RADIO RECEIVERS

V. BARKAN
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MIR PUBLISHERS MOSCOW

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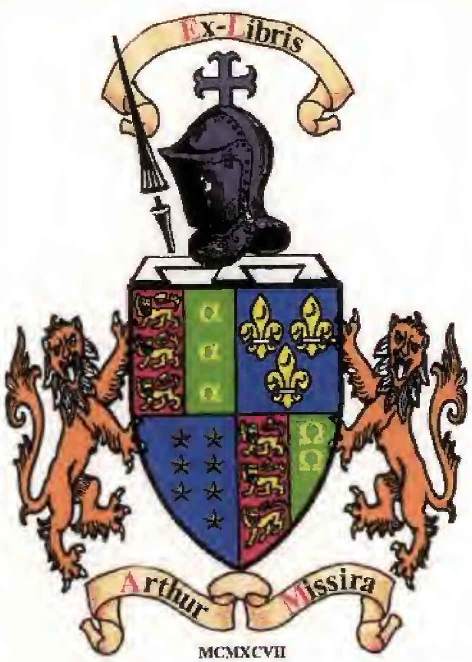

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В. Ф. Баркан, В. К. Жданов

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RADIO RECEIVERS

V. BARKAN, V. ZHDANOV

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CHAPTER I

INTRODUCTION

1. Uses for and Classes of Radio Receivers

Transmission of intelligence by radio is based on modulation. Modulation is a process by which the message to be transmitted is superimposed at the sending end of a radio link as a modulating signal (or simply, the signal) on a strong carrier wave, thereby changing the latter's amplitude, frequency or phase. The modulated carrier is radiated by a transmitting aerial as a wave of electromagnetic energy which propagates through space at the velocity of light.

At the point of reception, the modulated wave is picked up by a receiving aerial and is fed to the receiver input. In the receiver, the signal is separated from the radio-frequency carrier and drives the receiver load, which may be a speaker, a recorder, a cathode-ray tube, etc.

As an electromagnetic wave travels away from the transmitter it is weakened, or attenuated. This is why radio receivers should be capable of picking up relatively weak signals.

Uses for Radio Receivers. At present, radio serves a variety of purposes, such as communication, broadcasting, navigation, radar, and telecontrol.

Radio communication is the transmission and reception of messages without wires or waveguides. It includes communication by radio telegraph, radio telephone, radio teletype-writer, radio facsimile, and television. It is the only method of communication between stationary and mobile objects (such as from ship to shore, from ground to aircraft, from ship to ship, or from aircraft to aircraft, and more recently from ground to satellites or from satellites to ground, and between satellites, known as space communication).

Radio broadcasting is radio transmission for general reception, including speech, music and commercial television.

Radio navigation is the use of radio facilities for determining the position or direction or both of ships or planes.

Radar (which is an acronym for *Ra*dio *D*etection *a*nd *R*anging) is a technique for determining the range and bearings of objects (usually called *targets*) by the transmission of beamed high-power signals against reflective targets, the reception of the reflected signals, and the presentation of the resultant data on a dial or a cathode-ray display. Radar may be used for marine navigation, gun (fire) control, earth surveillance from the air, etc.

Telecontrol is a technique for control of machinery by radio.

Classes of Receivers. There are (a) communication and (b) broadcast receivers.

Communication receivers are used in point-to-point radio telephone and telegraph service, radio navigation systems, and radar.

Broadcast receivers are used for the reception of sound and visual programmes intended for the general public.

Communication receivers are usually classed according to operating principle, wavelength (or frequency), type of service, type of modulation, type of installation, and range of operation.

According to the operating principle, there are tuned radio-frequency (TRF), regenerative, superregenerative, and superheterodyne receivers.

According to wavelength (or frequency), there are long-wave (low-frequency), medium-wave (medium-frequency), short-wave (high-frequency), and ultra-short-wave (VHF, UHF, SHF, etc.) receivers. It may be added that although the trend has for some time been to drop the "wave" terms in preference to the "frequency" terms, the former are still in use in the Soviet literature on the subject.

According to the type of modulation, there are amplitude-modulated (AM) and frequency-modulated (FM) receivers.

According to the type of installation, there are stationary and mobile receivers.

According to the range of operation, there are long-distance, medium-range, and short-range receivers.

Review Questions

1. How is intelligence transmitted by radio?
2. What are the functions of radar?
3. What is the function of communication receivers?

2. Receiver Characteristics

The characteristics of importance to any receiver are: power output, output voltage, sensitivity, selectivity, bandwidth, frequency range, and fidelity.

Power Output and Output Voltage. The power output of a receiver is the power delivered to its load. It varies with the type of load which may be a speaker, a telegraph printer, an automatically controlled device, etc.

In television and radar receivers, in which a cathode-ray tube is the load, output voltage is specified instead of output power. Receiver output voltage usually ranges from a fraction of a volt to a few tens of volts.

Sensitivity. The sensitivity of a radio receiver is defined as the strength of the signal at its input required to produce a normal test output (which is a specified power at its load). The smaller the required input signal, the higher the receiver sensitivity. Sensitivity is expressed in microvolts.

For receivers (mainly, transistor) with built-in ferrite aerials the sensitivity is defined in terms of the minimum field intensity at the point of reception that produces the normal voltage applied to the load, and not in terms of the input signal. The sensitivity is then expressed in millivolts per metre of effective height of the aerial.

The sensitivity of a receiver is decided by the properties of each of its stages, but can be fully realised only if receiver noise at the output is lower than its signal. The signal-to-noise ratio specified for a particular type of receiver varies with the nature of the signal to be received.

