Chapter 7

VALVE AMPLIFIERS

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(In this chapter many of the diagrams are drawn with the screen and suppressor grid connections omitted for simplicity. When this occurs it may be assumed that normal supply connections are made, the suppressor grid being connected to cathode and the screen to a dropping resistor or potentiometer with bypass condenser).

1. INTRODUCTION

The types of voltage variations most frequently requiring amplification in radar are illustrated in Figs. 322 and 323. The variations shown in Fig. 322 (a) and (b) are commonly termed voltage (or current) pulses; those illustrated in Fig. 323 (a) are called

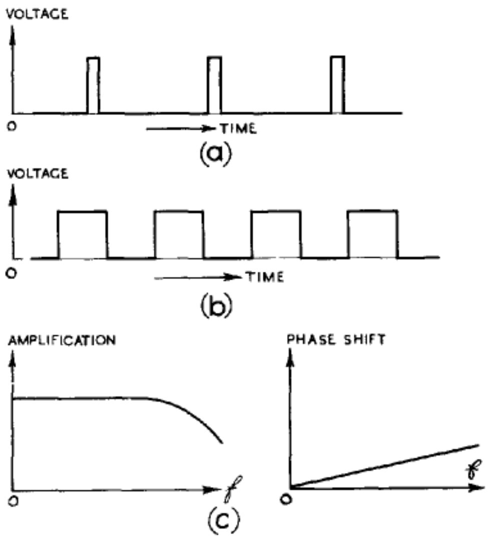

Fig. 322 - Typical voltage variations and the amplifier characteristics necessary for their amplification.

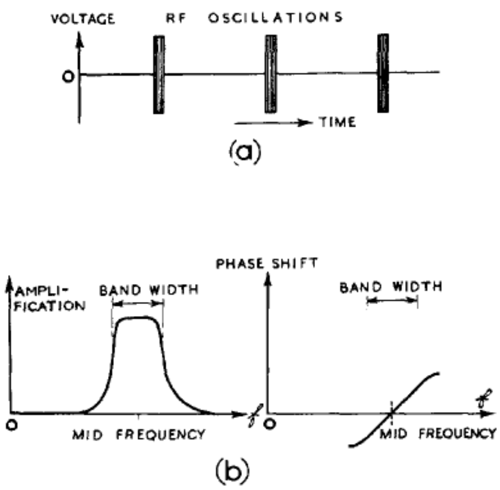

Fig. 323 - R F pulses and the amplifier characteristics necessary for their amplification.

RF pulses. Either of these types of voltage variation may be analysed into sinusoidal components covering a range of frequencies (Fourier analysis, Chap. 16 Sec. 1). In considering the behaviour of amplifiers it is sometimes desirable to consider the response of the circuit to these sinusoidal components. Alternatively, in the case of the pulses of the type shown in Fig. 322 it may be preferable to deal with the transient response of the amplifier to an idealised pulse without resorting to a detailed analysis.

Considering both pulses and RF pulses from the aspect of their sinusoidal components, the amplifiers required to handle such variations must possess characteristics of the types shown in Figs. 322 (c) and 323 (b) respectively.

2. TYPES OF DISTORTION

An amplifier is liable to produce three types of distortion, any one of which causes the form of the output voltage to be different from that of the input.

(i) Amplitude Distortion

This type of distortion is due to non-linearity of the valve characteristics over the operating range of grid and anode voltages. The effect is to introduce frequencies into the output which were not present in the input.

(ii) Frequency Distortion

This type of distortion is caused by unequal amplification of the components of different frequencies present in the input voltage.

(iii) Phase Distortion

This type of distortion occurs when the relative phases of the input components of different frequencies are not preserved in the output. However, if the phase-shift produced by the circuit is proportional to frequency all such components are delayed equally in time and the output, where there is no other distortion, is delayed, but distortionless.

3. CONVENTIONAL SYMBOLS AND EQUIVALENT CIRCUITS

Fig. 324 (a) shows the circuit arrangement of an amplifier with load z_l.

The following symbols are used :-

v_a is the potential difference between anode A and cathode K, and is reckoned positive if the anode is at the higher potential.
V_a is the steady component of v_a.
v_g is the potential difference between control grid G and cathode, and is reckoned positive if the grid is at the higher potential.
V_g is the steady (Bias) component of v_g. (V_g is usually negative).
V_B is the HT supply voltage.
i_a is usually (unless otherwise stated) taken to indicate the fluctuating component of the anode current, and is considered positive in the direction of conventional current flow, from anode to cathode.
I_a is the steady component of anode current.
v_i is the change of voltage at the grid (input voltage); i.e. the amount by which v_g exceeds V_g.
v_o is the change of voltage at the anode (output voltage); i.e., the amount by which v_a exceeds V_a.
v_s is the signal voltage applied to the input circuit (when different from v_g or v_i).
Capital letters A, G, K, used as subscripts, denote that the voltage concerned is measured from earth (or chassis) and not necessarily from cathode.
μ is the Amplification Factor of the valve, defined as the ratio of a small change in anode voltage to the corresponding small change in grid voltage provided the anode current in the valve remains unchanged.

R_a is the Slope Resistance (Differential or Output Resistance) of the valve, defined as the ratio of a small change in anode voltage to the corresponding small change in anode current provided the grid voltage of the valve remains constant.
G_m is the Mutual Conductance (Transconductance) of the valve, defined as the ratio of a small change of anode current to a small change of grid voltage, the anode voltage remaining constant.

If i_a, v_i and v_o are corresponding small changes in anode current, grid voltage and anode voltage respectively, the valve constants may be written :-

$G_m = \frac{i_a}{v_i}$, where
anode voltage is constant,

$\mu = -\frac{v_o}{v_i}$, where

anode current is constant

and
$R_a = \frac{v_o}{i_a}$, where
grid voltage is constant.

It can be shown that
$\mu = G_m R_a$

x x x <u>Proof of formula $\mu = G_m R_a$</u>

For any valve there is a relation between the anode current i_a, anode-cathode voltage v_a, and the grid-cathode voltage v_g.
We may denote this relation by the equation

$$f(i_a, v_a, v_g) = 0 \qquad \text{.......................(1)}$$

From the theory of partial derivatives we obtain

$$\frac{\partial f}{\partial i_a} di_a + \frac{\partial f}{\partial v_a} dv_a + \frac{\partial f}{\partial v_g} dv_g = 0 \qquad \text{..................... (2)}$$

Putting $di_a = 0$, $dv_a = 0$, $dv_g = 0$ in succession in (2), we have

$$-\mu = \frac{\partial v_a}{\partial v_g} = -\frac{\partial f}{\partial v_g} \Big/ \frac{\partial f}{\partial v_a},$$

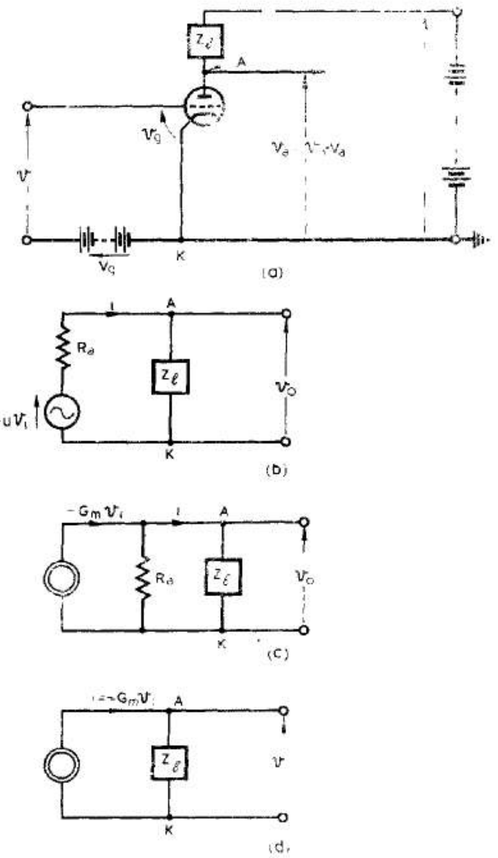

Fig. 324 - Equivalent circuits.

so that $\mu = \frac{\partial f}{\partial v_g} \Big/ \frac{\partial f}{\partial v_a}$;

$G_m = \frac{\partial i_a}{\partial v_g} = - \frac{\partial f}{\partial v_g} \Big/ \frac{\partial f}{\partial i_a}$,

and $R_a = \frac{\partial v_a}{\partial i_a} = - \frac{\partial f}{\partial i_a} \Big/ \frac{\partial f}{\partial v_a}$.

$$\text{Hence } G_m R_a = \frac{-\frac{\partial f}{\partial v_g}}{\frac{\partial f}{\partial i_a}} \cdot \frac{-\frac{\partial f}{\partial i_a}}{\frac{\partial f}{\partial v_a}} = \frac{\frac{\partial f}{\partial v_g}}{\frac{\partial f}{\partial v_a}} = \mu \; .$$

Stage Amplification and Equivalent Circuits

The expression for the voltage amplification provided by the stage is :-

$$m = \frac{v_o}{v_i}$$

If v_i, and hence v_o, vary sinusoidally then :-

$$m = - \frac{\mu z_\ell}{R_a + z_\ell} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (3)$$

In general m is complex; its magnitude will be denoted by $|m|$, and is equal to $\frac{\hat{v}_o}{\hat{v}_i}$.

The circuit of Fig. 324 (a) may be represented, from the point of view of fluctuating voltages, by the equivalent circuit shown in Fig. 324 (b). In this circuit the current i is given by

$$i = - \frac{\mu v_i}{z_\ell + R_a} ;$$

hence $\frac{v_o}{v_i} = \frac{i . z_\ell}{v_i} = - \frac{\mu z_\ell}{z_\ell + R_a} = m$, justifying the representation.

Equation (3) can be rearranged as follows :-

$$m = \frac{- \frac{\mu}{R_a} \cdot z_\ell \cdot R_a}{R_a + z_\ell}$$

$$= - G_m \frac{R_a \; z_\ell}{R_a + z_\ell} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4)$$

Equation (4) suggests the form of equivalent circuit shown in Fig. 324 (c). This is normally referred to as the Constant-Current Generator Equivalent Circuit whilst the other (Fig. 324 (b)) is termed the Constant-Voltage Generator Equivalent Circuit. The two circuits express the same relationship between v_i and v_o.

If the valve used is a pentode whose slope resistance is large compared with the load impedance the arrangement shown in Fig. 324 (c) simplifies to that shown in Fig. 324 (d).

If the simple equivalent circuits of Fig. 324 (b), (c) and (d) are compared with the actual circuit (a) it is seen that z_ℓ lies between A and K in the equivalent circuits and not A and HT+. This representation is possible because the impedance of the HT supply to alternating voltages is very low, so that, as far as such voltages are concerned, the upper end of the load is connected to cathode.

The fluctuating component of the anode current (Fig. 324(a)) is -i, i having the sign convention as shown in the figure.

4. BIAS FOR AMPLIFIERS

In using a valve as an amplifier, or for other purposes, it is common practice to use the current flowing through the valve, in conjunction with suitable circuits, to maintain the cathode at a fixed steady potential relative to the mean potential of the grid. The steady potential of the grid in the absence of any input voltage is usually earth potential, since the grid is normally connected to earth by means of a grid-leak resistor R_g and isolated from steady potentials on previous circuits by means of a condenser C: (Fig. 325). If a succession of rectangular voltage pulses is applied to C and R_g in series, the time-constant C R_g being of suitable duration, a succession of rectangular pulses is developed at the grid, having the same mean potential as the steady potential at the grid in the absence of the pulses. Assume that the succession of pulses at the grid of the amplifier is as shown in Fig. 326. The duration of each positive-going pulse is T_1 and the repetition period of the pulses is T_2,

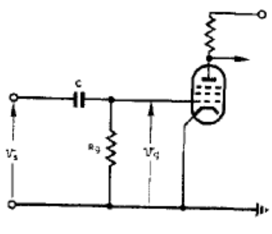

Fig. 325 - Common input network.

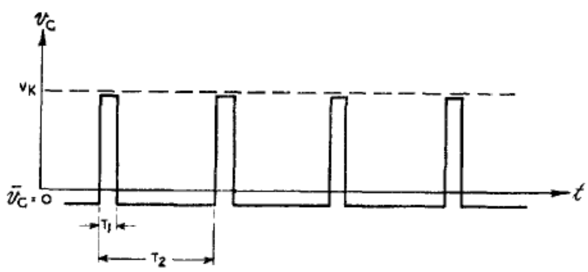

Fig. 326 - Choice of Bias (Input between grid and earth).

T_2 being much larger than T_1. Then if the cathode of the valve is at earth potential each positive-going pulse of voltage will carry the grid potential above that of the cathode and grid current will flow. This will introduce considerable distortion if the output impedance of the previous stage is at all large (see Chap. 9 Sec. 3).

In this case, therefore, it is necessary to make the steady potential of the grid more negative than the steady cathode potential i.e. a negative bias is required. The steady value of the cathode voltage should be chosen as illustrated in Fig. 326. If the succession of rectangular pulses is inverse to that shown in Fig. 326, i.e., the pulses are negative-going, it is not necessary, and may be inadvisable, for the valve to have an appreciable negative grid bias. In this case a negative grid bias might lead to the negative-going pulses cutting off the valve current.

If a negative grid bias is required it may be obtained by allowing the current flowing through the valve to pass through a resistor connected between the cathode and earth; (Fig. 327). The current in the valve varies with the potential at the grid, and unless negative feed-back for alternating currents is required (Sec.16) it is necessary to have a bypass condenser connected across the bias resistor; (Fig. 327). The condenser C_K charges so that the potential difference between its plates is $R_K \bar{I}$, where $\bar{I}$ is the mean current flowing through the bias resistor R_K. If the time-constant $C_K R_K \gg T_2$ the condenser cannot discharge appreciably within the repetition period T_2 of the applied pulses, and the cathode of the amplifier is maintained at a steady potential.

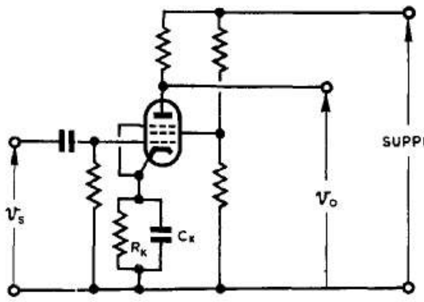

Fig. 327 - Use of cathode self-biasing circuit.

The method of using a cathode resistor in order to obtain negative grid bias has its limitations. Provided the valve is operated over the linear portion of its dynamic characteristic, variation of signal amplitude does not affect the mean valve current and the bias therefore remains constant. If the input voltage is allowed to vary over the curved portion of the characteristic, the mean current and hence the bias are affected. This means that the bias varies with signal amplitude. For Class B or C operation, therefore, some other system of biasing should be employed.

A convenient method of biasing a valve beyond cut-off or on the curved portion of the dynamic characteristic is illustrated in Fig. 328. Provided R_1 is small enough the variations of current through the valve are small compared with the current through R_1 and the bias remains approximately constant. The condenser C_K smooths out those fluctuations which still occur.

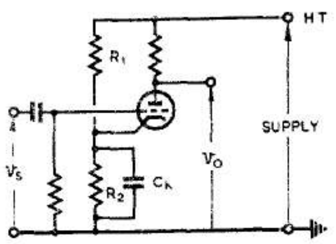

Fig. 328 - Circuit for class B or C biasing.

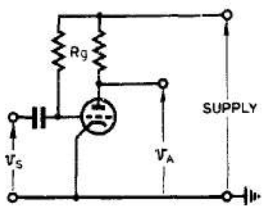

Fig. 329 - Connection providing small positive bias.

Alternatively, the cathode biasing arrangement may be dispensed with, and the grid leak connected to an additional source of negative potential. Although it is not always convenient to employ a separate bias supply voltage, this method has the advantage that the bias is independent of the amplifier valve current.

A few volts positive bias may be obtained by connecting the grid leak to the positive HT supply instead of to earth (Fig. 329). The flow of grid current through R_g should be just sufficient to maintain the grid voltage at the required bias level. For example, suppose the grid-current, grid-voltage conductance is 62 µA/volt, and the supply voltage is 250 volts. A grid leak of 2 MΩ is just sufficient to ensure that the grid is maintained at +2V relative to cathode. The grid current is 124 µA, which develops across the grid leak 248 volts.

This type of biasing arrangement is suitable for the amplification of negative-going pulses.

In some circuits it is necessary for the mean level of the input voltage to be positive with respect to earth, although the bias developed between grid and cathode may be zero or negative. Examples of this arise in current feedback circuits (Sec. 16) and in the limiter circuits of Chap. 9 Sec. 3.

These are illustrated in Figs. 330 (a) and (b) respectively. The steady voltage at B may be adjusted between the values zero and $\frac{R_2 V_B}{R_1 + R_2}$.

In the arrangement shown in Fig. 331 the steady voltage at B can be adjusted so that it has a value either positive or negative with respect to that of the cathode.

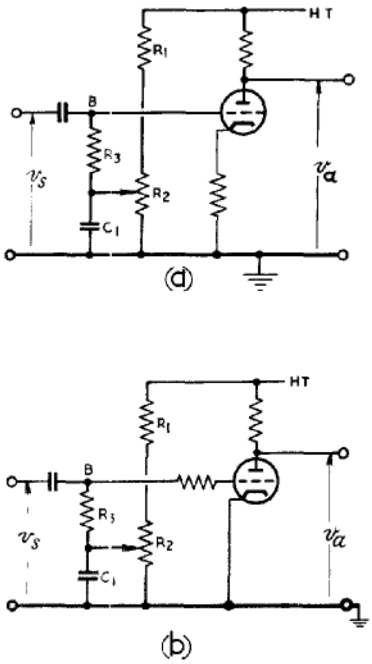

Fig. 330 - Arrangements for providing a positive mean level of input voltage with respect to earth.

