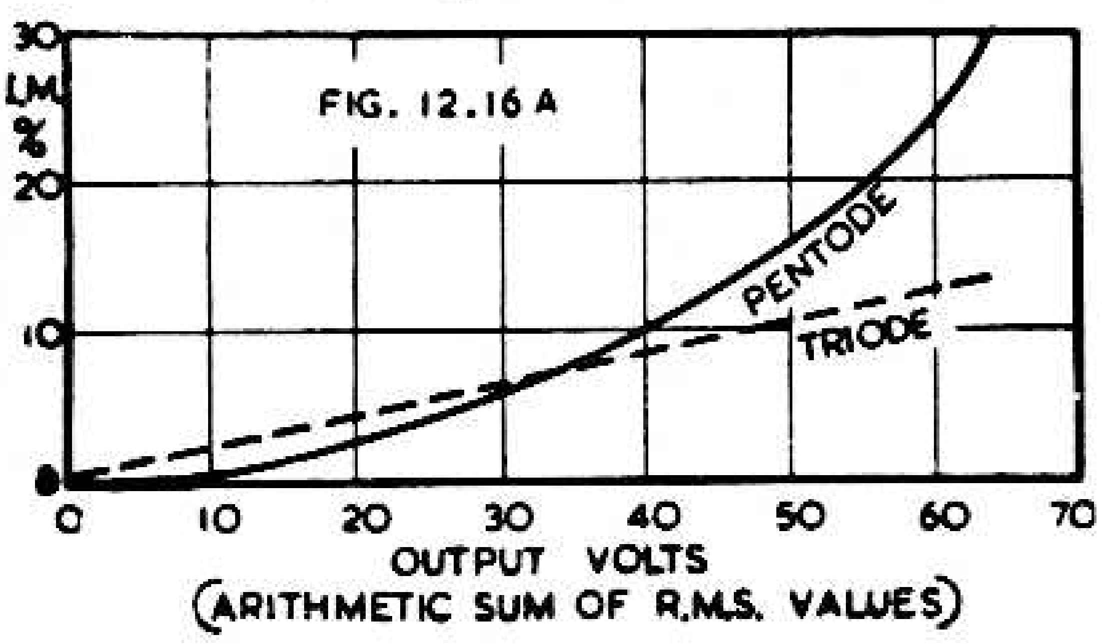


Fig. 12.16A. *Intermodulation distortion for type 6SJ7 with $E_{bb} = 250$ volts, $R_L = 0.25$ MΩ, $R_g = 1.0$ MΩ, in both triode and pentode operation.*

(x) Conversion factors with r.c.c. pentodes

(A) **The mutual conductance** may be derived from published data (generally plotted against E_{c1}) or if these are not available then it may be calculated as for triodes, Sect. 2(x). See also values quoted for 6J7 and 6SJ7 in (vii) above.

(B) **The plate resistance** may be estimated by the following method from published data. The usual data are for either equal plate and screen voltages (e.g. 100 V) or for 250 V and 100 V respectively. Unfortunately there is no general rule for calculating the effect of a change of plate voltage on the plate resistance, and the accuracy obtainable by graphical means is very poor. However, as some sort of guide, the plate resistance of type 6SJ7 is increased about twice, and of type 6J7 about 2.75 times, for a change from 100 to 250 volts on the plate, with 100 volts on the screen and − 3 volts grid bias.

The plate resistance of a r.c.c. pentode may vary from slightly below the published value at $E_b = E_{c2} = 100$ volts, to 6 or 8 times this value, depending on the load resistance and value of K. Typical values for types 6J7 and 6SJ7 are given below [see also table in (vii) above] :

$R_L = 0.1$ megohm	$r_p = 0.6$ to 2.1 megohms
$R_L = 0.25$ megohm	$r_p = 1.6$ to 3.7 megohms
$R_L = 0.5$ megohm	$r_p = 3.0$ to 6.0 megohms
	for $K = 0.78$ to 0.55 respectively.

There are considerable variations between different valves, and some may have plate resistances less than half these values, while others may have higher plate resistances.

(C) **Conversion factors applied to whole amplifiers**

If one set of operating conditions is available, it is possible to calculate others.

Given :	E_{bb}	R_L	R_k	R_s	R_g	A	E_0
	300 V	0.1 MΩ	450 Ω	0.5 MΩ	0.25 MΩ	82	81 V

Example :

(1) To calculate conditions for $R_L = 0.25$ megohm,

$E_{bb} = 300$ volts :—

$F_r = 2.5$ (i.e. resistance conversion factor).

$R_k = 2.5 \times 450 = 1130$ ohms.

$R_s = 2.5 \times 0.5 = 1.25$ megohms.

$R_g = 2.5 \times 0.25 = 0.75$ megohm.

E_0 will be approximately the same for the same distortion.

A is affected both by the load and the mutual conductance.

The load resistance factor is 2.5 ; the mutual conductance factor is $\sqrt[3]{I_{b2}/I_{b1}} = \sqrt[3]{1/2.5} = 0.74$. Therefore $A = 82 \times 2.5 \times 0.74 = 152$.

Note : If E_{bb} had been altered, the mutual conductance factor would have been $\sqrt[3]{E_{bb2}R_{L1}/E_{bb1}R_{L2}}$.

(2) If all resistors are multiplied by a factor F_r, leaving E_{bb} unchanged, the voltage gain will be increased approximately as tabulated :

F_r	0.4	0.5	0.75	1.0	1.5	2	2.5
Voltage gain	0.54A	0.63A	0.83A	A	1.43A	1.58A	1.85A approx.

In practice these may vary $\pm$ 10% or even more.

(xi) Equivalent circuit of r.c.c. pentode

The exact a.c. equivalent plate circuit of a r.c.c. pentode is given in Fig. 12.16B where the "constant current generator" circuit has been adopted. This could equally be applied to the triode case (Fig. 12.8) or vice versa. In other respects the triode and pentode equivalent circuits are identical [see Sect. 2(xii)]. Both the cathode and screen circuits are assumed to be adequately by-passed at all signal frequencies, or the effects to be separately calculated.

FIG. 12.16 B

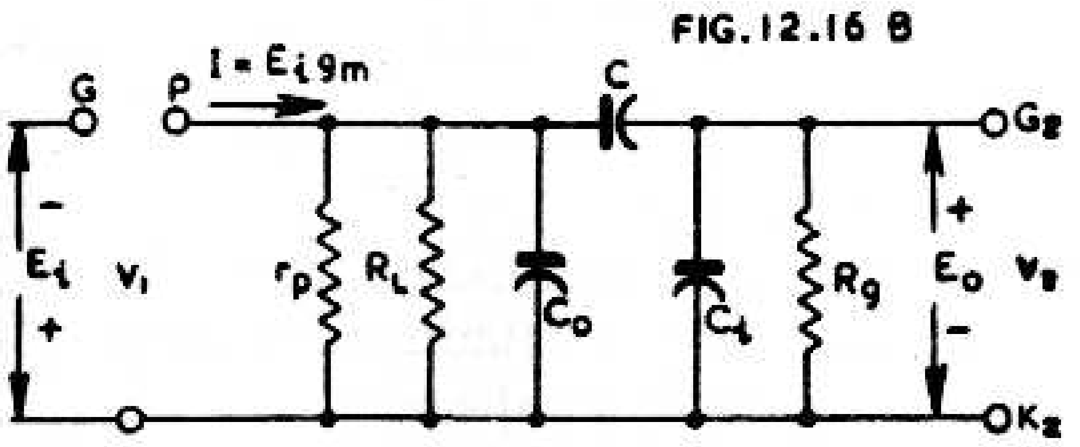


Fig. 12.16B. *Exact a.c. equivalent plate circuit of a r.c.c. pentode.*

(xii) Voltage gain and phase shift

Provided that the cathode and screen circuits are adequately by-passed at all signal frequencies, or that their voltages are obtained from low impedance sources, the voltage gain and phase shift will be the same as for the triode case [Sect. 2(xiii)]. In the case of a pentode it is usually possible to neglect the "Miller Effect" capacitance from the plate circuit.

The **low frequency response** [Sect. 2(xiii) ; eqn. (15)] will be slightly higher than for a triode with the same values of C_1, R_L and R_g owing to the higher value of r_p.

The **high frequency response** [Sect. 2(xiii), eqn. (16)] will normally be lower than that for a triode, owing to the higher value of R'' due again to the higher plate resistance. The valve output capacitance is usually greater for a pentode than for a triode, this being a further contributing factor. The high frequency response may be extended to higher frequencies by reducing the resistance of the plate load resistor, although at the cost of gain. A similar result may be achieved by the use of negative voltage feedback.

The effect of the screen and cathode by-passing has been covered in (iii) and (iv) above.

(xiii) Screen loadlines

(A) Exact method using I_{c2} versus E_{c2} characteristics

These characteristics (Fig. 12.17) only apply to a single value of supply voltage and plate load resistance. The screen loadline is a straight line drawn from A to B.
Point A : $E_{c2} = E_{cc2}$ (here 250 volts), $I_{c2} = 0$
Point B : $E_{c2} = 0$, $I_{c2} = E_{cc2}/R_s$ (here $250/1.5 = 167$ μA).

It is then necessary to transfer the values of E_{c2}, at the intersections of the E_{c1} curves and the loadline, to a second curve in which E_{c1} is plotted against E_{c2}. From the latter, the values of E_{c1} corresponding to the screen voltages of the dynamic characteristics (Fig. 12.14A) are determined, and transferred directly to the plate dynamic characteristics, and the screen loadline drawn as a smooth curve (Fig. 12.19).

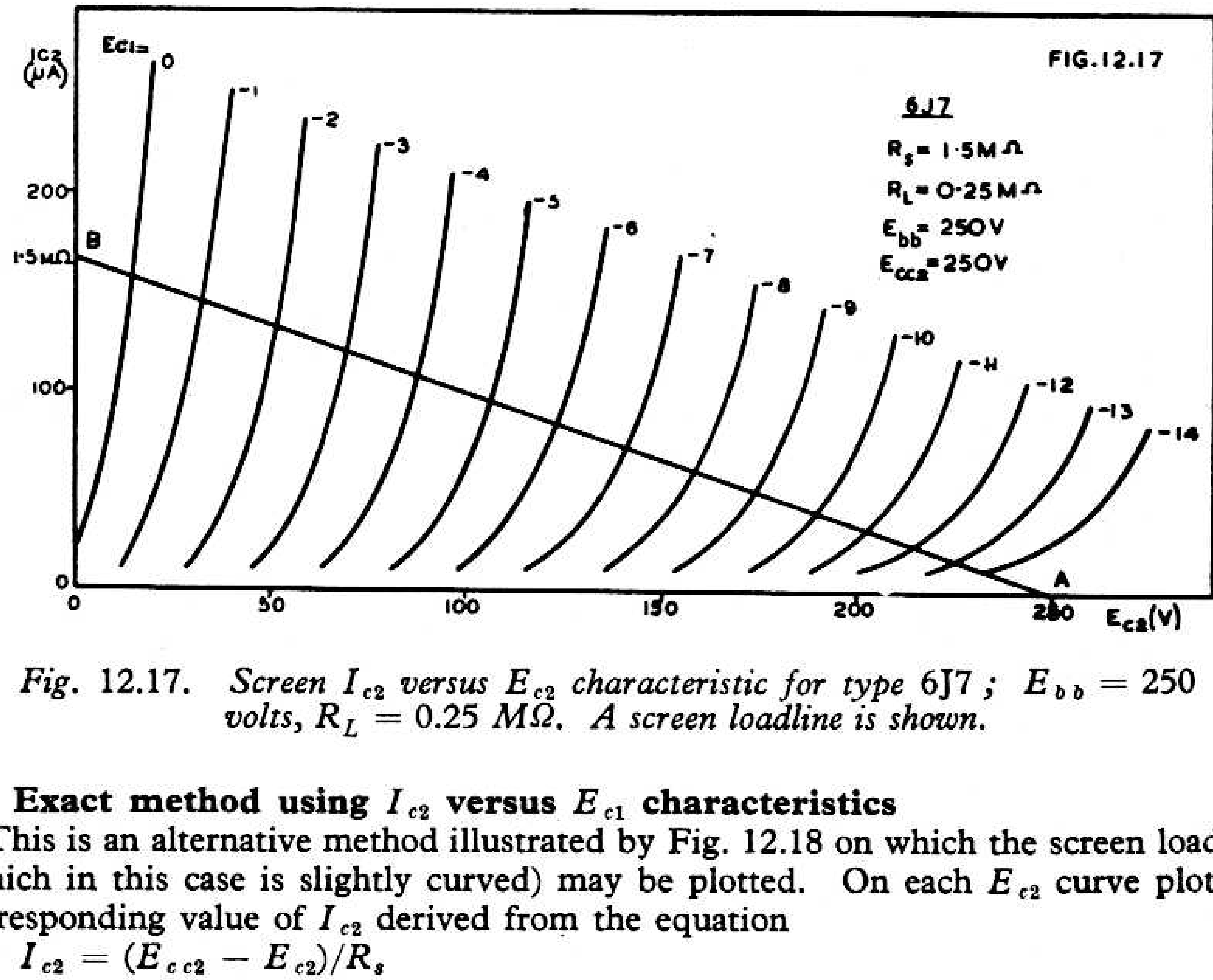


Fig. 12.17. *Screen I_{c2} versus E_{c2} characteristic for type* 6J7 ; $E_{bb} = 250$ *volts, $R_L = 0.25$ MΩ. A screen loadline is shown.*

(B) Exact method using I_{c2} versus E_{c1} characteristics

This is an alternative method illustrated by Fig. 12.18 on which the screen loadline (which in this case is slightly curved) may be plotted. On each E_{c2} curve plot the corresponding value of I_{c2} derived from the equation

$$I_{c2} = (E_{cc2} - E_{c2})/R_s \qquad (10)$$

Then draw a smooth curve (CD in Fig. 12.18) through the plotted points. The values of E_{c1} corresponding to the points of intersection are then transferred directly to the plate dynamic characteristic, and the loadline drawn as a smooth curve (Fig. 12.19).

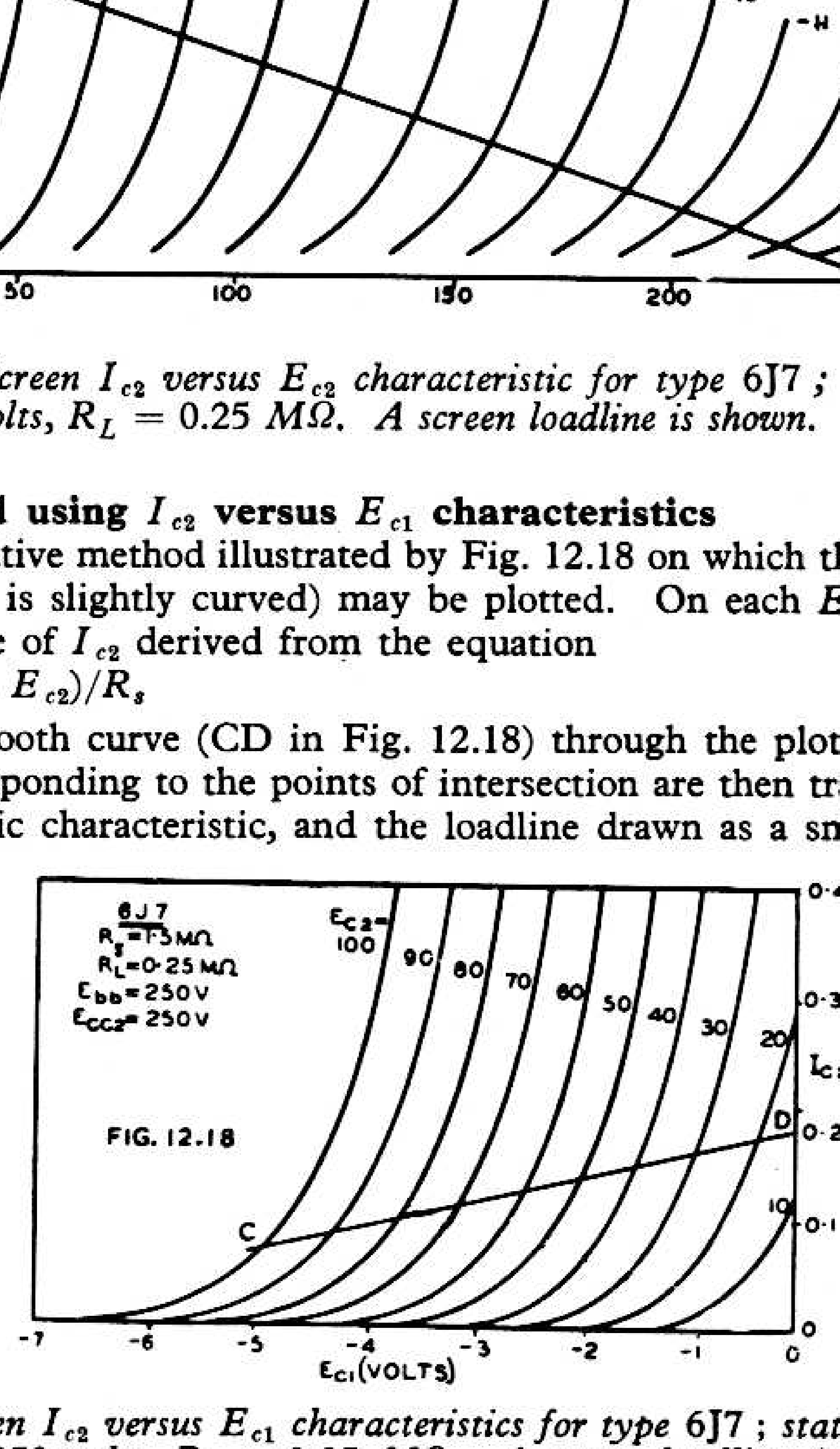


Fig. 12.18. *Screen I_{c2} versus E_{c1} characteristics for type* 6J7 ; *static curves are for $E_{bb} = 250$ volts, $R_L = 0.25$ MΩ. A screen loadline is shown.*

(C) Approximate method*

This is based on the assumption that the ratio of plate and screen currents is constant, such being only a rough approximation ; it is, however, good enough for many purposes. Using the conditions for Fig. 12.19 and assuming $I_b/I_{c2} = 4.0$ (as for type 6J7) and $R_s = 1$ megohm, we can draw up the table :

E_{c2}	$(E_{cc2} - E_{c2})$	I_{c2}	$I_b = 4I_{c2}$
100 V	150 V	0.15 mA	0.60 mA
80	170	0.17	0.68
60	190	0.19	0.76
40	210	0.21	0.84
20	230	0.23	0.92

The values of I_b are then plotted on the characteristics corresponding to the respective screen voltages (E_{c2}). The screen loadline (Fig. 12.19) is almost a straight line which may be extended to cut the horizontal axis at approximately

$$E_{c1} \approx -E_{cc2}/\mu_t \qquad (11)$$

and which may be shown to have a slope of approximately

$$m\mu_t g_{md}/(R_s g_{md} + m\mu_t) \qquad (12)$$

where μ_t = valve triode mu,
g_{md} = slope of plate dynamic characteristic at point of interest,
and m = ratio of plate to screen currents.

Alternatively, the point of intersection of the screen loadline with the horizontal axis may be determined by the point of cathode current cut-off on the "triode" characteristics (if available) where $E_b = E_{cc2}$.

The plate current at $E_{c1} = 0$ is given approximately by

$$I_{b0} \approx m g_{md} E_{cc2}/R_s g_{md} + m\mu_t \qquad (13)$$

The value of screen resistor to provide a plate current $I_b = KE_{bb}/R_L$ at a fixed grid voltage E_{c1} when $E_{cc2} = E_{bb}$ is given approximately by

$$R_s \approx \frac{m}{K}\left\{R_L - \frac{\mu_t R_L E_{c1}}{E_{bb}} - \frac{K\mu_t}{g_{md}}\right\} \qquad (14)$$

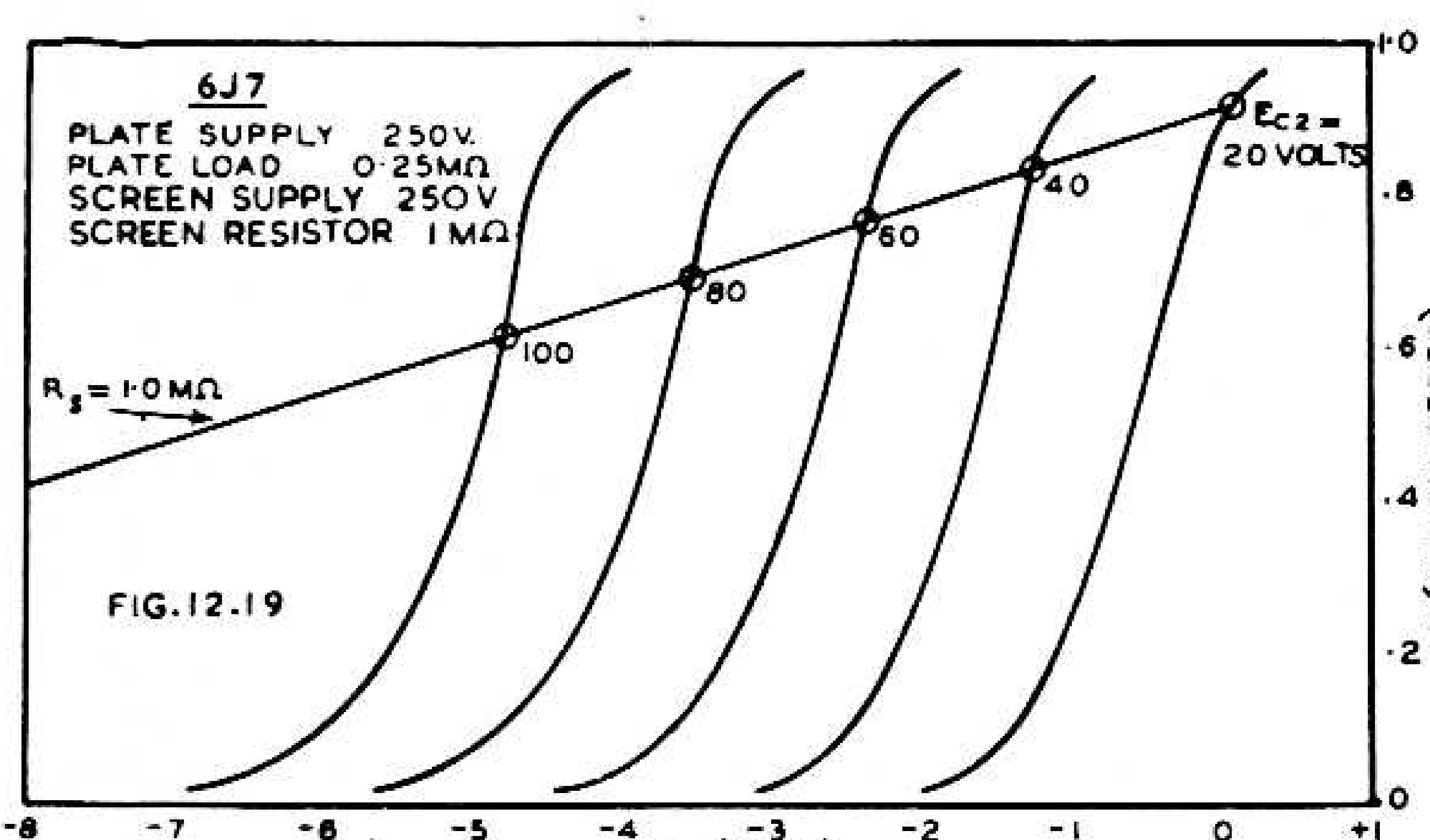


Fig. 12.19. *Plate dynamic characteristics for type* 6J7 *with screen loadline.*

Variations in ratio between plate and screen currents

Although considerable variations exist, particularly in the region of high plate current, there is a comparatively large region within which the ratio is constant within ± 5%, this being fortunately in the region most generally useful for amplification (e.g. shaded area in Fig. 12.20).