Bandwidth and Selectivity. The signal arriving at the input of a receiver contains a spectrum of frequencies which are due to modulation at the sending end. The width of the frequency spectrum is different for different types of modulation. An AM (amplitude-modulation) transmitter radiates the carrier frequency, f_0, and a whole gamut of what are called *side frequencies*, or *side bands* which extend from $f_0 - F_{max}$ to $f_0 + F_{max}$. Together, they make up the bandwidth of the transmitter. As is seen, in amplitude modulation the bandwidth is $2F_{max}$. In wide-band frequency modulation, the bandwidth is mainly determined by the frequency deviation Δf_{max} and is equal to $2\Delta f_{max}$.

Among other things, a good receiver should be capable of receiving the signal along with its side bands. In other words, it should perform satisfactorily within that band of frequencies, so that the natural relations between the amplitudes of the constituent frequencies will remain undistorted. This can be obtained only if a receiver has a constant sensitivity over that band, called its bandwidth.

On the other hand, a receiver should be able to discriminate against signals of frequencies differing from that of the desired signal. This is called the *selectivity* of a receiver.

The ideal *selectivity*, or *response*, *curve* (also called the *resonance curve*) is shown in Fig. 1.1*a*. A receiver with such a response curve, when tuned to a transmitting station, would readily pass the intelligence transmitted and reject all other signals.

The ideal response curve, however, is impossible to produce in practice. Normally, practical receivers have the sort of response shown in Fig. 1.1*b*. At f_0 is the resonant frequency of the receiver tuned circuits. The amount off resonance, Δf, is laid off as abscissa, and the ratio of the signal at a given frequency off resonance to the signal at resonance, Y, as ordinate.

As is seen, the frequencies removed from resonance contribute less and less to the signal. The band of frequencies which are still important to faithful reproduction of the signal make up the radio-frequency (r. f.) bandwidth of a receiver. The limiting frequencies are usually defined for a specified value of the ratio Y. In Fig. 1.1*b*, the bandwidth is for $Y = 0.7$.

The response curve provides a basis for comparison of receivers from the view-point of their selectivity. The smaller the ratio Y at a given amount off resonance, the weaker the signal at that frequency, and the greater the desired signal. For com-

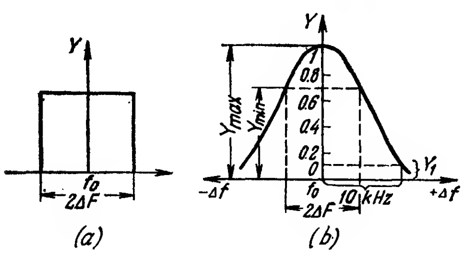


Fig. 1.1. Selectivity or response curves of radio receivers

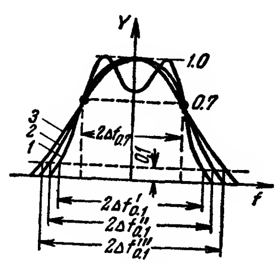


Fig. 1.2. Resonance curves of various oscillatory systems with similar bandwidth

munication and broadcast receivers the selectivity is specified at 10 kilohertz off resonance, because carrier frequencies are spaced that amount apart by international agreement. In Fig. 1.1*b*, $Y_1 = 0.1$ (20 db down) shows that the signal from an adjacent station is only one-tenth in strength compared with the desired signal. In modern communication and broadcast receivers, adjacent frequencies are attenuated from 10 to 1000 times (from 20 to 60 decibels down).

A wide bandwidth and a high selectivity are conflicting requirements—widening the former impairs the latter, and vice versa. A sort of balance between them can be struck by making the response curve approach the ideal square one.

Figure 1.2 shows the response curves for various types of r. f. circuits. They all have the same bandwidth at 0.7 level, or between 3 db points. Outside the bandwidth, however, they differ in slope. Obviously, there must be an additional criterion for comparison of these circuits. Such a criterion is the bandwidth ratio (or relative bandwidth), defined as the ratio of the bandwidth at the frequency of interest to that at 3 decibels down

$$K_{bw} = 2\Delta f / 2\Delta f_{3\mathrm{db}} = \Delta f / \Delta f_{3\mathrm{db}} \tag{1.1}$$

It is usual to specify the ratios of the 20-db and 40-db bandwidths (between 0.1 and 0.01 points, respectively) to the 3-db one (at 0.7 level)

$$K_{bw20\mathrm{db}} = \Delta f_{20\mathrm{db}} / \Delta f_{3\mathrm{db}}$$

and

$$K_{bw40\mathrm{db}} = \Delta f_{40\mathrm{db}} / \Delta f_{3\mathrm{db}} \tag{1.2}$$

The closer the value of K_{bw} to unity, the closer the response curve is to a square (ideal) one. Of the three resonance curves shown in Fig. 1.2, curve *1* comes closest to the ideal characteristic and has the lowest value of K_{bw}.

Fidelity of a Radio Receiver. This is a degree with which a radio receiver reproduces at its output the envelope of the modulated wave applied to its input.

The signal picked up by the receiver aerial goes through a succession of circuits which contain linear and non-linear elements. These elements cause distortion to the signal, so that a fully faithful reproduction is unfeasible.

Radio receivers are subject to frequency distortion, non-linear distortion, and phase distortion.

Frequency distortion is due to non-uniform amplification over the bandwidth of the receiver. This upsets the natural relation between the amplitudes of the harmonics contained in the composite signal. Frequency distortion is expressed in terms of the frequency distortion factor which shows how much the signal is attenuated at the boundary of the bandwidth.

In the selectivity curve of Fig. 1.1*b*, the distortion factor

$$M = Y_{min}/Y_{max}$$

shows how much the signal at the limiting frequency of the bandwidth is attenuated.

The a.f. section of a receiver cannot provide for uniform amplification of the signal either. As a result, the signals at the limiting frequencies of the bandwidth may be attenuated appreciably.

The attenuation of the upper frequencies in the signal may be as great as 50 per cent throughout a receiver.