Automatic bias (Slide-Back bias) may be provided by allowing grid current to flow into the coupling condenser in the input circuit of a valve amplifier. The condenser charges when the grid voltage is positive and, in the normal connection illustrated in Fig. 332 (a) discharges when the grid voltage is negative with respect to cathode. In the alternative circuit of Fig. 332 (b), the condenser is discharging continually through the grid leak, the bias adjusting itself so that the

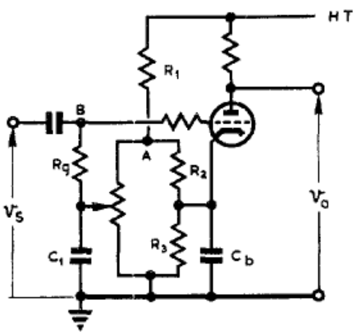

Fig. 331 - Circuit providing positive or negative bias.

more or less constant discharge rate is equal to the mean rate of charging. The principle of this process is dealt with more fully in Clamping Circuits, Chap. 12 Sec. 2. It is also the principle of the Cumulative Grid Detector. Apart from its use as a clamping arrangement or as a detector this method of biasing is most commonly encountered in RF Power Amplifiers and Oscillators. In these circuits distortion introduced by grid current flow is not important, since that due to the use of class C biasing is in any case very considerable. (This distortion is confined to the anode current and not to the output voltage since the anode circuit is normally highly selective and presents a low impedance to all but the first harmonic of the anode current). The principal advantage in such an application is that if the amplitude of the input voltage is reduced the bias is automatically reduced due to a decrease in grid current flow. This causes an increase in the Angle of Flow of anode * current so that the amplitude of the output voltage suffers less reduction than it would if a fixed bias were employed. It is possible for the output amplitude actually to increase as the input amplitude is decreased, due to this selfcompensating action. Where this "Automatic Gain Control" action is detrimental to the action of the amplifier some other method of biasing must be employed.

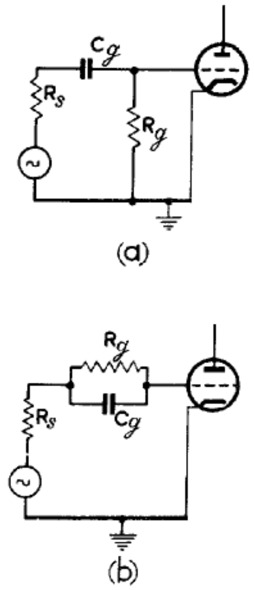

Fig. 332 - Slide-back biasing circuits.

As already pointed out, when slide-back biasing is used with Class C tuned amplifiers and oscillators the additional distortion produced by grid current flow is not normally important. However, there are occasions when the method is applied to resistance-loaded amplifiers and it may be necessary to reduce such distortion to a minimum. This can be arranged by ensuring that the output resistance of the circuit feeding the grid is small, and that the biasing condenser and grid leak are large. If T_1 is the duration of grid current flow, T_2 the period of the applied voltage, R_s the output impedance of the grid input circuit, R_i the input resistance of the valve when grid current flows, R_g the grid leak and C the bias condenser, the following conditions should hold if distortion is to be avoided :-

$$R_i \gg R_s, \; CR_g \gg T_2, \; C(R_i + R_s) \gg T_1.$$ (See Fig. 333).

The amount of bias depends on the ratio $\frac{R_g}{R_i+R_s}$. The larger the ratio the greater is the bias.

* The Angle of Flow is a measure of the duration of current flow in a valve during each cycle of a sinusoidal input voltage. Each period is represented by 360°, and the angle of flow is 360° for Class A, 180° for Class B and correspondingly less for Class C biasing).

If it is desired to introduce distortion without causing appreciable slide-back, either a short time-constant CR_g may be used, producing the type of distortion considered in Chap. 2 Sec. 3, or else the output resistance of the input circuit may be made large whilst the ratio $\frac{R_g}{R_i + R_s}$ is kept as low as possible. The limiting action introduced by this arrangement is discussed in Chap. 9 Sec. 3.

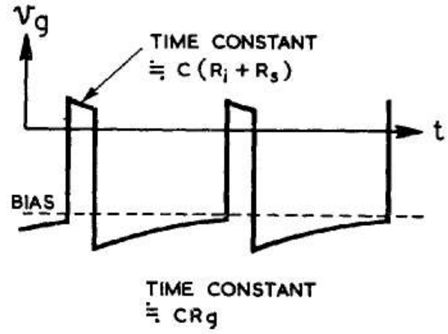

Fig. 333 - Distortion of rectangular pulses by grid current flow.

5. THE POTENTIAL OF THE SCREEN GRID OF A PENTODE USED AS AN AMPLIFIER

It is generally desired to maintain the screen grid of a pentode amplifier valve at a constant potential. If this potential were not constant but varied with alterations of current through the valve the amplification of the valve would be affected.

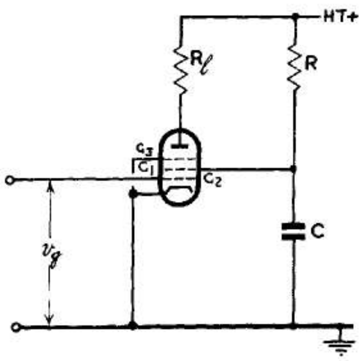

Fig. 334 - Screen supply circuit for class A operation.

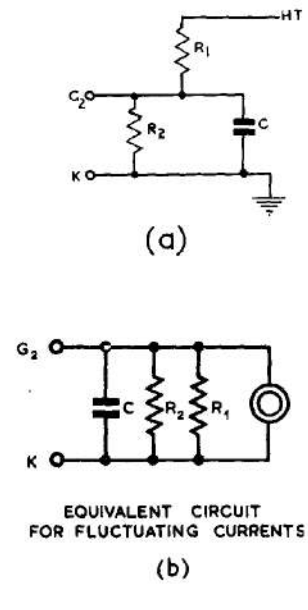

Fig. 335 - Screen supply circuit fpr class B or C operation.

The simplest method of supplying the screen with a suitable source of voltage is illustrated in Fig. 334. This suffers from the disadvantage of the cathode automatic biasing arrangement, namely that with other than Class A operation the valve current varies with signal amplitude, causing the mean screen potential to rise as the amplitude of the input signal is reduced. For such conditions the arrangement of Fig. 335 is to be preferred.

The condenser in Figs. 334 and 335 is adequate for smoothing out the fluctuation in screen potential provided the time-constant of the decoupling circuit is large compared with the period of the input voltage. In Fig. 334 this time-constant is CR. In Fig. 335 the resistors R_1 and R_2 are effectively in parallel so far as fluctuating currents are concerned, as illustrated at (b), so that the effective time-constant is $\frac{C R_1 R_2}{R_1 + R_2}$.

6. ANODE DECOUPLING OF AMPLIFIERS

It has so far been assumed that the HT supply circuit presents a negligible impedance to alternating currents. This is equivalent to the assumption that the time-constant of the supply is very long, and is justified if the smoothing condenser in the power pack is sufficiently large. In practice the voltage across the HT supply circuit varies to a certain extent in response to fluctuations of current passing through this circuit. These fluctuations are present in the current passing through a valve and so pass to the HT supply. A fluctuating voltage is, therefore, developed across the HT supply circuit and so is applied to other amplifier stages using this supply. This feedback of the fluctuating voltages is undesirable, as it may, particularly if applied to a high gain amplifier, cause continuous oscillations to be set up.

The method which is usually adopted to avoid the feedback described above is to prevent the variations of anode currents of the amplifier valves from flowing through the HT supply circuit, by the use of an anode decoupling circuit. Fig. 336 shows the anode decoupling circuit for one amplifier, the circuit consisting of a resistor R and condenser C. Provided the time-constant CR is long compared with the period of the applied voltage, the potential of the point A does not change appreciably with variations of the applied voltage. The point A is maintained at steady potential $R\bar{i}_a$ below HT level where $\bar{i}_a$ is the steady anode current of the valve about which the fluctuations of current occur. The introduction of the decoupling circuit necessitates an increase of HT potential by an amount $R\,\bar{i}_a$ if the mean anode potential is to remain the same as it was without the decoupling circuit.

Where a very considerable degree of decoupling is required it is usual to employ two or more C-R networks in cascade as shown in Fig. 337. By this arrangement a high degree of smoothing may be achieved without the use of prohibitively large resistors or condensers. For example, suppose that it is necessary to prevent more than 1% of the valve current fluctuations from reaching the supply circuits. This may be provided by a single circuit with time-constant T, say. The same effect may be obtained by the use of two circuits in cascade, each having a time-constant $\frac{T}{10}$, so that smaller resistors and condensers may be employed.

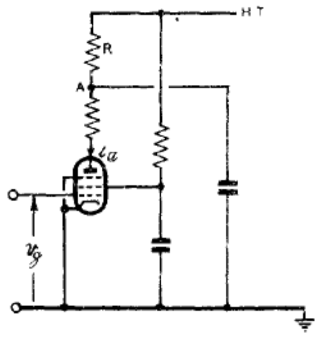

Fig. 336 - Anode de-coupling circuit.

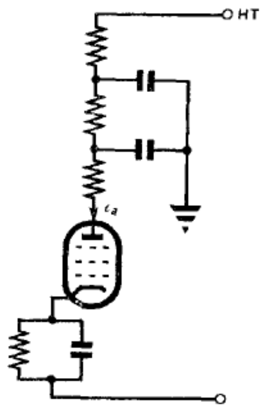

Fig. 337 - Decoupling networks in cascade.

7. MILLER EFFECT

If a triode is operating in an amplifier circuit the interelectrode capacitance C_{ga} of the valve between grid and anode provides coupling between the output and input circuits. This coupling affects the input admittance of the valve, the precise effect depending upon the type of load present in the output circuit.

Fig. 338 (a) shows a typical amplifier circuit, the interelectrode capacitances between grid and anode and between grid and cathode being denoted by C_{ga} and C_{gk} respectively.

The voltage across C_{ga}, as shown in the equivalent circuit of Fig. 338 (b) has an alternating component

$$v_i - v_o = v_i - mv_i$$
$$= v_i (1 - m).$$

Hence the current i_1 through C_{ga} is given by

$$i_1 = j\omega C_{ga} \cdot v_i (1 - m).$$

The current i_2 through C_{gk} is given by

$$i_2 = j\omega C_{gk} v_i.$$

Hence the total input current i_i is given by

$$i_i = j\omega v_i \left[C_{gk} + (1 - m) C_{ga} \right].$$

The input admittance is therefore

$$y_i = \frac{i_i}{v_i} = j\omega C_{gk} + j\omega C_{ga} (1 - m).$$

If θ is the phase displacement between the grid and anode voltages, the voltage amplification m can be expressed in the form

$$m = |m| \quad (\cos\theta + j\sin\theta)$$

where $|m|$ is the magnitude of the amplification.

Hence the input admittance can be expressed as

$$y_i = j\omega C_{gk} + j\omega C_{ga} \quad (1 - |m|\cos\theta - j\ |m|\ \sin\theta)$$
$$= \omega C_{ga}\ |m|\ \sin\theta \ + j\ (\omega C_{gk} + \omega C_{ga}\ (1 - \ |m| \ \cos\theta))$$

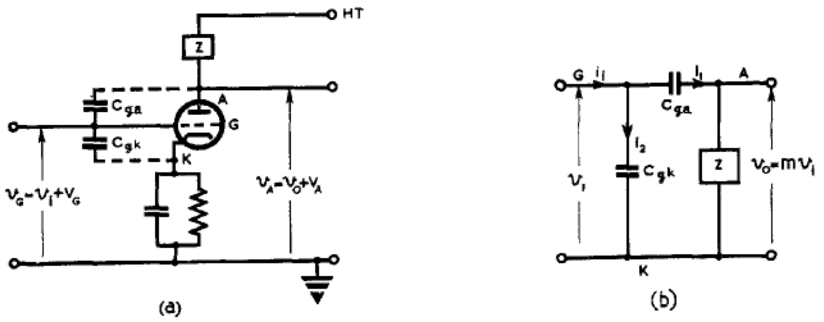

Fig. 338 - Miller Effect.

Provided the effective load (i.e. load impedance in parallel with stray and interlectrode capacitance) of the amplifier is purely resistive, the grid and anode voltages are approximately anti-phase and hence $\theta \doteqdot 180^\circ$. Under these circumstances the input admittance is a capacitive susceptance of value

$$\omega C_{gk} + \omega C_{ga}\ (1 + \ |m| \ ).$$

Consequently the effective input capacitance of the valve is equal to

$$C_{gk} + C_{ga} \quad (1 + \ |m| \ ).$$

For a pentode C_{ga} is very small (less than 0·5 per cent of C_{gk}). Consequently the effective input capacitance of a pentode employed in an amplifying circuit is likely to be considerably smaller than that of a triode ($C_{ga} \doteqdot C_{gk}$) in spite of larger values of m corresponding to the pentode, and of the additional capacitance introduced between grid and screen, C_{gs}.

If the load of an amplifier is inductive θ is greater than 180° and less than 270°. Hence $\sin\theta$ and $\cos\theta$ are negative and the input admittance is due to a negative conductance ($\omega C_{ga}\,|m|\,\sin\theta$) and a capacitance susceptance ($\omega C_{gk} + \omega C_{ga}\ (1 - |m|\ \cos\theta)$).

In the case of a valve employed in an HF amplifier circuit the anode load usually consists of a tuned circuit. For some frequencies off resonance this load is inductive; consequently the input conductance is negative and regeneration is likely to occur, possibly leading to continuous oscillations. If a triode is used, C_{ga}, and hence the value of the negative conductance, is large. Normally if a triode is used in an RF amplifier circuit neutralisation of the effect due to feedback via C_{ga} is necessary. Neutralisation may be performed if a voltage in the output circuit, equal in amplitude and anti-phase to the anode voltage, is impressed on the grid of the valve through a capacitance approximately equal to C_{ga}. Fig. 339 shows a circuit employing neutralisation.

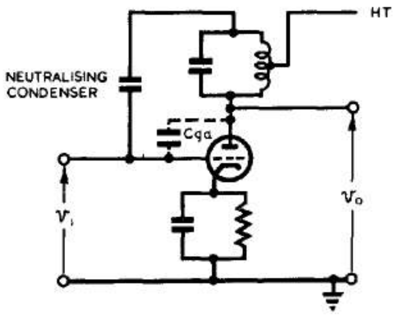

Fig. 339 - Neutralising circuit used in Power Amplifiers.

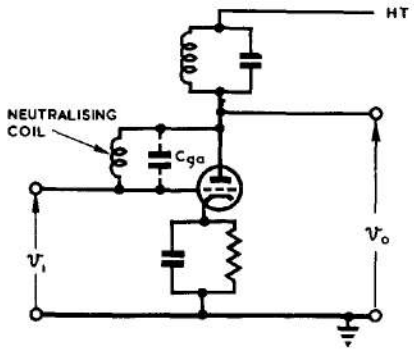

Fig. 340 - Neutralising circuit used in Radio Receivers.

Alternatively, if a coil of suitable inductance is connected between anode and grid so that it forms with C_{ga} a parallel circuit resonant at the frequency of operation, negligible feedback occurs; (Fig. 340).

If the load of the amplifier is capacitive, θ is greater than 90° and less than 180°. Hence $\sin\theta$ is positive. Under these circumstances the input conductance ($\omega C_{ga}\,|m|\,\sin\theta$) is positive, and damping is introduced into the input circuit of the amplifier.

8. RF AMPLIFICATION

The applications of RF amplifiers to the problems of radio reception are dealt with in Chap. 16. Here we are concerned only with the types of amplifying circuits which may be used.

These circuits are essentially the same whether used in the IF or RF stages of a receiver, the chief factor in design being the ratio of the bandwidth to the mid-frequency. The ideal amplification-frequency and phase-frequency characteristics for such an amplifier are illustrated in Fig. 341. Throughout the required band of frequencies, the amplification should be uniform and any phase-shift introduced should be proportional to $f - f_m$ where f is the frequency and f_m the mid-frequency of the band.

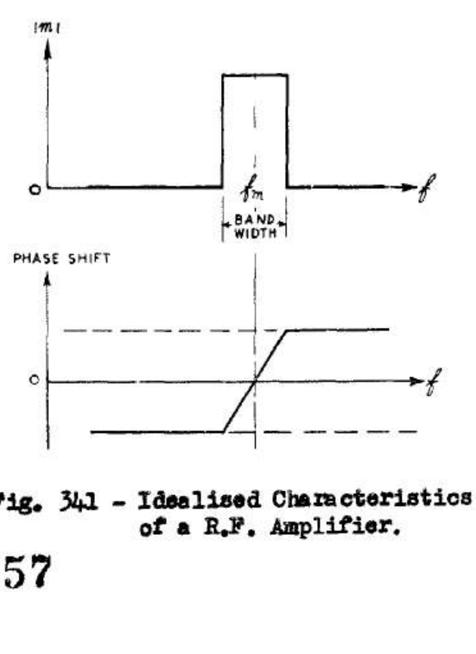

Fig. 341 - Idealised Characteristics of a R.F. Amplifier.

For amplifying RF pulses of about one micro-second duration the band-width is usually of the order of 2-4 Mc/s, whilst the mid-frequency may have any value from 9 Mc/s to 600 Mc/s. Amplification at frequencies higher than this is not normally attempted.

The most usual type of circuit is the simple tuned amplifier arrangement, illustrated in Fig. 342 (a). The equivalent circuit is shown at (b).

${}_oR_D$ is the dynamic resistance of the tuned circuit at resonance (Fig. 342 (c)).

${}_1R_D$ is the equivalent dynamic resistance of the amplifier load circuit, consisting of R_a, ${}_oR_D$ and R_ℓ in parallel.

It follows that at resonance the amplification $\frac{\hat{v}_o}{\hat{v}_i}$ is equal to $G_m\,{}_1R_D$.