The variations which occur are mainly caused by the variations in the ratio of plate to screen voltages. It is clear that they will be serious for E_b less than E_{c2} (above broken line in Fig. 12.20) but the variations become greater as E_b approaches E_{c2}. If it is desired to make accurate calculations involving an assumed constant value of m (= I_b/I_{c2}) it is desirable to maintain E_b/E_{c2} nearly constant. The use

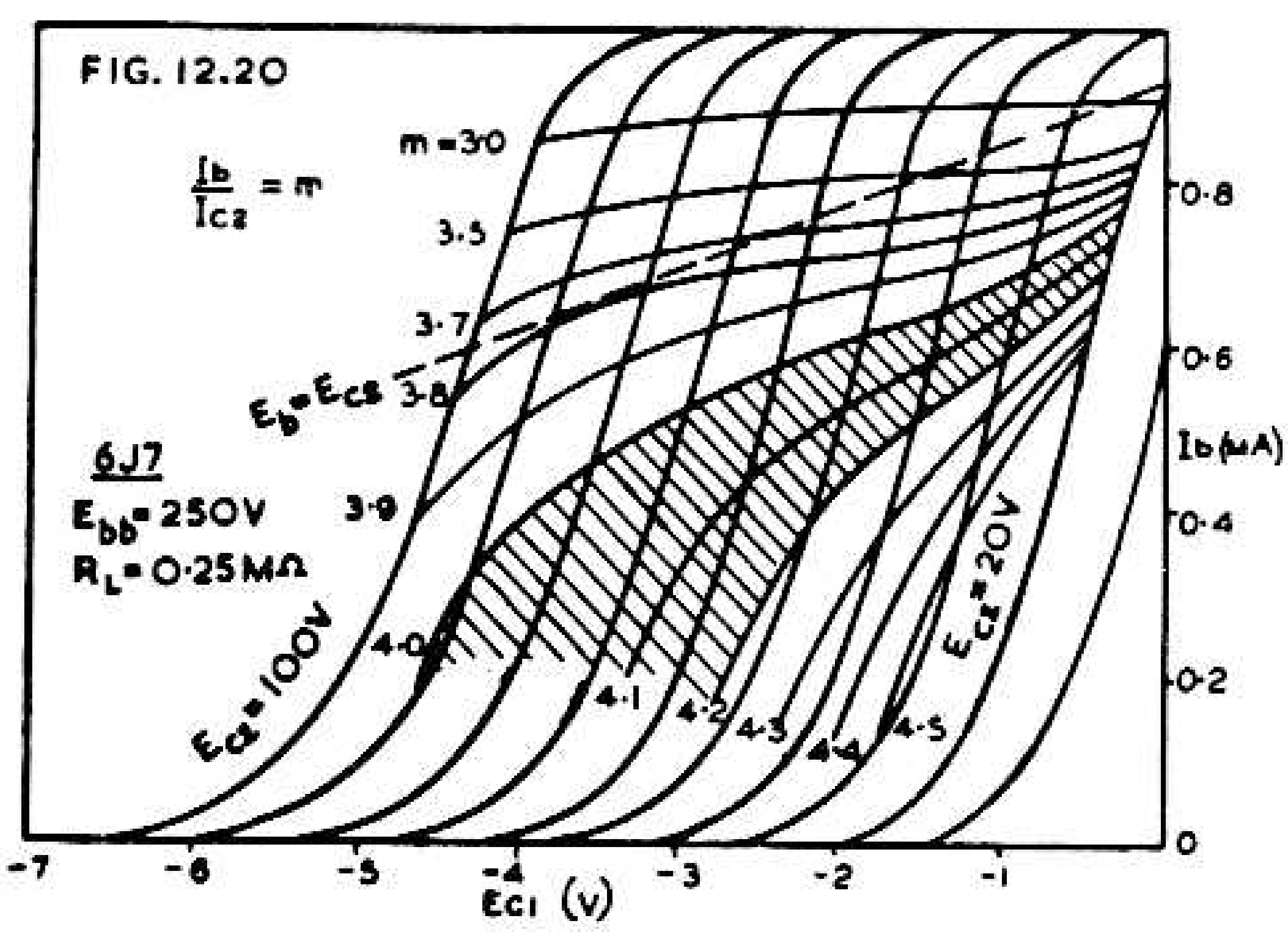
*A further approximate method is described in Ref. B12.

of a series screen resistor with a resistance equal to mR_s (or slightly above this value) assists in maintaining the constant current ratio.

(xiv) Combined screen and cathode loadlines and the effect of tolerances

(A) Cathode loadlines

The method normally adopted is an approximate one but very convenient, since it may be used directly with the plate dynamic characteristics. Provided that $I_b/I_{c2} = m =$ constant, the current through R_k will be $(m + 1) I_b$, so that it is necessary to use a " conversion factor " of $m/(m + 1)$ in respect to both the current and slope of the loadline. For example if $m = 4$, the loadline for a cathode bias resistor of 2000 ohms would have a slope of $- 1/2500$ mhos. The effect of any error in the value of m is minimized through the screen current being only a small fraction of the cathode current.



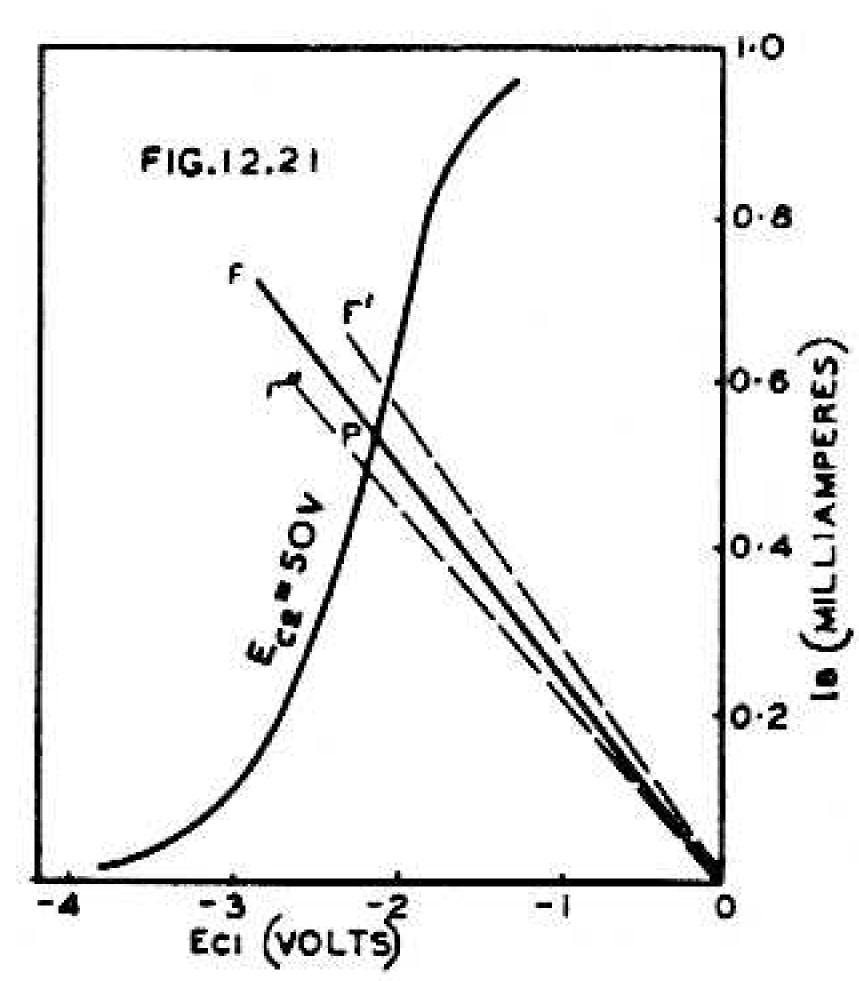


Fig. 12.20. *Plate dynamic characteristics for type* 6J7 *with curves of constant plate to screen current ratio. The shaded area has m within* $\pm$ 5%.
Fig. 12.21. *Cathode loadline (OF) with single plate dynamic characteristic for type* 6J7. *The broken lines show the limits of* $\pm$ 10% *tolerances in* R_k.

The method is illustrated in Fig. 12.21 where only one plate dynamic characteristic is shown, with a cathode bias loadline OF having a slope of $- m/(m + 1)R_k$. The static operating point is at P, the intersection of OF with the dynamic characteristic. Owing to the very large plate load resistors commonly employed, there is very little rise in plate current due to rectification effects, so that P may be regarded as the dynamic operating point.

The effect of $\pm$ 10% variation in R_k is illustrated by lines OF″ and OF′ respectively in Fig. 12.21.

It may readily be shown geometrically that $\pm$ 10% variation in R_k has less effect on I_b than $\pm$ 10% variation in E_{c1}.

(B) Cathode and screen loadlines and tolerances

When both cathode and screen resistors are used, the point of operation is the intersection of the two loadlines (Fig. 12.22). In this diagram it has been assumed that the screen loadline is a straight line, this being closely correct except at very low and high values of I_b. However, any practicable method for determining the screen loadline may be used—see (xiii) above.

The effect of $\pm$ 10% variation in the resistance of R_s is indicated by the broken lines, and the combined effect with $\pm$ 10% variation of both R_s and R_k is indicated by the region shown shaded. It is on account of these inevitable variations, together with valve variations, that it is inadvisable to operate with too low a nominal screen voltage and consequently very close to the grid current point.

(C) Tolerances in general

All three resistors R_L, R_s and R_k have partially self-compensating effects which enable fairly satisfactory results to be obtained with ordinary 10% tolerances in the resistors. If operation is required for minimum harmonic distortion at low level or

FIG. 12.22

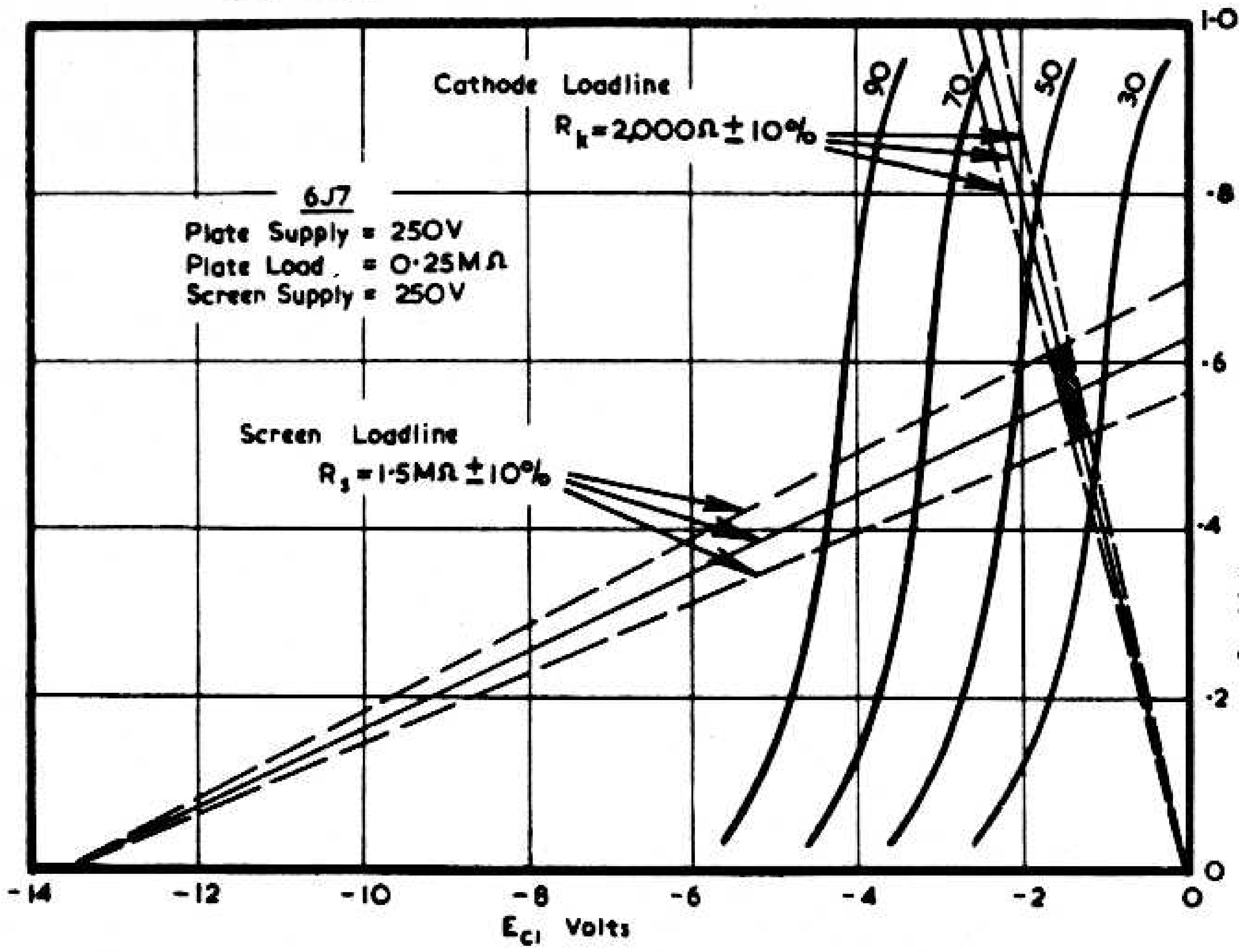


Fig. 12.22. *Plate dynamic characteristics of type* 6J7 *with screen and cathode loadlines, showing the effect of* ± 10% *tolerances in both* R_s *and* R_k.

for a high output voltage, the resistors may beneficially have closer tolerances or be selected so that high R_L, high R_s and high R_k go together, and similarly with low values.

The supply voltage may be varied by a ratio up to 2 : 1 in either direction without very serious effects in normal operation and without necessarily making any change in R_L, R_s or R_k.

(xv) Remote cut-off pentodes as r.c.c. amplifiers

Remote cut-off (" variable-mu ") pentodes may be also used satisfactorily as r.c.c. amplifiers, although they give higher distortion for the same voltage output, and should therefore be restricted to low level operation. However, they are quite satisfactory in ordinary radio receivers for the first a-f stage driving an output pentode. Operation is slightly more critical than with sharp cut-off pentodes, but a value of $K = 0.78$ is reasonably close for low level operation.

As a typical example, intermodulation distortion on type 6SF7 is tabulated for $E_{bb} = 250$ volts.

R_L	R_g	$E_0 = 10$	19	37	63 volts r.m.s.
0.25 MΩ	0.25 MΩ	1.8%	5.5%	17%	—
	0.5	1.5%	4.0%	13.5%	52%
	1.0	1.2%	3.0%	10.5%	34%
	Optimum value of K				
0.25 MΩ		0.8	0.78	0.73	0.63

(xvi) Multigrid valves as r.c.c. amplifiers

All types of multigrid valves may be used as r.c.c. amplifiers, and behave effectively the same as pentodes provided that the unused grids are connected to suitable fixed voltages. Operating conditions, including suitable values of K, are the same as for pentodes.

(xvii) Special applications

1. An article giving useful information on 28 volt operation is Ref. B4.

2. An article describing a special pentode operating as two triodes in cascade, giving gains up to 500 with a 45 volt plate supply, is Ref. B6.

(xviii) Comments on tabulated characteristics of resistance-coupled pentodes

Comments are as for triodes—Sect. 2(xiv). See Sect. 3(v), (vi)B and E for optimum operating conditions.

SECTION 4: TRANSFORMER-COUPLED VOLTAGE AMPLIFIERS

(i) Introduction (ii) Gain at the mid-frequency (iii) Gain at low frequencies (iv) Desirable valve characteristics (v) Equivalent circuits (vi) Gain and phase shift at all frequencies (vii) Transformer characteristics (viii) Fidelity (ix) Valve load-lines (x) Maximum peak output voltage (xi) Transformer loading (xii) Parallel feed (xiii) Auto-transformer coupling (xiv) Applications (xv) Special applications.

(i) Introduction

Transformer-coupled voltage amplifiers usually employ general purpose triode valves with plate resistances about 6 000 to 10 000 ohms. Valves having higher plate resistance require excessively large transformer inductances, while valves having lower plate resistances are only used in special applications, with transformers designed to handle the higher plate currents.

A typical transformer-coupled amplifier stage is shown in Fig. 12.23 where V_1 is the valve under consideration, with transformer T coupling it to the grid of the following valve (V_2). The stage gain is the voltage gain from the grid of V_1 to the grid of V_2. Push-pull transformers are covered in Section 6. Read also Chapter 5 Sections 1, 2 and 3.

FIG. 12.23

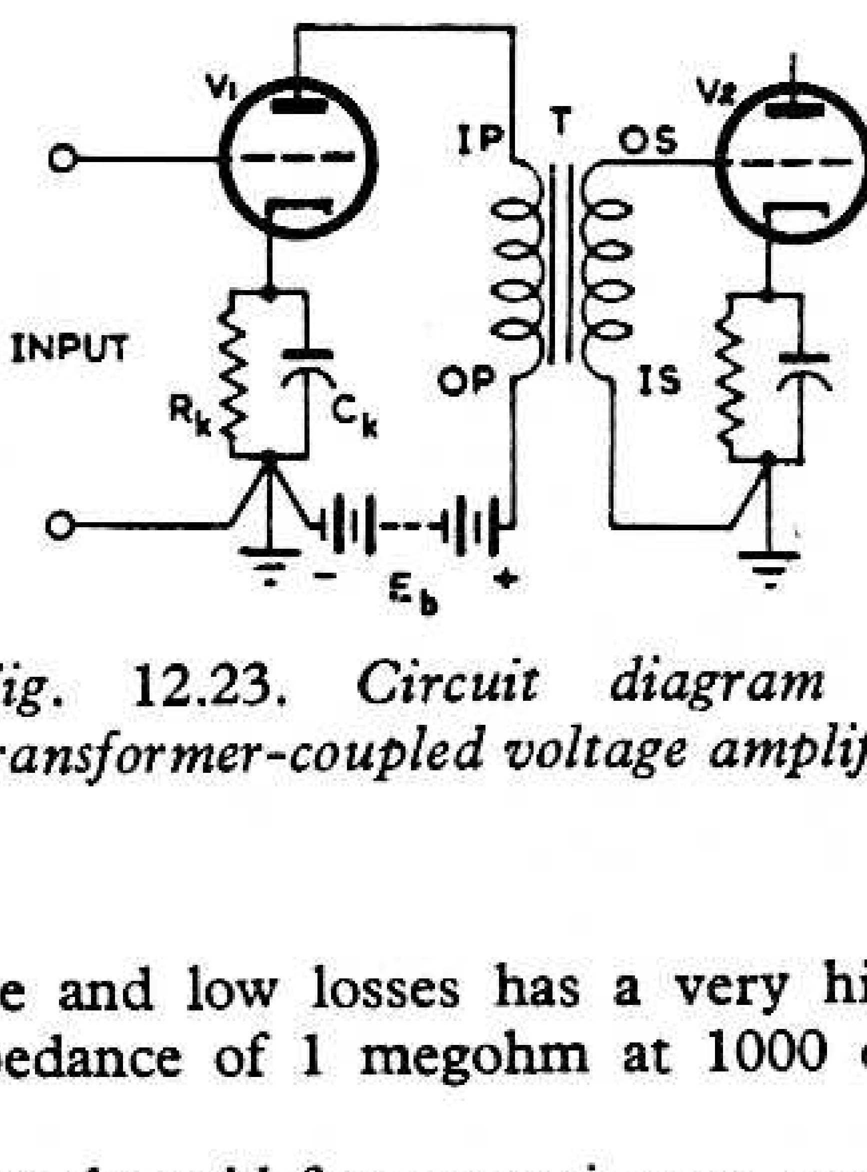


Fig. 12.23. Circuit diagram of transformer-coupled voltage amplifier

(ii) Gain at the mid-frequency

An unloaded transformer with high inductance and low losses has a very high impedance at the mid-frequency; an input impedance of 1 megohm at 1000 c/s is not uncommon.

With the circuit of Fig. 12.23, the voltage gain at the mid-frequency is very nearly

$$A_0 = \mu T \tag{1}$$

where μ = amplification factor of valve
T = turns ratio = N_2/N_1
N_1 = primary turns
and N_2 = secondary turns.

The gain with loaded transformers is dealt with in (xi) below.

(iii) Gain at low frequencies

At low frequencies the gain is reduced by the shunting effect of the primary inductance. The gain relative to that at the mid-frequency (A_0) is given in Chapter 5 Sect. 3(iii)a particularly eqn. (1).

(iv) Desirable valve characteristics

(a) Low plate resistance—gives better bass response.
(b) High amplification factor—gives higher gain.

(c) Low operating plate current—reduces the direct current through the transformer primary and thus increases the effective inductance compared with a valve having a higher plate current. If necessary, any valve may be overbiased so as to bring the plate current to a suitable value—5 mA is quite usual.

(d) High maximum plate voltage—only desirable if a high input voltage is required by the following stage (e.g. push-pull low mu triodes, or cathode follower).

(v) Equivalent circuits

See Chapter 5 Sect. 2, particularly Fig. 5.9, the input terminals of which are to be understood as being connected to a generator μe_g in series with r_p.

(vi) Gain and phase shift at all frequencies

A transformer tends to produce a peak in the gain/frequency characteristic due to the series resonance between the total leakage inductance and the distributed capacitance of the secondary. This peak may be made negligible by ensuring that

$$Q_s = \omega_1 L'/[r_1 + (r_2/T^2) + r_p] \leq 0.8 \qquad (2)$$

where Q_s = Q of resonant circuit in secondary

ω_1 = $2\pi \times$ frequency of resonance = $\sqrt{1/L'C_2T^2}$

L' = leakage inductance referred to primary

C_2 = secondary equivalent shunt capacitance

r_1 = d.c. resistance of primary

r_2 = d.c. resistance of secondary

and T = turns ratio = N_2/N_1.

Irrespective of the value of Q_s, there is a 90° phase angle shift (lagging) at the frequency of resonance which extends asymptotically to 180° at infinite frequency. The rate of change is more gradual for lower values of Q_s. It is on account of this phase shift that transformer coupling is avoided as far as possible with negative feedback, and that (when unavoidable) the secondary is heavily damped to give a low Q_s.

At low frequencies a transformer-coupled amplifier has the same attenuation and phase angle shift (leading) characteristics as a r.c.c. amplifier due to its grid coupling condenser alone (see Fig. 12.9).

See Refs. A12 (pp. 366-371 and curves Fig. 13) ; A13 (pp. 28-38).

(vii) Transformer characteristics

See Chapter 5 Section 3.

(viii) Fidelity

Distortion caused by the valve is usually very low in the vicinity of the mid-frequency with unloaded transformers of high impedance—about 1% to 2% second harmonic with maximum grid swing—but it increases as the transformer is loaded, up to about 5% second harmonic and 2% or 3% third harmonic. At frequencies below 1000 c/s the impedance steadily falls, while the loadline opens out into a broad ellipse at low frequencies. The valve distortion may be considerable at high output levels under these conditions [see Chapter 2 Sect. 4(vi)].

The transformer (core) distortion is usually quite appreciable and may cause serious intermodulation distortion [see Chapter 5 Sect. 3(iii)].

(ix) Valve loadlines

The loadline for the mid-frequency approximates to a resistive loadline [see Chapter 2 Sect. 4(i) and (ii)]. The procedure is identical to that for Fig. 2.23 except that the loadline AQB is at a smaller angle to the horizontal.

(x) Maximum peak output voltage

To determine the maximum peak output voltage for the mid-frequency, refer to the published curves (say type 6J5), determine I_b (say 5 mA) and E_b (say 250 V) ; mark Q at $E_b = 250$, $I_b = 5$ mA ; draw a horizontal loadline through Q ; determine E_{b0} for the grid current point ($E_c = -0.5$ V) on the loadline—here $E_{b0} = 65$ volts.

The maximum peak output voltage in this example is $(E_b - E_{b0})T = (250 - 65)$ $T = 185$ T volts $= 0.74\, E_b T$. This is reduced as the load becomes less, and a value of about $0.65\, E_b T$ is fairly typical.

(xi) Transformer loading

A resistor shunted across the secondary reduces the value of Q_s, thereby giving more uniform high frequency response, and also extends the low frequency response. A resistor shunted across the primary extends the low frequency response. If f_1 is the low frequency giving a specified attenuation with an unloaded transformer, then the same attenuation is reached at $0{\cdot}5 f_1$ when the total shunt load effective across the primary is equal to r_p, and at $(2/3)\, f_1$ with a shunt load of $2r_p$.

Loading results in higher valve distortion and secondary loading also results in higher transformer distortion. Loading is generally undesirable, although unavoidable in some applications ; it reduces the maximum peak output voltage and the gain.

The effective load on the valve is given by $R_L = R_1 + R_2/T^2$

where R_1 = resistance shunted across the primary
R_2 = resistance shunted across the secondary
and $T = N_2/N_1$ = turns ratio.

(xii) Parallel feed

Parallel feed (Fig. 12.24) avoids the direct plate current passing through the transformer primary and thereby increases its effective inductance, and decreases transformer distortion. Owing to the drop in R_L the average plate voltage will be con-

FIG. 12.24

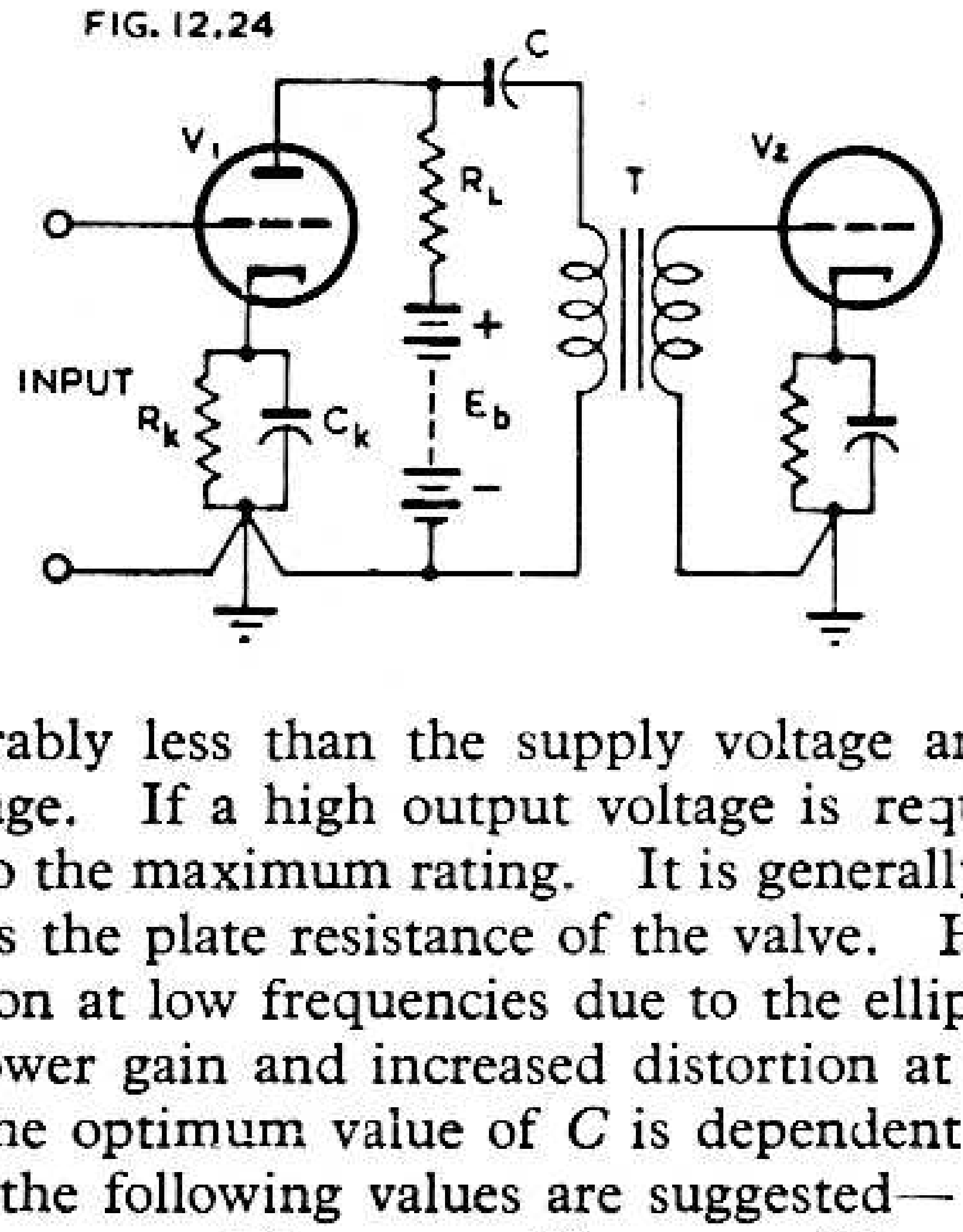


Fig. 12.24. Circuit diagram of parallel-fed transformer-coupled voltage amplifier.

siderably less than the supply voltage and this may restrict the maximum output voltage. If a high output voltage is required, the supply voltage may be increased up to the maximum rating. It is generally desirable to make R_L a resistance of 3 or 4 times the plate resistance of the valve. Higher values of R_L result in increased distortion at low frequencies due to the elliptical load-line. Lower values of R_L result in lower gain and increased distortion at all frequencies.