Non-linear distortion is due to the fact that on its way through a receiver the signal passes through circuit elements whose volt-ampere characteristics are not a straight line (valves, transistors, iron-cored inductors, etc.). Because of this, undesirable frequencies, not present in the input signal, might appear in the output, and its waveform might be distorted.

Phase distortion appears as a result of upsetting phase relations between the harmonic components of a non-sinusoidal signal, because of which the signal waveform differs from that of the input.

Non-linear and phase distortions are dealt with in detail in Chapter II.

Apart from the characteristics listed above, a radio receiver has to meet a number of requirements in respect to its construction. These above all are size and weight, reliability, mechanical and electrical strength, and ease of control.

Review Questions

1. Define the selectivity of a receiver.
2. Define the bandwidth of a receiver.
3. Which circuits of a receiver control its selectivity (frequency response)?
4. Describe the ideal response (selectivity) curve.

5. List the circuit elements responsible for frequency distortion in a receiver.
6. List the circuit elements responsible for non-linear distortion in a receiver.

3. Functional Units of a TRF Receiver

A general idea about operation of a receiver can be obtained by reference to its block diagram showing the various functional units.

We shall begin with the straight or tuned radio-frequency (TRF) receiver whose block-diagram appears in Fig. 1.3.

The input circuit of the receiver extracts the desired signal and attenuates other signals.

The radio-frequency (r.f.) amplifier amplifies the extracted signal and further attenuates the unwanted signals. The detector *D* converts the modulated radio wave into an audio-frequency signal. The detector may be a vacuum valve or a crystal diode.

The audio-frequency amplifier builds up the audio-frequency signal to deliver the power output or output voltage necessary for the operation of the terminal equipment (load).

A TRF receiver is not quite sensitive or selective, particularly on short and ultra-short waves.

The bandwidth $2\Delta F$ of a single tuned circuit and its quality-factor Q are related thus

$$2\Delta F = \frac{f_0}{Q} \tag{1.3}$$

where f_0 is the signal frequency.

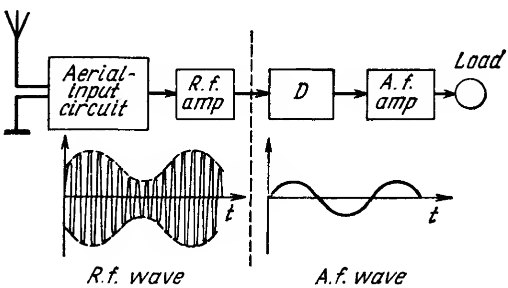


Fig. 1.3. Block-diagram of a TRF receiver

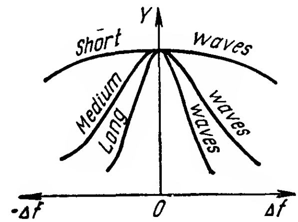


Fig. 1.4. Resonance curves of tuned circuits on different wave bands

The Q-factor of the tuned circuit may be made almost constant on all the bands. This will improve the bandwidth but the selectivity will be impaired. For instance, at $f_0 = 300$ kilohertz and $Q = 100$, the bandwidth is

$$2\Delta F = \frac{300}{100} = 3 \text{ kilohertz}$$

A tuned circuit possessing a similar Q at a frequency of $f_0 = 30{,}000$ kilohertz will have a bandwidth of 300 kilohertz. It should be noted that a TRF receiver has several circuits tuned to the signal frequency. Therefore, the selectivity and bandwidth of a practical TRF receiver considerably differ from the values indicated in the example. The example only gives a general idea of the cause for poor selectivity in TRF receivers. Actually, the selectivity in a commercial receiver is better and the bandwidth, narrower. However, on short and ultra-short waves the resonance curve of the receiver remains broad and the selectivity of the receiver is insufficient. Fig. 1.4 shows the resonance curves of tuned circuits on different wave bands. High selectivity and the required bandwidth may be obtained in superheterodyne receivers.

TRF receivers are seldom used on short and ultra-short waves. Receivers of this type find some application in television, where a wide bandwidth is used.

Review Questions

1. Name the main disadvantages of the TRF receiver.
2. What is the function of a detector in a receiver?

4. Functional Units of a Superheterodyne Receiver

In a superheterodyne receiver, the incoming r.f. signal is amplified and converted to a signal at what is called the *intermediate frequency*. The i.f. signal is then amplified and detected. Frequency conversion is effected in such a way that the envelope of the modulated wave is preserved. Since for any signal frequency there is one and the same intermediate frequency, it is easy to build the i.f. circuits of the requisite bandwidth and high selectivity. The block-diagram of a superheterodyne receiver is shown in Fig. 1.5.

The aerial-input circuit and radio-frequency amplifier have but little effect on receiver selectivity on short and ultra-short waves, and only improve receiver sensitivity.

The frequency changer consists of a mixer and a local oscillator. The latter is a low-power self-excited radio-frequency oscillator which generates a signal at a slightly different frequency from the r.f. signal, f_o.

Frequency conversion is obtained in the mixer which accepts two signals, the incoming r.f. signal f_s picked up by the aerial, and the local-oscillator signal f_o. The i.f. signal at the output of the frequency changer usually is a difference frequency, $f_o - f_s = f_i$.

The intermediate-frequency amplifier operates at a constant frequency and amplifies the i.f. signal to the value required for the normal operation of the detector. I.f. amplifiers are usually band-pass amplifiers whose resonance characteristic is nearly rec-

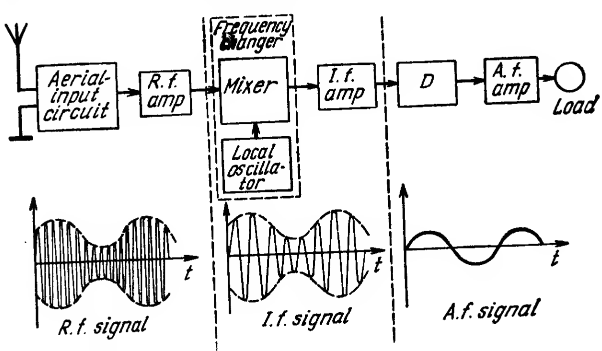


Fig. 1.5. Block-diagram of a superheterodyne receiver

tangular. In this way high selectivity can be obtained at uniform amplification within the bandwidth.