The amplification at frequencies other than the resonant frequency is proportional to the impedance of the resultant circuit (Fig. 342 (b)) and follows the normal response curve for a parallel tuned circuit. The selectivity, or Q for this curve is Q_1, and is related to Q_o of the undamped tuned circuit by the equation

$$\frac{Q_1}{{}_1R_D} = \frac{Q_o}{{}_oR_D} .$$

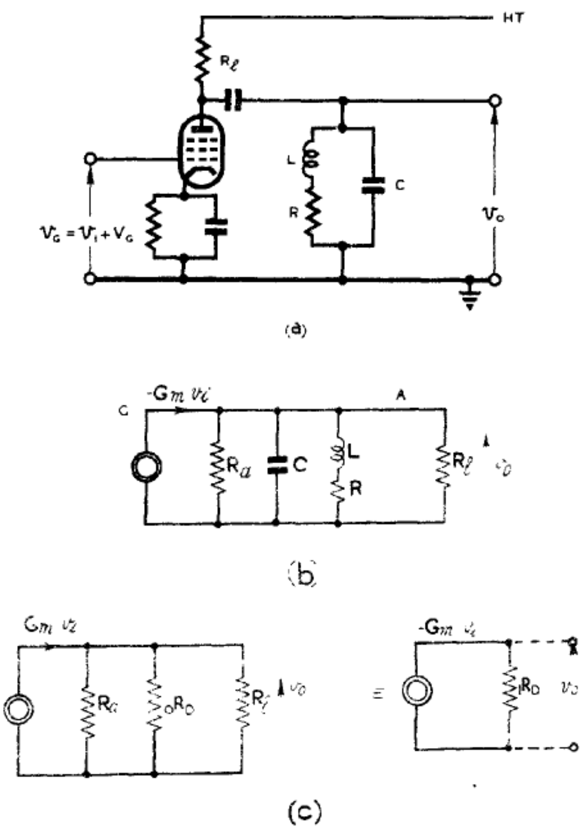

Fig. 342 - R.F. amplifier circuit.

For amplifiers with bandwidth exceeding about 1 Mc/s the input capacitance of the succeeding stage together with the self-capacitance of the RF coil provides sufficient capacitance for resonance. At such frequencies Permeability Tuning may be employed. This consists of inserting a movable core either of copper or of powdered iron. The eddy currents set up in the copper reduce the effective inductance of the tuned circuit. The high permeability of the iron core, powdered to prevent losses due to eddy currents, increases the effective inductance.

At frequencies of the order of 200 Mc/s and upwards lumped L-C circuits are often replaced by tuned lines, usually coaxial with plunger tuning. At frequencies above about 30 Mc/s the input resistance of a valve amplifier is small, so that the equivalent circuit of Fig. 342 (b) is shunted by the low input resistance of the succeeding stage. Normally R_a and ${}_oR_D$ (Fig. 342 (c)) are of the order of 1 MΩ and 100 KΩ respectively; R_ℓ is usually only a few thousand ohms, this fact by itself accounting for wide bandwidth and low gain. The low value of the input resistance further heightens this effect. This resistance decreases with frequency (see Sec. 25) so that the higher the frequency the wider is the bandwidth and the lower the gain.

At lower frequencies where the input resistance is not important, higher gains can be achieved by the use of an RF choke in place of R_ℓ.

An alternative circuit for an RF amplifier is the series-fed arrangement shown in Fig. 343. In the absence of a low input resistance, higher amplification can be obtained from this circuit that from that of Fig. 342 (a) since R_g is normally much greater than R_ℓ.

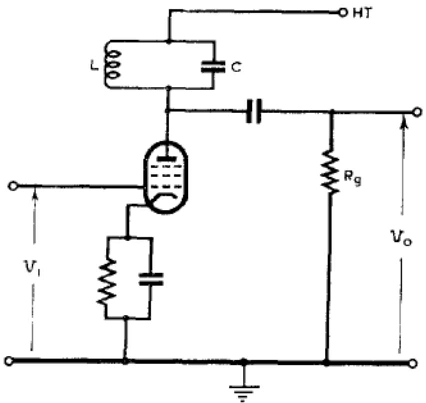

Fig. 343 - Alternative (series-fed) R.F. amplifier circuit.

Because of the heavy damping of the tuned circuit at high frequencies it is essential to use a valve with a large mutual conductance if appreciable amplification is to be provided. Practical values for the G_m of typical RF pentodes are 6·5 mA/V (CV1091) and 8·5 mA/V (CV1065).

Substantially uniform amplification and linear phase-shift over a wide band of frequencies is often obtained by band-pass coupling: Fig. 344; (See also Chap. 1 Sec.21). If the resonant frequencies of the two tuned circuits are equal the pass band is determined by the coefficient of coupling.

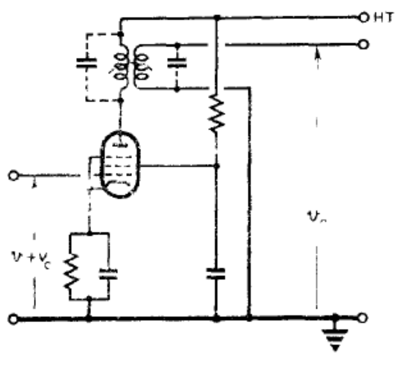

Fig. 344 - Band-pass Circuit.

In some cases double-tuned transformers are employed, the circuit appearing as in Fig. 344, but primary and secondary circuits being tuned to different frequencies on either side of the mid-frequency of the amplification band. The coupling in this case is not so tight as that for the former circuit, and some loss of amplification occurs. The advantage lies in ease of adjustment, since with less tight coupling the tuning of primary and secondary circuits is less interdependent.

A common type of circuit arrangement consists of several pairs of stages using single tuned circuits, each stage of a pair being tuned to a frequency differing from the mid-frequency by an amount equal but in the opposite direction to the other. Such an arrangement is known as Staggered Tuning. Thus with five stages where the intermediate frequency is 7 Mc/s, the first and third stages might be tuned to 5·5 Mc/s, the second and fourth to 8·6 Mc/s and the fifth to 7 Mc/s. The overall response curve then indicates uniform amplification and linear phase-shift over a frequency band of about 3 Mc/s. (Fig. 345).

In general, for a given number of stages employing single tuned circuits it is much more economical to obtain the necessary bandwidth by means of staggered tuning than by resistive damping alone, the latter arrangement resulting in very low gain, and the response curve being far from ideal. For ease of lining-up the amplifier, i.e., adjustment of the various tuned circuits, it is preferable that all the stages should be tuned to the same frequency. Staggered tuning is next in order of simplicity, and the use of double-tuned transformers or tightly coupled circuits is still more complicated, due to interaction between primary and secondary circuits.

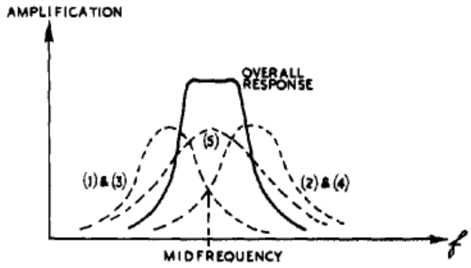

Fig. 345 - Response curve: staggered tuning.

VIDEO-FREQUENCY AMPLIFICATION

9. General Considerations

In a normal broadcasting receiver it is common to refer to post-detector amplification as Audio-Frequency amplification, since the signal to be amplified can be analysed into Fourier components with frequencies in the audible range from 50 c/s to about 4000 c/s. The detected signal of a radar receiver, on the other hand, normally consists of a train of rectangular pulses of short duration (1 to 3 μs) and with a repetition frequency of the order of 400 to 4000 c/s.

Fourier analysis shows that a succession of rectangular pulses can be considered as consisting of sinusoidal components covering a range of frequencies which in theory is infinite. In practice, frequencies up to a few megacycles only are important, and higher components may be ignored. Signal voltages resolvable into components within this range of frequencies are also common in television practice, and since they constitute the intelligence which goes to make the picture they are usually referred to as Video-Frequency signals in contrast to the Audio-Frequency signals of sound-reproducing systems.

This nomenclature has been carried over to radar, and the amplification of voltage pulses is referred to as Video-Frequency Amplification.

A video-frequency amplifier should provide uniform amplification with phase-shift proportional to frequency for voltages of frequencies up to several megacycles. It is difficult in practice to avoid frequency and phase distortion in such an amplifier, and, as discussed below, it is necessary to be content with low gain in order to obtain a satisfactory response.

It is usual for a video-frequency amplifier to have a resistive load, since this has, theoretically, uniform response to all frequencies. In practice stray capacitance in parallel with the resistance may interfere with the performance.

Amplitude distortion is always present in amplifiers, although it can be reduced in magnitude by careful circuit design. Some video-frequency amplifiers are operated under such conditions that non-linear portions of the valve characteristics are used deliberately. (See Chap. 9).

10. The Amplification of an Instantaneous Change of Voltage

If an instantaneous change of voltage (Fig. 346 (a)) is applied to the grid of a valve, which is biased so that it operates on the linear part of its dynamic characteristic, (assuming the load to be purely resistive, as shown in Fig. 347), then the anode current in the valve varies in a manner similar to that of the applied voltage (Fig. 346 (b)).

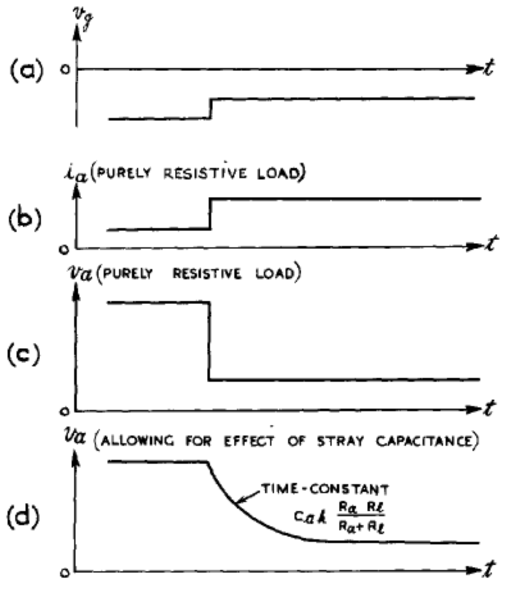

Fig. 346 - Effect of stray capacitance on output voltage.

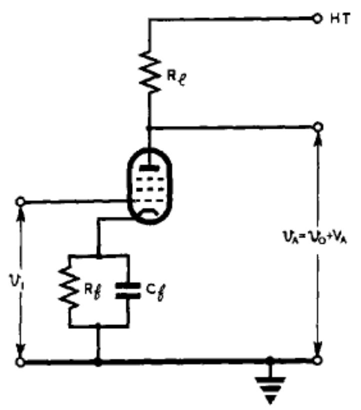

Fig. 347 - Video-frequency amplifier.

The variation of anode potential is exactly inverse to that of the anode current (Fig. 346 (c)). In fact, however, there are usually stray capacitances between anode and earth, so that the anode load consists effectively of a resistor and condenser in parallel (Fig. 348).

The equivalent circuit of the amplifier shown in Fig. 348(a) is given at (b). To consider the response of this circuit to an instantaneous change v_i of the input voltage between grid and cathode we rearrange this in the constant-voltage form as shown in Fig. 348(c) (See Chap. 1 Sec. 21). The response of this circuit to an instantaneous change of voltage has already been dealt with in Chap. 2. In this case the effective generator voltage is $- G_m R_1 v_i$, where $R_1 = \frac{R_a R_\ell}{R_a + R_\ell}$, and the response is shown in Fig. 346 (c). The time-constant of the exponential change of voltage at the anode is $C_{ak} R_1$.

The change of anode potential corresponds more and more closely to the change of grid potential as the time-constant $C_{ak} R_1$ is diminished, by reducing either R_ℓ or the capacitance C_{ak}. Reduction of R_ℓ, however, results in a decrease in amplification.

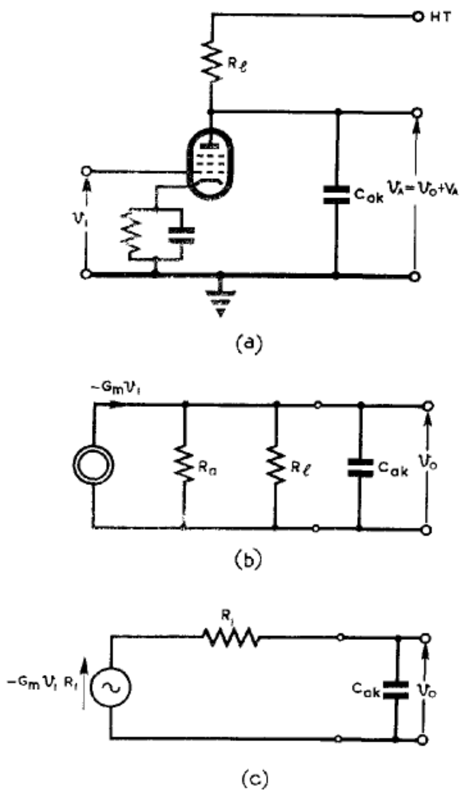

Fig. 348 - Effect of stray capacitance on video-frequency amplification.

If the output from valve 1 (Fig. 349 (a)) is applied to valve 2 through a capacitance C and resistance R_g, with C_{gk} representing the input capacitance of valve 2, then the equivalent circuit for valve 1 has the form shown in Fig. 349 (b). The voltage v_o produced across the circuit between $_2$G and earth is applied to the grid of valve 2. Since the current through the circuit changes very rapidly (ideally instantaneously) the condenser C, which is assumed to have a capacitance large compared with the stray capacitances, can neither charge nor discharge appreciably. Therefore, so far as rapid changes in the circuit are concerned, the effective equivalent circuit is as shown in Fig. 349 (c). The voltage v_o falls exponentially by an amount

$$\frac{R_1 R_g}{R_1 + R_g} \cdot G_m \hat{v}_i ,$$

with a time-constant

$$\frac{R_1 R_g}{R_1 + R_g} (C_{ak} + C_{gk}).$$

In practice, the capacitance C_{gk} is usually much greater than C_{ak} and is therefore the controlling factor in this time-constant. C_{gk} is assumed to include any input capacitance of valve 2 arising from the Miller effect (Sec. 7). The condenser C now discharges exponentially through R_g and R_1 in series (Fig. 349 (b)) with time-constant $C(R_g + R_1)$ until the total fall in anode voltage becomes $R_1 G_m \hat{V}_i$. Meanwhile the grid voltage rises towards zero. The total changes of voltage at the anode of valve 1 and the grid of valve 2 are illustrated in Fig. 350.

If C is not very large compared with C_{gk} these two capacitances have a potential dividing effect, and the amplitude of the initial fall in voltage at the grid of the second valve is reduced. (See Chap. 2 Sec.5). This is likely to occur only when a short time-constant coupling circuit is used for the purpose of distorting the pulse-shape.

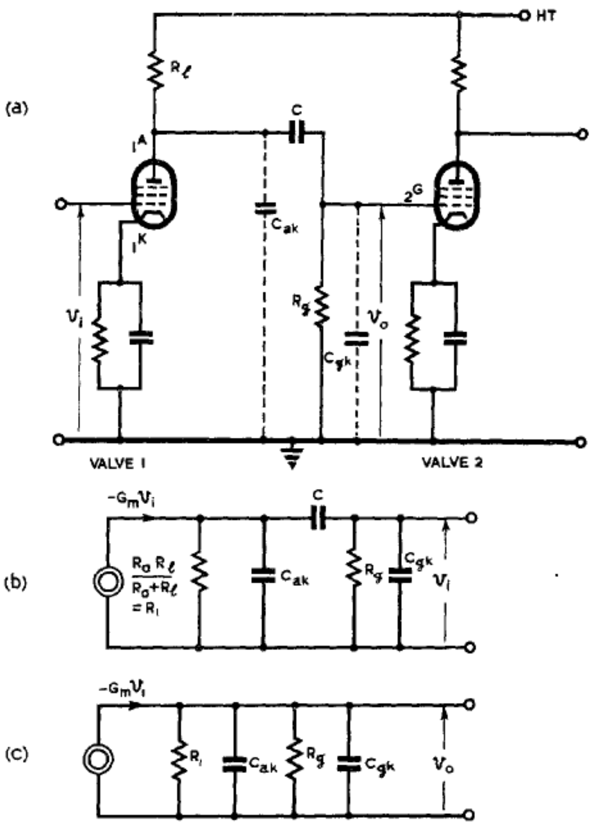

Fig. 349 - Two-stage video-frequency amplifier.

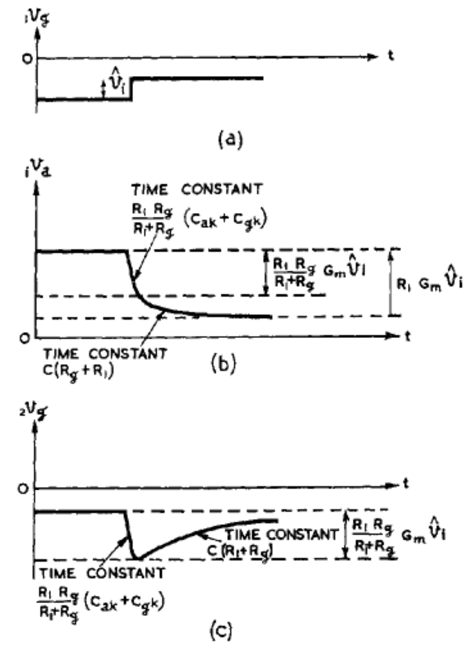

Fig. 350 - Effect of stray and coupling capacitances on video-frequency amplification.

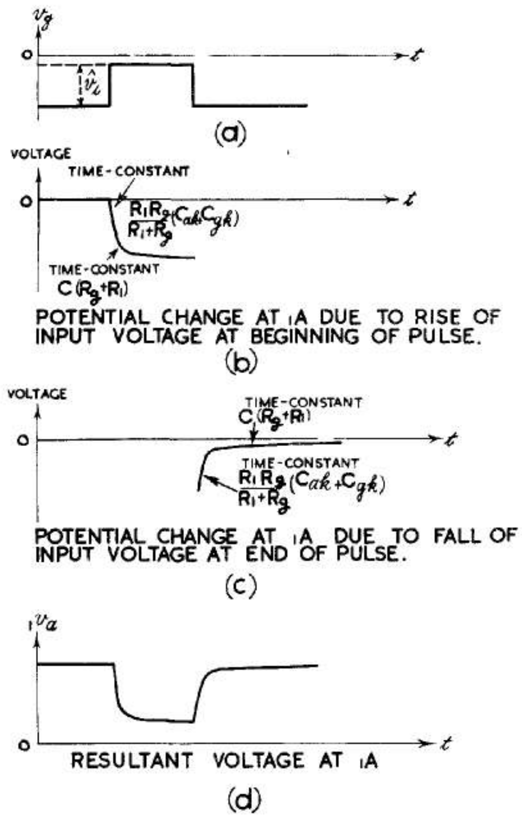

Fig. 351 - Response of a video-frequency amplifier to a rectangular pulse at the grid.