The optimum value of C is dependent upon the transformer primary inductance, and the following values are suggested—

L =	10	20	30	50	100	150 henrys
C =	4.0	2.0	2.0	1.0	0.5	0.5 μF.

These values of capacitance are sufficiently high to avoid resonance at an audible frequency. Use is sometimes made of the resonance between C and the inductance of the primary to give a certain degree of bass boosting. By this means a transformer may be enabled to give uniform response down to a lower frequency than would otherwise be the case. It should be noted that the plate resistance of the valve, in parallel with R_L, forms a series resistance in the resonant circuit. The lower the plate resistance, the more pronounced should be the effect [see Chapter 15 Sect. 2(iii)C].

It is frequently so arranged that the resonant frequency is sufficiently low to produce a peak which is approximately level with the response at middle frequencies, thereby avoiding any obvious bass boosting while extending the frequency range to a maximum.

For mathematical analysis see Ref. A13 pp. 38-41.

In all cases when making use of any resonance effects involving the inductance of the transformer primary, it is important to remember that this is a variable quantity.

Not only are there considerable variations from one transformer to another, but there are large variations of inductance caused by the a.c. input voltage (see Chapter 5).

The series resonant circuit presents a low impedance to the valve at the resonant frequency, thus tending to cause serious distortion, particularly when the valve is being operated at a fairly high level. Fot these reasons the resonance method is not used in good design.

(xiii) Auto-transformer coupling

An " auto-transformer " is a single tapped inductance which is used in place of a transformer. Fig. 12.25 shows a parallel-fed auto-transformer coupled amplifier. The auto-transformer may be treated as a double-wound transformer (i.e. with separate primary and secondary) having primary turns equal to those between the tap and earth, and secondary turns equal to the total turns on the inductance. A step-up or step-down ratio may thus be arranged. An ordinary double-wound transformer may be connected with primary and secondary in series (with sections aiding) and used as an auto-transformer, but capacitance effects between windings may affect the high-frequency response of certain types of windings.

The inductance between the tapping point and earth should be the same as for a normal transformer primary. With the parallel-fed arrangement of Fig. 12.25 the plate current does not flow through the inductance, but an alternative arrangement is to omit the parallel-feed and to add a grid coupling condenser and grid resistor for V_2.

FIG. 12.25

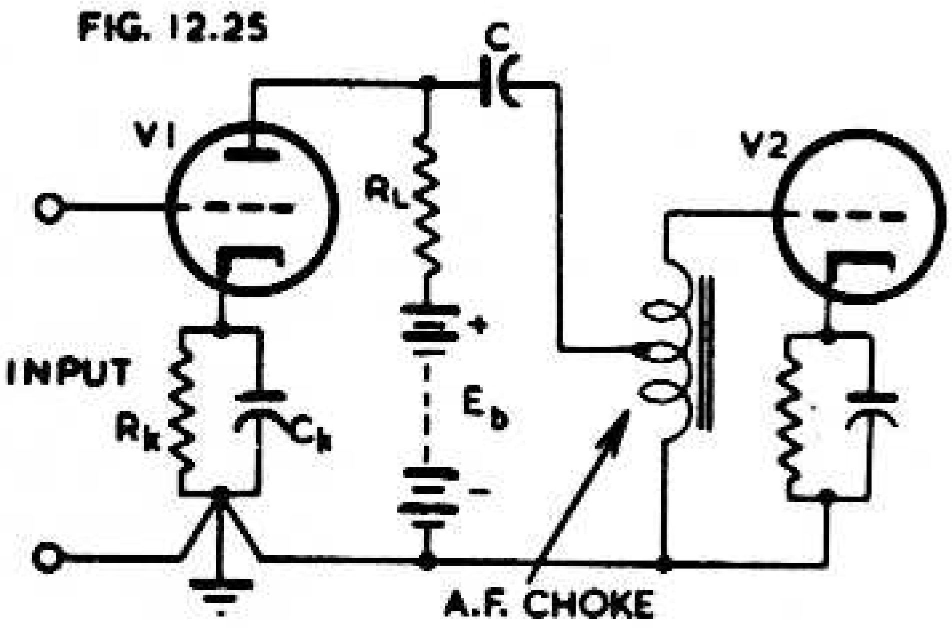


FIG. 12.26

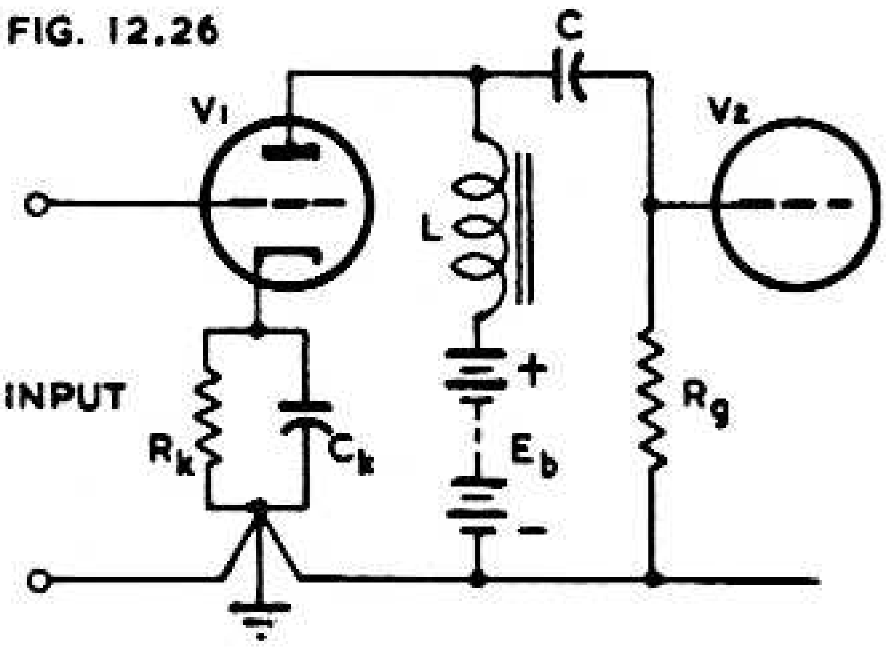


Fig. 12.25. *Circuit diagram of parallel-fed auto-transformer coupled voltage amplifier.*
Fig. 12.26. *Choke-capacitance coupled voltage amplifier.*

(xiv) Applications

The cost of a transformer having linear response over a wide frequency range is considerable and, since equally good response may generally be obtained by a very simple resistance-coupled amplifier, the transformer is only used under circumstances where its particular advantages are of value. Some of these are—

(1) High output voltage for limited supply voltage,

(2) Stepping up from, or down to, low-impedance lines,

(3) When used with split or centre-tapped secondary for the operation of a push-pull stage, and

(4) When a low d.c. resistance is essential in the grid circuit of the following stage.

(xv) Special applications

(a) **Cathode loading** : V_1 in Fig. 12.25 may be connected as a cathode follower either with the transformer primary in the cathode circuit or with parallel feed (see Chapter 7 Sect. 2(i)G).

(b) **28 volt operation** : see Reference B4.

SECTION 5 : CHOKE-COUPLED AMPLIFIERS

(i) Performance (ii) Application.

(i) Performance

A typical choke-capacitance coupled amplifier is shown in Fig. 12.26. The operation and design are similar to those of a transformer-coupled amplifier (Fig. 12.23) with a transformer ratio 1 : 1 except that C must be designed as in a r.c.c. amplifier to avoid additional low frequency attenuation. An amplifier of this type produces a higher maximum output voltage than a r.c.c. amplifier but less than that with a step-up transformer.

(ii) Application

It is occasionally used with valves having rather high plate current, for which a suitable transformer may not be available. It was also used with tetrode valves of old design, the inductance being several hundred henries, shunted by a resistance of about 0.25 megohm.

SECTION 6 : METHODS OF EXCITING PUSH-PULL AMPLIFIERS

(i) Methods involving iron-cored inductors (ii) Phase splitter (iii) Phase inverter (iv) Self-balancing phase inverter (v) Self-balancing paraphase inverter (vi) Common cathode impedance self-balancing inverters (vii) Balanced output amplifiers with highly accurate balance (viii) Cross coupled phase inverter.

Normally we begin with a single-sided amplifier, and then at some suitable level a stage may be inserted having a single input and a push-pull output. In a radio receiver such a stage usually immediately precedes the output stage, but in more ambitious amplifiers there may be several intermediate push-pull stages. In this section we consider the methods of exciting push-pull amplifiers.

(i) Methods involving iron-cored inductors

(A) **Tapped secondary transformer** (Fig. 12.27A)

This needs little explanation except that the input transformer step-up ratio (primary to half secondary) does not usually exceed 1 : 2. The fidelity is largely dependent upon the quality of the transformer. This method may be used with almost any type of amplifier, and the arrangement illustrated is merely typical. For example, fixed bias operation, or operation with triodes in place of pentodes could equally well be adopted.

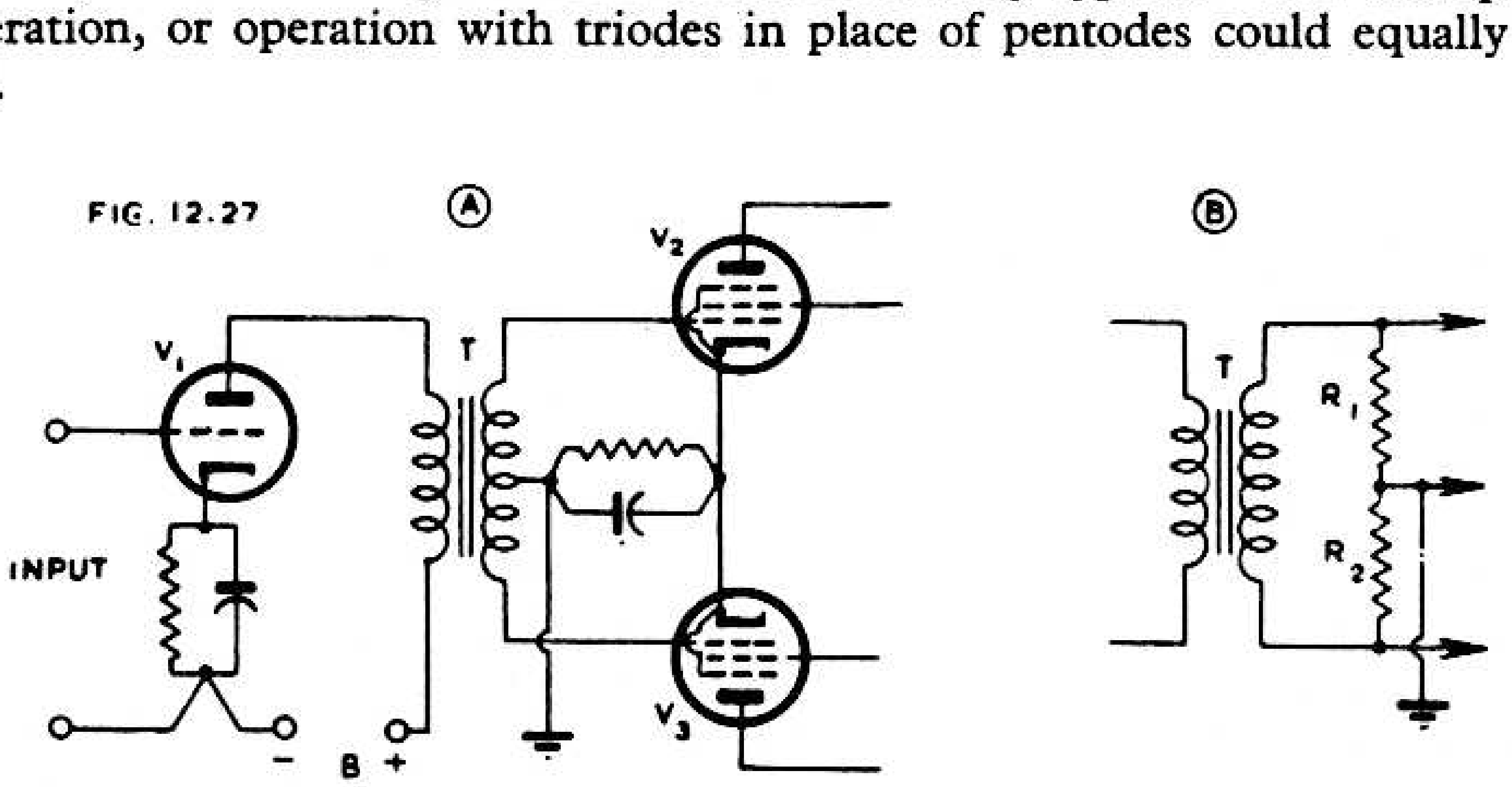


Fig. 12.27. *(A) Triode valve (V_1) followed by transformer (T) having centre-tapped secondary and exciting the grids of the push-pull stage (V_2, V_3). (B) Modified form with non-centre-tapped transformer, using two resistors to provide an equivalent centre-tap.*

(B) Centre-tapped resistor across secondary (Fig. 12.27B)

This is an alternative arrangement which does not require centre-tapping of the transformer secondary. In this case an ordinary transformer with a single secondary winding is used, and a centre-tapped resistor ($R_1 = R_2$) is connected across the secondary. The resistors cause a load to be reflected into the plate circuit of V_1 which is equal to $(R_1 + R_2)/T^2$ where $T = N_2/N_1$. For example if $R_1 = R_2 = 0.1$ megohm, and $T = 3$, the load reflected into the plate circuit will be 200 000/9 = 22 000 ohms. This load is lower than some triodes are capable of handling without noticeable distortion, and it might be necessary to increase R_1 and R_2, the limit being set by the maximum grid circuit resistance permitted with valves V_2 and V_3. In addition, this arrangement gives greater distortion than the centre-tapped transformer when the valves are slightly over-driven and pass grid current.

(C) Centre-tapped choke (Fig. 12.28)

This is an alternative method sometimes used, involving parallel feed. The effective voltage gain from primary to half secondary is 1/0.5, indicating a loss as compared with a transformer. It is difficult to obtain perfect balance between the two sides.

(D) Choke-coupled phase inverter—see page 355.

(ii) Phase splitter

(A) This is an excellent method which is also self-balancing. Its principal characteristics are given in Chapter 7 Sect. 2(ii)B and Fig. 7.25. Perfect balance* is obtained at low frequencies if the plate and cathode load resistors are equal, while commercial tolerances in the following grid resistors have only a small effect on the balance. General purpose triode valves are normally used, and the gain is about 0.9 to each side of the output. The input resistance is of the order of 10 megohms, and the harmonic distortion is extremely low.

FIG. 12.28

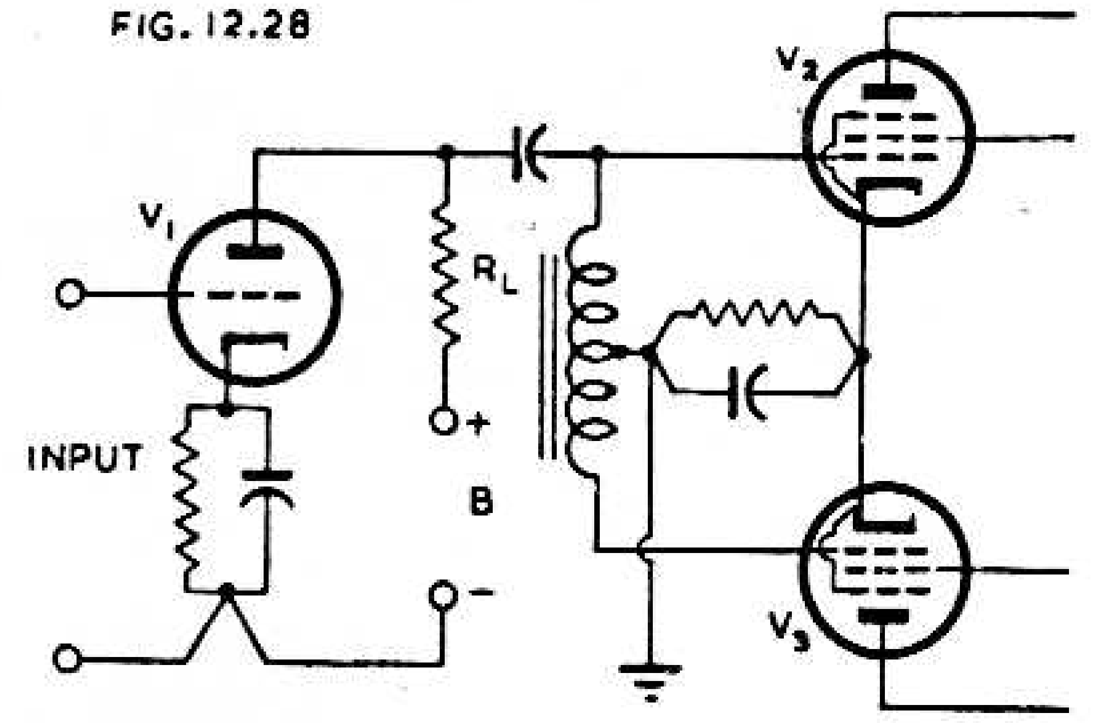


FIG. 12.29 A

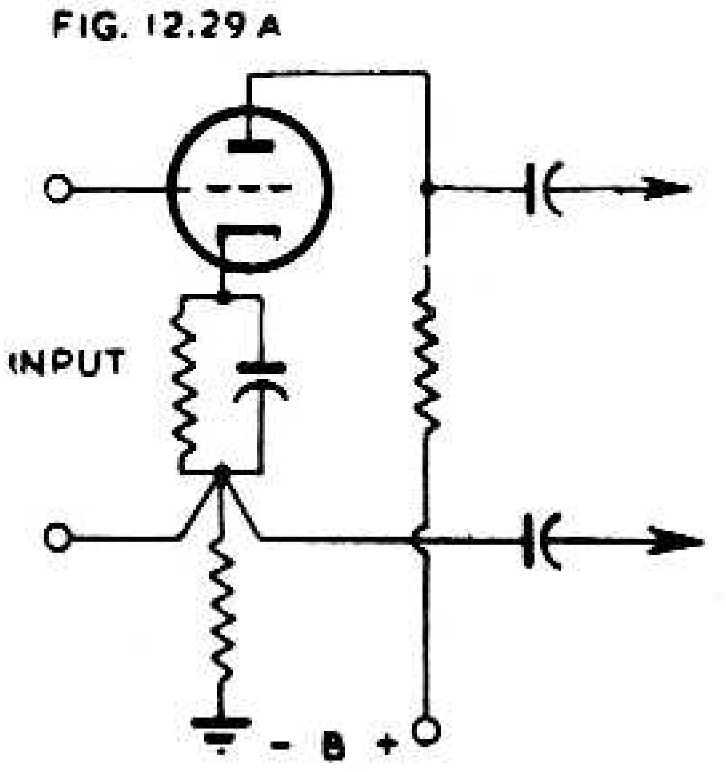


Fig. 12.28. *Triode valve (V_1) with parallel feed and centre-tapped choke exciting the grids of the push-pull stage.*

Fig. 12.29A. *Modified form of phase splitter giving full gain without degeneration. The input is from grid to cathode and cannot be earthed. The cathode resistor is by-passed to provide full gain.*

An analysis of the balance with equal load resistors and shunt capacitors is given in Ref. C27, indicating that under these conditions the balance is perfect at all frequencies. However, in practice, the total shunt capacitances across the two channels differ slightly and there is a slight (and generally negligible) unbalance at high frequencies; see also page 330.

If a high gain amplifier is placed between the phase splitter and the output stage, hum may be troublesome. Part of the hum is due to the difference of potential between the heater and cathode. This may be reduced by operating the heater of the phase splitter from a separate transformer winding which may be connected to a suitable point in the circuit at a potential approximating that of the cathode.

It is normal practice to assume a maximum r.m.s. output voltage (grid-to-grid)

*In an experimental test it is important to reverse the phase of the input from the B.F.O. when the valve voltmeter is moved from one output to the other.

of $0.18E_{bb}$ for less than 2% total harmonic distortion, equivalent to 0.25 E_{bb} peak-to-peak output. The distortion drops rapidly as the output voltage is decreased. With a plate supply voltage of 400 volts to the phase splitter, the output is sufficient to excite push-pull Class A 2A3 valves operating with 250 volts on their plates and $-$ 45 volts bias.

A complete circuit with 3 stages incorporating a phase splitter and negative feedback is shown in Fig. 7.42.

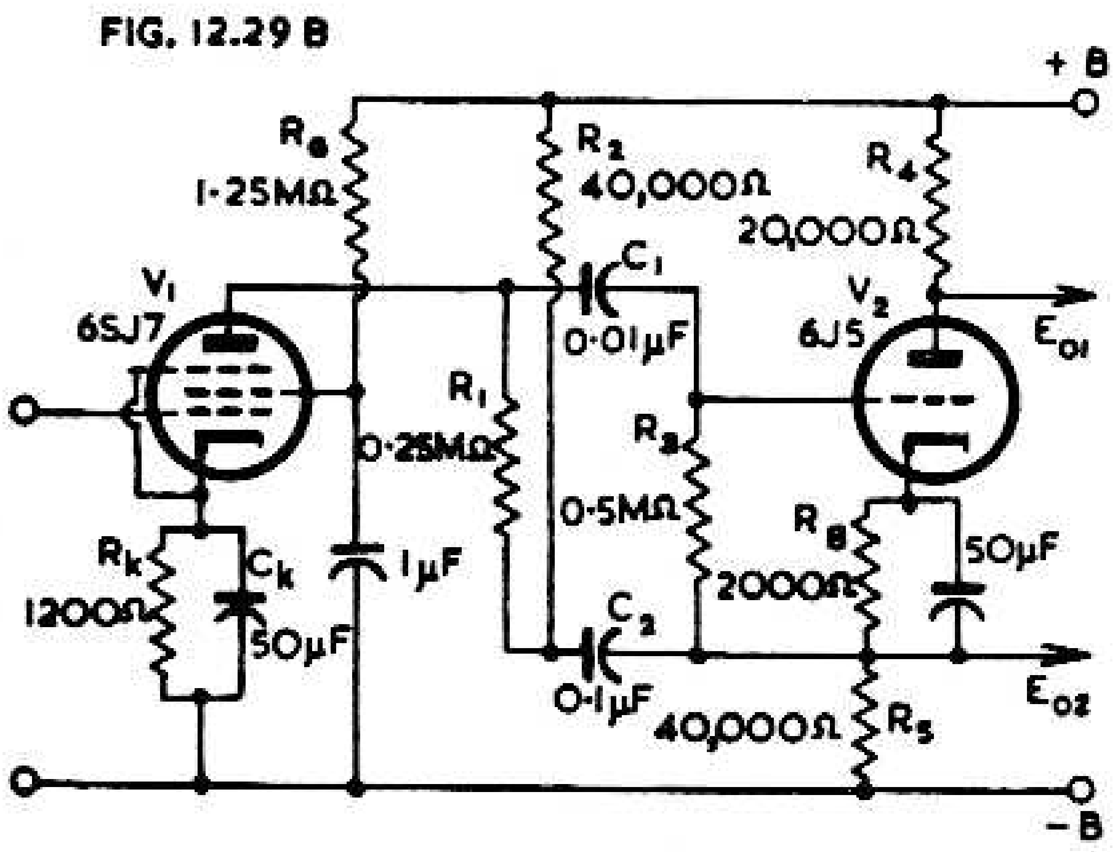
(B) **A modified form of phase splitter** (Fig. 12.29A) gives the full gain without degeneration, but the input is floating and cannot be earthed. For this reason it cannot generally be used with a pickup, although it may be applied to a radio receiver. In the latter application, the valve may be a duo-diode triode performing detection and 1st a-f stage amplification ; a.v.c. may be operated with some complication. This circuit is particularly prone to suffer from hum, owing to the high impedance from cathode to earth and the high gain. The hum may be minimized by adjusting the potential on the heater to approximately that of the cathode.

References (C) 1, 3, 7, 11, 12.

(C) **It is possible to apply positive feedback** from a tapping on the cathode resistor, through a coupling resistor to the unbypassed cathode of the preceding r.c. pentode, and thus increase the gain. A direct-coupled version is shown in Fig. 7.51A. See Chapter 7 Sect. 2(xi).

(D) **Another modification** (Fig. 12.29B) also gives high gain not from the phase splitter itself but from the preceding stage. It makes use of the high input resistance of the phase splitter as the dynamic load on V_1, thereby increasing its gain. The full analysis of its operation is given in Reference C18 and the following is a summary

The effective cathode load on V_2 (apart from the cathode resistor R_8) is R_2 and R_5 in parallel, i.e. 20 000 ohms, which is the same as R_4. The cathode resistor R_8 is by-passed in order to increase the input resistance.