The remaining units (the detector and audio-frequency amplifier) operate in a superheterodyne receiver just as they do in a TRF receiver.

In addition to high selectivity and sensitivity, superheterodyne receivers show better overall performance.

The superheterodyne principle is the basis of modern receivers.

Review Questions

1. What characteristics of a receiver are improved by frequency conversion?
2. What is the function of the i.f. amplifier?
3. What is the function of the local oscillator in a superheterodyne receiver?

5. Receiver Circuit Analysis

In the study of various receiver circuits it is necessary to establish mathematical relations between the circuit parameters and express them as equations, convenient for calculation.

Analysis of the amplifier stage has as its objective to establish relationships between the voltages and currents at its output and input.

These relationships may be presented either graphically by sets, or families, of input and output characteristic curves, or analytically by suitable equations.

The graphical method is ordinarily used to establish the quiescent (no-signal) operation of a stage component or element and also to investigate its large-signal operation.

The analytical method is used in analysis of the stage operation, its physical properties, and in the derivation of design equations from which one can calculate the parameters of the stage elements.

In the general case, an amplifier stage may be represented by a network with two input and two output terminals (Fig. 1.6).

Fig. 1.6. A two-port

Such networks are called *four-terminal networks* or *two-ports*, each pair of terminals serving as a *port*.

The physical properties of a two-port may be expressed in terms of four electrical quantities, input voltage v_1, input current i_1, output voltage v_2, and output current i_2.

The four quantities may be related in the following way:

$$\text{I.}\quad \left.\begin{aligned} v_1 &= z_{11}i_1 + z_{12}i_2 \\ v_2 &= z_{21}i_1 + z_{22}i_2 \end{aligned}\right\} \tag{1.4}$$

where v_1, i_1 = input voltage and current of the two-port
v_2, i_2 = output voltage and current of the two-port
z_{11} = input impedance with the output port open ($i_2 = 0$)
z_{22} = output impedance with the input port open ($i_1 = 0$)
z_{12} = reverse transfer impedance (that is, from output to input) with its input port open-circuited, such that the output current produces v_1
z_{21} = forward transfer impedance (that is, from input to output) of the network with its output port open-circuited, such that the input current produces v_2.

These are the *z-parameters* of a two-port. Since they are determined with the input and output ports open-circuited and have dimensions of impedance, another name for them is the *open-circuit impedance parameters*. Sometimes, the impedances are pure resistances. Then,

$$\begin{aligned} z_{11} &= R_{11} \\ z_{21} &= R_{21} \\ z_{12} &= R_{12} \\ z_{22} &= R_{22} \end{aligned}$$

$$\text{II.}\quad \left.\begin{aligned} i_1 &= y_{11}v_1 + y_{12}v_2 \\ i_2 &= y_{21}v_1 + y_{22}v_2 \end{aligned}\right\} \tag{1.5}$$

where y_{11} = input admittance of the two-port with its output port short-circuited ($v_2 = 0$)
y_{22} = output admittance of the two-port with its input port short-circuited ($v_1 = 0$)
y_{12} = reverse transfer admittance of the two-port with its input port short-circuited ($v_1 = 0$)
y_{21} = forward transfer admittance of the two-port with its output port short-circuited ($v_2 = 0$).

These are the *y-parameters* of two-ports. Since they are determined with the input and output ports short-circuited and have dimensions of admittance, they are called the *short-circuit admittance parameters*.

$$\text{III.}\quad \left.\begin{aligned} v_1 &= h_{11}i_1 + h_{12}v_2 \\ i_2 &= h_{21}i_1 + h_{22}v_2 \end{aligned}\right\} \tag{1.6}$$

where h_{11} = input impedance of the two-port with the output port short-circuited ($v_2 = 0$)

h_{22} = output admittance of the two-port with its input port open-circuited ($i_1 = 0$)

h_{21} = forward current gain of the two-port with its output port short-circuited ($v_2 = 0$)

h_{12} = reverse voltage gain or voltage feedback factor of the two-port with its input port open-circuited ($i_1 = 0$).

These are the *h*-, or *hybrid, parameters*, so called because of their scrambled nature—some are determined with the ports short-circuited, and the others with the ports open-circuited; some are dimensionless, one has the dimensions of an impedance, and one those of an admittance.

The values of the network parameters can be found from the characteristic curves, measured with suitable devices, or taken from a data book, which is usually done in the case of transistors.

The choice of a particular set of parameters and related equations is decided by the amplifying element on hand and by the objective of the analysis.

One set of parameters can be converted to any other by reference to Table 1.1, where

$$\left.\begin{aligned} \Delta_z &= z_{11}z_{22} - z_{12}z_{21} \\ \Delta_y &= y_{11}y_{22} - y_{12}y_{21} \\ \Delta_h &= h_{11}h_{22} - h_{12}h_{21} \end{aligned}\right\} \tag{1.7}$$

TABLE 1.1.

	z		y		h	
z	z_{11}	z_{12}	y_{22}/Δ_y	$-y_{12}/\Delta_y$	Δ_h/h_{22}	h_{12}/h_{22}
	z_{21}	z_{22}	$-y_{21}/\Delta_y$	y_{11}/Δ_y	$-h_{21}/h_{22}$	$1/h_{22}$
y	z_{22}/Δ_z	$-z_{12}/\Delta_z$	y_{11}	y_{12}	$1/h_{11}$	$-h_{12}/h_{11}$
	$-z_{21}/\Delta_z$	z_{11}/Δ_z	y_{21}	y_{22}	h_{22}/h_{11}	Δ_h/h_{11}
h	Δ_z/z_{22}	z_{12}/z_{22}	$1/y_{11}$	$-y_{12}/y_{11}$	h_{11}	h_{12}
	$-z_{21}/z_{22}$	$1/z_{22}$	y_{21}/y_{11}	Δ_y/y_{11}	h_{21}	h_{22}

Before any of the parameter equations may be applied to a practical circuit, however, the latter should be reduced to some simple form, known as an *equivalent circuit*. Consider some of the principles underlying the use of equivalent circuits.