11. The Amplification of a Rectangular Voltage Pulse

The response of a video-frequency amplifier to a rectangular voltage pulse applied between grid and cathode may be found by considering the effect of applying a sudden change of voltage followed after an interval by another sudden change of equal amplitude but of opposite polarity.

If the pulse shown in Fig. 351 (a) is applied to the circuit shown in Fig. 349 the resultant changes in potential at the anode of valve 1 and the grid of valve 2 are as shown in Figs. 351 and 352 respectively. The waveform of Fig. 351 (d) is obtained by adding to the initial potential of the anode of valve 1 the changes of potential at the anode due in turn to the negative-going and positive-going changes illustrated in Fig. 351 (b) and (c). The waveform of Fig. 352 (c) is obtained by adding to the initial potential of the grid of valve 2, the potential changes of the grid due in turn to the negative-going and positive-going changes illustrated in Fig. 352 (a) and (b).

The time-constant $\frac{R_g\ R_1}{R_1 + R_g} \cdot (C_{ak} + C_{gk})$ determines the rates of rise and fall of the output pulse, whilst the time-constant $C.\ (R_1 + R_g)$ determines the constancy of the output voltage during the period of the pulse. The potential produced at the grid of valve 2 approximates most closely in form to the potential applied to the grid of valve 1 if the time-constant $C(R_1 + R_g)$ is very long compared with the duration of the input pulse and if the time-constant $\frac{R_g \cdot R_1}{R_g + R_1}\ (C_{gk} + C_{ak})$ is

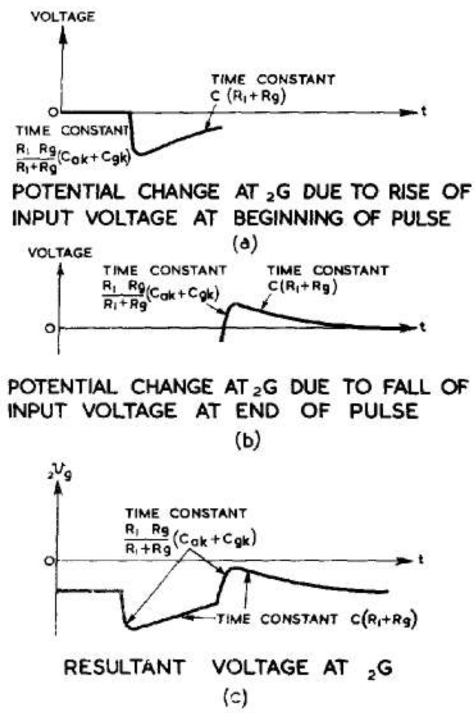

Fig. 352 - Response produced at the grid of valve 2 by rectangular pulse at grid of valve 1.

very short compared with the duration of the input pulse. It is assumed, as is normally the case, that $C \gg C_{gk}$.

The pulse developed between the grid and cathode of valve 2 is in the opposite sense to that applied between the grid and cathode of valve 1. This corresponds to the change of phase of 180° obtained when a sinusoidal voltage is amplified by a valve with a purely resistive load.

The amplification of the applied voltage pulse by valve 1 is

$$-\frac{G_m \,.\, R_1 R_g}{R_1 + R_g}$$

Generally, R_g is made large compared with R so that the total amplification is approximately $G_m \,.\, R_1$, or, with a pentode, $G_m R_\ell$, since $R_a \gg R_\ell$.

The amplification of the circuit as a whole is increased by increasing R_ℓ, but if this is done the time-constant $\frac{R_g \,.\, R_1}{R_g + R_1} . (C_{ak} + C_{gk})$ is increased so that the rise and fall of the output pulse

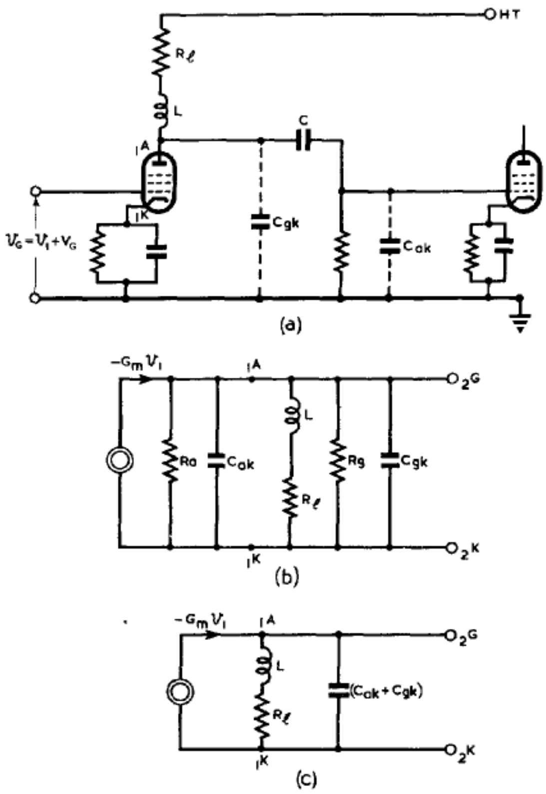

Fig. 353 - Use of compensating coil in anode circuit.

are less rapid. For example suppose $R_g \gg R_\ell$, and $C_{gk} + C_{ak} = 20$ pF. If the rise or fall of the output pulse is to be completed in $\frac{1}{10}$ μs, then 5 x (time-constant of rise or fall of output voltage) $< \frac{1}{10}$ μs; i.e.,

$$5 \cdot \frac{R_g \cdot R_1}{R_g + R_1} (C_{ak} + C_{gk}) < \frac{1}{10} \mu s$$

or

$$5 \cdot R_1 \cdot 20.10^{-12} < \frac{1}{10} .10^{-6};$$

therefore $\quad R_1 < 1000$ ohms

or in the case of a pentode $R_\ell < 1000$ ohms.

For a pentode of mutual conductance 4 mA/volt the amplification $G_m R_\ell$ cannot be greater than 0·004 x 1000 = 4.

For a given value of R_ℓ, the rise and fall of the output pulse can be made more rapid by introducing a coil of inductance L in series with the anode load resistor as shown in Fig. 353 (a). The equivalent circuit for rapid changes of voltage is then as shown in Fig. 353 (b). If R_a and R_g are large compared with R_ℓ, the circuit of Fig. 353 (b) approximates to that of Fig. 353 (c). The circuit may act as an over-damped "ringing circuit". If L is suitably chosen the response of the anode circuit when shock excited by a sudden change of valve current is more rapid than that of a circuit consisting of R_ℓ and $(C_{ak} + C_{gk})$ alone.

(See Chap. 1, Sec. 21).

If the condition

$$L = \frac{1}{2} R_\ell^2 \ (C_{ak} + C_{gk})$$

is satisfied, it can be shown that the time taken for the output pulse to rise or fall is $\frac{3\pi}{8} R_\ell \ (C_{ak} + C_{gk})$, compared with $5 R_\ell \ (C_{ak} + C_{gk})$ for the circuit without the coil. Fig. 354 shows the fall of anode voltage, compared with that of the uncompensated circuit.

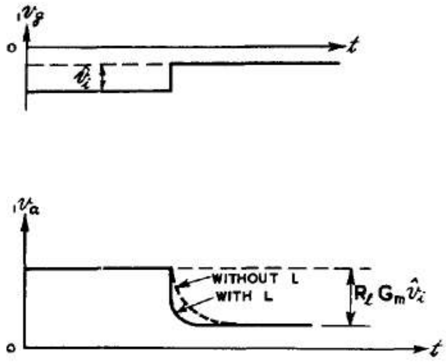

Fig. 354 - Effect of compensating coil in anode circuit.

If the rise or fall of the output pulse is to be complete in $\frac{1}{10}$ μs, then

$$\frac{3\pi}{8} \cdot R_\ell \cdot 20 \text{ pF} = \frac{1}{10} \ \mu s$$

Hence, R_ℓ = 5000 ohms approximately,

and $L = \frac{1}{2} R_\ell^2 \ (C_{ac} + C_{gc}) = 250 \ \mu H.$

For a pentode of mutual conductance 4 mA/volt the amplification $G_m R_\ell$ in equal to 0·004 . 5000 = 20.

The addition, therefore, of a coil of small inductance in series with the anode load of a pulse amplifier allows the amplification of the circuit to be increased without loss of steepness of the rise and fall of the output pulse.

12. The Frequency Spectrum Approach to Video-Frequency Amplification

The response of a video-frequency amplifier to voltage pulses may be calculated by means of methods discussed in Chap. 16 Sec. 1. The Frequency Spectrum of the input voltage is found by harmonic analysis, and then the effect of the amplifier circuit on the phase and amplitude of the components at the various frequencies is calculated. The mathematical technique involved is complicated. It is, however, comparatively simple to discover the conditions which produce distortion even if the determination of the exact form of the output voltage is difficult.

The variation of amplification and phase-shift with Frequency for the circuit of Fig. 349 (a), shown in its equivalent form at (b), may in general be formed by dividing the frequency range into three parts.

(i) Over the mid-frequency region the shunt susceptance $\omega(C_{ak} + C_{gk})$ is sufficiently small and the series susceptance ωC sufficiently large to be negligible in comparison with the conductances. Hence the equivalent circuit takes the form shown in Fig. 355 (a).

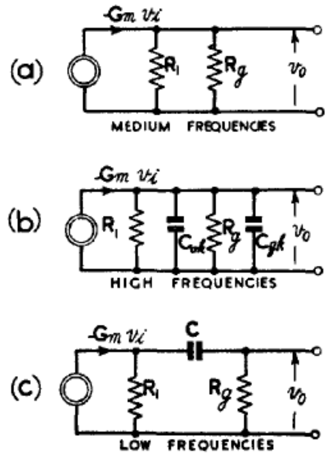

Fig. 355 - Equivalent circuits for video-frequency amplifier.

(ii) At high frequencies the series susceptance is still negligible but the shunt susceptance is comparable with the conductance;(b).

(iii) At low frequencies the shunt susceptance is negligibly small, but the series susceptance is comparable with the conductance;(c).

At medium frequencies the amplification is

$$G_m \frac{R_1 \; R_g}{R_1 + R_g}$$

and the phase-shift by which the input leads the output is 180°.

At high frequencies the amplification is approximately $G_m \dfrac{R_1 \; R_g}{R_1 + R_g}$ $\cos \phi_h$ and the phase-shift is $180^\circ + \phi_h$, where ϕ_h is given by

$$\tan \phi_h \doteqdot \omega \frac{R_1 \; R_g}{R_1 + R_g} (C_{ak} + C_{gk}).$$

For example, the input leads the output by 225° where $\omega = \dfrac{R_1 \; R_g}{R_1 \; R_g \; (C_{ak} + C_{gk})}$ so that $\tan \phi_h = 1$.

The amplification is $\frac{1}{\sqrt{2}}$ times that at medium frequencies, representing a reduction in voltage gain by 3 db.

At low frequencies the amplification is approximately $G_m \dfrac{R_1 \; R_g}{R_1 + R_g} \cdot \cos \phi_\ell$ and the phase-shift is $180° - \phi_\ell$, where ϕ_ℓ is given by

$$\tan \phi_\ell \doteq \frac{1}{\omega C \;\; (R_1 + R_g)}$$

For example, the input leads the output by 135° when $\omega = \dfrac{1}{C \; (R_1 \; + R_g)}$

so that $\tan \phi_\ell = 1$. As before, the amplification is $\frac{1}{\sqrt{2}}$ times that at medium frequencies. The overall amplification-frequency and phase-frequency characteristics for a typical video-frequency amplifier are shown in Fig. 356, plotted to a logarithmic frequency scale.

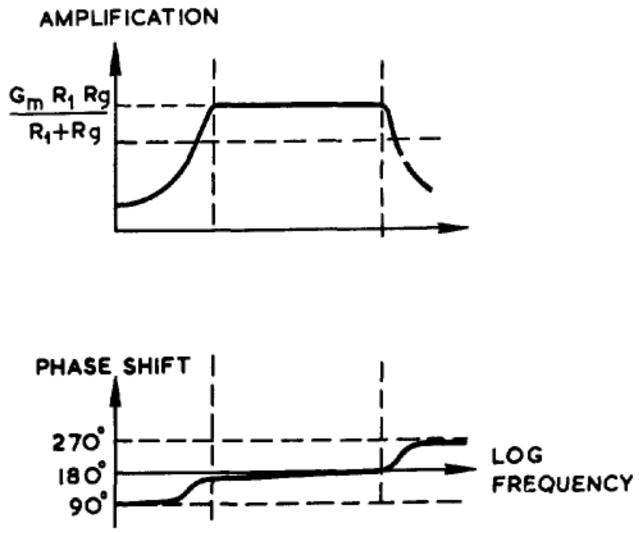

Fig. 356 - Amplification and phase shift characteristics for a video-frequency amplifier.

13. <u>The Use of Pentodes in Video-Frequency Amplification.</u>

Pentodes are normally preferred to triodes in video-frequency amplifiers because of the greater amplification obtainable. This is due to the fact that R_1, which equals $\dfrac{R_a \;\; R_\ell}{R_a + R_\ell}$, can be made much larger for a pentode with an R_a of the order of megohms, than for a triode, for which R_a is a few thousand ohms. The time-constant for rapid changes in anode voltage is thereby made larger by the increase in R_1. However, a pentode normally has a smaller input capacitance than a triode, so that when two stages of amplification are considered the increase in the time-constant due to the larger value of R_1 for the first stage is offset by the reduction in input capacitance of the second stage. Thus the use of pentodes enables greater amplification to be obtained without an increase of distortion, compared with that due to triodes.

NEGATIVE FEEDBACK AMPLIFICATION

14. General Principles

A feedback amplifier is one in which a fraction, or all, of the output voltage or current of an amplifier is fed back via a coupling network to the input circuit of the amplifier.

Consider the case of an undistorting amplifier, with a feedback network which also produces no distortion. The feedback is termed positive if the magnitude of the input voltage is increased, and negative if it is decreased by the feedback connection. In both cases the properties of the amplifier are modified as regards amplification, input and output impedances and distortion.

In general, some form of distortion is invariably produced either by the amplifier or by the feedback network, and it is convenient to consider the nature of the feedback for sinusoidal input voltages at various frequencies. Usually over some range of frequencies the fed-back voltage is largely in quadrature with the input, whilst it is possible for positive feedback (regeneration) to occur in a circuit in which negative feedback predominates over the principal frequency range.

Suppose a sinusoidal voltage v_s is applied to the input terminals (1) and (2) of the amplifier shown in Fig. 357. Let v_i be the voltage developed between terminals (3) and (4). If m is the amplification of the amplifier alone, the output voltage v_o is equal to $m\,v_i$. Since a fraction β of v_o is fed back to the input, the voltage v_i is given by $v_s + \beta v_o$ and the output v_o produced by this input is $m(v_s + \beta v_o)$.

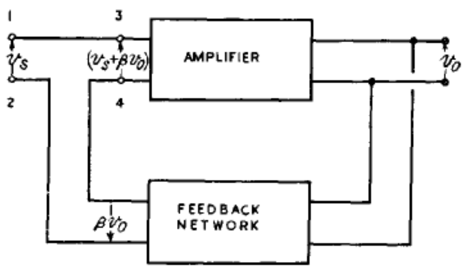

Fig. 357 - Schematic diagram showing feedback.

Therefore $m(v_s + \beta v_o) = v_o$

so
$$v_o = \frac{m v_s}{1 - \beta m} \qquad \text{.........................(1)}$$

In general m and β are both complex, amplification with and without feedback being accompanied by phase-shift which varies with frequency. In the particular case where β and m are real, as, for example where the feedback network is resistive and the amplifier is a video-frequency amplifier operating over its mid-frequency range, we may deduce simple expressions for the overall amplification and the effect on amplitude distortion.

In such a case, equation 1 gives

$$\frac{v_o}{v_s} = \frac{m}{1-\beta m} \text{ (where m is negative) } \ldots\ldots\ldots\ldots\ldots(2)$$

For negative feedback β is positive and the amplification is reduced in the ratio $\frac{1}{1-\beta m}$. If m is very large (2) may be written

$$\frac{v_o}{v_s} = \frac{1}{\frac{1}{m} - \beta} \quad \text{where } \frac{1}{m} \text{ is negligible compared with } \beta.$$

Hence

$$\frac{v_o}{v_s} \doteqdot -\frac{1}{\beta} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(3)$$

It follows that if $\frac{1}{m}$ is negligible compared with β, the amplification depends only on β and is practically independent of the amplification produced by the amplifier itself, without feedback. In other words, the output voltage is substantially independent of the characteristics of the valve. The result is obtained only at the expense of amplification.

For example, suppose a pentode amplifier, with an amplification of 500 (without feedback) is used with a feedback circuit having $\beta = +\frac{1}{20}$.

Then m = -500 so that

$$\frac{v_o}{v_s} = \frac{-500}{1 - \left(\frac{1}{20}\right)(-500)}$$

$$= -\frac{500}{26}$$

$$= -19.2$$

$$\doteqdot -\frac{1}{\beta}.$$

Without feedback it would not be possible to maintain a uniform amplification over the whole of the useful portion of the dynamic characteristic, since m would vary considerably in practice. With feedback quite large variations in m may be permitted, the overall amplification remaining approximately equal to $\frac{1}{\beta}$.

In general, provided m is large and a suitable value for β is chosen, amplitude distortion can be made negligible by the use of negative feedback.

It is usual to distinguish between two types of negative feedback, viz, voltage and current feedback. In the former the voltage fed back is proportional to the voltage developed across the amplifier load, whereas in the latter it is proportional to the current delivered to the load. Whilst these produce similar effects for a constant resistive load, they react in different ways to non-resistive or changing loads.

15. Voltage feedback (Parallel Feedback)

The simplest way of achieving voltage feedback is to connect, in parallel with the load R_ℓ of an amplifier, a potentiometer consisting of two resistors of values R_1 and R_2. These resistances should be large so that the effective value of the load R_ℓ is not modified, and they should be connected in series with a negative supply of steady voltage so that the mean positive potential of the anode is offset and the grid of the amplifier valve suitably biased. Fig. 358 shows the circuit arrangement. The feedback voltage is taken from the junction of the two resistors R1 and R2 and is in series with the applied voltage v_s.