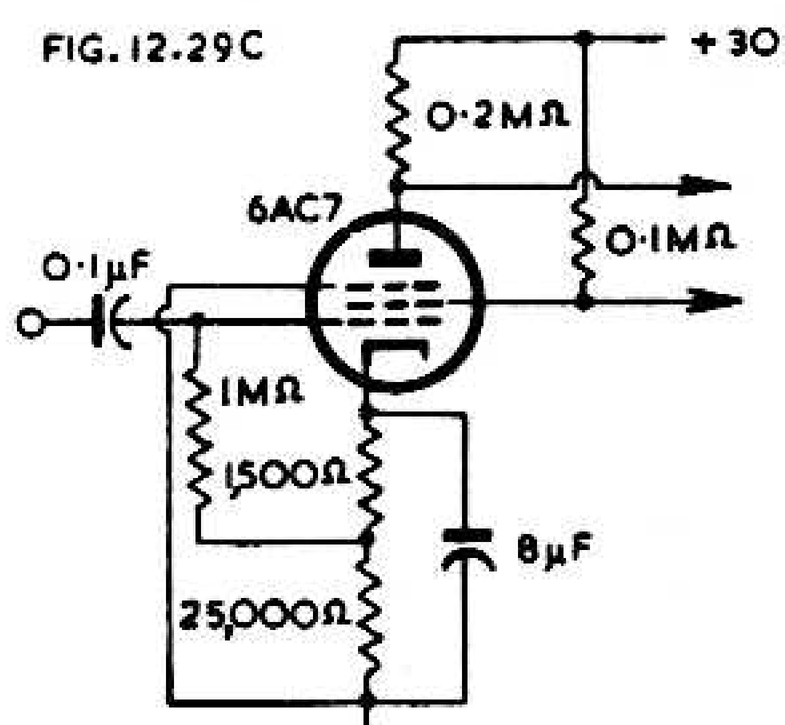


Fig. 12.29B. *Phase splitter* (V_2) *using high input impedance to increase the gain of the preceding stage* (V_1) *by about six times.*

Fig. 12.29C. *Phase splitter using pentode with unbypassed screen and suppressor grid.*

The d.c. load resistance in the plate circuit of V_1 is $R_1 + R_2 = 0.29$ megohm, while the dynamic load is the input resistance of V_2, i.e. [see Chapter 7 Sect. 2(ii)B]

$$r_i' = R_g/(1 - A')$$

where $R_g = R_1R_3/(R_1 + R_3) = 167\,000$ ohms

and $A' \approx 0.9$,

so that in this example, $r_i' = 1.67$ megohms.

The gain of V_1 is given by

$$A_1 = \frac{g_m}{(1/r_p) + (1/r_i')} = \frac{950}{1/4 + 1/1.67} = 1120,$$

which is about 6 times the gain under normal conditions.

The circuit is nearly balanced if $R_4 = R_c$ and

$$\frac{E_{01}}{E_{02}} = \frac{\mu_2 R_g - R_c}{\mu_2 R_g + r_{p2} + R_c}$$

where $R_c = R_2R_5/(R_2 + R_5)$.

The out-of-balance, being about 1.2% in this example, is negligible.

(E) **A further type makes use of a heptode (mixer) valve** in which unequal load resistors are placed in both plate and screen circuits, the push-pull output being taken from plate and screen. The input is taken to grid No. 3.

Ref. C20.

(F) **A further modification makes use of a pentode with an unbypassed screen.** The two output voltages are taken from plate and screen, and the suppressor grid is maintained at a negative potential with respect to the cathode (Fig. 12.29C and Ref. C23).

(iii) Phase inverter (Fig. 12.30A)

This is a popular arrangement with twin triode valves, either general purpose or high-mu. It is not self-balancing, and requires individual adjustment for accurate balance both during manufacture and after the valve has been replaced. It is slightly out of balance at very low frequencies owing to the two coupling condensers operating in the lower channel, but C_2 may be made larger than C_1 if desired. It gives a gain (to each channel) equal to the normal gain of one valve.

If it is preferred to avoid individual balancing, the value of R_2 is given by $R_2 = (R_1 + R_2)/A$ where A is the voltage gain of valve V_2. If R_1 and R_2 both have $\pm$ 10% tolerances, the maximum possible out of balance will be nearly 20% due to the resistors alone, plus valve voltage gain tolerances.

Separate cathode resistors, each by-passed, are helpful in reducing valve gain tolerances, but require independent cathodes. If a common cathode resistor is used, it may be unbypassed, thus introducing negative feedback for out-of-balance voltages. The hum level is quite low.

N.B. This circuit was originally named Paraphase, but the latter name covers a large number of different circuit arrangements and cannot therefore be used to distinguish one from another.

References C1, C3, C7, C11, C14.

FIG. 12.30 A

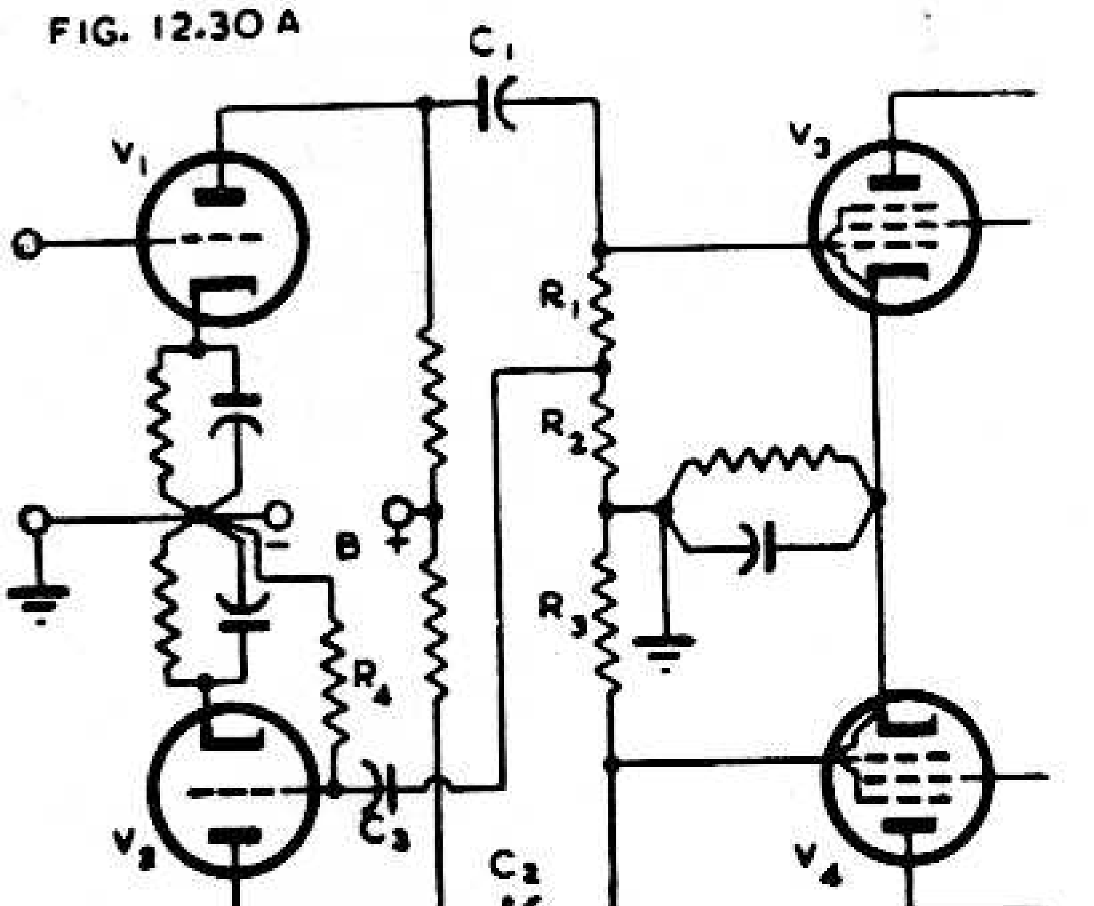


FIG. 12.30 B

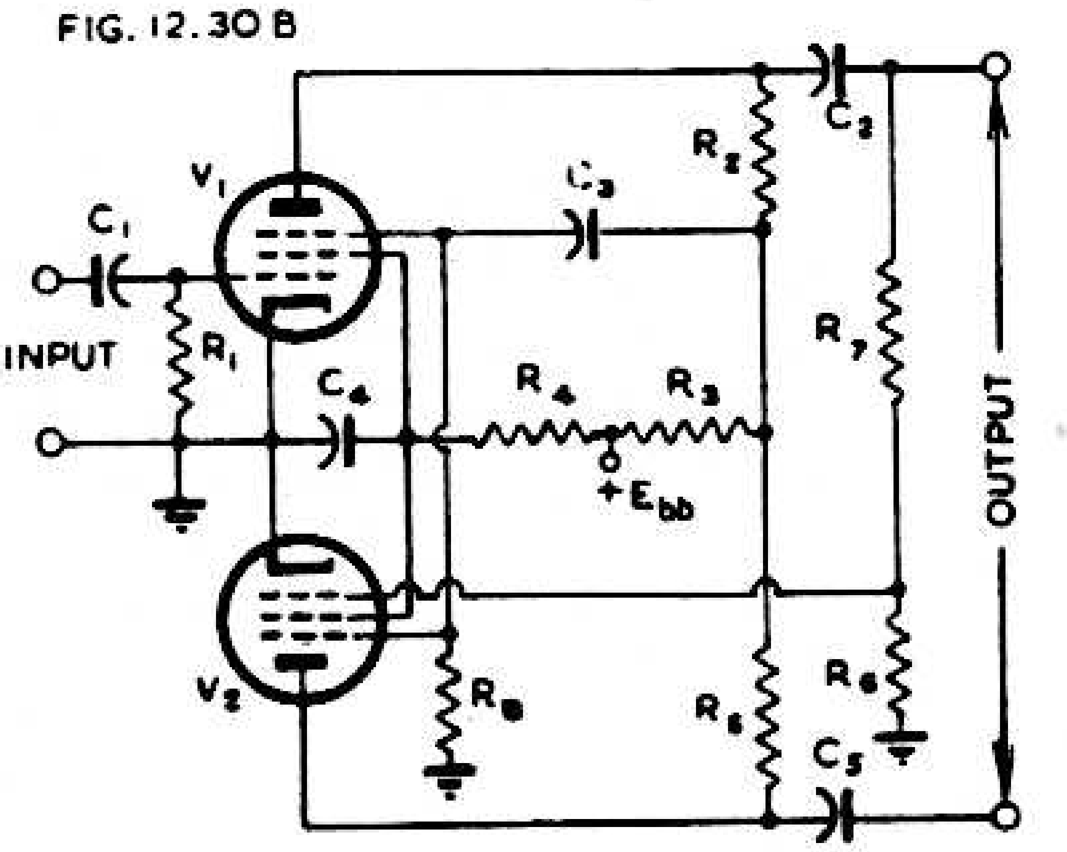


Fig. 12.30A. *Conventional form of phase inverter in which V_1 excites V_3, and V_2 excites V_4, the grid of V_2 being connected to a tapping on the grid resistor of V_3.*
Fig. 12.30B. *Phase inverter with pentodes using the suppressor grids for self-balancing (Ref. C19).*

(iv) Self-balancing phase inverter (Fig. 12.30B)

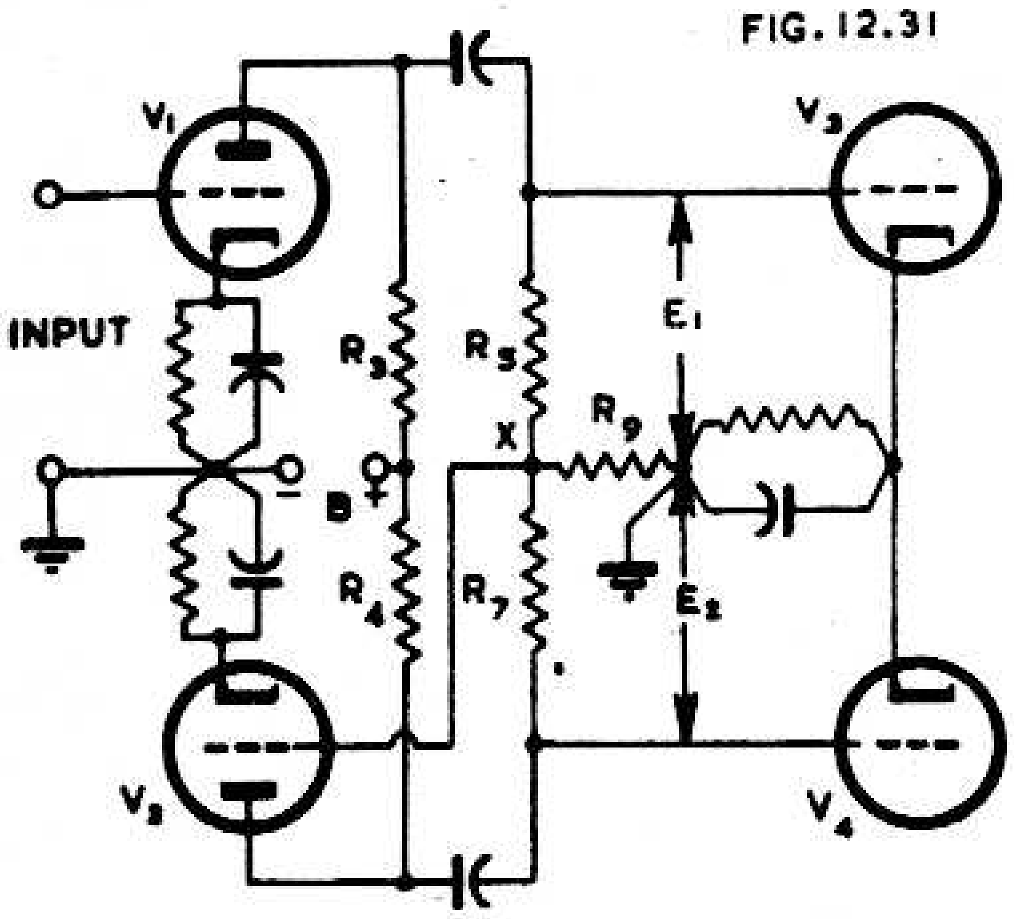
In this circuit V_1 and V_2 are pentodes, and any unbalanced voltage appears across the common plate resistance R_3 and is fed to both suppressors through the blocking condenser C_3, thus causing degeneration in the valve producing the larger signal output, and regeneration in the other. Ref. C19.

(v) Self-balancing paraphase inverter

(A) **Floating paraphase** (Fig. 12.31)

This circuit is, to a considerable extent, self-balancing thereby avoiding any necessity for individual adjustment except in cases where a very high accuracy in balancing is required.

In order to visualize the operation of this circuit consider firstly the situation with V_2 removed. Resistors R_5 and R_9 in series form the load on valve V_1, and the voltage at the point X will be in proportion to the voltage at the grid of V_3. When V_2 is replaced, the voltage initially at point X will cause an amplified opposing voltage to be applied to resistors R_7 and R_9. If resistor R_7 is slightly greater than R_5, it will be found that the point X is nearly at earth potential. If the amplification of V_2 is high, then R_7 may be made equal to R_5 and point X will still be nearly at earth potential. The point X is therefore floating, and the circuit a true Paraphase ; the derivation of the name " Floating Paraphase " is obvious.



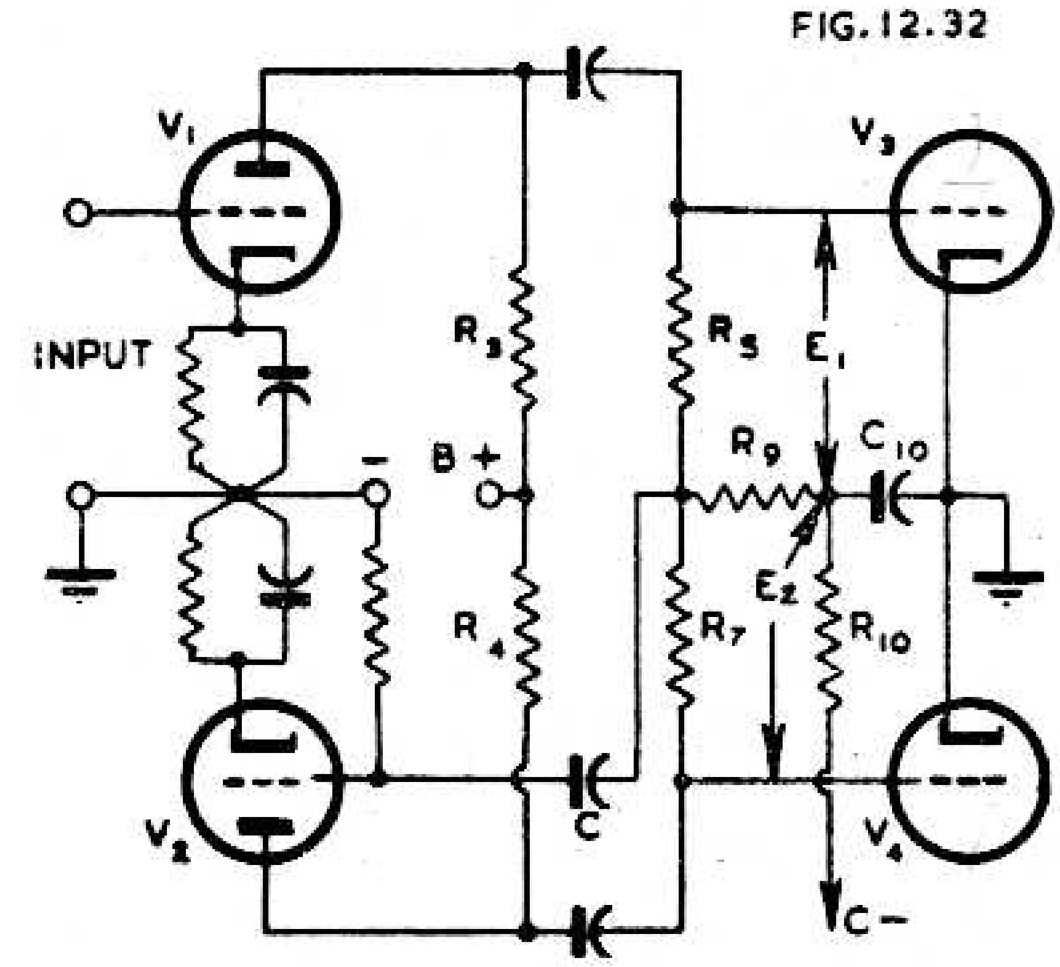


Fig. 12.31. *The Floating Paraphase self-balancing phase inverter with cathode bias.*
Fig. 12.32. *The Floating Paraphase circuit with fixed bias in the following stage.*

The degree of balance is given by

$$\frac{E_1}{E_2} \approx \frac{R_5}{R_7} + \frac{1}{A_2}\left(1 + \frac{R_5}{R_7} + \frac{R_5}{R_9}\right) \qquad (1)$$

where A_2 = voltage gain of V_2 into plate load resistor R_4 and following grid resistor R_7.

If $R_5 = R_7 = R_9$, then $E_1/E_2 = 1 + 3/A_2$ (2)

If V_2 is type 6J5 (or half type 6SN7-GT) $R_4 = 0.1, R_5 = R_7 = R_9 = 0.25$ megohm, then $A_2 = 14$ and $E_1/E_2 = 1.21$ which is too high to be acceptable. In such a case R_7 may be increased to, say, 0.3 megohm giving $E_1/E_2 = 1.03$.

If V_2 is type 6SQ7 with $R_4 = R_5 = R_7 = R_9 = 0.25$ megohm, then $A_2 = 48$ and $E_1/E_2 = 1.06$ which is generally acceptable.

If V_2 is a pentode (e.g. type 6J7) with $R_4 = R_5 = R_7 = R_9 = 0.25$ megohm, then $A_2 = 104$ and $E_1/E_2 = 1.03$, which is very close.

The gain from the grid of V_1 to the grid of V_3 is slightly greater than the gain with $R_9 = 0$.

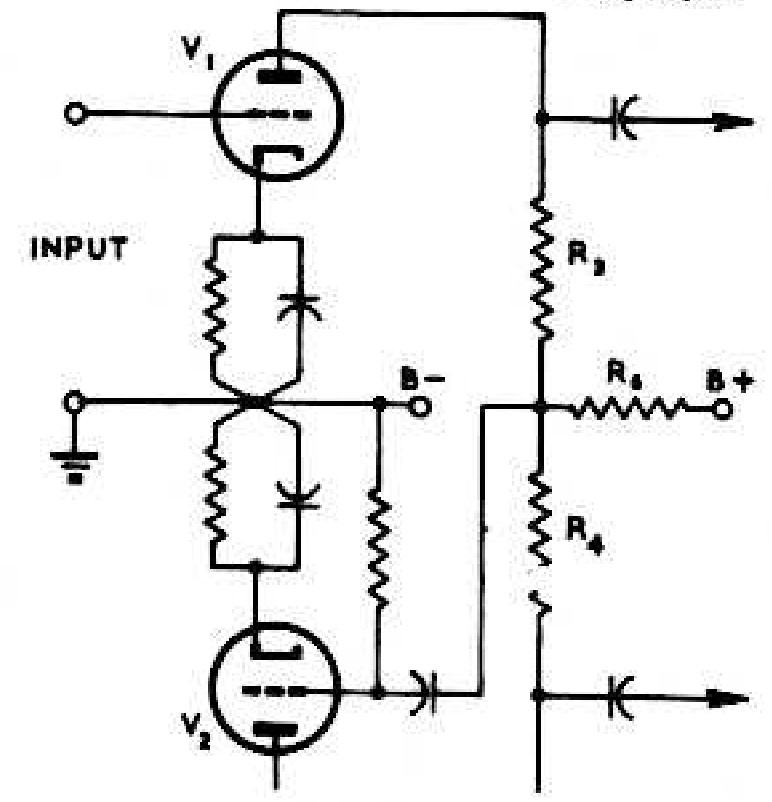


Fig. 12.33. *Common plate impedance self-balancing phase inverter.*

If fixed or partially-fixed bias is employed, it is necessary to couple the grid of V_2 to point X through a suitable condenser (C in Fig. 12.32). In addition, a hum filter (R_{10}, C_{10}) may be required, because most partially-fixed bias sources contain appreciable hum voltage ; any hum voltage appearing across the grid resistor of V_2 is amplified by V_2 and V_4. References C12, C14, C16.

(B) Common plate impedance (Fig. 12.33)

This follows the same principle as the Floating Paraphase, except that the common impedance is in the d.c. plate circuit instead of in the shunt a.c. (following grid) circuit. Here similarly

FIG. 12.34

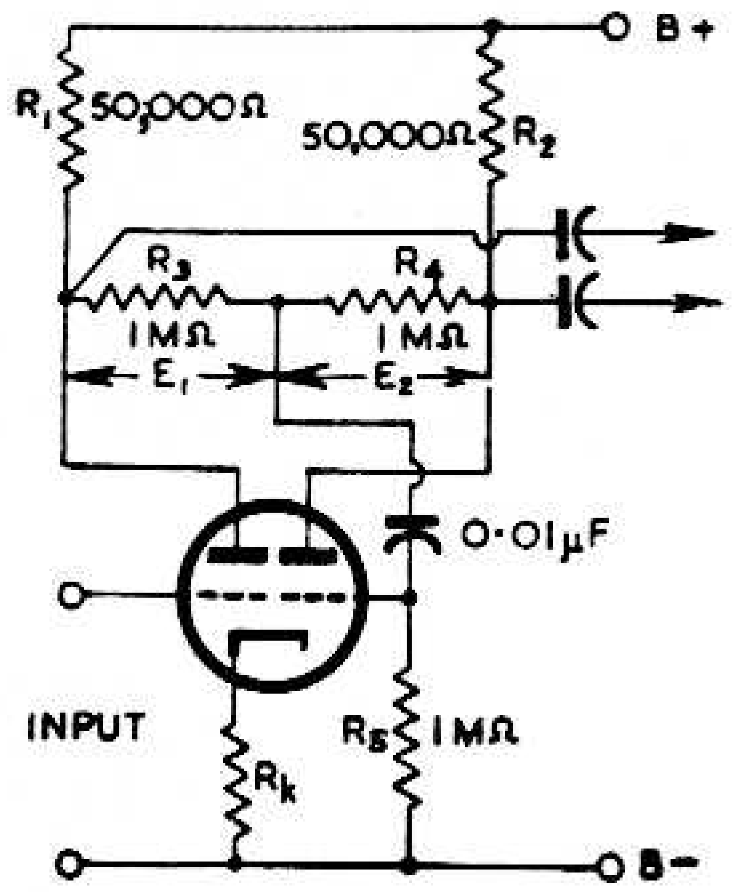


Fig. 12.34. *"See-saw" self-balancing phase inverter (Ref. C9).*

$$\frac{E_1}{E_2} = \frac{R_3}{R_4} + \frac{1}{A_2}\left(1 + \frac{R_3}{R_4} + \frac{R_4}{R_6}\right) \qquad (3)$$

while if $R_3 = R_4 = R_6$, then $E_1/E_2 = 1 + 3/A_2$ (4)
where A_2 = voltage gain of V_2 into plate load resistance R_4. Ref. C12.

(C) See-saw self-balancing phase inverter (Fig. 12.34)

This is merely another form of the common plate impedance circuit, with two separate resistors.

When $R_3 = R_4 = R_5$,

then $E_1/E_2 = 1 + 3/(A_2 + 3)$ (5)

For perfect balance,

$R_4 = R_3\ [1 + 6/(2A_2 - 3)]$. (6)

It is desirable for A_2 to be greater than 40, thereby giving an out-of-balance less than 7%. References C9, C10, C17.

(D) Modified see-saw self-balancing phase inverter (Fig. 12.35)

This is very similar to the original form, except that it eliminates the loading of the grid resistor R_5 (Fig. 12.34). As a result the balance is improved :

When $R_3 = R_4$, then

$E_1/E_2 = 1 + 2/(M + 2)$ (7)

while for perfect balance, $R_4 = R_3\ [1 + 2/(A_2 - 1)]$ (8)

References C9, C10.