Thévenin's Theorem. This theorem, widely used in circuit analysis, states that any circuit made up of a linear two-port with an emf E at its input and a load impedance Z at its output (Fig. 1.7*a*), may be replaced by an equivalent circuit containing an equivalent generator (Thévenin equivalent) of emf $\dot{V}$ having internal impedance Z', and the same load Z (Fig. 1.7*b*).

In this statement, the linear two-port may contain linear inductances, capacitances, and resistances (L, C, R), connected in any manner; $\dot{V}$ is defined as the voltage at the output terminals *ab* of the two-port with the load disconnected. The internal impedance of the equivalent generator is defined as that seen looking into the output port *ab*, with the input *cd* short-circuited.

Then by Ohm's law and from the equivalent circuit, the load current $\dot{I}$ is

$$\dot{I} = \frac{\dot{V}}{Z' + Z} \tag{1.8}$$

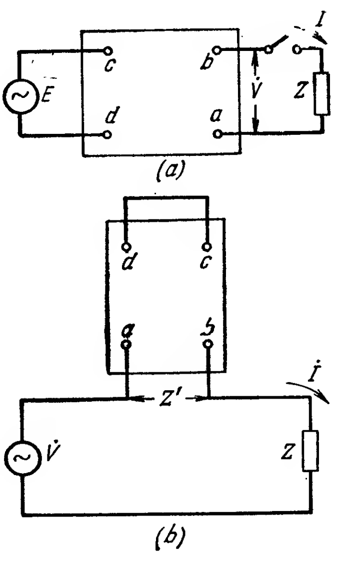


Fig. 1.7. Explaining Thévenin's theorem

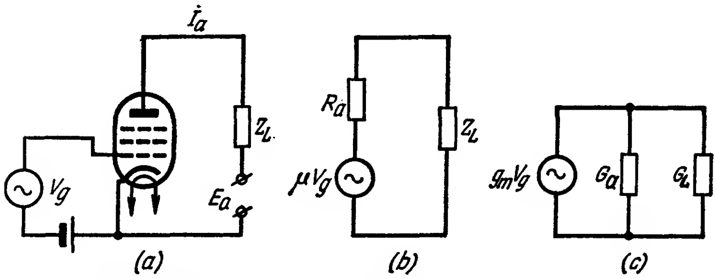


Fig. 1.8. Valve circuit and its equivalent

When we deal with a valve circuit, the valve is assumed to be an a.c. generator. Accordingly, the valve is substituted by an equivalent generator developing an emf μV_g and having an internal resistance R_a. Hence, the circuit of Fig. 1.8*a* may be represented by the equivalent circuit of Fig. 1.8*b*.

Ohm's law also gives the a.c. component of the anode current:

$$\dot{I}_a = \frac{\mu V_g}{R_a + Z_L} \tag{1.9}$$

As an alternative, the valve may be reduced to what is known as the equivalent current-generator circuit, also called the Norton equivalent, $g_m V_g$ (Fig. 1.8*c*). In this case, the anode (a.c.) resistance of the valve, R_a, is replaced with its anode (a.c.) conductance G_a, defined as $G_a = 1/R_a$. The load, too, is represented by a conductance G_L.

If in the with-signal operation the load impedance is comparable with the anode resistance of the valve, preference should

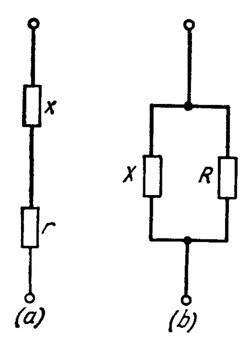


Fig. 1.9. Series and parallel circuits

be given to the Thévenin equivalent (the equivalent voltage-generator circuit). If the load impedance is only a few tenths or even a few hundredths of R_a, the equivalent current-generator circuit (the Norton equivalent) is more convenient.

The same reasoning applies to the representation of a transistor amplifier stage by its equivalent circuit.

Series to Parallel and Parallel to Series Conversion. Series-connected resistances and reactances may be replaced by an equivalent parallel connection, and vice versa (Fig. 1.9).

In such conversions, the reactive elements remain unchanged, i.e. $x = X$, and the resistive elements are recalculated. Thus, when changing over from circuit a to circuit b

$$R \cong \frac{x^2}{r} \tag{1.10}$$

and going back from b to a

$$r = \frac{X^2}{R} \tag{1.11}$$

Review Questions

1. When is Thévenin's theorem used?
2. Where may the Thévenin equivalent (the Norton equivalent) be used?
3. Develop an equivalent circuit containing a Thévenin generator for a transistor stage.

CHAPTER II

AUDIO-FREQUENCY VOLTAGE AMPLIFIERS

6. General

An audio-frequency amplifier is intended for amplification in the frequency range from 16 to 20,000 hertz.

Audio-frequency amplifiers are used in radio receivers, long-distance wire communication, radio rediffusion systems, automatic process control, remote control, and many other branches of engineering.

A generalized block-diagram of an amplifier appears in Fig. 2.1. The input voltage source V_g may be the detector of a receiver, a microphone, a pick-up adapter, the play-back head of a tape-recorder, a photo-cell, a thermocouple, or other sources of a.f. signals.