The feedback constant β is given by

$$\frac{R_2}{R_1 + R_2}$$

and the voltage fed back will be 180° out-of-phase with the applied voltage if the load R_ℓ is purely resistive. Therefore, if the load is purely resistive and $\frac{1}{m}$ is negligible compared with β, the output voltage depends entirely on R_1 and R_2 and is quite independent of valve characteristics, supply variations, and the magnitude of R_ℓ. The amplification is substantially independent of the frequency of the sinusoidal components of the input voltage, and negligible phase shift is introduced into the output.

The effect of the feedback connection on the valve performance may be considered by replacing the amplifier, with feedback, by an equivalent valve amplifier without feedback, having the same anode load but with valve constants $\bar{G}_m$, $\bar{\mu}$, and $\bar{R}_a$ in place of G_m, μ, and R_a. These are defined as follows. If i_a, v_s and v_o are corresponding small changes in anode current, input voltage and anode voltage respectively,

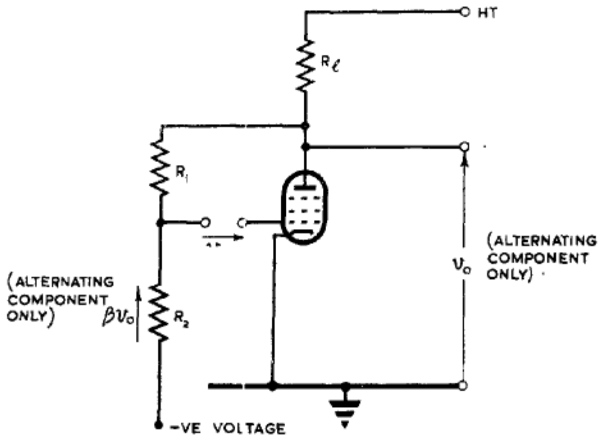

Fig. 358 - Voltage feedback circuit.

$$\bar{G}_m = \frac{i_a}{v_s}, \text{ where anode voltage is constant;}$$

$$\bar{\mu} = -\frac{v_o}{v_s}, \text{ where anode current is constant;}$$

$$\bar{R}_a = \frac{v_o}{i_a}, \text{ where input voltage is constant.}$$

From Fig. 358, the small change βv_o in feedback voltage is in series with v_s so that the small change v_i in grid voltage is given by

$$v_i = v_s + \beta v_o \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (4)$$

If the anode current is constant, i_a is zero and $v_o = -\mu v_i$

$$= -\mu (v_s + \beta v_o).$$

$$\therefore \quad v_o \ (1 + \mu\beta) = -\mu v_s$$

$$\therefore \quad \bar{\mu} = -\frac{v_o}{v_s} = \frac{\mu}{1 + \mu\beta.} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots (5)$$

If the anode voltage is constant, v_o is zero and so, therefore, is βv_o. Hence from (4) $v_s = v_i$, and $\bar{G}_m = \frac{i_a}{v_s} = \frac{i_a}{v_i} = G_m$ ..... (6)

Since $\mu = G_m R_a$, $\bar{\mu} = \bar{G}_m \bar{R}_a$, and $G_m = \bar{G}_m$, it follows that

$$\frac{\bar{R}_a}{R_a} = \frac{\bar{\mu}}{\mu} = \frac{1}{1 + \mu\beta}$$

$$\text{i.e. } \bar{R}_a = \frac{R_a}{1 + \mu\beta} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots (7)$$

Hence the amplifier with valve constants G_m, μ, R_a, and with a feedback factor β behaves like an amplifier without feedback but with constants G_m, $\frac{\mu}{1 + \mu\beta}$ and $\frac{R_a}{1 + \mu\beta}$. Since β is positive for negative feedback $\bar{\mu} < \mu$ and $\bar{R}_a < R_a$, both being reduced by a factor $\frac{1}{1 + \mu\beta}$.

The equivalent circuit is shown in Fig. 359 (a).

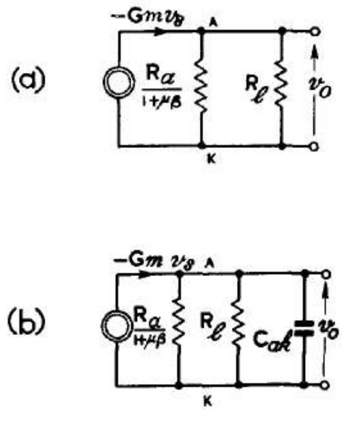

Fig. 359 - Equivalent circuits for voltage feedback amplifier.

In practice it is not possible to make the load purely resistive since there is always some capacitance C_{ak} between the anode and cathode of the amplifier. The equivalent circuit, incorporating this capacitance, is shown in Fig. 359 (b).

The time-constant of this circuit is $\frac{C_{ak} \ R_\ell \ R_a}{R_a + R_\ell + \mu\beta R_\ell}$.

If, as for a pentode, $R_a \gg R_\ell$, this reduces to $\frac{C_{ak} \ R_\ell}{1 + G_m \beta R_\ell}$.

This reduction in the output time-constant increases the fidelity with which sudden changes of input voltage are reproduced at the output. This is illustrated in Fig. 360. For example, with a pentode amplifier having $G_m = 4$ mA/V, $R_\ell = 25{,}000\ \Omega$ and $\beta = \frac{1}{10}$, the time-constant is reduced by a factor

$$\frac{1}{1 + 0{\cdot}004 \cdot 0{\cdot}1 \cdot 25{,}000} = \frac{1}{11}$$

by the use of voltage feedback. To obtain this result the amplification has been reduced in the same ratio.

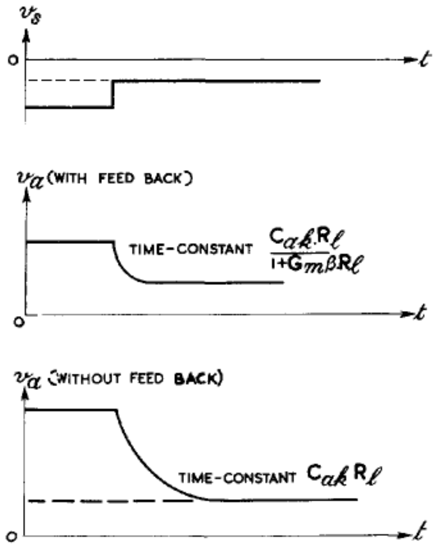

Fig. 360 - Effect of voltage feedback on response of output circuit.

In the circuit so far considered the feedback network has not discriminated between different frequencies. It is often required that feedback should occur for signals within some range of frequencies but not for signals outside this range. The commonest instance of this arises when it is not necessary to consider the amplification of steady voltages. In this case a large blocking condenser can be introduced in series with the feedback potentiometer, as shown in Fig. 361 (a), and the negative bias for the control grid may be provided by a cathode biasing circuit as indicated. This obviates the necessity for providing an additional negative supply voltage to offset the anode voltage.

A circuit which amplifies alternating voltages without feedback but which, owing to negative feedback, provides very little amplification of steady voltages, is illustrated in Fig. 361 (b). The condenser C which is effectively in parallel with R_2 for alternating voltages bypasses current fluctuations so that except for very low frequencies the fraction β of the output voltage which is fed back to the input is small. Also this fraction βv_o is, at high frequencies, in quadrature with v_s. For steady voltages the fraction $\beta = \frac{R_2}{R_1 + R_2}$ is made large so that the amplification, approximately $\frac{1}{\beta}$, is small.

So far we have ignored the problem of applying the voltage v_s to the input circuit of the voltage feedback amplifier. Since this voltage is applied between terminals neither of which is at a fixed potential relative to earth it cannot be obtained direct from a generator having one terminal earthed. One method of overcoming the difficulty is to use a transformer, as shown in Fig. 362.

The use of a transformer tends to introduce phase and frequency distortion into the input voltage, and its use may be avoided by arranging the circuit as shown in Fig. 363 (a). The equivalent network showing how the voltages are added is given in Fig. 363 (b). R' is the output resistance of the generator, together with any additional series resistance which may be required.

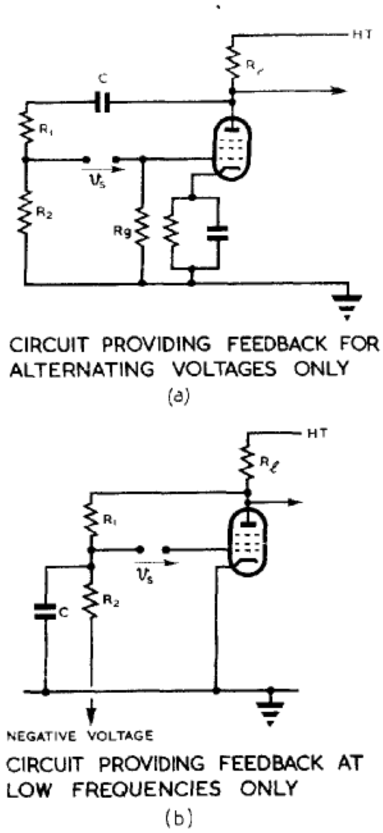

Fig. 361 - Circuit providing feedback (a) for alternating voltages only and (b) at low frequencies only.

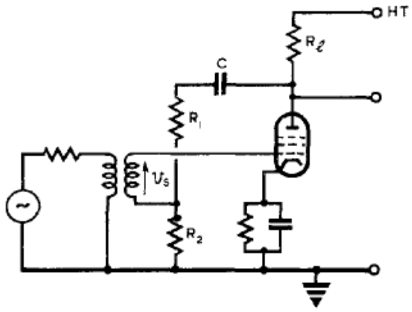

Fig. 362 - Voltage feedback: Transformer input circuit.

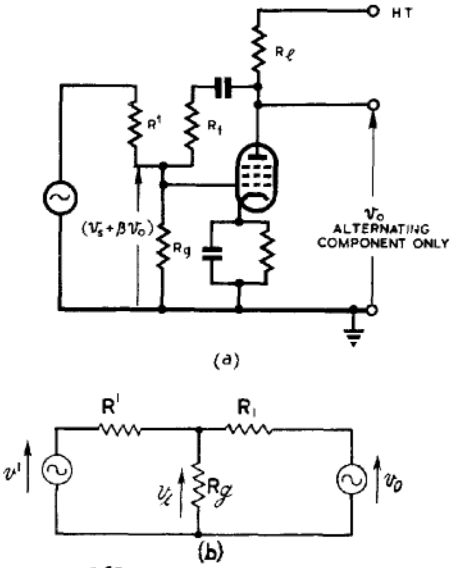

Fig. 363 - Voltage feedback: use of adding network.

The voltage v_i developed across R_g and applied between grid and cathode is given by

$$v_i = \frac{\frac{v'}{R'} + \frac{v_o}{R_1}}{\frac{1}{R'} + \frac{1}{R_g} + \frac{1}{R_1}} \qquad \text{(See Chap. 1 Sec. 12).}$$

If we write $\frac{1}{R'} + \frac{1}{R_g} = \frac{1}{R_2}$, this reduces to

$$v_i = \frac{v' \; R_1 \; R_2}{R' \; (R_1 + R_2)} + \frac{v_o \; R_2}{R_1 + R_2} \,,$$

$$= v_s + \beta v_o;$$

where $\beta = \frac{R_2}{R_1 + R_2}$,

and $$v_s = \frac{v' \ R_1 \ R_2}{R' \ (R_1 + R_2)}.$$

Thus this adding network produces the effect of the arrangement of Fig. 358 without introducing any special problems arising from earthing the generator.

16. Current Feedback (Series Feedback)

To provide negative feedback proportional to the output current it is usual to include a resistor in the cathode circuit of the valve as in Fig. 364 (a). An increase of current through the resistor R_f resulting from a rise of grid voltage causes the cathode voltage also to rise. The effective input voltage developed between the grid and cathode is therefore reduced, i.e., the resistor gives rise to negative feedback.

As for voltage feedback we shall consider the effect of the feedback arrangement in terms of the equivalent amplifier having the same anode load and input and output voltages, but with valve constants $\overline{G_m}$, $\overline{\mu}$, and $\overline{R_a}$.

The feedback fraction β, is given by

$$\beta = \frac{R_f}{R_\ell + R_f}.$$

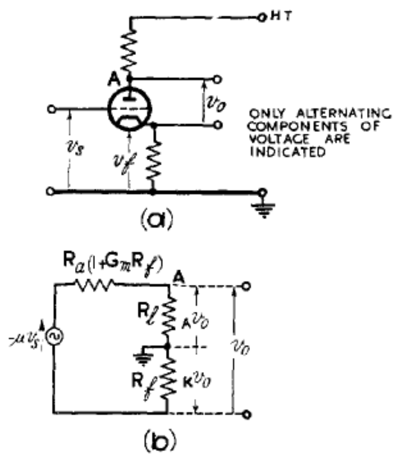

Fig. 364 - Current feedback amplifier.

Let i_a, v_s and v_o represent corresponding small changes in anode current, input voltage, and anode-cathode voltage respectively. The feedback voltage v_f is then equal to $R_f \ i_a$, so that the voltage v_i developed between grid and cathode is given by

$$v_i = v_s - v_f$$
$$= v_s - R_f \ i_a \ \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \ (8)$$

If the anode-cathode voltage is constant, v_o is zero and $i_a = G_m \ v_i$

$$= G_m(v_s - R_f \ i_a) \ \ldots. \text{ from } (8).$$

Hence $i_a \ (1 + G_m \ R_f) = G_m \ v_s$

and $$\overline{G_m} = \frac{i_a}{v_s} = \frac{G_m}{1 + G_m \ R_f} \ \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \ (9)$$

If the anode current is constant, i_a is zero and so, therefore, is v_f. Hence,

from (8), $v_s = v_i$ and $\bar{\mu} = -\frac{v_o}{v_s} = -\frac{v_o}{v_i} = \mu$ .... (10)

Since $\mu = G_m R_a$, $\bar{\mu} = \bar{G}_m \bar{R}_a$, and $\mu = \bar{\mu}$,

$$\frac{\bar{R}_a}{R_a} = \frac{G_m}{\bar{G}_m} = 1 + G_m R_f$$

i.e. $\bar{R}_a = R_a (1 + G_m R_f)$ .... (11)

Hence the amplifier with valve constants G_m, μ and R_a and with a feedback resistor R_f behaves like an amplifier without feedback but with constants $\frac{G_m}{1 + G_m R_f}$, μ, $R_a (1 + G_m R_f)$.

$\bar{G}_m < G_m$ and $\bar{R}_a > R_a$, the one being reduced, and the other increased, by the factor $(1 + G_m R_f)$. The equivalent circuit is shown in Fig. 364 (b).

In this circuit we are interested not only in the voltages developed between anode and cathode, but also in the voltages between anode and earth and between cathode and earth since, in various applications, any of these may be used as output voltages. These will be denoted by v_o, ${}_Av_o$ and ${}_Kv_o$ respectively. From Fig. 364(b) we obtain

$$v_o = \frac{R_\ell + R_f}{R_a (1 + G_m R_f) + R_\ell + R_f} \cdot (-\mu v_s)$$

$$= \frac{R_\ell + R_f}{R_a + \mu R_f + R_\ell + R_f} \cdot (-\mu v_s)$$

$$= \frac{R_\ell + R_f}{R_a + R_\ell + (\mu + 1) R_f} (-\mu v_s).$$

Provided μ is large so that μR_f is much greater than the other terms in the denominator, we have

$$v_o \doteqdot \frac{R_\ell + R_f}{R_f \ \mu} (-\mu v_s)$$

$$\doteqdot -\frac{v_s}{\beta}.$$

Hence $m \doteqdot \frac{v_o}{v_s} \doteqdot -\frac{1}{\beta}$ .... (12)

Since ${}_Av_o = \frac{R_\ell}{R_\ell + R_f} v_o$

$= (1 - \beta) v_o$ .... (13)

we deduce that

$$\frac{A^{v}o}{v_s} \doteqdot -\frac{(1-\beta)}{\beta}.$$

Similarly, $K^{v}o = -\beta v_o$

$$\doteqdot -\beta\left(-\frac{v_s}{\beta}\right)$$

so that

$$\frac{K^{v}o}{v_s} \doteqdot 1 \quad \text{.................................... (14)}$$

In the case of current feedback the increase in the output resistance of the valve, and hence in the output resistance of the amplifier, increases the effect of stray capacitance in parallel with the anode load. The increase in the output time-constant can be offset by the addition of a suitable capacitance in parallel with R_f, thereby increasing the high-frequency response.

Where the amplifier has a choke load instead of a resistor, as, for instance, in a circuit for producing magnetic deflection currents for a CRT, the increase in the output resistance of the valve due to current feedback is an advantage since it reduces the time-constant of the output circuit.

In the above consideration of current feedback we have assumed that the valve is a triode. If a pentode is used two points should be noted. The screen grid should normally be decoupled to cathode and not to earth, otherwise the effective screen potential varies and the amplification is reduced. Also the current through the cathode feedback resistor consists of screen, as well as anode, current. If the variations in screen current are neglected the effect will be the same as if the resistor R_f were assumed to have a value larger than it actually has.

In all cases of current feedback difficulties arise if the cathode potential is allowed to differ substantially from that of earth, with the possibility of insulation breakdown between cathode and heaters. This difficulty may be overcome by using special heater windings carefully insulated from earth.

17. Bias for Feedback Amplifiers

Where no blocking condenser is inserted in the feedback network, an amplifier employing negative voltage feedback normally requires a large negative bias to offset the positive potential at the anode. This may be provided either by a separate negative supply or by means of a cathode bias resistor, suitably decoupled.

If a blocking condenser is used the bias required is no different from that needed by an ordinary amplifier.

The steady voltage developed by the direct current of the valve through the cathode feedback resistor automatically provides a bias voltage for a current feedback amplifier. This bias may be just sufficient to ensure that the valve is operated over the desired portion of its dynamic characteristic, but generally is either too large or too small.

Fig. 365 (a) illustrates the case in which the mean feedback voltage $R_f \bar{i}_a$, where $\bar{i}_a$ is the mean anode current, is just sufficient to provide the correct bias.