(vi) Common cathode impedance self-balancing inverters

(A) The Schmitt phase inverter

This circuit has been described in Chapter 7 Sect. 2(viii)A. It is very useful with twin triode valves, and exact balance may be obtained by suitably proportioning the two load resistors. A slightly modified practical form is shown in Fig. 12.36. The gain with this type of circuit is only about half the normal amplification. References C7, C12.

FIG. 12.35

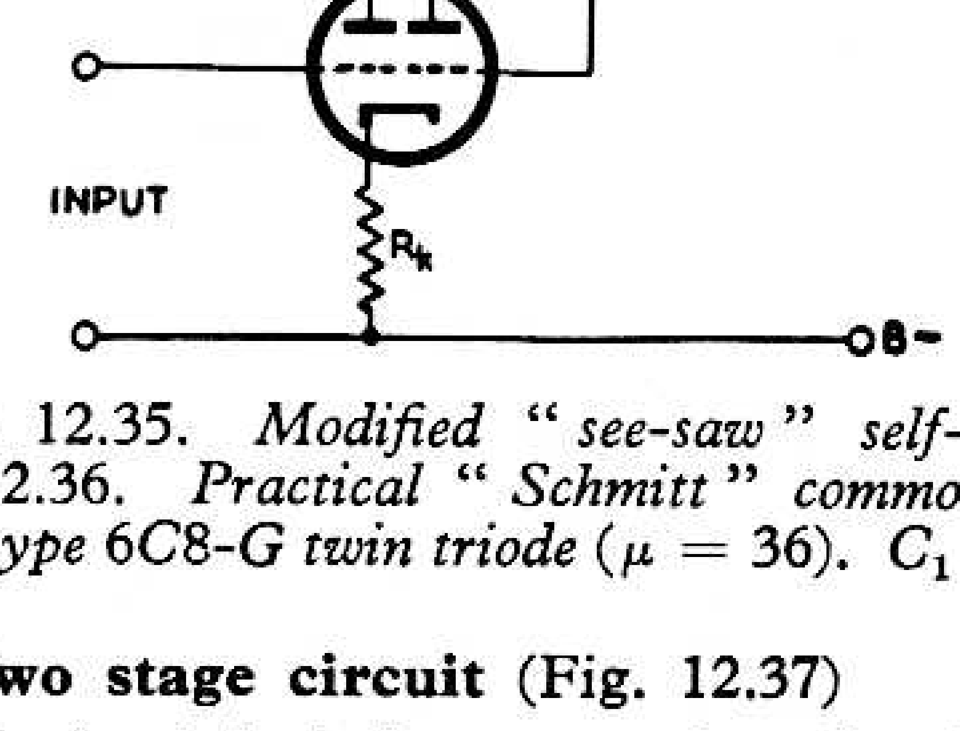


FIG. 12.36

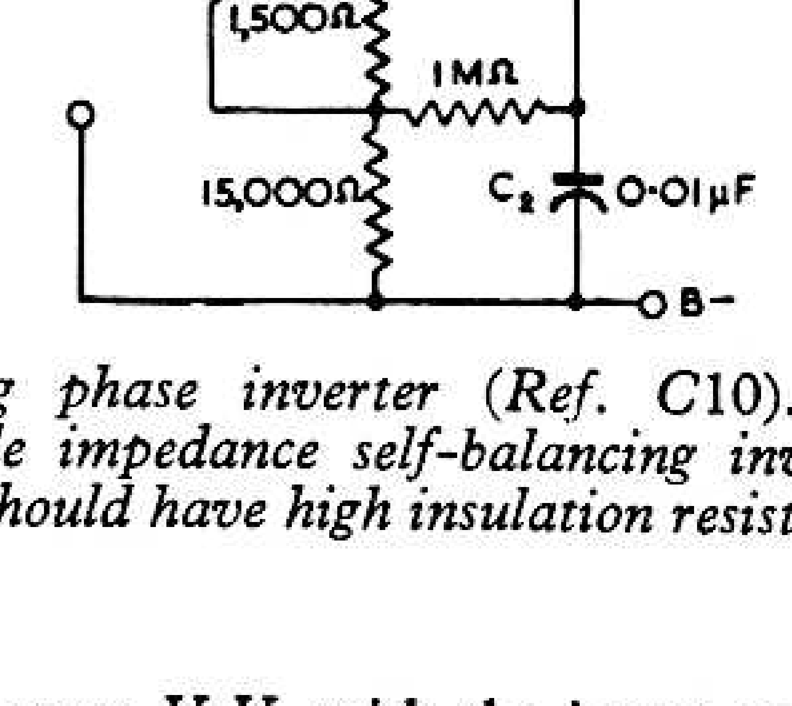


Fig. 12.35. *Modified "see-saw" self-balancing phase inverter (Ref. C10).*
Fig. 12.36. *Practical "Schmitt" common cathode impedance self-balancing inverter using type* 6C8-G *twin triode* (μ = 36). C_1 *and* C_2 *should have high insulation resistance.*

(B) Two stage circuit (Fig. 12.37)

This circuit includes a normal push-pull triode stage V_1V_2 with the input applied to V_1 grid only. The grid of V_2 is excited from the common cathodes of the following stage (V_3V_4) through blocking condenser C. Thus the out-of-balance voltage across R excites V_2. No analysis appears to have been published.

(vii) Balanced output amplifiers with highly accurate balance

For certain specialized applications it is necessary to have highly stable and accurate balance. This may be achieved in various ways including

(a) the use of two or more inverters in cascade, the later ones operating with push-pull input.
(b) the use of an additional valve to amplify the out-of-balance voltage before applying it for correction (Ref. C13).
(c) the use of the so-called " phase compressor " following the inverter—see Sect. 7(v).

FIG. 12.37

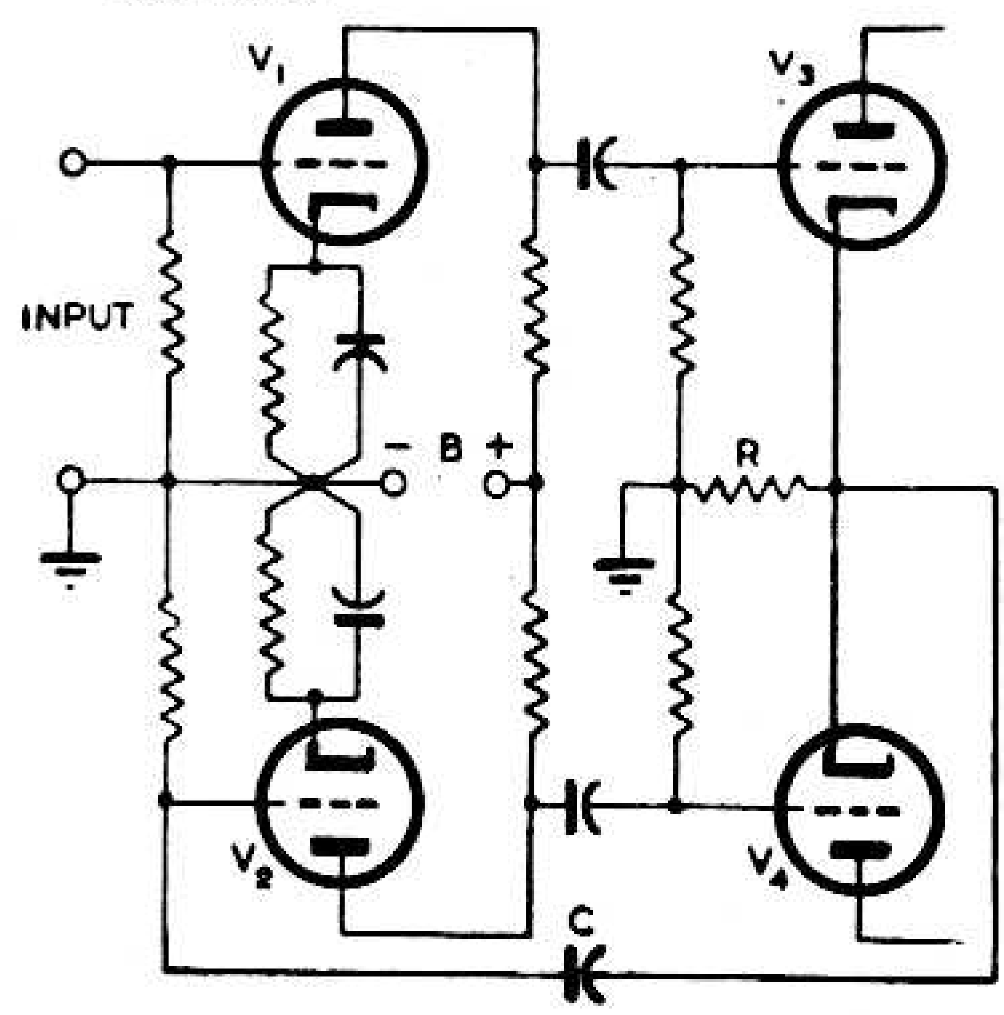


Fig. 12.37. *Two stage self-balancing phase inverter with common cathode impedance.*

(viii) Cross coupled phase inverter

The cross coupled phase inverter (Ref. C25) employs two twin triodes, the circuit being that of the second and third stages of Fig. 15.43A. In this application the single input may be connected between either of the two grids and earth, the other being connected to earth. The hum level from the plate supply is low because hum voltages balance out.

SECTION 7 : PUSH-PULL VOLTAGE AMPLIFIERS

(i) Introduction (ii) Cathode resistors (iii) Output circuit (iv) Push-pull impedance-coupled amplifiers—mathematical treatment (v) Phase compressor.

(i) Introduction

A push-pull voltage amplifier stage is one having push-pull (3 terminal) input and push-pull output. Two separate valves (or one twin valve) are required. The two valves are each treated as for a single-ended amplifier, whether resistance- or transformer-coupled.

(ii) Cathode resistors

A common cathode resistor may generally be used and should not normally be bypassed. Separate cathode resistors are beneficial in that they aid in the correct adjustment of the operating point when this is at all critical, such as

(a) r.c.c. pentodes when extremely low distortion is required*,
(b) any r.c.c. valves operating near to the overload point,
(c) transformer-coupled triodes when it is desired to reduce to a minimum the out-of-balance direct current.

(iii) Output circuit

The voltage from plate to plate does not include any appreciable even harmonic distortion owing to the cancellation between the two valves, but it does include odd harmonics. The voltage from either plate to earth, however, includes even harmonics. If such a stage is r.c. coupled directly to the grids of a following similar push-pull stage, there is no benefit obtained in the form of reduced distortion. For the latter achievement it is necessary to use either a push-pull transformer with a

*Alternatively a similar result may be achieved by adjusting the screen resistors.

centre-tapped primary, or the so-called " phase compressor "—see (v) below. The same also holds true when feeding into a single-ended stage.

Many push-pull transformers are designed to operate with an out-of-balance plate current not exceeding about 1 or 2 mA, thus necessitating matched valves, or adjusted bias, or parallel feed.

(iv) Push-pull impedance-coupled amplifiers—mathematical treatment (Fig. 12.38)

Such a 6 terminal amplifier involves four gain factors

$$A = e_2/e_1 \quad \text{for} \quad e_1' = 0 \qquad (1)$$
$$A' = e_2'/e_1' \quad \text{for} \quad e_1 = 0 \qquad (2)$$
$$\gamma = -e_2'/e_1 \quad \text{for} \quad e_1' = 0 \qquad (3)$$
$$\gamma' = -e_2/e_1' \quad \text{for} \quad e_1 = 0 \qquad (4)$$

In a perfectly balanced amplifier $A = A'$ and $\gamma = \gamma'$.

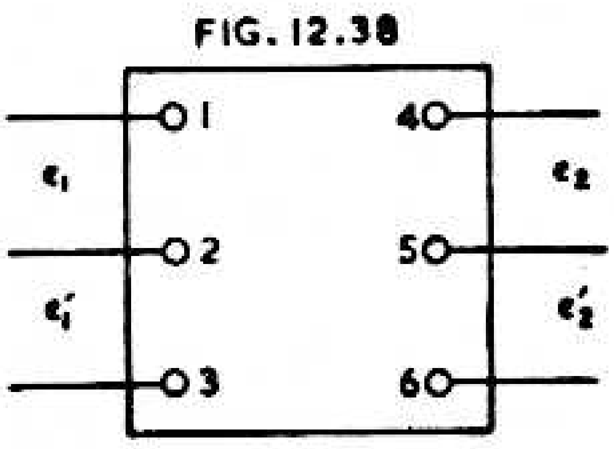


Fig. 12.38. *6 terminal push-pull amplifier.*

If there is no cross-coupling (e.g. from a cathode resistor common to both sides) $\gamma = \gamma' = 0$.

In a linear amplifier—

$$e_2 = Ae_1 - \gamma' e_1' \qquad (5)$$
$$e_2' = A'e_1' - \gamma e_1 \qquad (6)$$
$$\text{Differential gain} = (e_2 - e_2')/(e_1 - e_1') \qquad (7)$$
$$= \tfrac{1}{2}(A + A' + \gamma + \gamma') \qquad (8)$$
$$\text{In-phase gain} = (e_2 + e_2')/(e_1 + e_1') \qquad (9)$$
$$= \tfrac{1}{2}(A + A' - \gamma - \gamma') \qquad (10)$$
$$\text{Inversion gain} = (e_2 - e_2')/\tfrac{1}{2}(e_1 + e_1') \qquad (11)$$
$$= A - A' + \gamma - \gamma' \qquad (12)$$
$$\text{Differential unbalance} = (e_2 + e_2')/(e_1 - e_1') \qquad (13)$$
$$= \tfrac{1}{2}(A - A' - \gamma + \gamma') \qquad (14)$$

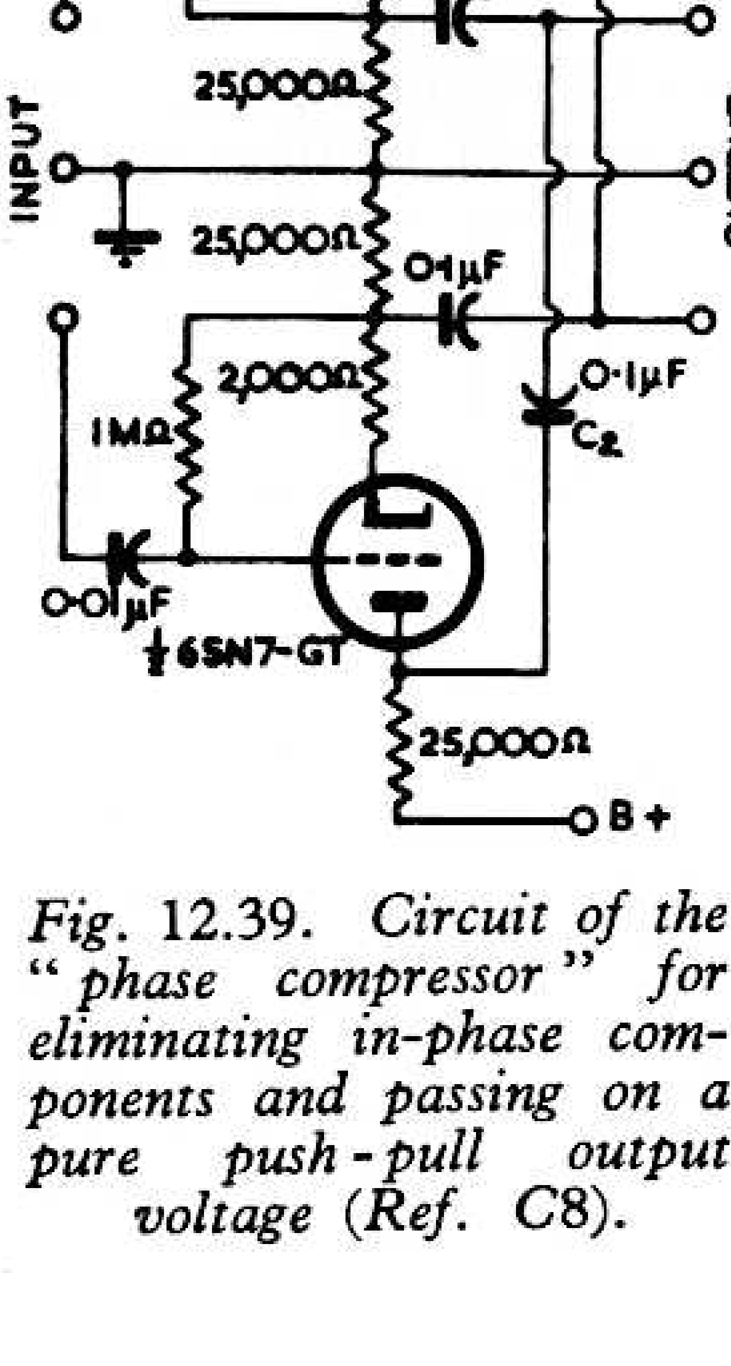


Fig. 12.39. *Circuit of the " phase compressor " for eliminating in-phase components and passing on a pure push-pull output voltage (Ref.* C8).

In an amplifier without cross-coupling, $\gamma = \gamma' = 0$:

Differential gain $= \tfrac{1}{2}(A + A')$ = average gain of two sides.

In-phase gain = differential gain.

Inversion gain $= A - A'$ = difference between gains of two sides.

Differential unbalance $= \tfrac{1}{2}(A - A')$.

See also pages 573-574 for the theory of push-pull amplification based on the expansion of the valve characteristic into an infinite series.

(v) Phase compressor

The so-called " phase compressor " is rather a misnomer since its function is to eliminate the in-phase components (e.g. even harmonics) from a nominally push-pull output. In this application it operates in very much the same manner as a transformer. The circuit is given in Fig. 12.39 and is a push-pull phase splitter with the addition of capacitances C_1 and C_2 from each plate to the opposite output terminal. It is these condensers which attenuate in-phase components. The gain for push-pull input is approximately 0.9 from input to output.

Even if the input signal is imperfectly balanced, the output is still balanced. This circuit, unlike a transformer, does not remove hum from the B+ line, but reduces it to about half at the output terminals. Ref. C8.

SECTION 8 : IN-PHASE AMPLIFIERS

(i) Cathode-coupled amplifiers (ii) Grounded-grid amplifiers (iii) Inverted input amplifiers (iv) Other forms of in-phase amplifiers.

(i) Cathode-coupled amplifiers

These are described in Chapter 7 Sect. 2(viii)B and have many forms for special applications. In Fig. 7.46 the output from terminal B is in phase with the input. See references Chapter 7 Sect. 6(G).

(ii) Grounded-grid amplifiers

In these, the input voltage is applied to the cathode, the grid is earthed, and the output is taken from the plate, being in phase with the input. Driving power is required, so that it is not strictly a voltage amplifier.

(iii) Inverted input amplifiers

In Fig. 7.18 the input is from grid to plate and the output voltage is in phase with the input.

(iv) Other forms of in-phase amplifiers

There are many other forms too numerous to mention.

SECTION 9 : DIRECT-COUPLED AMPLIFIERS

(i) Elementary d-c amplifiers (ii) Bridge circuit (iii) Cathode-coupled (iv) Cathode follower (v) Phase inverter (vi) Screen coupled (vii) Gas tube coupled (viii) Modulation systems (ix) Compensated d.c. amplifiers (x) Bridge-balanced direct current amplifiers (xi) Cascode amplifiers.

(i) Elementary d-c* amplifiers

A direct-coupled amplifier is one in which the plate of one stage is connected to the grid of the next stage directly, or through a biasing battery or equivalent. It usually receives the plate, screen and grid voltages from sources which do not include any reactances such as filter or by-pass condensers. If this condition is fulfilled, it may amplify down to zero frequency without attenuation or phase shift ; it may also be used to amplify direct voltages.

FIG. 12.40

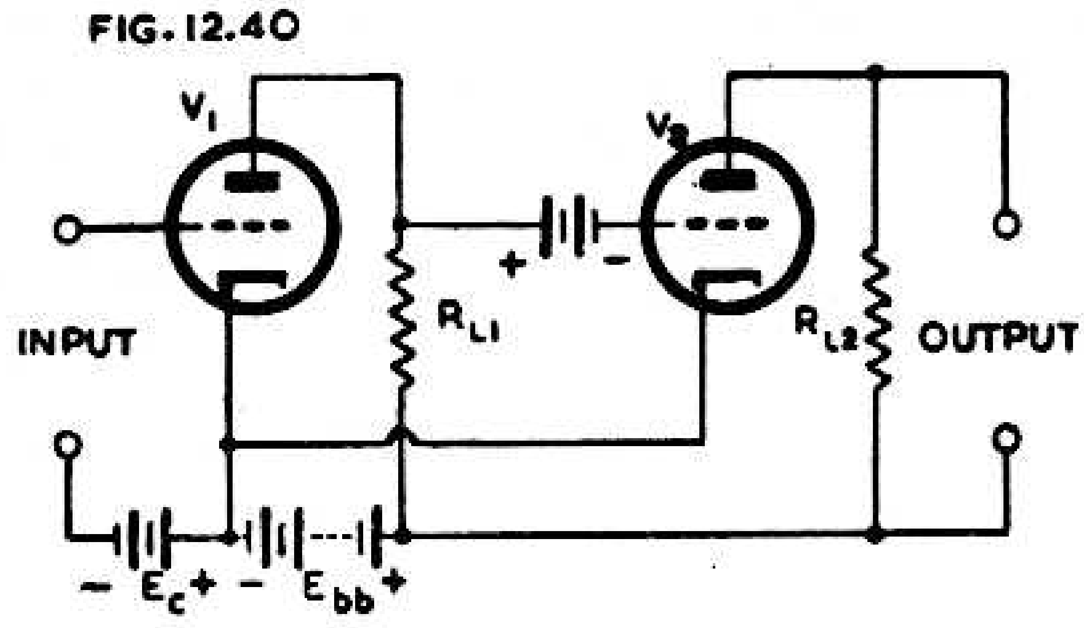


Fig. 12.40. *Elementary direct-coupled amplifier.*

FIG. 12.41

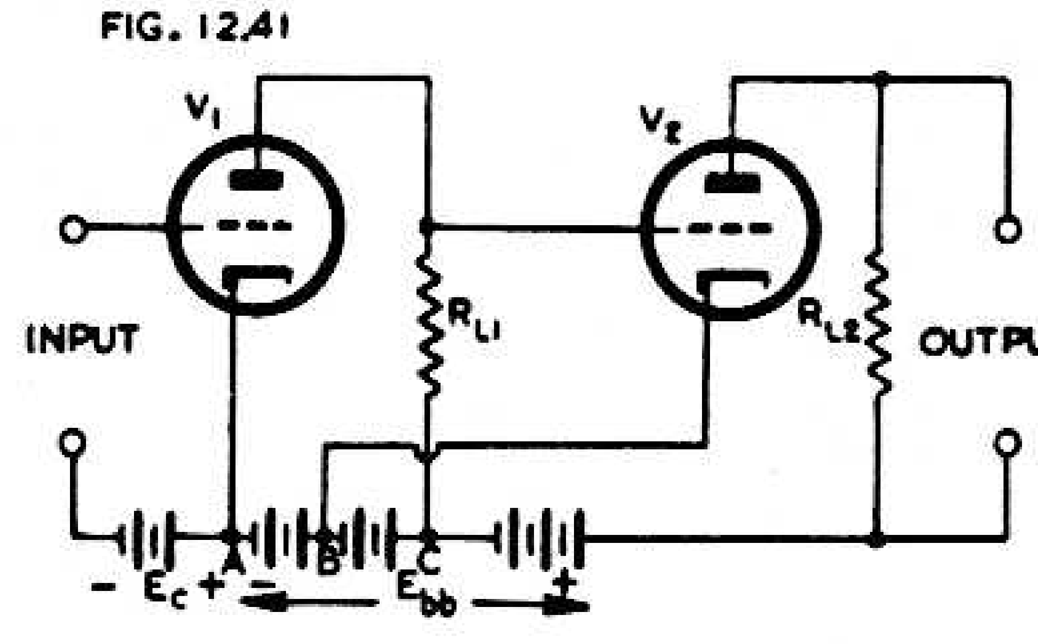


Fig. 12.41. *Direct-coupled amplifiers without " hot " grid bias battery.*

The most elementary form is Fig. 12.40 in which the plate of V_1 is coupled to the grid of V_2 through a bias battery to provide the correct grid bias. As it stands it is of little practical use, owing to the " hot " bias battery.

A more practical circuit is Fig. 12.41 in which all the batteries may be at earth potential. Tapping point B has to be adjusted to give the correct bias between grid

*The abbreviation d-c amplifier is used in this Handbook to indicate a direct-coupled amplifier.

and cathode of V_2. The voltage drop across R_{L1} will normally be more than half the voltage between A and C.

A circuit requiring only a single source of voltage is Fig. 12.42 ; the two dividers are desirable to avoid interaction (degeneration) between the two stages unless the bleed current is very high. The output terminal is returned to a point on the voltage divider having the same potential.

Another circuit requiring only a single source of voltage is Fig. 12.43 ; this has only a single voltage divider. The voltage drop across R_{k2} must equal the plate voltage of V_1 minus the grid bias of V_2. There will be degeneration caused by the unby-passed cathode resistors, which may be avoided by a push-pull arrangement with common cathode resistors for both stages.