An a.f. amplifier may in turn be loaded into a speaker, a pair of headphones, the recording head of a tape recorder, an oscilloscope, a next stage in an amplifier, and the like.

The input voltage of an amplifier may vary from a split microvolt to several volts, depending on the signal source used. For example, a photo-cell generates a voltage of a few microvolts. Electrodynamic microphones generate anywhere from one to three millivolts. A pick-up adapter should deliver 100 to 500 millivolts. The detector of a receiver supplies from a few tenths of a volt to several volts.

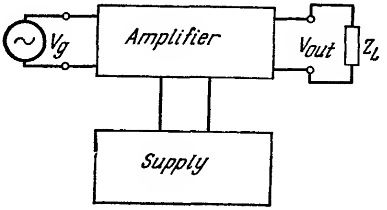


Fig. 2.1. Block-diagram of a single-stage amplifier

The output voltage of an amplifier may likewise range from a few tenths of a volt to hundreds of volts, and the power, from a few milliwatts to hundreds or even thousands of watts (kilowatts).

Amplification is essentially the control of a secondary energy source delivering its power into the load circuit by the energy from a primary audio-frequency source.

An amplifier is a device in which the power delivered to the load is many times the power expended for control at its input.

Present-day electronics offers a variety of devices which can accomplish amplification. These are vacuum valves, transistors, magnetic and ferroelectric amplifiers.

Consider the operating principle of amplifiers based on a vacuum valve and a transistor.

Referring to Fig. 2.2, the grid of the valve accepts $V_g = V_i$, that is, an a.f. input voltage. The alternating electric field that the input voltage establishes between the grid and cathode causes marked variations in the anode current. This is another way of saying that an alternating component develops in the anode current of the valve. On passing through the high-ohmic load impedance Z_L, the a.c. anode current produces a voltage drop V_L across the load (or the output voltage V_o) varying at the signal frequency. When the value of Z_L is sufficiently high, the a.c. output (load) voltage will be many times the a.c. input (grid) voltage.

Thus, the low-voltage source in the grid circuit controls the current in the high-resistance anode circuit which contains a local anode supply source.

The ratio of the a.c. output voltage V_o to the a.c. input voltage V_i is called the stage voltage gain:

$$K = V_o/V_i = V_L/V_g$$

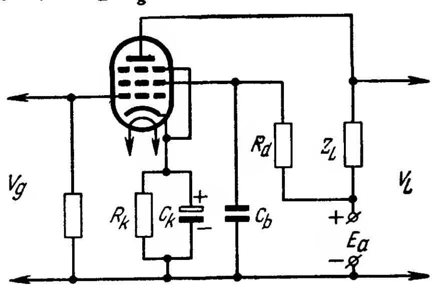


Fig. 2.2. Pentode amplifier circuit

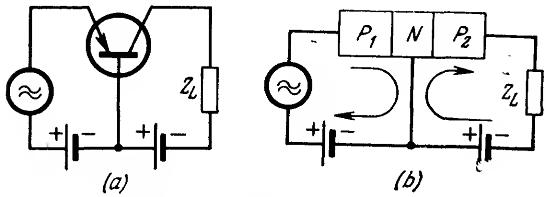


Fig. 2.3. Transistor amplifier

The part of the anode circuit between the cathode and the negative terminal of the anode supply source applies a fixed negative bias voltage to the valve grid.

The cathode resistor R_k passes only the direct component of the anode and screen-grid current. The alternating component is bypassed around R_k by a bypass capacitor C_k of low reactance. R_d in the screen-grid circuit is a dropping resistor which absorbs the excess anode supply voltage. C_b bypasses the alternating component of the screen-grid current to the cathode (via C_k).

Figure 2.3*a* shows the circuit schematic of a transistor amplifier stage. For its operation, a transistor, as an amplifying element, depends on the injection of minority carriers from one region to another In the *P*-regions (the emitter, P_1, and the collector, P_2), the majority carriers are holes (Fig. 2.3*b*). In the *N*-region (the base, *N*) the majority carriers are electrons. The source voltage connected in the emitter circuit forward-biases the P_1-N junction, and the electric field established at the junction sweeps holes from the emitter region into the base region, while the electrons are swept from the base region into the emitter region. The resistance of the P_1-N junction is drastically reduced, and a current begins to flow in the emitter circuit.

The carriers transported from one region to another are the minority for the latter. The transport of carriers from one region into another is called the injection of minority carriers.

The source voltage connected in the collector circuit reverse-biases the N-P_2 junction. With the emitter lead open-circuited, the resistance of N-P_2 junction is very high. In such a case, an electric current can flow only due to the transport of minority carriers caused by thermal agitation. This collector leakage current, as it is called, does not usually exceed a few microamperes.

When the emitter circuit is completed, the injection of minority carriers, already mentioned, takes place.

The base region is very thin, usually 20 to 30 microns. On the other hand, the distance that minority carriers are able to cover before recombination may be several times greater (120 to 130 microns). Therefore, a sizeable proportion of the minority carriers injected into the base region move into the collector region and fill the depletion (transition) region, considerably increasing its conductivity.

Existent transistors are built so that up to 95 to 99 per cent of all minority carriers moving from the emitter region (P_1) into the base region (N) are swept by the electrostatic field at the N-P_2 junction into the P_2 region, giving rise to a collector current which is 95 to 99 per cent of the current in the emitter circuit.