Fig. 365 (b) shows the use of an additional cathode bias resistor R_K suitably by-passed, for the case in which $R_f \bar{I}_a$ is not large enough to provide the correct bias.

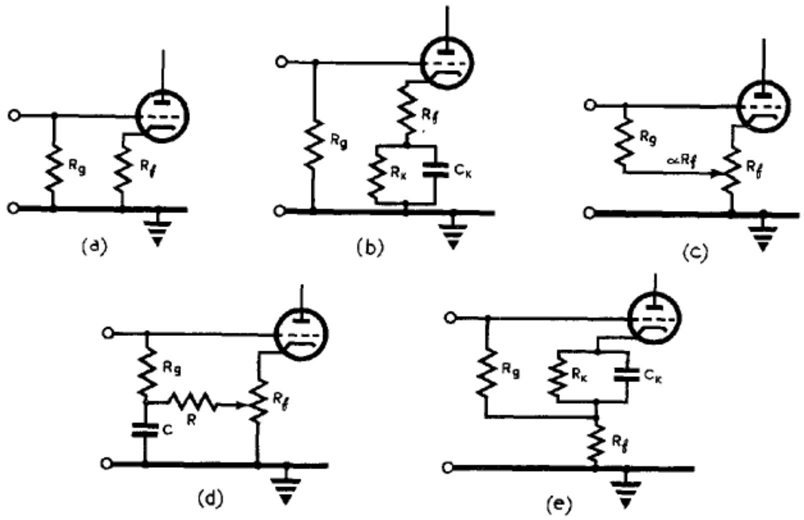

Fig. 365 - Current Feedback: biasing arrangements.

In the arrangement of Fig. 365 (c) $R_f \bar{i}_a$ provides more bias than is necessary. A point on R_f is chosen at which the mean potential, $R_f \bar{i}_a$ below cathode potential, has the correct value, and the lower end of the grid leak is connected to this point. If the output impedance of the generator feeding the grid circuit is large enough to be comparable with R_g, additional feedback to the grid circuit will occur due to this arrangement. This effect may be substantially eliminated by the insertion of a decoupling condenser C connected to the lower end of the grid leak (d). The additional resistor R is necessary to provide the steady voltage connection but must be of sufficiently large value to prevent the lower portion of R_f from being bypassed.

The network of Fig. 365 (e) may be used to provide any degree of negative bias irrespective of the size of the feedback resistor. This method is not economical if a large bias is required.

Sometimes voltage and current feedback are used in the same valve circuit. By this method the output voltage and current can both be made substantially independent of valve characteristics. No additional biasing problems are introduced if the voltage feedback network contains a blocking condenser.

Fig. 366 (a) illustrates the simple case in which voltage and current feedback both operate for steady as well as alternating voltages. The voltage $R_f \bar{i}_a$ is just sufficient to offset the positive voltage at the grid due to the feedback potentiometer, and to provide the correct bias.

In the circuit of Fig. 366 (b) feedback occurs for very low frequencies only. The condenser C ensures that alternating input voltages are effectively developed between grid and cathode so that no appreciable feedback occurs. R_g is connected to a suitable biasing point on R_f.

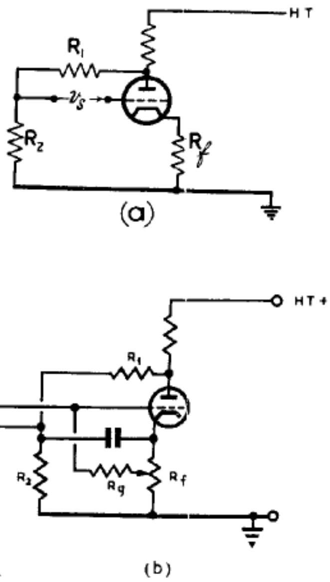

Fig. 366 - Biasing arrangements for circuits using combined current and voltage feedback.

THE CATHODE FOLLOWER

18. The Properties of a Cathode Follower.

In this type of current feedback amplifier the output is taken from the cathode circuit and there is no anode load. The output is the voltage developed across the cathode load by the fluctuating component of the cathode current, which in a triode is the same as that of the anode current. The presence of negative current feedback causes the cathode follower to have the following main properties.

(1) A voltage amplification which may be approximately equal to, but is always less than, unity.

(2) A low output resistance.

(3) A small input capacitance.

Fig. 367 (a) shows a typical cathode follower circuit. A resistor may be included in the anode circuit to ensure that there is the correct steady potential on the anode for normal operating conditions, but this resistor is usually by-passed to earth by means of a large condenser so that there are no fluctuations of anode potential.

The response of the cathode follower circuit to alternating voltages may be derived from the results of Sec. 16, or may be obtained independently in the following manner.

Fig. 367 (b) shows an equivalent circuit for the arrangement of Fig. 367 (a). From (a) we obtain $v_i = v_s - v_o$, whilst from (b)

$$v_o = - \frac{R_f}{R_f + R_a} \cdot (-\mu v_i)$$

$$= \frac{R_f}{R_f + R_a} \cdot \mu (v_s - v_o).$$

i.e. $v_o\,(R_f + R_a) = \mu R_f\ v_s\ -\ \mu R_f\, v_o.$

$$\text{Hence } v_o = \frac{\mu\ R_f\ v_s}{R_a + \mu R_f + R_f}.$$

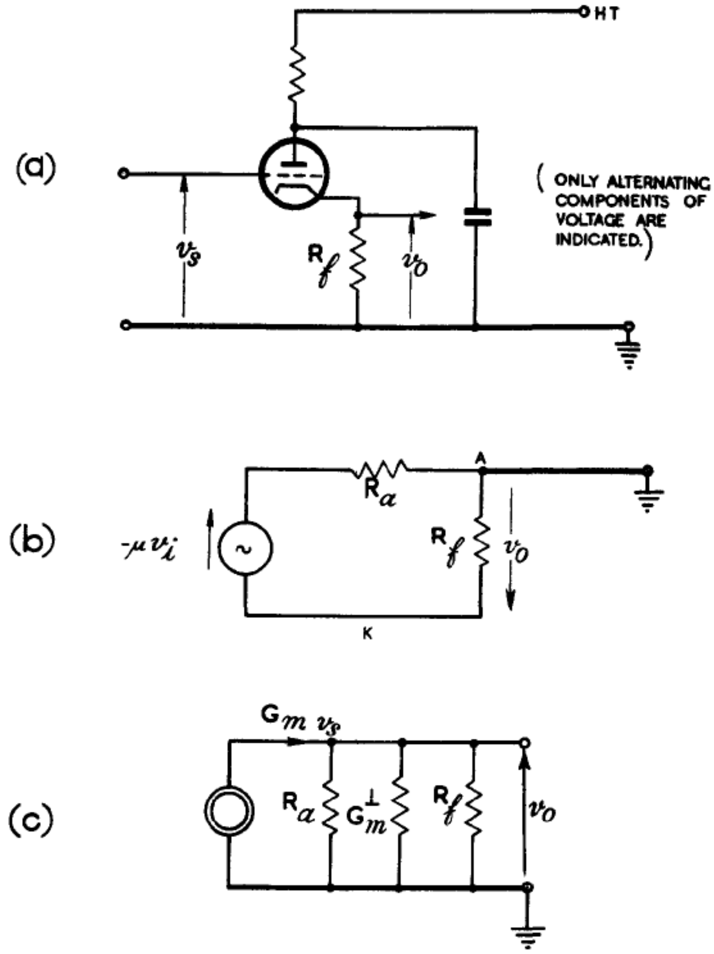

Fig. 367 - Cathode follower.

Writing $\mu = G_m R_a$, and dividing numerator and denominator by $R_a R_f$, we obtain

$$v_o = \frac{G_m\ v_s}{\frac{1}{R_f} + G_m + \frac{1}{R_a}} \qquad \text{.........................} \quad (15)$$

If R_f and R_a are both much greater than $1/G_m$, $v_o \doteqdot v_s$.

Equation (15) shows that the equivalent circuit may be drawn as indicated in Fig. 367 (c). Since G_m is normally a few milliamps per volt whereas R_a is several thousand ohms the output resistance of the cathode follower presented to the load R_f is approximately $\frac{1}{G_m}$, with a value equal to a few hundred ohms. This is very much less than the output resistance of a conventional amplifier, and is of a value suitable for matching to a cable or to an artificial line.

Since the output resistance of the circuit is small, the time-constant due to stray capacitance C in parallel with the load is also small so that there is less likelihood of distortion from this cause than in a normal amplifier. If, however, the rate of change of input voltage is rapid enough, the presence of capacitance in the output circuit prevents the cathode voltage from following that of the grid instantaneously. This may cause grid current to flow if the input is positive-going, or may cut off the valve current if it is negative-going. In the latter case the output resistance of the valve becomes infinite and the output time-constant of the circuit becomes CR_f.

As an example, consider a triode valve operating as a cathode follower under the following conditions.

Slope of dynamic characteristic	=	2 mA/V
Maximum value of grid-cathode voltage	=	0 volts
Minimum value of grid-cathode voltage	=	-10 volts
Maximum value of anode current	=	22 mA
Cathode feedback resistor	=	10 kΩ

Assuming linearity, the minimum value of the anode current is 22 - 2.10 mA = 2 mA. For this value, the grid-cathode voltage is -10 volts and the cathode-earth voltage is 10,000 . 0·002 = 20V.

When the grid-cathode voltage is zero, the anode current is 22 mA and the cathode-earth voltage is 10,000 . 0·022 = 220V.

Hence the output voltage varies from 20 to 220 volts while the input varies from 10 to 220 volts.

The mean grid-cathode voltage is -5 volts, and if the bias is obtained by connecting the grid leak to a point on the cathode feedback resistor, the connection should be made at a point approximately 400ohms from the cathode. Since the mean anode current is $\frac{22 + 2}{2}$ = 12 mA, the bias is then -0·012, 400 $\doteqdot$ -5V.

If in this example the grid voltage were allowed to exceed 220 volts grid current would flow even if the output time-constant were zero. If the grid voltage were allowed to fall below 10 volts the valve would operate over the lower bend of the dynamic characteristic and the anode current might possibly be cut off.

The relative unimportance of Miller effect in a cathode follower, causing it to present a much smaller input capacitance than that of a similar valve in an ordinary amplifier circuit, will now be considered. The relevant interelectrode capacitances are shown in Fig. 368 (a) and the equivalent circuit is given in Fig.368 (b). Since the anode is at earth potential so far as alternating voltages are concerned, C_{ga} is in effect across the input terminals. By an analysis similar to that adopted in Sec. 7, for the circuit of Fig. 338, we obtain

$$y_i = \omega C_{gk} |m| \sin \theta + j\omega \left[C_{ga} + C_{gk}(1 - |m| \cos \theta) \right].$$

If the effective load of the cathode follower is purely resistive the cathode and grid voltages are in phase so that $\theta = 0$. The input admittance is then a capacitive susceptance of value $\omega\left[C_{ga} + C_{gk}(1 - m)\right]$. Consequently the effective input capacitance of the circuit is equal to $C_{ga} + C_{gk}(1 - m)$. Since $m \doteqdot 1$, this normally reduces to C_{ga}. Even if the load impedance is not purely resistive, the maximum value which the input capacitance can acquire is $C_{ga} + C_{gk}$.

The input conductance, when the load is not purely resistive is given by $\omega C_{gk}\,|m|\,\sin\theta$.

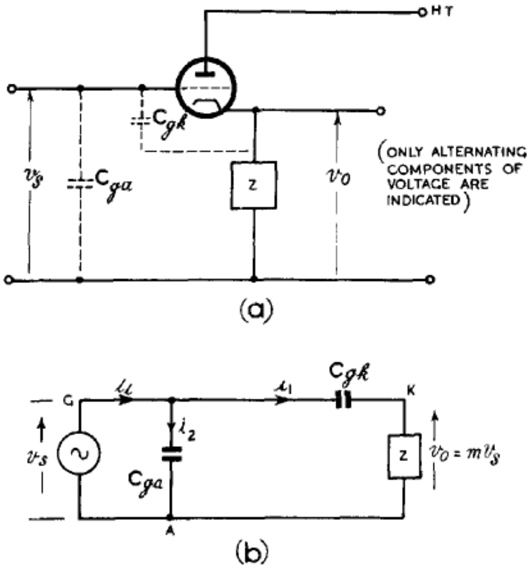

Fig. 368 - Miller effect in a cathode follower.

If the effective load is inductive, $\sin\theta$ is positive and the input conductance has a damping effect on the input circuit. If the effective load is capacitive, $\sin\theta$ is negative and the circuit is regenerative; this condition may give rise to continuous oscillations.

19. Applications of a Cathode Follower

Because the output resistance of a cathode follower is normally much lower than that of a conventional amplifier, the cathode follower is useful for feeding voltages into a relatively low impedance load such as that presented at video-frequencies by the input capacitance of a coaxial cable. Such a cable is often used for carrying voltage pulses from one part of a radar equipment to another.

A cathode follower circuit is also useful for driving a non-linear circuit, to minimise the distortion due to this non-linearity. A typical example of a non-linear circuit is the input circuit of an amplifier in which grid current flows for part of the time. The input resistance, i.e., the resistance between grid and cathode of an amplifier, is usually of a high value, but if grid current flows, it is considerably reduced. If an amplifier is fed from the output of a cathode follower the input resistance of the amplifier shunts the output resistance of the cathode follower (constant current circuit, Fig. 367 (c)). However, since the output resistance of the cathode follower is usually very small, the effect when the input resistance of the amplifier changes from a high to a lower value is not large. Consequently the amount of distortion introduced by non-linearity of the input circuit of the amplifier is small. An additional advantage, under the circumstances described above, is that the cathode of the cathode follower can often be directly connected to the grid of the amplifier, since the steady voltage at the cathode is usually of a low value. If this arrangement is possible, the use of a coupling condenser and grid leak resistor, usually essential in the input circuit of an amplifier, is avoided. If there is no coupling condenser or grid leak resistor, the possibly undesirable effects of slide-back bias due to grid current are avoided. Another common instance of the use of a cathode follower to avoid distortion due to a non-linear load arises in feeding the deflector plates of a CRT when deflector plate current flows.

If it is required to feed the output of an amplifier into a load circuit which has a large input capacitance a cathode follower is particularly useful. The output from the amplifier is taken to the input circuit of the cathode follower, and the output of the cathode follower is fed to the load. The small input capacitance of the cathode follower is in shunt with the output of the amplifier so that the time-constant is small and the potential of the anode of the amplifier (and of the grid of the cathode follower) can change rapidly.

The output resistance of the cathode follower is in shunt with the input capacitance of the load, so that the output time-constant also is small and the load voltage can change rapidly. A cathode follower used in this way may be regarded as an impedance transformer, i.e., the input circuit presents a high impedance whilst the output circuit presents a low impedance.

Since it can be arranged that the cathode potential of a cathode follower rises to almost the same value as the grid potential, it is possible to connect the grid of the cathode follower directly to a point at a high positive potential, without causing grid current to flow. This means that the use of a coupling condenser and grid leak resistor may be avoided. A cathode follower may, therefore, be connected across a circuit, the operation of which would be upset by a resistive loading. For example, suppose that a condenser is charged through a constant-current device so that the voltage across the condenser rises linearly with time. If there is resistance in parallel with the condenser this linearity is affected. A direct-coupled cathode follower may be used to transfer the linear voltage change across the condenser to another part of the circuit, since it does not appreciably shunt the condenser with resistance; (Fig. 369).

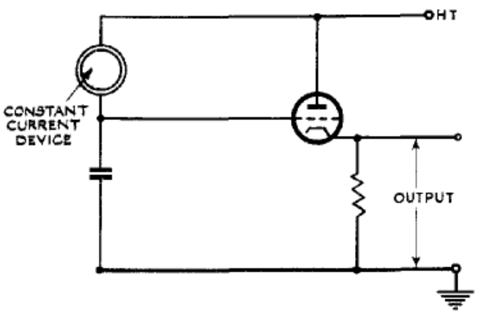

Fig. 369 - Use of direct-coupled cathode follower to minimise loading on a time-base generator.

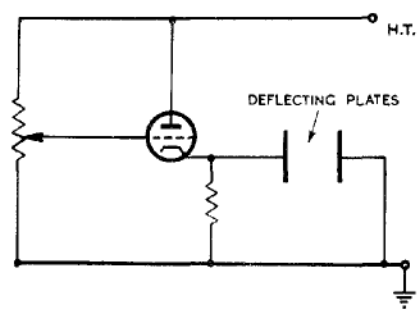

Fig. 370 - Use of a direct-coupled cathode follower to minimise loading on a potentiometer.

It is often necessary to arrange that the potential on the slider of a potentiometer (Fig. 370) varies linearly with the position of the slider. This cannot happen if there is shunt resistance between the slider and one or other of the ends of the potentiometer. If, however, the slider is directly coupled to the grid of a cathode follower, the voltage of the cathode of this valve varies linearly with the position of the slider on the potentiometer. In Fig. 370 the cathode of the cathode follower is shown to be directly connected to a deflecting plate of a cathode ray tube. This arrangement makes it possible to apply to the deflector plate a shift potential which varies linearly with the position of the slider on the potentiometer.

20. THE CATHODE INPUT (OR GROUNDED GRID) AMPLIFIER

The Cathode Input Amplifier is complementary to the cathode follower and is, therefore, considered at this point. An important feature of this circuit is its low input resistance, which makes it suitable for terminating a long cable to which it may be matched. The same property makes the circuit useful as an RF amplifier, since it helps to prevent regeneration. The circuit has the comparatively high output resistance of a normal amplifier.

The circuit arrangement is shown in Fig. 371 (a).

This circuit differs from that of a normal amplifier in that the negative HT supply lead is connected to the grid of the valve instead of to the cathode so that the anode current flows through the input circuit. The grid is normally earthed, and the input voltage is applied to the cathode. The anode and cathode voltages are in phase, since when the cathode voltage is reduced the current through the valve increases and the anode voltage falls also.

Fig. 371 (b) shows an equivalent circuit. The resultant EMF in the circuit is $-(\mu + 1)\, v_i$ so that the output voltage developed across R_ℓ is given by

$$v_o = -\frac{R_\ell}{R_a + R_\ell}(\mu + 1)\, v_i$$

and the amplification by

$$m = \frac{v_o}{v_i} = -\frac{(\mu + 1)\, R_\ell}{R_a + R_\ell}.$$

The amplification is slightly greater than that due to a normal amplifier using the same valve and anode load.