FIG. 12.42

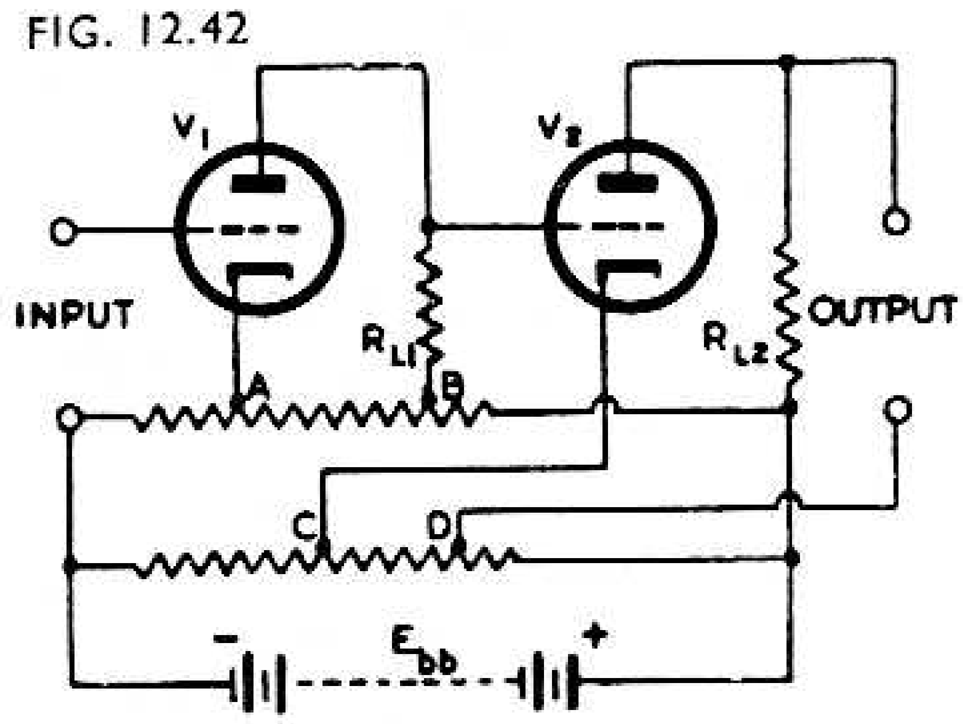


FIG. 12.43

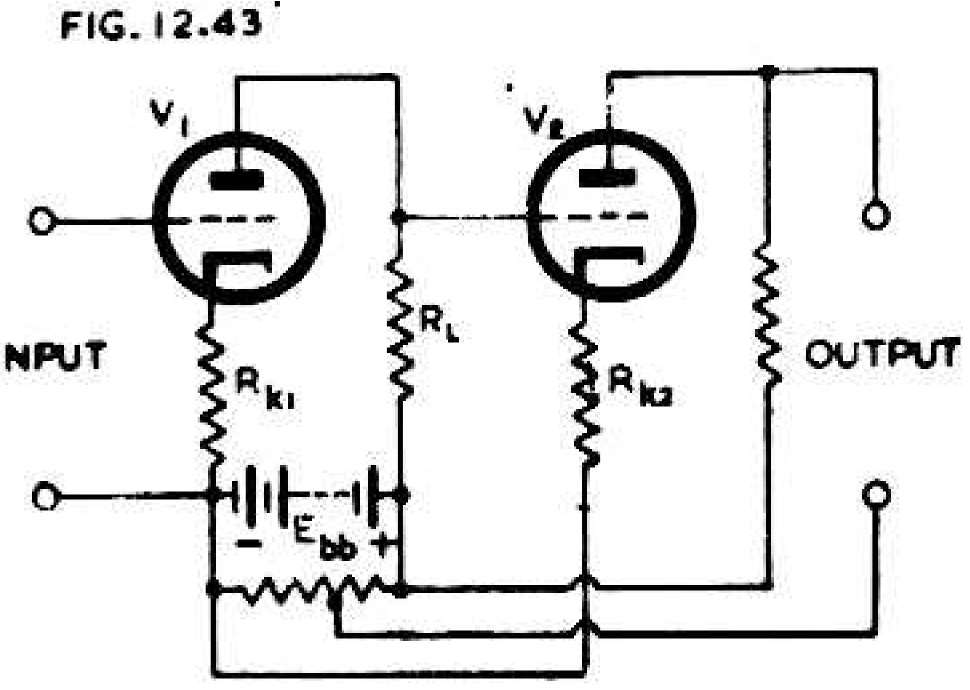


Fig. 12.42. *Direct-coupled amplifier with only one source of voltage* (E_{bb}).
Fig. 12.43. *Direct-coupled amplifier with one source of voltage and one voltage divider.*

Pentodes may be used, if desired, in all these circuits by making suitable provision for the screen voltages. In Fig. 12.43 the screen of V_1 may be taken to a tap on R_{k2}, provided that either the screen is by-passed to cathode, or R_{k2} is by-passed ; the amplifier would then be limited to audio frequencies only (Ref. D40).

When the first amplifier stage is a pentode, its load resistor may be increased to values much greater than those conventionally used, provided that sufficient negative voltage feedback is applied to secure an acceptable high frequency response (e.g. Ref. D40).

A pentode may also be used with another pentode as its plate resistor. By this means a gain of several thousand times may be obtained, but this is only useful in electronic measuring instruments.

Such circuits (Figs. 12.40—12.43) are generally limited to two stages. If any increase is made, there is distinct danger of slow drift occurring in the direct plate current, due to variations in battery voltages and valve characteristics. These may be minimized by voltage regulators and controlled heater voltage or current, or may be avoided by one of the special methods described below (viii to x). See also Refs. D36, D39.

Negative feedback may be applied to any d-c amplifier in the normal manner.

(ii) Bridge circuit

The bridge circuit (Fig. 12.44) may be used with any number of stages in cascade from a single B supply. The basic design equations are :

$$R_2 = R_1(E_b + E_3)/[E_1 - E_b(1 + R_1/R_p)] \quad (1)$$

$$R_3 = R_1(E_2 - E_3)/[E_1 - E_b(1 + R_1/R_p)] \quad (2)$$

where R_p = the d.c. plate resistance of $V_1 = E_b/I_b$
and E_b = the direct plate voltage.

Typical operation with 6SJ7 pentode as V_1

$E_{bb} = 400$ volts, $E_1 = E_2 = 200$ volts, $E_b = 100$ volts, $E_3 = 0$, $R_1/R_p = 0.5$. Amplification is 71.4% of that as a r.c.c. amplifier. A two stage amplifier using type 6J7 pentodes has an amplification of 100 per stage with fixed voltages for bias and screen supplies, or with push-pull connection.

Negative feedback may be applied to this type of circuit.

References D8, D15.

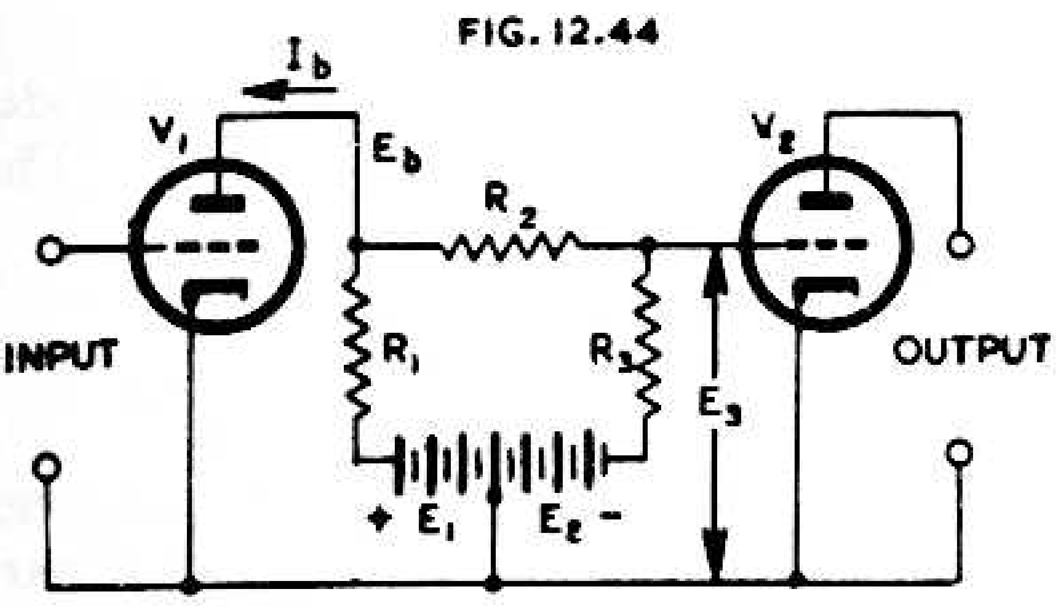


Fig. 12.44. *"Bridge" circuit, direct-coupled amplifier.*

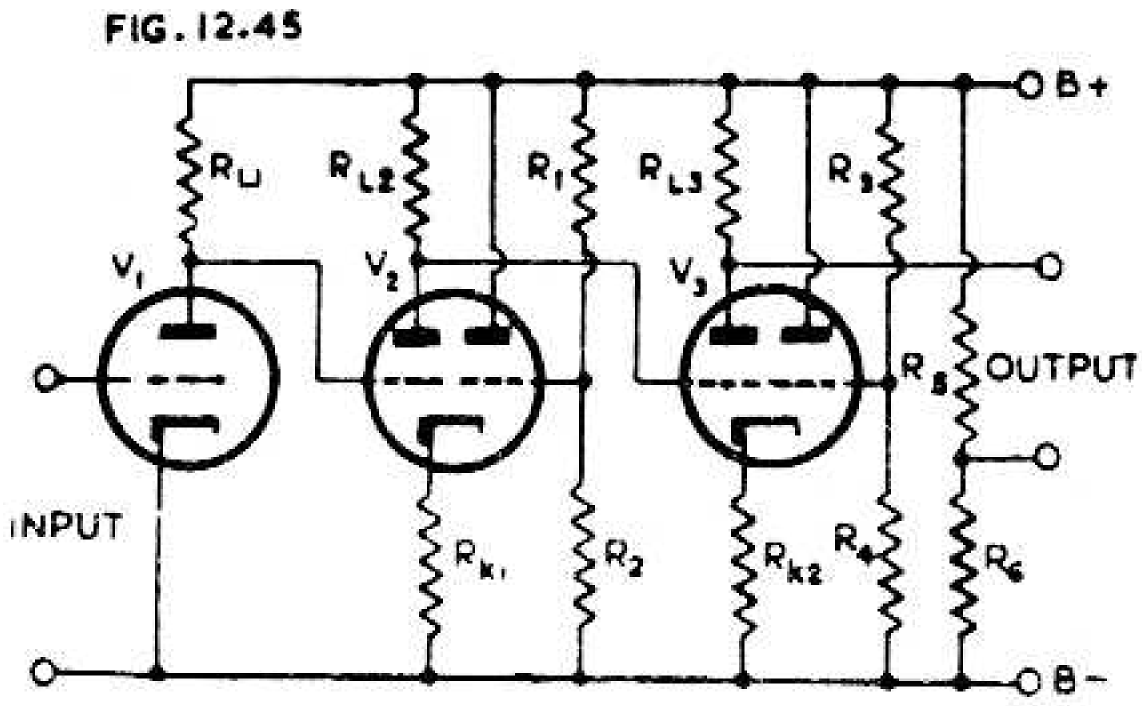


Fig. 12.45. *Three stage d-c amplifier with V_2 and V_3 as cathode-coupled twin triodes.*

(iii) Cathode-coupled

The fundamental form of a cathode-coupled amplifier is covered in Chapter 7 Sect. 2(viii)B.

Fig. 12.45 shows a conventional d-c single triode (V_1) followed by two twin triodes as d-c cathode-coupled amplifier stages. For example, V_2 has one triode as a cathode follower with its grid at a fixed voltage from the voltage divider R_1R_2, while the other triode operates as an amplifier but sharing the common cathode resistor R_{k1}.

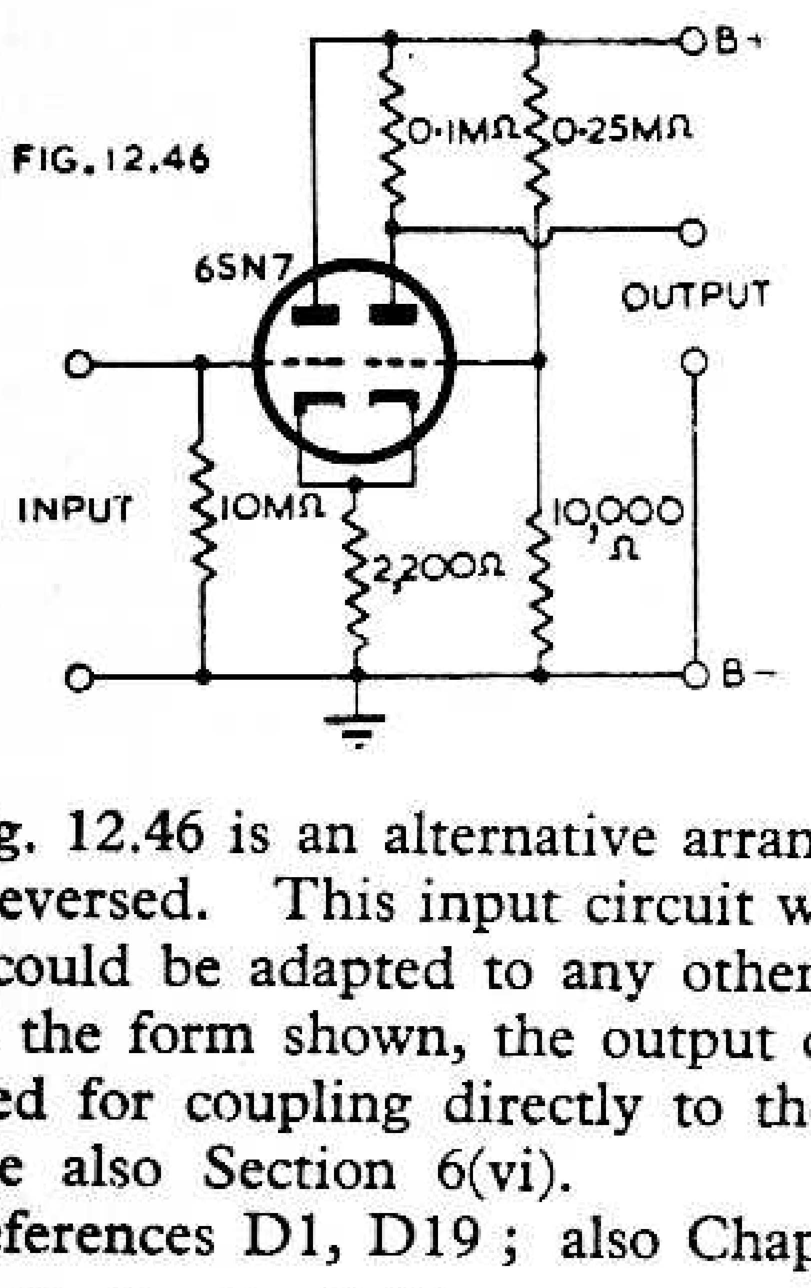


Fig. 12.46. *Direct-coupled amplifier incorporating twin triode with relative positions opposite to those in Fig.* 12.45 *(extracted from Ref.* D19).

Fig. 12.46 is an alternative arrangement in which the positions of the two triodes are reversed. This input circuit was primarily for use with a phototube (Ref. D19) but could be adapted to any other requirement.

In the form shown, the output circuit has a direct potential difference, being intended for coupling directly to the screen of the 6V6 following.

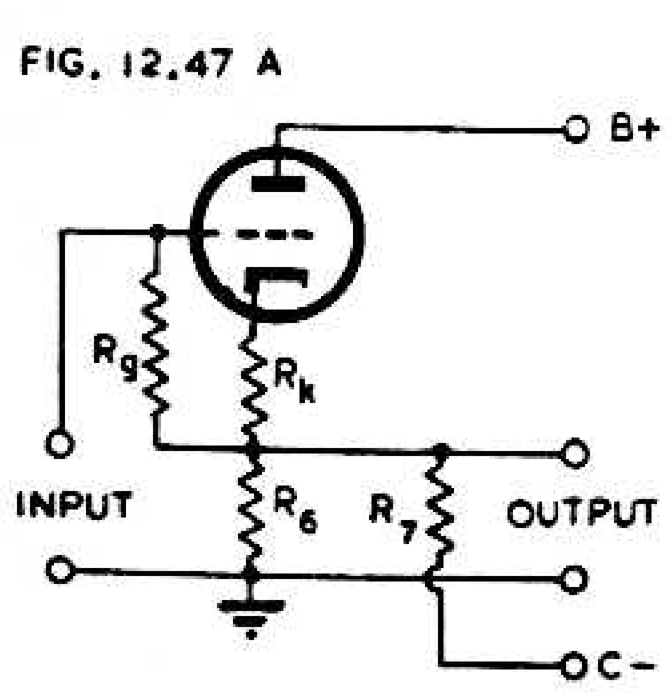
See also Section 6(vi).

References D1, D19 ; also Chapter 7 Refs. (G).

(iv) Cathode follower

The cathode follower may also be used as a d-c amplifier. One circuit is Fig. 12.47A in which the total cathode load is $R_k + R_6R_7/(R_6 + R_7)$. The value of R_7 is equal to E_c/I_b so as to eliminate the undesired direct voltage across R_6.

References D15, D36.



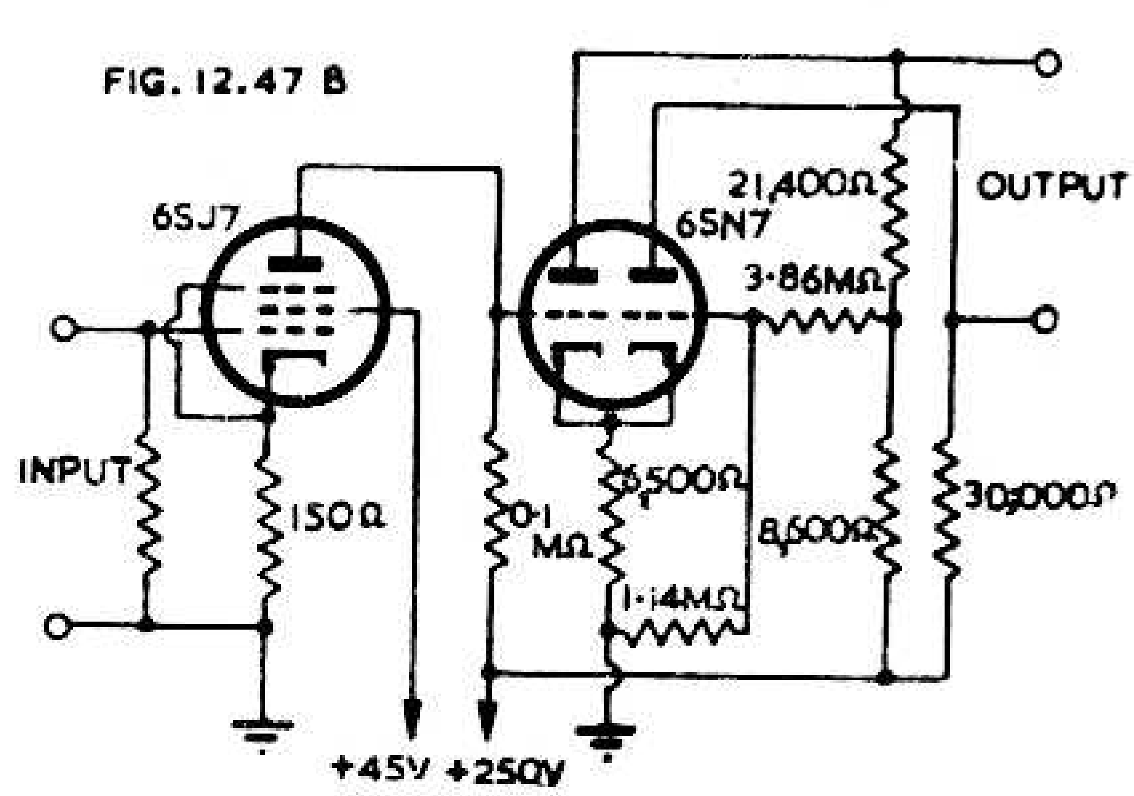


Fig. 12.47A. *Cathode follower used as d.c. amplifier (Ref.* D15).
Fig. 12.47B. *Direct-coupled amplifier with r-c pentode exciting phase inverter (Ref. D*14).

(v) Phase inverter

A direct coupled amplifier having in the first stage a resistance coupled pentode, and in the second stage a twin triode phase inverter, is shown in Fig. 12.47B. The voltage gain is 67 db, with uniform gain up to 12 000 c/s.

References D14, D15.

(vi) Screen-coupled

The preceding stage may be directly-coupled to the screen of a cathode follower pentode (Fig. 12.48). This circuit has a voltage gain of 30 db with 0.5% total harmonic distortion at 0.85 volt peak output. Output terminal A has a d.c. potential of $-$ 1.5 volts, which may be used as bias for the following stage.

Screen-coupled cathode followers are stable, with a wide-band frequency response, but the distortion is higher than with normal operation owing to the non-constant ratio of plate to screen currents. About 85% of this distortion can be cancelled by a push-pull arrangement.

References D14, D19.

(vii) Gas tube coupled

Fig. 12.49 shows the simplest form of gas tube coupling in which a gas tube (GT) provides the desired voltage drop from the plate of V_1 to the grid of V_2. The gas tube here must be a glow tube or neon lamp, a voltage regulator tube being unsuitable because its d.c. plate resistance is of the same order of magnitude as the plate load resistor. The values of R_{L1} and R_g must be carefully selected to meet the various limitations.

FIG. 12.48

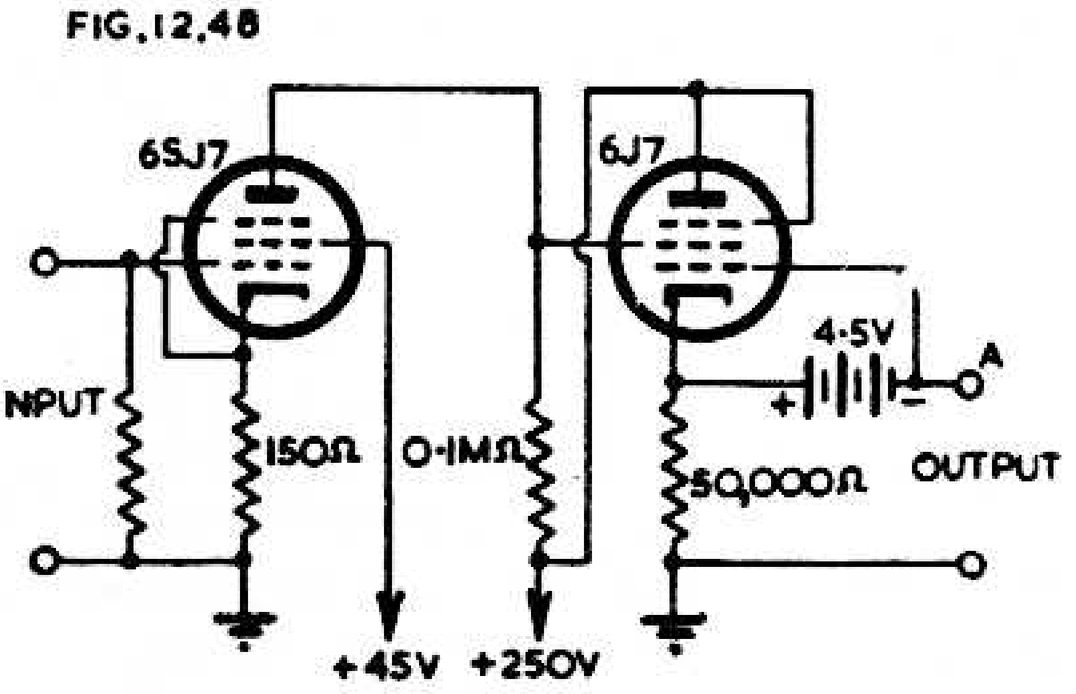


FIG. 12.49

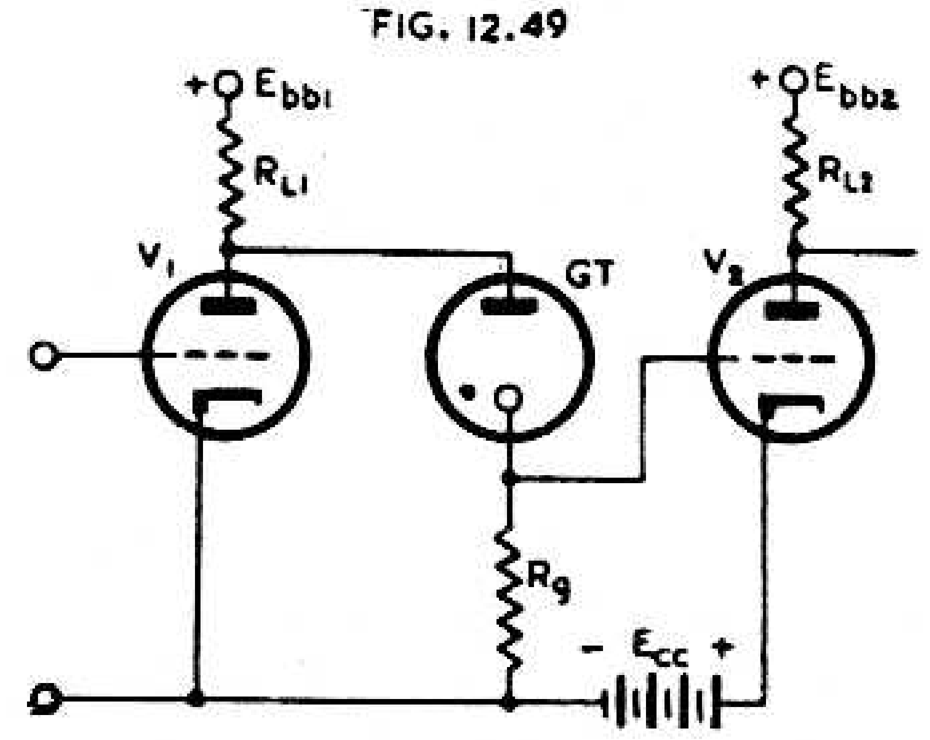


Fig. 12.48. *Two stage amplifier with the plate of the first stage directly-coupled to the screen of the second stage (Ref. D14).*

Fig. 12.49. *Two stage amplifier with gas tube coupling from the plate of the first stage to the grid of the second stage.*

Fig. 12.50 is an improved circuit in which an additional valve (6J5) is used as a cathode follower with its cathode impedance composed of the voltage regulator tube and the resistor R_g. With this arrangement there is no d.c. load on the first amplifier and the input impedance of the cathode follower is so high that it does not affect the gain of the first stage. Almost all the signal voltage drop occurs across R_g. The design of the first stage is independent of the d.c. resistance of the V.R. tube, but R_g must be much larger than the dynamic impedance of the V.R. tube. Negative feedback for improved stability is provided by R_k in the second stage.