Thus, a low-voltage current source in the low-resistance emitter circuit produces nearly as great a collector current flowing in a high-resistance circuit and having a voltage, E_{cb}, higher than E_{eb}. If the voltage source in the emitter circuit is an audio-frequency one, it will give rise to an alternating component in that circuit, and nearly as great a current will flow in the collector circuit. On passing through the load impedance $Z_c = Z_L$, it will produce a voltage drop $V_c = V_o$ across it, which will be several times the signal (input) voltage $V_s = V_i$. The ratio of the output, or load, voltage to the input, or signal, voltage is called the voltage gain of a single-stage transistor amplifier:

$$K_v = V_o/V_i = V_c/V_s$$

Main Characteristics of an Amplifier. In order to evaluate the various physical properties of an amplifier, the following basic characteristics are used:

(1) stage gain;
(2) maximum rated power;
(3) bandwidth;
(4) dynamic power range;
(5) efficiency;
(6) distortion.

The *stage gain* K is one of the most important characteristics of an amplifier.

It is customary to distinguish between voltage gain and power gain.

The *voltage gain* is numerically equal to the ratio of the voltage at the amplifier output to the voltage applied to the input

$$K = \frac{V_{out}}{V_{in}}$$

The *power gain* is the ratio of the power at the output of an amplifier to the power applied to its input

$$K_p = \frac{P_{out}}{P_{in}}$$

It may be shown that the total gain of a multi-stage amplifier is equal to the product of its separate stage gains

$$K_t = K_1 K_2 K_3 K_4$$

The gain may vary from several units to hundreds of thousands, or even several millions.

For practical purposes, the gain of an amplifier is often expressed in logarithmic units, decibels (db). Voltage gain in decibels in terms of relative gain is

$$K_{db} = 20 \log_{10} \frac{V_{out}}{V_{in}} = 20 \log_{10} K$$

Power gain in decibels is

$$K_{pdb} = 10 \log_{10} \frac{P_{out}}{P_{in}} = 10 \log_{10} K_p$$

Decibel units for gain are especially convenient in analysis of multi-stage amplifiers. The total voltage gain in decibels is equal to the sum of the gains of the individual stages

$$K_{tdb} = 20 \log_{10} K_1 K_2 K_3 \ldots = 20 \log_{10} K_1 + 20 \log_{10} K_2$$
$$+ 20 \log_{10} K_3 + \ldots = K_{1db} + K_{2db} + K_{3db} + \ldots$$

In exactly the same way, the total power gain in decibels is equal to the sum of the gains of the individual stages

$$K_{pdb} = K_{p1db} + K_{p2db} + K_{p3db} + \ldots$$

The *power output* of an amplifier is the maximum power delivered by the amplifier to the load at specified signal distortion. It usually varies from a fraction of one watt to several watts.

The *bandwidth* of an amplifier is the difference between the limiting frequencies, within which the gain of the amplifier has a specified value. Ordinarily, the gain should remain constant within ± 3 db (± 30 per cent) over the bandwidth.

The bandwidth of an audio-frequency amplifier is determined by its application.

In radio telephony it extends from 300 to 2,500 hertz. In order to reproduce a chorus or an orchestra the bandwidth must be at least 100 to 5,000 hertz.

In test amplifiers, the bandwidth usually embraces the whole audio-frequency sprectrum, i.e. 16 to 20,000 hertz.

In special-purpose amplifiers, the bandwidth is sometimes limited to several tens or hundreds of hertz.

The *dynamic power range* represents the span of power levels from minimum to maximum. Within this range the electrical properties of the amplifier must fall within specified limits. For instance, the quality of reproduction may be considered satisfactory if the ratio of the maximum to minimum power or voltage is 60 db.

The *amplifier efficiency* shows whether the amplifier is sufficiently economical in operation. The electrical efficiency η of the individual stages of an amplifier is equal to the ratio of the power output P to the power P_0 drawn from the anode supply source

$$\eta = \frac{P}{P_0}$$

Distortion due to an amplifier is the change in the signal waveform during its passage through the amplifier.

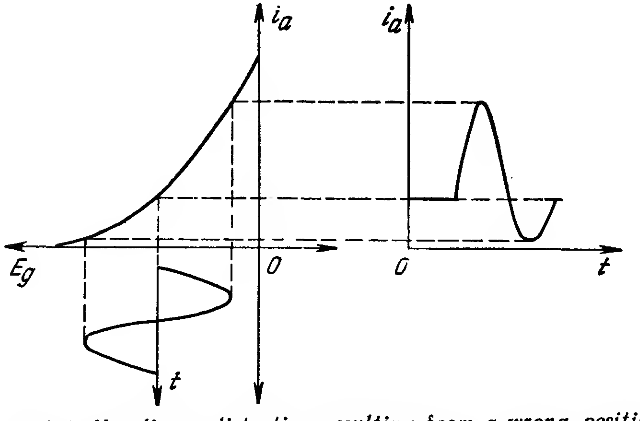


Fig. 2.4. Non-linear distortion resulting from a wrong position of the operating point on the valve characteristic

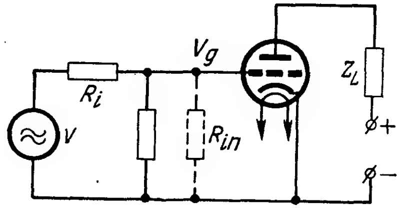


Fig. 2.5. Grid circuit of an amplifier

As already noted, there exists non-linear, frequency, and phase distortion.

Non-linear distortion is caused by the non-linear behaviour of the circuit elements, such as valves or transformers, through which the signal has its path. As a result, the variation of output current is not directly proportional to that of input voltage. This may happen when the operating point of a valve has been positioned on a curved (non-linear) region of its characteristic. Referring to Fig. 2.4, i_a varies in a different manner from e_g.

Non-linear distortion also occurs when the grid of a valve is allowed to draw current. As is seen in Fig. 2.5, during the negative half-cycle of the input voltage there is no grid current. The input resistance R_{in} of the valve at this moment is sufficiently great in comparison with the internal resistance R_i of the source, which makes it possible to disregard its effect upon the input voltage.