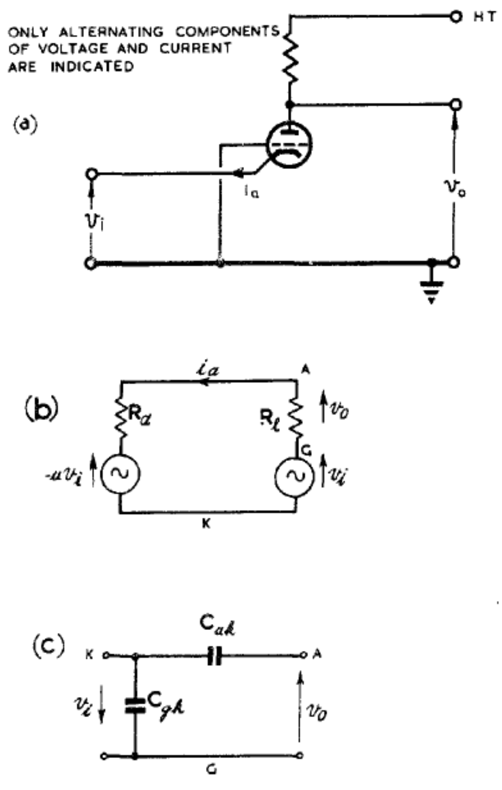

Fig. 371 - Cathode input amplifier.

The alternating component i_a of the anode current is given by

$$i_a = \frac{(\mu + 1)\, v_i}{R_a + R_\ell}.$$

Since this current flows through the input circuit, in phase with v_i, the input resistance R_i of the amplifier is given by

$$R_i = \frac{v_i}{i_a} = \frac{R_a + R_\ell}{\mu + 1}.$$

If $R_a \gg R_\ell$ and $\mu \gg 1$, this reduces to

$$R_\ell \doteqdot \frac{R_a}{\mu} = \frac{1}{G_m}.$$

The input circuit of a cathode input amplifier contributes towards the output power, the fraction supplied being $\frac{1}{\mu}$ of the power derived from the HT supply. This makes the circuit particularly useful as a power amplifier.

Miller effect is of little importance since it occurs through feedback via C_{ak} and this capacitance is usually very small. (Fig. 371 (c)). This fact enables triodes to be used without neutralising. Since the input resistance of the amplifier in any case is small, the current fed back from the output circuit via C_{ak} must be very large before regeneration occurs.

PARAPHASE AMPLIFIERS

21. General

The purpose of a Paraphase Amplifier is to amplify an unbalanced input voltage so as to produce a balanced output (See Chap. 3 Sec 1). This could be accomplished by means of a transformer with the primary winding earthed at one end and the secondary winding earthed at its electrical centre, as illustrated in Fig. 372. Since the design of transformers for undistorted pulse amplification is complicated the employment of a paraphase amplifier is normally preferable. Single-valve and two-valve amplifiers are described.

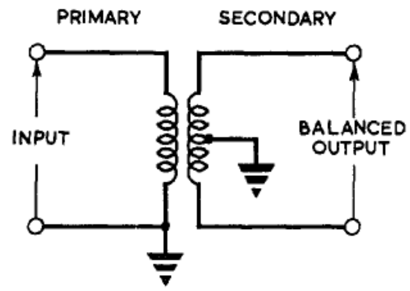

Fig. 372 - Transformer used for paraphase operation.

22. Single-Valve Paraphase Amplifiers

The simplest type of paraphase amplifier employs a single valve, the input for which is a fraction $\frac{1}{|m|}$ of the available applied voltage. This fraction is amplified $|m|$ times, and inverted, so that the output voltage is equal to the applied voltage and of opposite polarity. Fig. 373 shows the circuit arrangement.

The potentiometer formed by R_1 and R_2 provides the fraction

$$\frac{1}{|m|} = \frac{R_2}{R_1 + R_2}$$

of the applied voltage v_s.

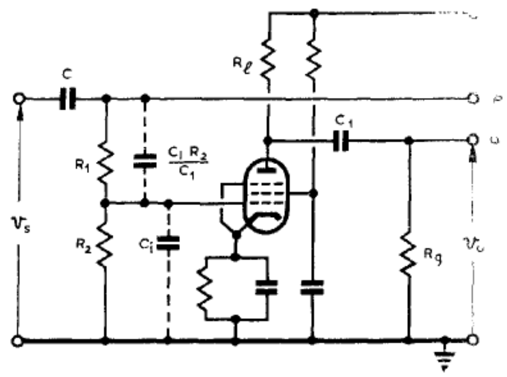

Fig. 373 - Simple one-valve paraphase amplifier.

The condensers C and C_1 eliminate the steady components applied and ouput voltages respectively, so that both v_s and v_o have a mean value zero and $v_o = - v_s$.

Distortion of the input voltage due to input capacitance C_i may be minimised by adding capacitance between the grid and the point P (Fig. 373) so that the time-constants of both portions of the potentiometer are equal.

One disadvantage of this circuit is that any variation in the valve characteristics or supply voltages affects the valve amplification and unbalances the outputs. The amplifier may be made linear and stable by introduding negative feedback. The simplest way of achieving this is by the introduction of a resistor R_f in the cathode circuit (Fig. 374). If a pentode is used in this circuit, the inequality between corresponding changes of anode and cathode currents makes the balance dependent on the ratio between these currents being constant. A triode cannot always be used, in place of the pentode, owing to the larger input capacitance due to Miller effect. Consequently it is advantageous to use voltage feedback. In the circuit shown in Fig. 375, the voltage feedback is introduced by means of a potentiometer, consisting of the resistors R_1 and R_g. The fraction of the input voltage which is actually applied to the valve is determined by the relative values of the resistors R' and R_g. It is found that provided the ratio $\frac{R_1}{R_g}$ is small and the amplification of the valve circuit is large then paraphase amplification is obtained when $R' = R_1$.

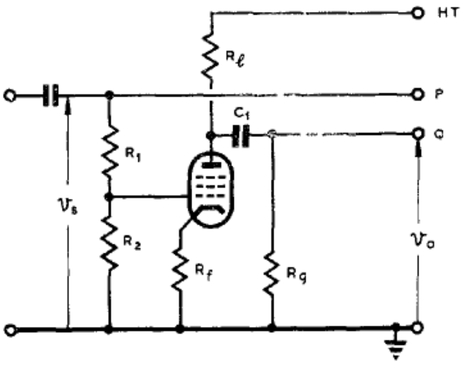

Fig. 374 - Paraphase amplifier with current feedback.

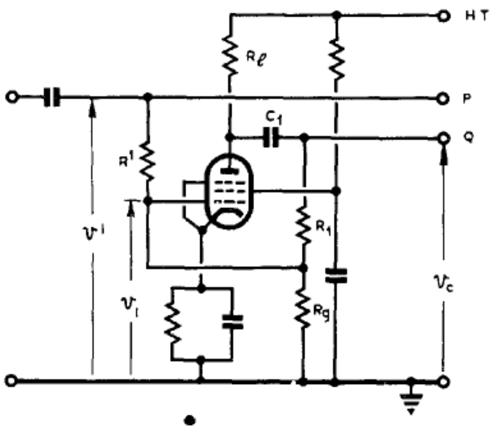

Fig. 375 - Paraphase amplifier using voltage feedback.

*** **Analysis**

This circuit is essentially the same as that of Fig. 363(a). For that circuit we obtained in Sec. 15 the results

$$v_i = v_s + \beta v_o$$

where $$\beta = \frac{R_2}{R_1 + R_2},$$

$$v_s = \frac{v' \cdot R_1 \; R2}{R' \; (R_1 + R_2)} = v' \frac{R_1}{R'} \beta,$$

and $$R_2 = \frac{R_g \; R'}{R_g + R'}.$$

In this case we want $v_o = - v'$.

Then, since $v_o = m v_i$, where m is the amplification of the valve (without feedback),

$$v_o = m \, (v_s - \beta v')$$

$$\text{i.e.} - v' = m \left(\frac{v' \; R_1 \beta}{R'} - \beta v'\right)$$

$$= m \, v' \, \beta \left(\frac{R_1}{R'} - 1\right).$$

Hence $\frac{R_1}{R'} - 1 = -\frac{1}{m\beta}$, where m is negative.

Provided $-m\beta$ is very large, $\frac{R_1}{R'} \doteq 1$ for the output to be correctly balanced. β is large provided R_1 and R' are small compared with R_g, so that the condition $R_1 \doteq R'$ holds if the amplification is large and $\frac{R_1}{R_g}$ is small.

An alternative method of obtaining a paraphase output, using a single valve, depends on the fact that in a valve with purely resistive anode and cathode loads the anode and cathode potentials vary in opposite phase. In the circuit shown in Fig. 376 the anode and cathode loads are equal in value. In the absence of grid current flow, changes in valve current due to changes of input voltage produce equal voltage changes across the two equal loads. A paraphase output is, therefore, developed between the anode and cathode of the valve.

The circuit is a particular case of an amplifier with current negative feedback, so that the amplification characteristics of the circuit are linear and stable.

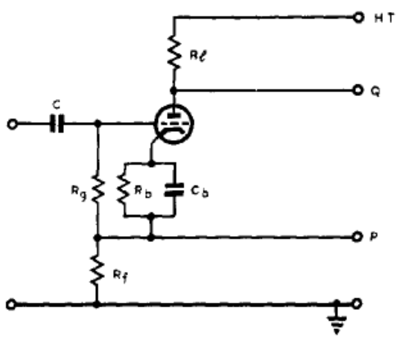

Fig. 376 - Paraphase amplifier using current feedback.

Since the potential change at the cathode cannot be greater than the change at the input, and is equal and opposite to the potential change at the anode, the overall amplification of the stage is not greater than 2.

The output resistance of the valve when viewed from the anode is different from that viewed from the cathode. When stray capacitances are considered the time--constants of anode and cathode circuits are generally different so that when the input signal consists of pulses the distortion is different in the two outputs.

Fig. 376 shows the valve as a triode. There is no serious disadvantage in using a triode, since, though Miller affect is present, that part of the input admittance, which is due to the grid-anode interelectrode capacitance, is not large since the amplification is small, so that, like the cathode follower, the circuit has a small input capacitance. If a pentode is used the resistance of the cathode load has to be made somewhat smaller than that of the anode load since the change of cathode current for a given change of applied voltage is somewhat larger than the corresponding change of anode current.

*** Analysis

The results (12), (13) and (14) of Sec. 16 give approximate values for the amplification of the circuit of Fig. 376 if R_ℓ is made equal to R_f. This makes $\beta = \frac{1}{2}$ so that $m \doteq -2$. The analysis of Sec. 16 gives the exact value as

$$m = -\mu \cdot \frac{2R_\ell}{R_a + (\mu + 2) R_\ell}$$, so that the amplification produced at each output is

$$\frac{\mu R_\ell}{R_a + (\mu + 2) R_\ell}.$$

23. Two-Valve Paraphase Amplifiers

In the single-valve paraphase amplifiers described in Sec. 22, the valve acts as a phase reverser, but in no case does the magnitude of each output voltage exceed that of the applied voltage. If a greater output is required in paraphase two valves may be used as described below.

The most obvious manner in which two valves may be used is by introducing, before the circuit of Fig. 373, a single stage of amplification. This arrangement gives the circuit shown in Fig. 377. It is normal, though not essential, for the values of the anode load and of the mutual conductance of the second valve to be the same as those of the first valve. Then, as shown in Sec. 22 in reference to Fig. 373, the condition for a paraphase output is

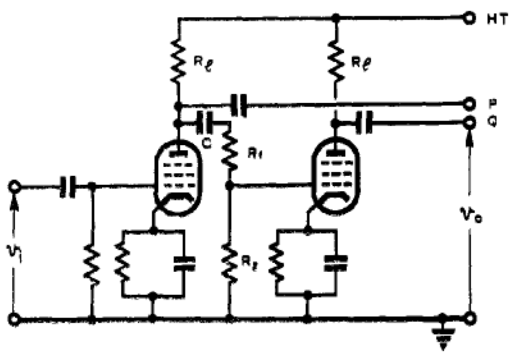

Fig. 377 - Simple two-valve paraphase amplifier.

$$\frac{m \cdot R_2}{R_1 + R_2} = -1,$$

where m is the amplification of each valve circuit. Each of the amplifying stages introduces distortion as a result of the non-linearity of the valve characteristics and as a result of interelectrode and stray capacitances. However, though the first stage may give as its output a distorted version of the input voltage, it is distortion introduced by the second stage which prevents perfect paraphase. The distortion introduced and the dependence of the circuit on valve characteristics can, of course, be reduced by using negative feedback in each of the amplifying stages. A condenser may also be introduced in parallel with resistor R_1 in order to avoid the distortion which would otherwise be introduced by the input capacitance of the second valve, as explained in Sec. 22.

An amplifying stage may also be introduced before the circuit of Fig. 375. This type of circuit is often known as the Floating Paraphase Amplifier, and is shown in Fig. 378. This first stage can be considered as a normal amplifier, though its amplification is modified by the network of resistors R', R_g and R_1, the input resistance of which is in parallel with the load resistor. This effect is negligible provided the input resistance of the network is large compared with the load resistance. In the second amplifier stage, the variations of potential of the point 0 provide the input voltage to the second valve, and these variations depend on the changes in potential at the anodes of the two valves, which are antiphase.

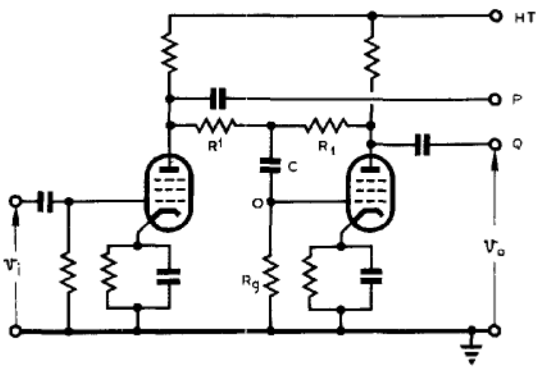

Fig. 378 - Floating paraphase amplifier.

Usually the resistor R' is made equal to the resistor R_1 and it can be seen that perfect paraphase is then impossible, since equal and opposite changes of potential at the two anodes produce zero variation of potential of the point O, and, therefore, no input voltage to the second valve. In this arrangement the output from the second valve is less than that from the first valve by just that amount necessary to produce at the point O variations which, after amplification, give the output from the second valve. As in the case discussed in Sec. 22, provided $\frac{R_1}{R_g}$ is small, and the amplification of the second stage is large, the changes in potential at the two anodes can be regarded as equal in amplitude. There is negative voltage feedback operating on the second stage so that the operation of the second stage is stable, more or less independent of valve characteristics, and produces little distortion.

The conditions for exact paraphase working are the same as those given in Sec. 22 for the circuit of Fig. 375.

A further method of obtaining a paraphase output voltage is provided by the Cathode Inversion Circuit (or cathode-coupled paraphase amplifier) shown in Fig. 379 (a). It consists of two valves coupled together by the common cathode resistor R_5, which is usually of several thousand ohms and is connected to a large constant negative potential - V. The values of the resistor R_5 and the potential -V (about -200 volts) are adjusted so that in the absence of any applied voltage the valves are both biased in Class A. The anode load resistors R_1 and R_2 are usually made equal. The input voltage is applied between the grid of the first valve and earth, and the potential of the other grid is maintained constant, in this case at earth potential.

Initially, if there is no applied voltage, both valves are biased by an amount equal to the difference between -V and the voltage developed across R_5 by the flow of the combined valve currents.

Consider an actual circuit with $R_5 = 5\ k\Omega$, and the slope of the dynamic characteristic of each valve equal to 1 mA/volt. An increase of one mA through R5 raises the potential of K by 5 volts, decreasing the cathode current of valve 2 by 5 mA. This must be accompanied by an increase of 6 mA in valve 1, due to an increase of 6 volts between its grid and cathode, and therefore of 11 volts between its grid and earth. Thus any input voltage is distributed between the two valves in the ratio 6 : 5, so that with equal anode loads the circuit is unbalanced in this ratio. The larger the value of R_5 the more nearly balanced are the output voltages; at the same time the voltage -V must be made more negative so that the mean valve current is unchanged.

Since the grid base of either valve is unlikely to be more than about 15 volts, the change of potential at K is small compared with the voltage across R_5 (100V to 200V). Hence the current through

R_5 is approximately constant so that increases in the current of one valve are accompanied by approximately equal decreases in the current of the other.

Instead of the lower end of R_5 being connected to a negative potential, bias may be provided for the grids by connecting them to a source of constant positive potential. This is usually more convenient, but considerably diminishes the effective HT voltage, developed between the positive HT line and cathode.

Since approximately half the input voltage is developed between the grid and cathode of each valve the overall amplification of the circuit is approximately the same as that for a single valve of the same type with the same anode load ($R_1 = R_2$).

The equivalent circuit for Fig. 379 (a) is shown at (b). In this circuit $v_s = {}_1v_i - {}_2v_i$, where v_s is the applied voltage and ${}_1v_i$, ${}_2v_i$ the input voltages between the grid and cathode of valves 1 and 2 respectively. Similarly ${}_1v_o$ and ${}_2v_o$ are the alternating components of the output voltages measured between the anodes and earth.

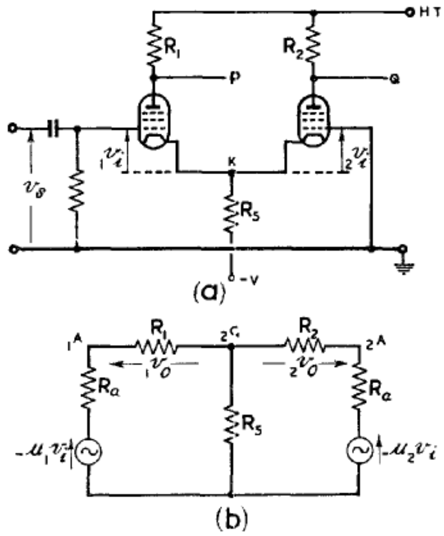

Fig. 379 - Cathode-coupled paraphase amplifier.