The gas tube introduces noise, hence should not be used in low-level amplifiers.

For design, see Reference D13.

(viii) Modulation systems

Although not strictly direct-coupled amplifiers, they may be used in many applications. The d.c. signal to be amplified modulates a carrier wave, and after sufficient amplification the modulated wave is detected to obtain the amplified signal. In a modified arrangement, the input signal is interrupted or " chopped."

References D2, D8, D9, D11, D20, D21, D22, D28.

FIG. 12.50

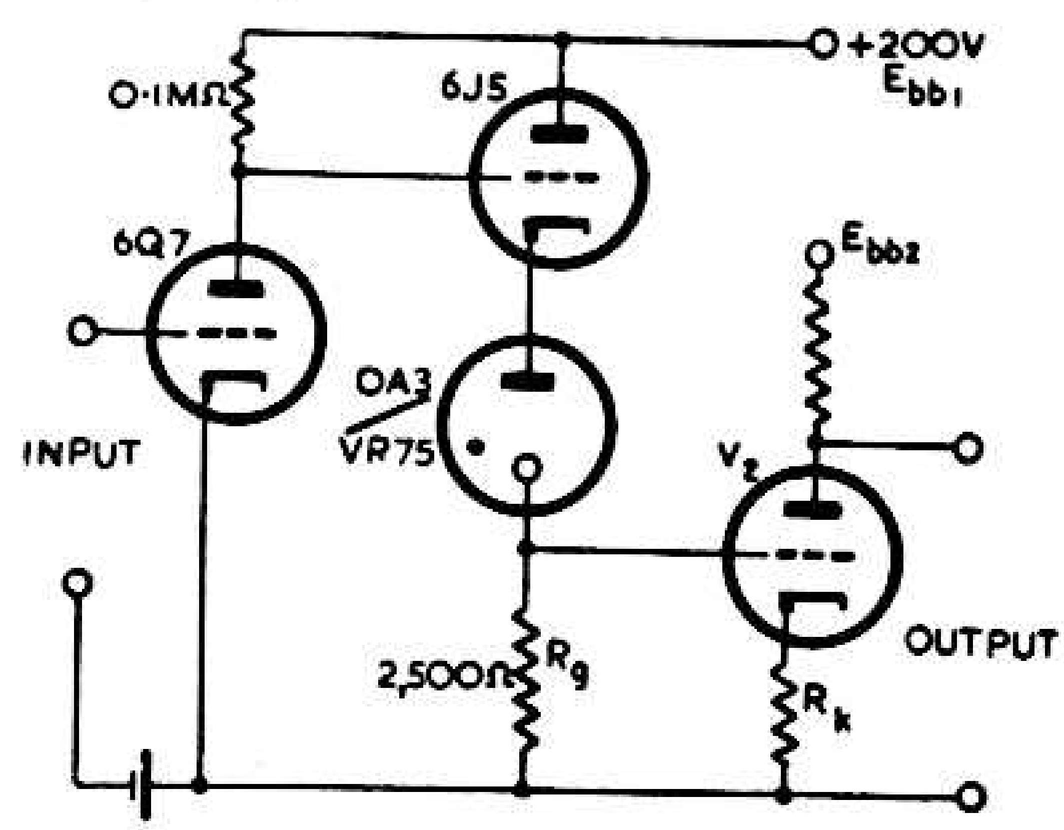


Fig. 12.50. Improved two stage amplifier with gas tube coupling, in which a cathode follower is introduced for better performance.

(ix) Compensated direct current amplifiers

In these, some variable characteristic of the amplifying valve is balanced against the same variations in another valve, or against a different characteristic of the same valve.

(A) Cathode compensation

A typical circuit is Fig. 12.51A and makes use of a twin triode with common cathode resistance. This largely compensates for contact potential drift, and provides a stable amplifier provided that an accurately regulated power source is used. Valves with common cathodes are also used (e.g. 6SC7).

A diode-triode or diode-pentode valve with a common cathode may also be used (Refs. D36, D39).

Both cathode and B supply compensation may be obtained by returning the lower input terminal of Fig. 12.51A to a tapping point ($+ E_b/A$) on the voltage divider (broken line) instead of to earth (Ref. D8).

Other circuits used are series balance, and cascode series balance (Ref. D39); also cathode coupled phase splitter with single ended output (Refs. D1, D39).

References D1, D4, D8, D15, D23, D24, D36, D39.

FIG. 12.51A

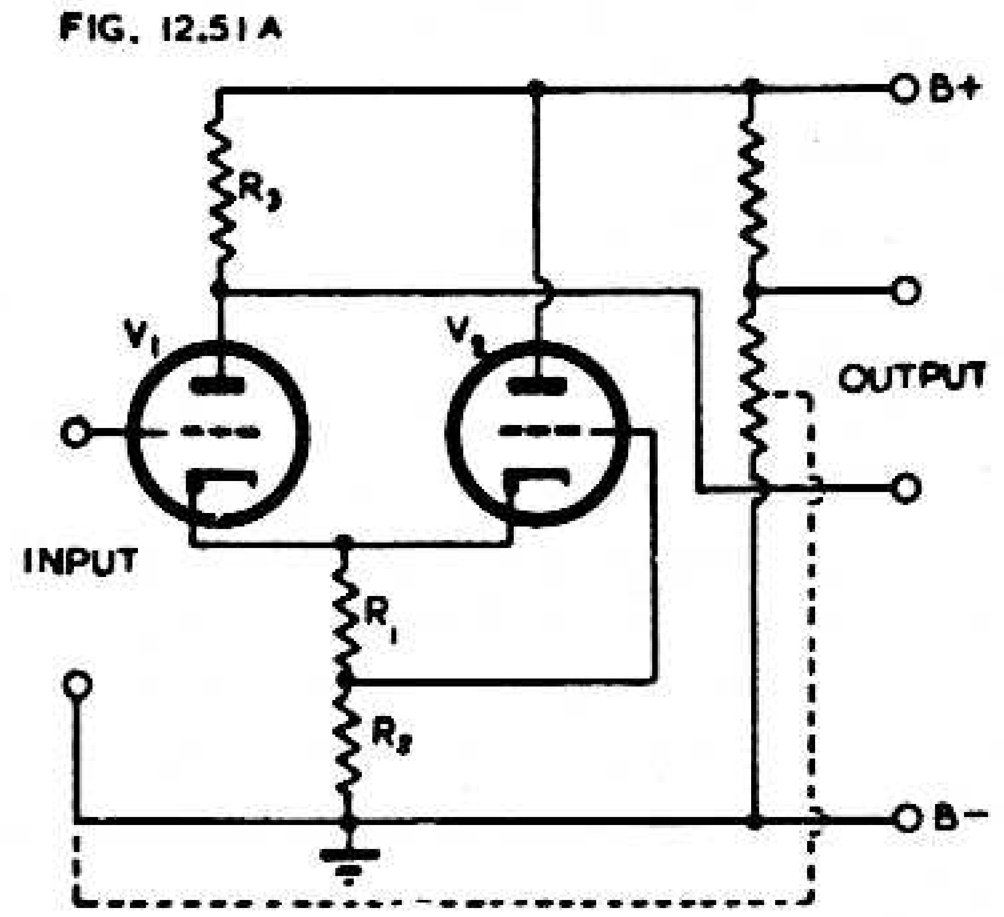


Fig. 12.51A. Cathode-coupled twin-triode used as d.c. amplifier with cathode compensation.

(B) Compensation for filament and plate voltages

This is used in the electrometer tube circuit, and has low drift but cannot be cascaded. Ref. D8.

(C) Compensation for emission

This can be obtained by a circuit using a pentagrid valve. Refs. D4, D8.

(D) Push-pull operation

A degree of compensation is provided by any push-pull amplifier An alternative form is a push-pull circuit in which one half only of each stage is used as an amplifier, and the other half as a dummy to reduce drift (Ref. D25).

(x) Bridge-balanced direct current amplifiers

With this type of direct current amplifier the regulation of plate and filament supplies usually becomes unnecessary. These are normally used only in laboratory instruments. Ref. D8.

(xi) Cascode amplifiers

The cascode amplifier fundamentally consists of two triodes connected in series (Fig. 12.51B). The usual arrangement in practice is to provide a fixed positive voltage for the grid of V_1.

A cascode amplifier may be considered as a single valve having the characteristics μ', g_m', r_p'. The load into which V_2 works is given by

$$\frac{r_p + R_L}{\mu + 1}$$

where μ, g_m and r_p are the characteristics of both V_1 and V_2.

The amplification of V_2 is therefore given by

$$A' = \frac{\mu R_L}{r_p + (r_p + R_L)/(\mu + 1)} \qquad (3)$$

$$= \frac{1}{\dfrac{\mu + 2}{g_m(\mu + 1)R_L} + \dfrac{1}{\mu(\mu + 1)}} \qquad (4)$$

Eqn. (4) may be compared with the ordinary form for expressing amplification, namely

$$A = \frac{1}{\dfrac{1}{g_m R_L} + \dfrac{1}{\mu}}$$

and it will be seen therefore that from equation (4)

$$\mu' = \mu(\mu + 1) \qquad (5)$$

$$g_m' = g_m(\mu + 1)/(\mu + 2) \qquad (6)$$

and therefore $r_p' = \dfrac{\mu'}{g_m'} = (\mu + 2)r_p$ (7)

It is preferable to make R_L very much greater than r_p in order to avoid distortion in V_2 due to the low load resistance into which it works ; a value of $2\mu r_p$ is satisfactory

For example, consider a twin triode with $\mu = 20$ and $r_p = 10\,000$ ohms under resistance-coupled conditions. A suitable value for the load resistance is $2\mu \times 10\,000 = 400\,000$ ohms—say 0.5 megohm.

From eqn. (5) : $\mu' = 20 \times 21 \times 500\,000 \div 510\,000 = 412$

From eqn. (6) : $g_m' = \mu/r_p = 20/10\,000 = 2000$ μmhos.

From eqn. (7) : $r_p' = 21 \times 10\,000 \times 500\,000 \div 510\,000 = 206\,000$ ohms.

A high-mu triode would show even higher values of amplification factor and plate resistance, resembling those of a sharp cut-off r-f pentode.

Curves have been drawn for some typical twin triodes operating as cascode amplifiers ; they resemble the curves of pentodes except that the rounded knee has been replaced by a nearly straight, sloping line (Ref. D35).

One special application is as a low-noise r-f amplifier (Refs. D33, D35).

Another application is as a voltage stabilizer (Refs. D32, D34).

The cascode amplifier has been used as a direct current amplifier responding to zero frequency (Ref. D35).

Two cascode amplifiers have been used in a " floating paraphase " push-pull amplifier operating with single-ended input, to deliver a balanced output. In this application, two high-mu twin-triode type 6SL7-GT valves were used each as a cascode amplifier, to deliver an output of about 30 volts peak, each side. Only one coupling capacitor was used in the whole stage, thus simplifying the design of the feedback circuit (Ref D41).

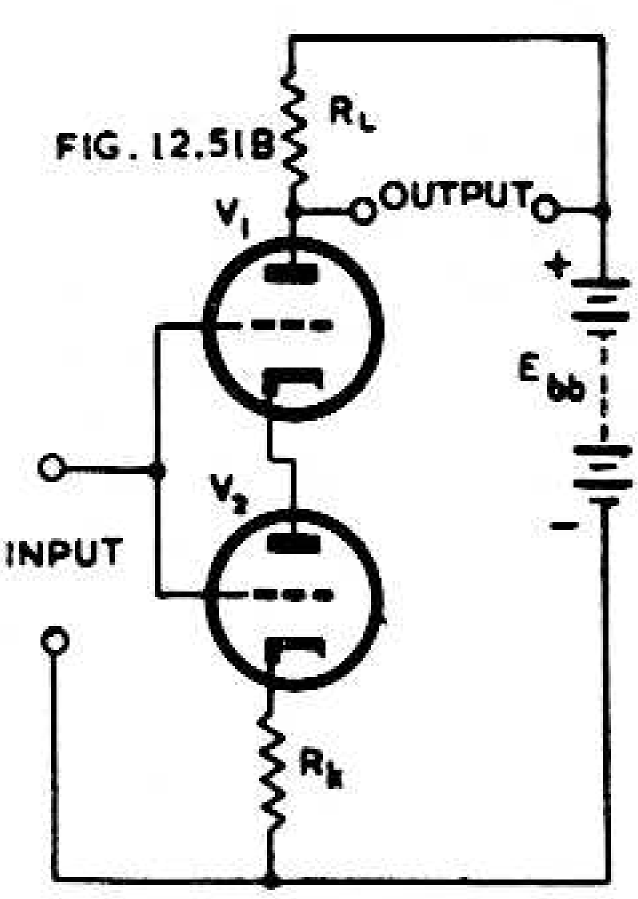


Fig. 12.51B. *Fundamental circuit of cascode amplifier.*

References to cascode amplifiers : D32, D33, D34, D35, D36, D41.

References to direct-coupled amplifiers (general) Refs. (D).

Ref. D36 is particularly valuable as it gives a detailed examination of the whole subject, including all causes of " drift." See also Chapter 2 Sect. 2(vii) Drift of characteristics during life, and (viii) Effect of heater voltage variation.

SECTION 10 : STABILITY, DECOUPLING AND HUM

(i) Effect of common impedance in power supply (ii) Plate supply by-passing (iii) Plate circuit decoupling (iv) Screen circuit decoupling (v) Grid circuit decoupling (vi) Hum in voltage amplifiers.

(i) Effect of common impedance in power supply

Every form of power supply has some impedance—even a dry battery has appreciable internal resistance, particularly when partially discharged. This is represented by the resistance R_4 in Fig. 12.52.

When two circuits operating at the same frequency have an impedance common to both there is coupling between them, and the phase relationships may be such that the coupling is either regenerative or degenerative. In the former case instability may result.

A two-stage resistance coupled a-f amplifier has degenerative coupling through the common power supply since the plate currents are out of phase.

A three stage r.c.c. amplifier (Fig. 12.52) has the signal plate currents of the first and third stages in phase, and the total signal current through R_4 is $I_{p1} + I_{p3} - I_{p2}$. Since $(I_{p1} + I_{p3})$ is greater than I_{p3}, the resultant signal current through R_4 will be in the direction of I_{p3}, thus causing across R_4 a signal frequency voltage drop E_4. As a result, the signal voltage applied to the grid of V_2 through R_{L1} and C_1 will be in phase with the normal signal on the grid of V_2, thus giving positive feedback. If the gain of the amplifier is high, there may be sufficient positive feedback to cause oscillation. This effect may be prevented by the use of decoupling.

On the subject of stability, see Chapter 7 Sect. 3 (stability with feedback), also Ref. E1 (requires a high standard of mathematics).

FIG. 12.52

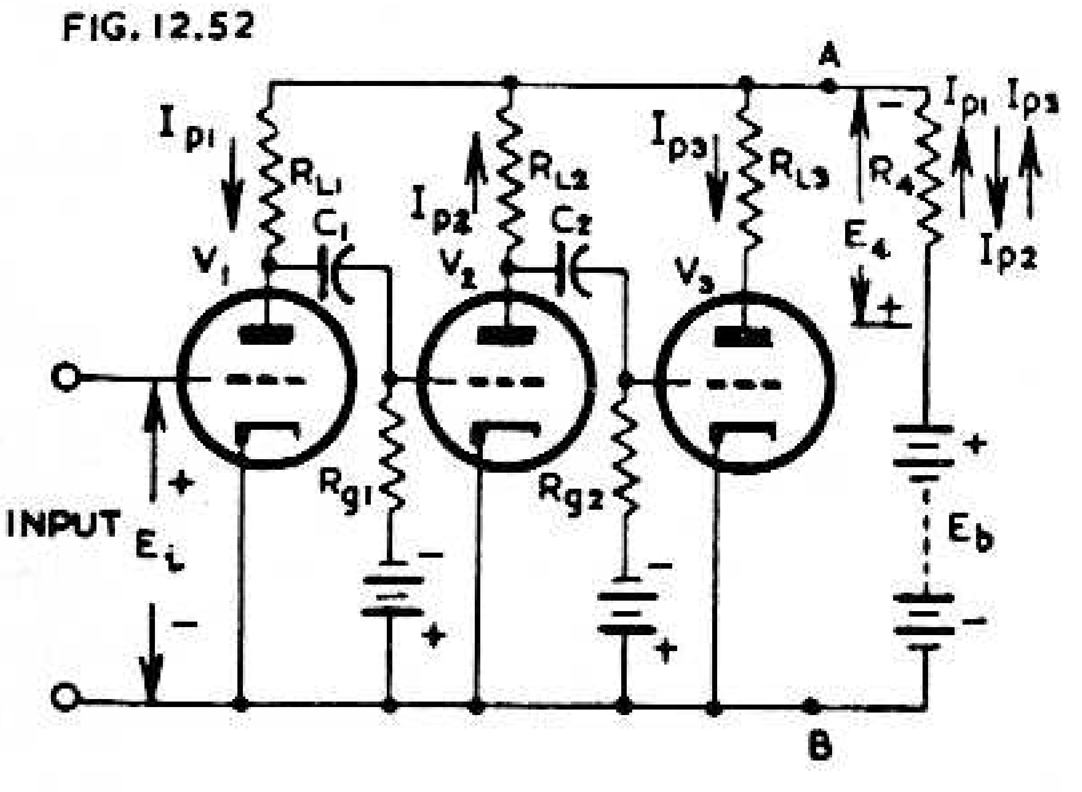


Fig. 12.52. Three stage amplifier demonstrating the effects of impedance (R_4) in the power supply.

FIG. 12.53

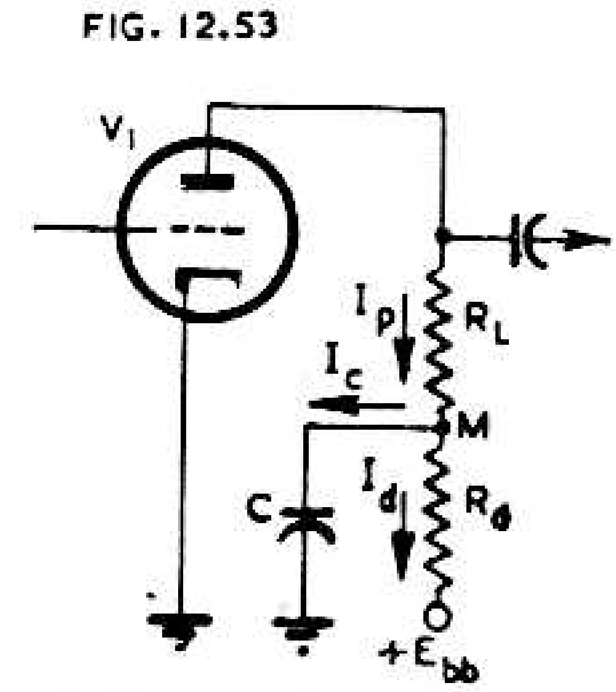


Fig. 12.53. Plate circuit decoupling.

(ii) Plate supply by-passing

The simplest form of decoupling, but one having limited usefulness, is a large capacitance across the power supply (points A and B in Fig. 12.52). This reduces the effective power supply impedance to a low value except at very low frequencies, but it may not be sufficient to prevent low frequency instability (" motor-boating ") and frequently requires to be supplemented by plate circuit decoupling.

This by-pass capacitance also fulfils a useful purpose in that it completes the circuit for signal frequencies without appreciable signal currents passing through the B battery or power supply.

Electrolytic condensers are generally used ; if paper condensers are used they should be of a type having low inductance. Either type of condenser may need, in certain rare cases, to be shunted by an additional condenser with a small value of capacitance.

(iii) Plate circuit decoupling

The most popular method is illustrated in Fig. 12.53 where R_d is the decoupling resistance and C the decoupling capacitance. The signal plate current I_p divides between the path through C and the path through the B supply in the ratio

$$\frac{I_c}{I_d} = \frac{R_d + R_4}{X_c} \quad (1)$$

where R_4 = resistance of B supply
and $X_c = 1/\omega C = 1/(2\pi f C)$.

If X_c is very much less than $R_d + R_4$ then almost the whole of the signal current will pass through C, thus reducing the coupling through the B supply. If the value of R_4 is unknown, it may be neglected as an approximation, since R_d is usually much greater than R_4.

The decoupling circuit comprising C and R_d is actually a resistance-capacitance filter [for theory see Chapter 4 Sect. 8(ii)]. The frequency at which the current divides equally between the two paths is given by

$$f_1 \approx 1/2\pi R_d C \quad (2)$$

A normal minimum value of $R_d C$ is $10\,000 \times 10^{-6}$, giving a time constant 0.01 second and $f_1 \approx 16$ c/s. Typical combinations for this value of RC are :

R_d	5000	10 000	20 000	40 000	100 000	ohms
C	2	1	0.5	0.25	0.1	μF

Higher values of C may, of course, be used if desired ; these higher capacitances are often necessary in pre-amplifiers where several stages operate from the same power supply.

The total d.c. load resistance in the plate circuit is $R_L + R_d$ ($+ R_4$ if desired), and the quiescent operating conditions should be based on this value. On the other hand, the dynamic (a.c.) load is approximately R_L.

The dynamic operating conditions (gain, voltage output etc.) may be determined by referring to the published data for a supply voltage the same as that for point M (Fig. 12.53). The voltage from M to earth is given by

$$E_M = E_{bb} - I_b R_d \quad (3)$$

where $I_b = KE_{bb}/R_L$
and the value of K (for low level operation) may be taken as approximately
$K = 0.75$ for general purpose triodes and pentodes
$K = 0.6$ for high-mu triodes
[for more exact values see Sects. 2 (vi) for triodes and 3 (vi) for pentodes].

A good general value of R_d for most purposes is one fifth of R_L. If the stage is operating at low level, R_d may be increased up to about the same resistance as R_L to give better decoupling. If the stage is operating at high level, R_d should be reduced as much as possible provided that sufficient decoupling can be maintained. This may be assisted by increasing C, but the cost and size of paper condensers may set a limitation. Two stages of decoupling can sometimes be used to advantage. Electrolytic condensers should be used with caution, since their leakage currents are appreciable and they tend to cause noise if used in low level stages. They should never be used with values of R_d above 50 000 ohms, and very much lower values are desirable.

The plate decoupling circuit of Fig. 12.53 has the effect of increasing the bass response, since at extremely low frequencies the total plate load becomes very nearly $(R_L + R_d)$. If V_1 is a triode, and R_L is greater than 5 r_p, then the effect is slight. If V_1 is a pentode, or a triode with R_L less than 5 r_p, the effect may be appreciable. The increase of gain and the phase angle are both identical in form (except for the sign) with the loss of gain and phase angle caused by the cathode bias resistor and by-pass condenser—see Section 2 (iii) for triodes and 3 (iv) for pentodes ; also Ref. E3 for curves.

The limiting increase in gain is given by

$$\left|\frac{A'}{A}\right| = 1 + \frac{1}{(R_L^2/R_d)(1/r_p + 1/R_g) + R_L(1/r_p + 1/R_g + 1/R_d)} \quad (4)$$

or if $r_p \gg R_L$ as for pentodes,

$$\left|\frac{A'}{A}\right| \approx 1 + \frac{R_g R_d}{R_L(R_L + R_d + R_g)} \quad (5)$$

or if the effect of R_g may be neglected,

$$\left|\frac{A'}{A}\right| \approx 1 + \frac{r_p R_d}{R_L(R_L + R_d + r_p)} \tag{6}$$

Example : Pentode with $R_L = 0.25$, $R_d = 0.05$, $R_g = 1$ megohm.
$A'/A \approx 1.15$.

This increase of gain at low frequencies may be exactly cancelled by a suitable choice of cathode bias resistor and by-pass condenser (Refs. E3, E5) provided that R_L and R_d are small compared with r_p and R_g : for triodes—

$$R_k = R_d/R_L g_m \text{ and } C_k R_k = C R_d \tag{7}$$

and then $A = g_m r_p R_L/(r_p + R_L)$. (8)

This generally gives a value of R_d greater than R_L, and is of limited usefulness.

A special circuit giving **plate decoupling without decoupling condensers,** suitable for use at low frequencies, is shown in Fig. 12.54 (Ref. E4). The plate currents of the last two valves, which are in phase opposition, flow through separate impedances Z_1, Z_2, of such value that the back e.m.f.'s developed across them and the common impedance Z_3 of the plate supply balance out at the junction points J1, J2. The screens and plates of the preceding valves are supplied from J2 and J1 through pairs of resistances R, R and R_1, R_1 in order to balance out undesirable ripple.

The need for decoupling is less when **push-pull operation** is used, although there is still the possibility of motor-boating arising from the effect on earlier stages of plate supply voltage variations caused by the final stage. The latter are small when pure Class A operation is employed.

Chokes may be employed instead of resistors for decoupling, and are frequently used when the plate currents are large—e.g. power amplifiers.

It is desirable in all cases to reduce the internal resistance of the power supply, so that even at very low frequencies there may be no tendency towards the production of relaxation oscillations (" motor-boating "). Power supplies having good regulation have low effective internal impedance ; thermionic valve type voltage regulators are highly desirable for special applications and have the feature of retaining low internal impedance characteristics down to the lowest frequencies, if correctly designed.

Motor-boating in transformer-coupled amplifiers may frequently be cured by reversing the transformer connections on either primary or secondary. This has the effect of providing degeneration instead of regeneration, and may have an adverse effect on the frequency response (Ref. E3).