Analysis shows that the condition for true paraphase working, assuming that the valves have identical characteristics, is

$$(\mu + 1)\ \frac{(R_2 - R_1)}{R_a + R_2} = \frac{R_1}{R_5}$$

Provided $\mu \gg 1$ and $R_a \gg R_2$, this gives

$$R_2 - R_1 \doteqdot \frac{R_1}{G_m\ R_5}$$

Hence good paraphase working with $R_1 = R_2$ is possible only if $G_m R_5 \gg 1$.

On putting $R_1 = R_2$ in the expressions for ${}_1v_o$ and ${}_2v_o$ it may be shown that the overall amplification $m = \dfrac{{}_1v_o - {}_2v_o}{v_s} = -\dfrac{\mu R_1}{R_a + R_1}$ as it would be for a single valve with the same characteristics and the same anode load as used for either valve in the paraphase circuit.

The output voltages depend on the valve characteristics, no stabilising effect being introduced by the cathode load R_5. Such an effect can be obtained by the use of negative feedback separately for each of the valves. This may be achieved, for example, by introducing resistors R_3 and R_4 as shown in Fig. 380. By applying a star-delta transformation (Chap.1, Sec.13) to the network formed by R_3, R_4 and R_5 the equivalent network formed by R_6, R_7 and R_8 is obtained (Fig. 381).

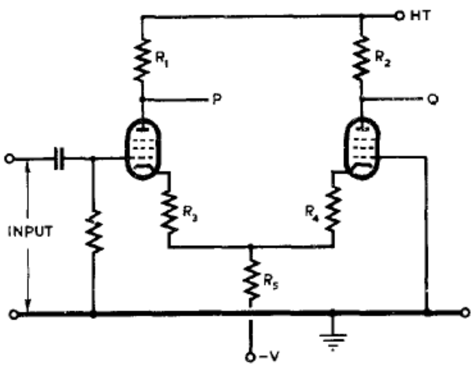

Fig. 380 - Cathode-coupled paraphase amplifier with current feedback (Y-network).

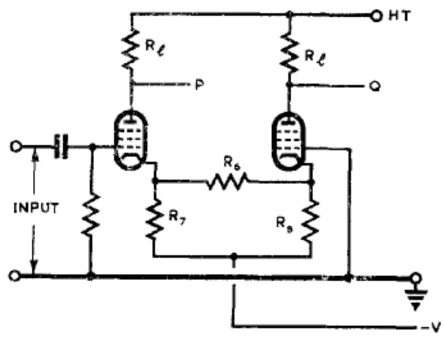

Fig. 381 - Cathode-coupled paraphase amplifier with current feedback (∇ -network).

The relations between these networks may be written :-

$$R_6 = \frac{R_3 R_4 + R_4 R_5 + R_5 R_3}{R_5},$$

$$R_7 = \frac{R_3 R_4 + R_4 R_5 + R_5 R_3}{R_4},$$

and $$R_8 = \frac{R_3 R_4 + R_4 R_5 + R_5 R_3}{R_3}.$$

This is a practical variant of the cathode inversion circuit.

In practice the presence of stray capacitance across the output terminals, and from each anode to earth, affects the rapidity of response of the amplifier to rapidly changing input voltages. The output impedance of the circuit is comparable in magnitude with that of a normal amplifier using a valve and load similar to those of one amplifier of the paraphase circuit, so that the output time-constant is not substantially altered by the paraphase connection. This is not necessarily true if negative feedback is used.

The cathode inversion circuit may be used to supply a paraphase voltage change to a pair of deflecting plates of a cathode ray tube, and the output of the circuit can be taken direct to the plates without the inclusion of coupling condensers. In this case paraphase shift potentials can be obtained and varied in amplitude by alteration of the steady potential of the grid of the second valve, since this results in amplitude changes of steady potential in opposite directions at the anodes of the two valves. The use of a separate "shift" network is avoided. If a variable resistor is included in the cathode circuit, in series with R_5, variation of the resistance causes changes of potential in the same direction at the anodes of both valves. The variable resistor may, therefore, be used as a control to correct astigmatic distortion of the cathode ray tube. If this cathode inversion circuit is used to provide shift voltages and correction for astigmatic distortion, the potential of the final anode of the cathode ray tube should normally be about the same as the mean potential of the anodes of the two valves.

The cathode inversion amplifier may be used as a sum-and-difference device and as such finds frequent application in control and computing circuits.

Suppose that separate input voltages are applied to the grids of the two valves. Then the change of voltage developed between the two anodes is proportional to the difference between the input voltages, whilst the voltage developed at the cathode is proportional to the sum of the two input voltages. This assumes that the anode loads are equal and that the two valves have similar characteristics.

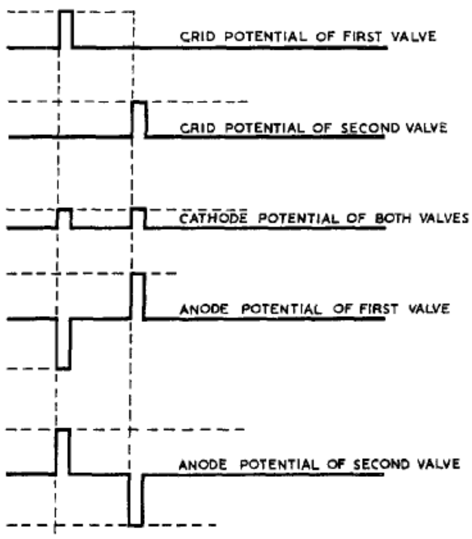

Fig. 382 - Use of a cathode-coupled paraphase circuit as an adding (subtracting) device.

The cathode inversion circuit is often used to provide a paraphase output consisting of the result of adding two separate input pulses occurring at different times. Fig. 382 shows the result of applying positive-going pulses at different instants to the grids of the two valves.

LIMITATIONS OF THE USE OF VALVES AT HIGH FREQUENCIES

24. General

Amplifier and oscillator circuits employing negative-grid valves operate efficiently at frequencies up to a few hundreds of megacycles. As the frequency is increased efficient operation is more difficult to obtain. At frequencies of 1000 Mc/s or more the inherent difficulties are so great that amplification is not usually attempted. Further, at these frequencies negative-grid valves are inefficient in oscillator circuits and special valves (i.e. Magnetrons and Klystrons) are normally used as oscillators.

The difficulties experienced with amplifier and oscillator circuits employing negative-grid valves at the higher frequencies may be considered as due to :-

(i) the finite time required for electrons to traverse the valve (transit time),

(ii) limitations imposed by interelectrode and stray capacitance, inductance of valve leads and resistive and radiation losses.

25. Transit Time Effects

In RF Amplifiers the most series consequence of a finite transit time is damping of the input circuit. At any frequency of operation there is at any instant, owing to the finite transit time, a difference in the number of electrons approaching and receding from the grid of the amplifier valve. The electrons forming the valve current at any instant induce a positive charge on the grid. If the grid voltage is increasing there is an excess of electrons approaching the grid, causing an increase of positive charge at the grid, i.e. there is a resultant current flowing into the grid from the external input circuit. If the grid voltage is decreasing there is an excess of electrons receding from the grid and this leads to a resultant current flow from the grid to the external circuit. Consequently an alternating current flows in the input circuit. At low frequencies of operation the transit time is small compared with the period of oscillation and the instant of zero grid current practically coincides with that of maximum grid voltage. The grid current leads the applied grid voltage by approximately 90° and there is practically no loss of power in the input circuit. There is however, a small increase in the grid-cathode capacitance in the valve when a signal is applied to the grid; this is unlikely to be of practical importance. At higher frequencies of operation the phase angle between the applied voltage and grid current is reduced, since there is an appreciable lag between the application of the grid voltage and the corresponding flow of current in the valve. Consequently a resistive component is introduced into the input impedance of the valve so that there is a loss of power in the input circuit. An approximate relation between the input resistance and transit time can be obtained as follows :-

let the grid-cathode voltage variation be given by :-

$$v_g = \hat{v}_g \sin\omega t.$$

If τ_k = transit time of the cathode-grid space and τ_a = transit time of the grid-anode space the current flowing towards the grid is given by :-

$$i_1 = G_m \hat{v}_g \sin\omega(t - \tau_k),$$

while the current flowing from the grid is given by

$$i_2 = G_m \hat{v}_g \sin\omega(t - \tau_k - \tau_a).$$

Consequently an alternating grid current flows given by :-

$$i_1 - i_2 = 2 G_m \hat{v}_g \cos\omega\left(t - \tau_k - \frac{\tau_a}{2}\right) \sin\omega\frac{\tau_a}{2}$$

$$= 2 G_m \hat{v}_g \sin\omega\frac{\tau_a}{2} \left(\cos\omega t \cos\omega\left(\tau_k + \frac{\tau_a}{2}\right) + \sin\omega t . \sin\omega\left(\tau_k + \frac{\tau_a}{2}\right)\right).$$

The magnitude of the component of current in phase with v_g is given by :-

$$\hat{i}_r = 2 G_m \hat{v}_g \sin\frac{\omega\tau_a}{2} \sin\omega\left(\tau_k + \frac{\tau_a}{2}\right)$$

whilst that of the current in quadrature with v_g is given by :-

$$\hat{i}_c = 2\, G_m\, \hat{v}_g \sin \frac{\omega \tau_a}{2} \quad \cos \omega \left(\tau_k + \frac{\tau_a}{2}\right).$$

If the ratio of each transit time to the period of operation is small i.e. if $\omega\tau_k$ and $\omega\tau_a$ are small,

$$\hat{i}_r \doteqdot G_m\, \hat{v}_g\, \omega^2 \left(\tau_a \tau_k + \frac{\tau_a^2}{2}\right)$$

$$\text{and } \hat{i}_c \doteqdot G_m\, \hat{v}_g\, \omega\, .\, \tau_a\, .$$

Hence the input resistance R_i is given by :-

$$R_i = \frac{1}{G_m\, \omega^2 \left(\tau_a \tau_k + \frac{\tau_a^2}{2}\right)}$$

and the increased input capacitance when the amplifier is operating is given by :-

$$C_i = G_m\, \tau_a.$$

The transit times are dependent on the linear dimensions of the valve and on the grid and anode voltages. The loss of power at the grid is used in increasing the mean electron velocity and appears as heat at the anode. The increase in input capacitance is small and can usually be neglected. However R_i, which damps the input tuned circuit, is inversely proportional to the square of the operating frequency, and hence the input conductance is proportional to the square of the frequency.

Consider the case of an amplifier circuit, the valve of which is a CV1091. The values of R_i at different frequencies are

100,000ohms	at	10 Mc/s
4,000ohms	at	50 Mc/s
250ohms	at	200 Mc/s.

Thus if the dynamic resistance of the input tuned circuit is, say, 100 kΩ the damping introduced is moderate at 10 Mc/s but prohibitive at 200 Mc/s unless special measures are taken. Hence, while damping of the input circuit is negligible at low frequencies it is considerable at frequencies above about 50 Mc/s.

Input damping can be reduced by decreasing the transit times; i.e. by reducing the spacing of the electrodes of the valve. The electrode areas must also be reduced so that the interelectrode capacitances of the valves are not increased. Extremely small spacing is the main criterion of Acorn valves, but the percentage of rejects in manufacture is high, and other valves (CV1091 and CV1065) with rather larger spacing are in general use. The grid-cathode transit time can be reduced by increasing the effective voltage at the grid. This means, however, that in order to maintain space-charge limitation, the emission density of the cathode must be high. Analysis shows that the cathode emission density required to ensure space charge limitation for a given ratio of transit time to operating period is proportional to the cube of the operating frequency.

Typical figures are as follows :-

If the transit time of the grid-cathode space must not be greater than a quarter of the operating period, then for a grid-cathode spacing of 1mm. the minimum current densities required at various frequencies are

2×10^{-6} A/sq cm at 10 Mc/s

250×10^{-6} A/sq cm at 50 Mc/s

0.016 A/sq cm at 200 Mc/s

2 A/sq cm at 1000 Mc/s.

In a valve oscillator the phase difference between grid voltage and anode current due to an appreciable transit time makes it difficult to adjust the relative phases at anode and grid for efficient operation. This does not apply to a valve amplifier, in which anode and grid circuits are independent of each other.

A further effect of transit time is an increase in anode dissipation. This may be of importance in power amplifier and oscillator circuits, in which the output power is limited by permissible anode dissipation. The Door-Knob valve is designed to give useful power output at frequencies as high as 600 Mc/s. The electrode spacing is small, and 35 watts can be dissipated at the anode since it is fitted with three radiation fins and enclosed in a relatively large glass envelope (of door-knob shape).

26. Circuit Limitations

(i) Frequency limitation

The values of the interelectrode capacitances and of the inductance of the internal leads of a valve operating as an oscillator fix the upper limit of frequency. If, however, the valve leads form part of a transmission line system of total length ℓ, this frequency limitation is removed. For example, if the end of the transmission line is short-circuited, the resonant frequency of the circuit formed by the line and the interelectrode capacitance is given by

$$\tan \frac{2\pi \ell f}{c} = \frac{1}{2\pi f C R_o}$$

(See Chap. 4 Sec. 14 for the effective inductance L_ℓ)

where $c = 3 \times 10^{10}$ cms/sec

C = interelectrode capacitance

R_o = characteristic resistance of the line.

For a given frequency f the permissible length of the line can be increased by decreasing R_o. Hence a valve designed for use with coaxial lines (which are of low characteristic impedance), and whose electrodes and leads inside the envelope are integral parts of the lines, can be used for operation at very high frequencies.

In an oscillator the maintenance of continuous oscillations depends on the transference of energy from some source of power to a tuned circuit. In the case of a parallel tuned circuit fed with energy from a parallel negative-resistance source, the dynamic resistance of the tuned circuit must be greater than the magnitude of the negative resistance for oscillations to build up. The dynamic resistance $\frac{L}{CR}$, where R is the series resistance of the coil L, may be made large only if C is kept sufficiently small. Hence it is necessary to keep interelectrode and stray capacitances to a minimum. This is still true when resonant lines are used as tuned circuits, since normally the line acts as an inductive susceptance resonating with interelectrode and other unavoidable capacitances.

The same problem does not arise in RF amplifiers above about 50 Mc/s since, owing to the high conductance of the input circuit due to transit time, cathode lead inductance, etc., the unavoidable resistance in parallel with the tuned circuits is much smaller than the undamped dynamic resistance of the latter. Since we may therefore neglect series damping owing to the low value of R_p, the parallel damping resistance, the Q of the total parallel circuit may be written

$$Q = R_p\sqrt{\frac{C}{L}}\,.$$

It may therefore be advantageous to make $\frac{C}{L}$ large in order to make the circuit sufficiently selective.

(ii) Input damping

The inductance of the internal cathode lead of a valve, which forms part of the input circuit of an amplifier, may bring about damping of the input circuit.

This inductive reactance in the cathode lead introduces feedback and modifies both input and output characteristics of the amplifier. In particular a component of the input current in phase with the applied voltage is generated, thus adding to the input conductance of the amplifier.

Fig. 383 shows the input circuit of an amplifier. An external voltage $v_i = \hat{v}_i \sin\omega t$ is applied to the grid, and a voltage v_L is developed across the cathode-lead inductance L_K. The cathode current is the sum of i_i and i_a, where i is the current through the grid-cathode capacitance C and i_a is the space current of the valve. Assuming the valve to be a pentode in which the slope resistance is very large compared with the anode load, we may write

$$i_a \doteqdot G_m v_i\,.$$

It follows from the figure, that

$$v_s = v_i + v_L\,;$$

hence $v_s = v_i + j\omega L(i_i + G_m v_i)$

or $$v_s = \frac{-ji_i}{\omega C} + j\omega L\left(i_i - \frac{G_m j i_i}{\omega C}\right);$$

i.e.,

$$v_s = \frac{i_i}{\omega C}\left[\omega L\, G_m + j\,(\omega^2 LC - 1)\right].$$

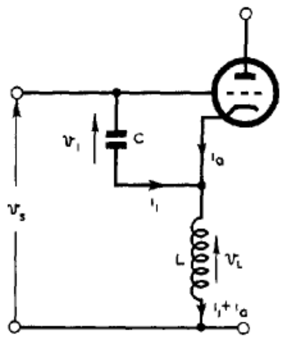

Fig. 383 - Effect of Cathode Lead Inductance.

Therefore the input admittance of the amplifier is given by

$$\frac{i_i}{v_s} = \frac{\omega C \, [\omega L \; G_m - j \; (\omega^2 LC - 1)]}{(\omega L \, G_m)^2 + (\omega^2 LC - 1)^2}$$

Provided the operating frequency is not too high, then

$$\omega^2 LC \ll 1$$

and $\omega LG_m \ll 1$,

and the real part of the admittance (i.e., the conductance G_K) is given by

$$G_K \doteqdot G_m \omega^2 LC.$$

Thus the effect is that of placing in parallel with the input a resistance

$$R_K = \frac{1}{G_m \omega^2 LC}$$

The input conductance $\frac{1}{R_K}$ is proportional to the mutual conductance and the square of the frequency just as the input conductance which results from transit time. The two effects are of the same order of magnitude, and the separation of the two causes is difficult.

Suppose, for example, that G_m = 5 mA/volt, C = 5 pF and L = 0·05 μH, and the frequency of operation f is such that

$$\omega^2 LC \ll 1$$

and $\omega L \, G_m \ll 1$.

[$\omega^2 LC \doteqdot 1$ if f $\doteqdot$ 300 Mc/s and $\omega L \, G_m \doteqdot 1$ if f = 4000 Mc/s]

Take f = 50 Mc/s.

Then the value of $R_K = \frac{1}{0{\cdot}005 \, . \, (2\pi \, . \, 50 \, . \, 10^6)^2 . \, 0{\cdot}05 . 10^{-6} . 5 . 10^{-12}}$

$$\doteqdot 8000 \text{ ohms.}$$

Since the pinch construction of supporting electrode leads in valves involves long leads, so that L is large, a decided decrease of input conductance is obtained if such a construction is replaced by short straight electrode leads mounted in a flat disc of glass. This latter type of base construction is used in the CV1091. A marked decrease of effective cathode lead inductance is obtained if such a lead consists of multiple wires in parallel as in the CV1136.

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