FIG. 12.54

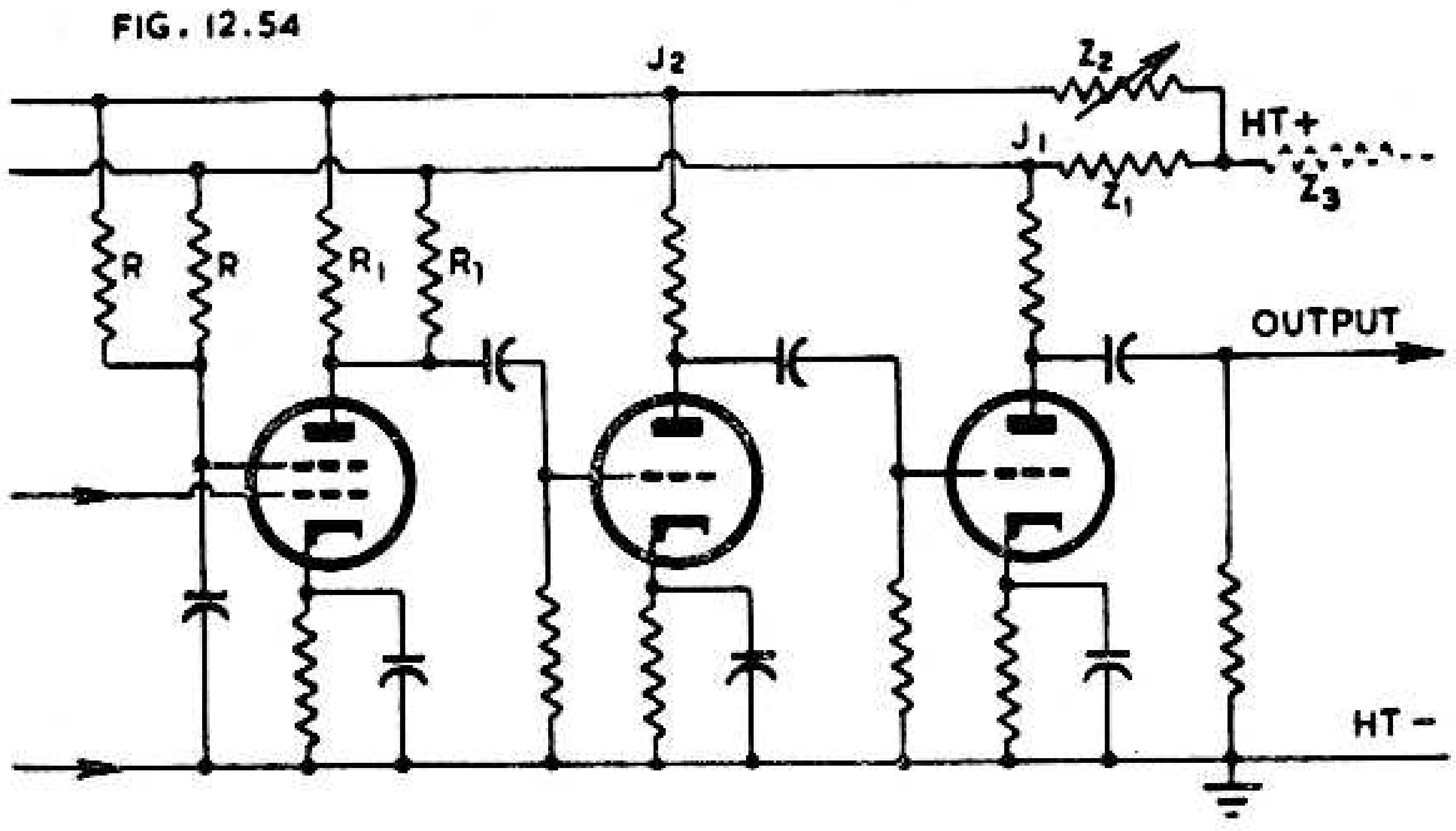


Fig. 12.54. *Special circuit giving plate decoupling without decoupling condensers (Ref. E4).*

(iv) Screen circuit decoupling

Resistance-coupled pentodes in multistage amplifiers should always have their screens separately decoupled from B+. By this means any impedance in the screen supply may be prevented from causing instability, and the hum is much reduced.

(v) Grid circuit decoupling

Fig. 12.55 is a method of applying grid-circuit decoupling with transformer input and cathode bias. The condenser C acts as a by-pass to resistors $R + R_k$ and if its reactance is small at the lowest frequencies to be amplified, a negligible portion of the signal voltage will occur across C at all signal frequencies. This method requires a smaller value of C than for cathode bias in the ratio $R_k : (R + R_k)$. This circuit is only effective with transformer coupling.

A similar arrangement for fixed bias is shown in Fig. 12.56, which may be used with any form of coupling.

FIG. 12.55

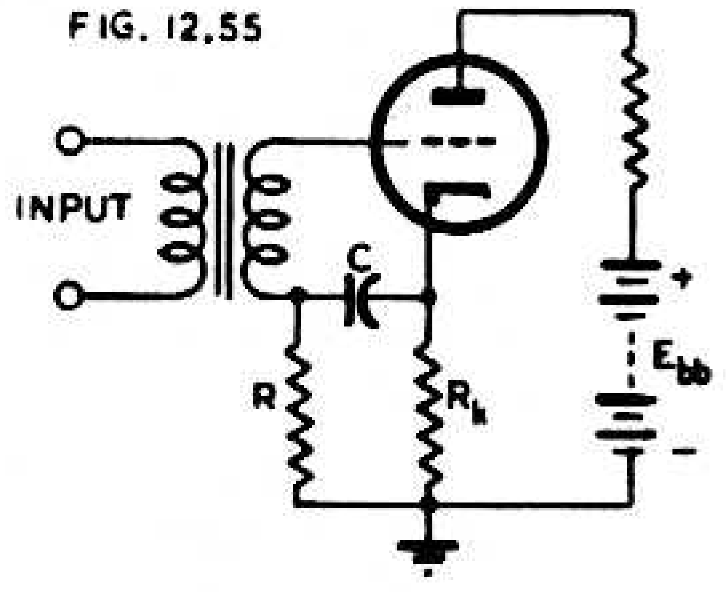


Fig. 12.55. *Grid circuit decoupling with transformer input and cathode bias.*

FIG. 12.56

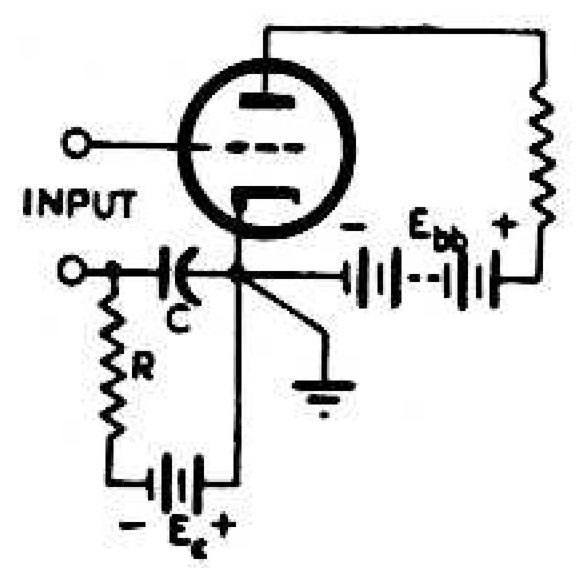


Fig. 12.56. *Grid circuit decoupling with fixed bias.*

(vi) Hum in voltage amplifiers

(A) Hum from plate supply

The plate supply is generally filtered sufficiently for the final stage, but additional filtering is usually necessary for the earlier stages. This is usually provided by the plate decoupling circuits, which may be made more effective than demanded by stability in order to reduce the hum.

In a normal r.c.c. amplifier, such as Fig. 12.1, the hum voltage passed on to the following stage is given by

$$E_h' = E_h r_p/(R + r_p) \qquad (9)$$

where E_h = hum voltage across the plate power supply
E_h' = hum voltage across R_{g2}
r_p = effective plate resistance of V_1, taking into account any feedback
and $R = R_L R_{g2}/(R_L + R_{g2})$.

If r_p is very much less than R, the hum voltage passed on to the following stage is much less than E_h. On the other hand, if r_p is greater than R, then E_h' may approach E_h.

With a transformer-coupled amplifier (Fig. 12.23) the corresponding expression becomes, in vector form,

$$E_h' = E_h\mathbf{Z}/(\mathbf{Z} + r_p) \qquad (10)$$

where E_h' = hum voltage across transformer primary
and $\mathbf{Z}$ = input impedance of transformer at hum frequency.
Normally $\mathbf{Z} \gg r_p$, and E_h' approaches E_h.

In the case of parallel-feed (Fig. 12.24) the expression becomes, in vector form,

$$E_h' = E_h\mathbf{Z}'(\mathbf{Z}' + R_L) \qquad (11)$$

where $\mathbf{Z}' = r_p\mathbf{Z}/(r_p + \mathbf{Z}) \approx r_p$.

The hum passed on to the following stage is normally only a small fraction of E_h.

The hum voltages of successive stages are usually out of phase, thus resulting in some cancellation, but this is very slight if the same degree of filtering is provided for all stages. In such a case the hum of the first stage is the only one to be considered.

For amplifiers using **negative feedback,** see Chapter 7 Sect. 2(ix).

(B) **Hum from the screens** may also be reduced by screen decoupling. The hum from the screen is out of phase with the hum from the plate, but the screen hum predominates. For amplifiers using negative feedback, see Chapter 7 Sect. 2(ix).

(C) **Hum from the grid bias supply** may be reduced in some cases by grid decoupling. This hum voltage may be either in phase, or out of phase, with the plate circuit hum, depending on the source.

(D) Hum neutralization

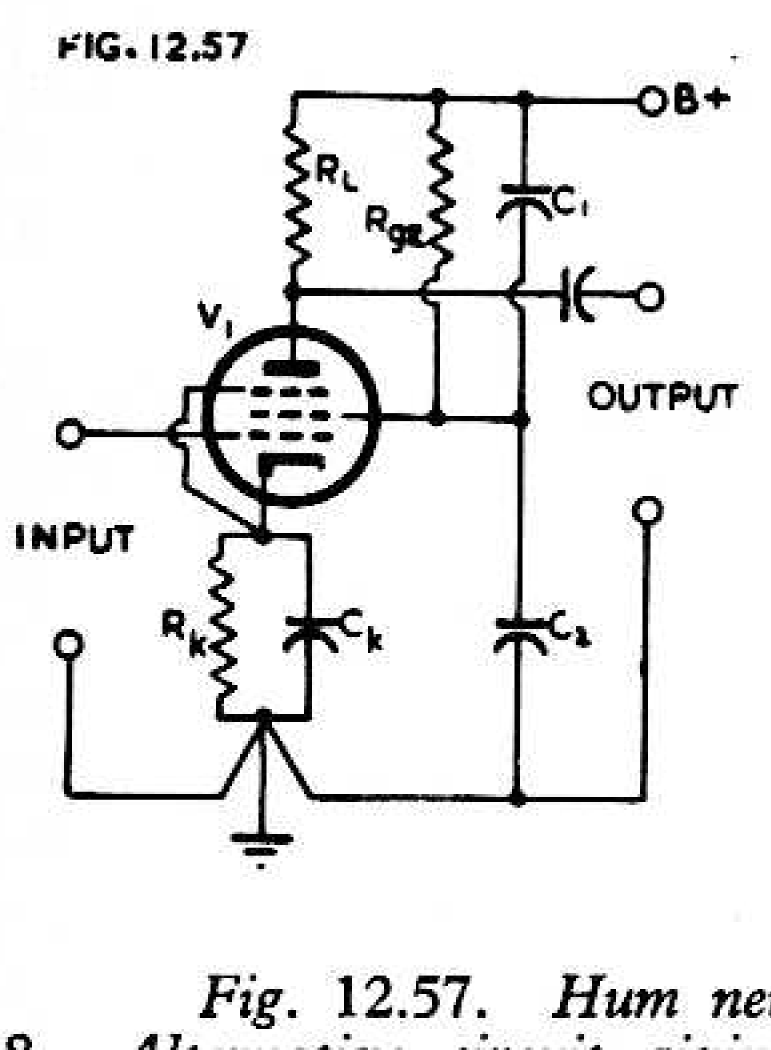
In a r.c.c. pentode the hum voltages from the plate and screen are out of phase, so that it is possible to neutralize the hum by a circuit such as Fig. 12.57. Here the screen is fed from a series resistance R_{g2}, and the hum voltage is applied to the screen by the capacitance divider C_1C_2. If V_1 is type 6J7 with $R_L = 0.25$ and $R_{g2} = 1.5$ megohms, minimum hum is obtained when $C_1 = 0.05$ and $C_2 = 0.5\ \mu$F, the hum then being about 14% of that without neutralization. C_1 and C_2 may have $\pm$ 10% tolerances in quantity production. For perfect neutralization it is necessary to balance both resistive and capacitive elements.

An alternative form (Ref. E2) is shown in Fig. 12.58 in which, for perfect neutralization at all frequencies,

$$\frac{C_k}{C_1} = \frac{R_{L1}}{R_g}\left(1 + \frac{R_g}{r_{p1}}\right) \text{ and also } \frac{C_3}{C_1} = \frac{R_k}{R_g}\left(1 + \frac{R_{L1}}{r_{p1}}\right) \qquad (12)$$

where r_{p1} = plate resistance of V_1 (preceding stage) and the other values are marked on the diagram. This circuit gives neutralization both for hum and for low frequency regeneration.

There are other methods which may also be used for hum neutralization or reduction (see Chapter 31 Sect. 5).



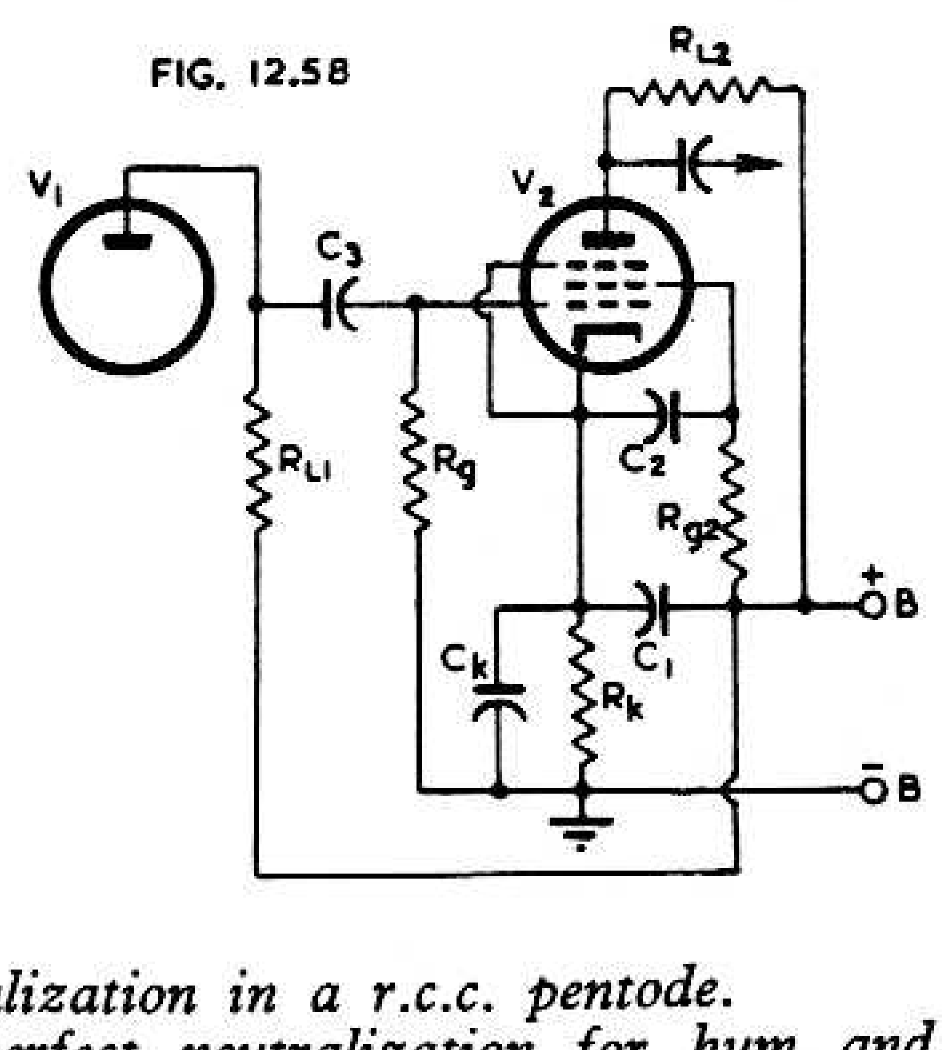


Fig. 12.57. *Hum neutralization in a r.c.c. pentode.*
Fig. 12.58. *Alternative circuit giving perfect neutralization for hum and for low frequency regeneration at all frequencies (Ref. E2).*

(E) Hum caused by inductive coupling

The most common cause is an a-f transformer or choke, which may be placed in an electromagnetic field. This effect may be minimized by altering the position of the transformer or choke, but cannot usually be eliminated entirely. The best procedure is to avoid using the transformer, to remove the cause of the electromagnetic field, or to employ a properly shielded transformer.

Some valves exhibit hum when they are placed in a strong field—the cure is obvious.

See also Chapter 18 Sect. 2(iii) on pre-amplifiers, and Chapter 31 Sect. 4 on hum.

(F) Hum caused by electrostatic coupling

This may be almost entirely eliminated by suitable placing of the mains and rectifier leads, and by electrostatic shielding of all low-level wiring (especially the first grid lead) and by shielding the valves and other susceptible components. See also Chapter 18 Sect. 2(iii) on pre-amplifiers.

(G) Hum due to heater-cathode leakage or emission

When a valve is operated at a fairly high input level, it is often sufficient to earth one side of the heater supply and to use a large by-pass condenser (40 μF or more) with cathode bias, or an earthed cathode with fixed bias.

At a somewhat lower input level, hum due to heater-cathode leakage or emission may be reduced as required, by one or more of the following devices—

1. Centre-tapped heater supply, with tap earthed.
2. Centre-tapped heater supply, with tap connected to a fixed positive or negative voltage (say 15 or 20 volts bias).
3. Potentiometer across heater supply, with moving arm connected to earth and adjusted for minimum hum.
4. Potentiometer across heater supply, with moving arm connected to a positive or negative voltage (not more than 50 volts maximum) selected experimentally to give minimum hum, in conjunction with adjustment of the potentiometer.

For further information see Chapter 18 Sect. 2 (pre-amplifiers) and Chapter 31 Sect. 4 (hum).

SECTION 11 : TRANSIENTS AND PULSES IN AUDIO FREQUENCY AMPLIFIERS

(i) Transient distortion (ii) Rectangular pulses.

(i) Transient distortion

A transient is a complex wave which does not repeat periodically; it may be analysed into a fundamental and harmonic frequencies (see Chapter 6 Sect. 8). If an a-f amplifier contains a tuned circuit, e.g. in a compensated wide band amplifier, or in a tone control stage, distortion of transients can occur if the damping of the tuned circuit is too slight. The effect is actually shock excitation of the tuned circuit; this may be reduced to acceptable proportion if Q does not exceed 0.7 (Ref. A12 p. 428).

(ii) Rectangular pulses

The design of special pulse amplifiers is beyond the scope of this handbook, but amplifiers are frequently tested with a rectangular input voltage (" square wave ") and their performance under these conditions is of interest.

FIG. 12.59

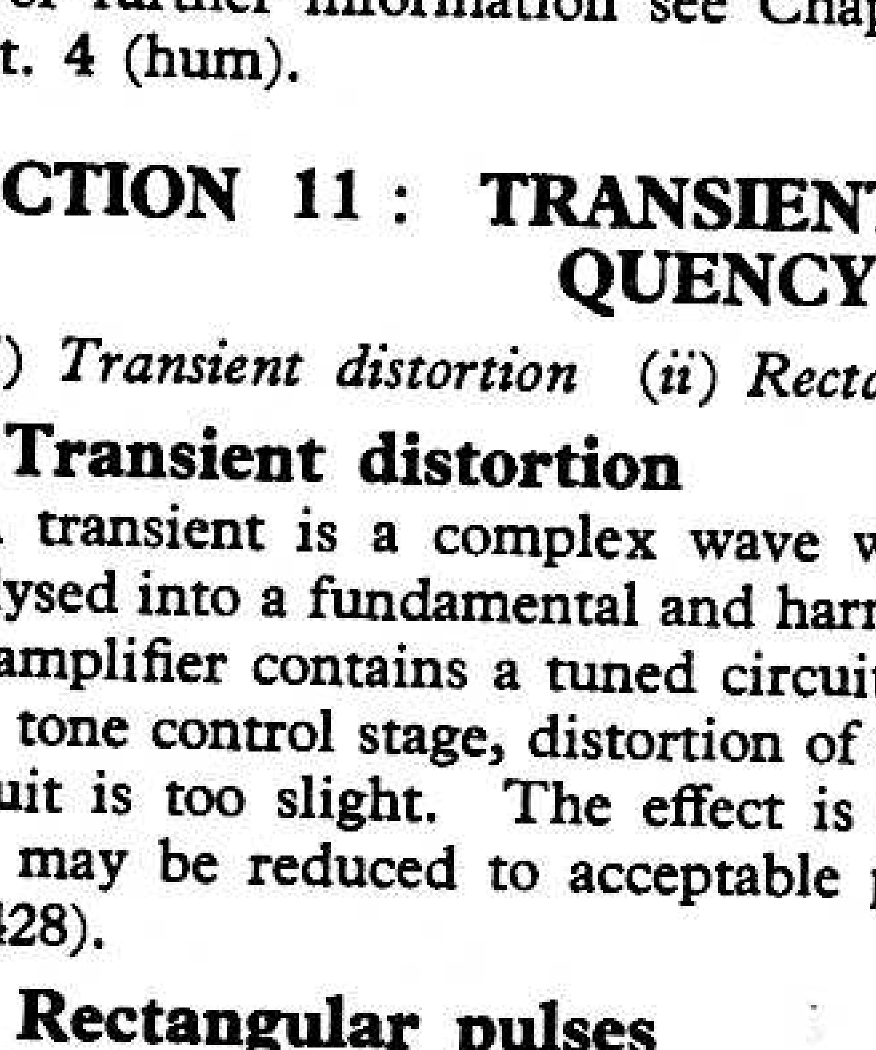


Fig. 12.59. A single resistance-capacitance coupling and its response to a rectangular waveform.

In the case of a single r.c. coupling (Fig. 12.59), the voltage across R is given by

$$e = E\epsilon^{-t/RC} \qquad (1)$$

where E = amplitude of pulse voltage input

$R = R_g + r_pR_L/(r_p + R_L) \approx R_g$

and t = time measured from commencement of pulse, in seconds.

If t is small compared with RC, we may write the equation as the approximation (Ref. F1)

$$e \approx (1 - t/RC)E \qquad (2)$$

or $E/(E - e) \approx RC/t =$ (say) X (3)

the error being negligible if t is not greater than 0.1 RC.

The value of the time constant RC to give any desired value of X at the end of a pulse is given by

$$RC \approx TX \qquad (4)$$

where T = time length of pulse,

and $X = E/(E - e)$,

provided that X is not less than 10.

For example, if X is required to be 20 (i.e. the amplitude of the square top falls by 5% at the end of the pulse) and if $T = 0.01$ second, then $RC \approx 0.2$ second. If $R = 0.5$ megohm, then $C = 0.4\ \mu F$.

If the pulses are repeated periodically with the length of pulse equal to the time between pulses, and if the applied voltage is zero during the period between pulses (as Fig. 12.59),

then $RC \approx X/2f_1$ (5)

where f_1 = frequency in cycles per second.

With sine-wave input, the frequency (f_0) at which the response of a single-stage r.c.c. amplifier falls by 3 db is given by

$f_0 = 1/(2\pi RC)$ (6)

But with " square-wave " input, $RC \approx X/2f_1$ (7)

where f_1 = frequency of square wave in cycles per second.

Therefore $f_0 = f_1/\pi X$ (8)

By using a square wave input, and noting on a C.R.O. the frequency f_1 at which there is (say) 10% drop at the end of the pulse (i.e. $X = 10$), it is possible to calculate the frequency for 3 db attenuation with sine-wave input:

$f_0 \approx f_1/10\pi$ [for $X = 10$] (9)

Vice versa, by measuring the frequency f_0 at which the sine-wave response is 3 db below that at the mid-frequency, it is possible to calculate the frequency f_1 at which the square wave shows a specified drop at the end of the pulse:

$f_1 = X\pi f_0$ (10)

where $X = E/(E - e)$.

Equations (9) and (10) may also be used in connection with multi-stage amplifiers.

SECTION 12 : MULTISTAGE VOLTAGE AMPLIFIERS

(i) Single-channel amplifiers (ii) Multi-channel amplifiers.

(i) Single channel amplifiers

Almost any desired number of single r.c.c. stages may be connected in cascade, if adequate provision is made for decoupling. A practical limit is reached when the noise and hum from the first stage become excessive (see Chapter 18 Sect. 2—pre-amplifiers).

The total voltage gain of the amplifier (A) is given by

$A = A_1 \times A_2 \times A_3 \times \ldots$

where A_1 = voltage gain of first stage, etc.

For decibel calculations see Chapter 19 Sect. 1.

The attenuation in decibels below the mid-frequency gain (A_0) is given, for any frequency f, by the sum of the attenuations in decibels of the individual stages at the same frequency.

It is normal practice to design such an amplifier with a flat, or nearly flat, response over the desired frequency range, and to introduce one or more stages for either fixed tone compensation or manual tone control, or both (see Chapter 15). These stages usually give very limited gain at the mid-frequency.

(ii) Multi-channel amplifiers

Amplifiers having 2 or 3 channels are sometimes used. For example, a 3 channel amplifier may be preceded by a frequency dividing network such that each channel only amplifies a limited range of frequencies—low, middle and high. Each channel may have its own attenuator, with an additional attenuator for the whole amplifier (see Chapter 18). The three outputs may feed three separate power amplifiers, or they may be recombined to form a tone compensating amplifier (see Chapter 15).

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