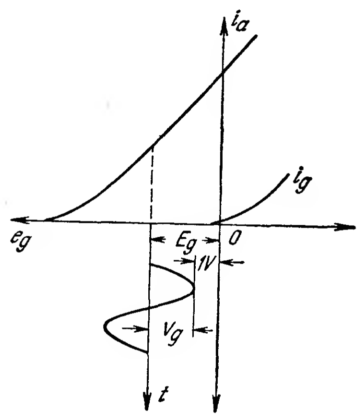


Fig. 2.6. Operation without grid current

During the positive half-cycle of the input voltage the grid draws a current. Due to this, the input resistance of the valve is sharply reduced; this brings about an increased voltage drop across the internal resistance of the supply source and also a decrease in the input voltage of the valve.

The amplitude of the grid voltage (V_g) is noticeably less during the positive half-cycles than during the negative ones. As a result, the sinusoidal signal is distorted already at the amplifier valve input. In order to avoid grid current i_g, the negative bias voltage E_g applied to the valve grid should differ from the amplitude of the input voltage by at least one volt (Fig. 2.6).

$$E_g + V_g \leqslant -1 \text{ volt or } |E_g| - V_g > 1 \text{ volt} \tag{2.1}$$

Non-linear distortion also occurs when an amplifier employs iron-core components operating close to saturation (Fig. 2.7). When the number of ampere-turns is sufficiently large, the operating point may happen to be at the end of the linear region of the magnetisation curve of the transformer used as the anode load. Magnetic-flux change $\Delta\Phi$ and, consequently, change of the emf of induction in the secondary will not be proportional to anode current change. If the alternating component of the anode current is sinusoidal, the waveform of a. c. voltage at the transformer output will be distorted.

From the examples given above it is apparent that in all cases the waveform of the output signal is distorted by high harmonics.

The severity of non-linear distortion may be found by determining the amplitude of the first and higher harmonic components. Once these values are known, the non-linear distortion factor will be given by the following equation:

$$\gamma = \frac{\sqrt{I_2^2 + I_3^2 + \dots}}{I_1} \tag{2.2}$$

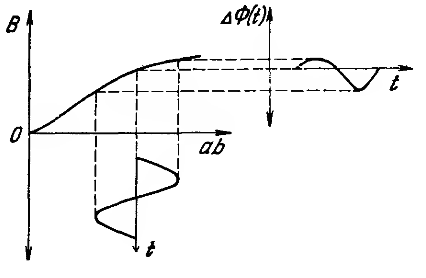


Fig. 2.7. Non-linear distortion in a circuit with iron-core components

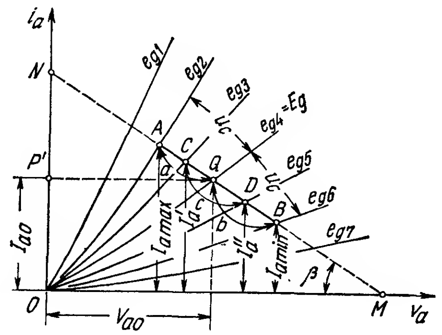


Fig. 2.8. Determining non-linear distortion from the dynamic load line

The amplitudes of a. c. components may be found analytically by resolving the distorted signal into Fourier's series. However, in this case it is necessary to know the exact functional relationship between the anode current and the valve voltages.

The non-linear distortion factor γ may be determined with an accuracy sufficient for practical purposes directly from the dynamic (a. c.) load line of the valve, $i_a = f(e_a)$.

Figure 2.8 gives a family of anode characteristics and the dynamic (a. c.) load line of a triode. The load line *MN* passes through the operating point *Q* which fixes quiescent bias voltage E_{g0} and quiescent anode voltage V_{a0}.

The slope of the characteristic is determined by

$$\tan\beta = \frac{1}{R_L}$$

Points *M* and *N* can be found by plotting them graphically:

$$ON = OQ' + Q'N$$
$$OQ' = I_{a0}$$
$$Q'N = \frac{V_{a0}}{R_L}$$

Having located point *N*, a straight line is drawn through it and through the *Q*-point until it intersects the anode voltage axis at point *M*.

The non-linear distortion factor γ is usually expressed in terms of straight-line segments *a*, *b* and *c*:

$$a = AQ$$
$$b = QB$$
$$c = CD$$

Segments AQ and QB are equal to the segment representing the change in grid voltage $\pm V_g$. Segments CQ and QD are equal to the segment representing the change in the grid voltage, $\pm V_g/2$

$$CD = CQ + QD$$

The equations for computing the harmonics of anode current and non-linear distortion follow.

The amplitude of the first-harmonic anode current is

$$I_{a1} = \frac{1}{3}(I_{a\,max} - I_{a\,min})\frac{a+b+c}{a+b} \tag{2.3}$$

the ampitude of the second-harmonic anode current

$$I_{a2} = \frac{1}{4}(I_{a\,max} - I_{a\,min})\frac{a-b}{a+b} \tag{2.4}$$

the amplitude of the third-harmonic anode current

$$I_{a3} = \frac{1}{6}(I_{a\,max} - I_{a\,min})\frac{a+b-2c}{a+b} \tag{2.5}$$

the average anode current

$$I_{a\,av} = I_{a0} + \frac{1}{4}(I_{a\,max} - I_{a\,min})\frac{a-b}{a+b} \tag{2.6}$$

the second-harmonic distortion is

$$\gamma_2 = \frac{1}{2}\,\frac{a-b}{a+b} \tag{2.7}$$

the third-harmonic distortion is

$$\gamma_3 = \frac{1}{2}\,\frac{a+b-2c}{a+b+c} \tag{2.8}$$

The limits of non-linear distortion depend on the application of the amplifier. In test amplifiers the factor γ usually does not exceed 1 or 2 per cent. In ordinary broadcast amplifiers and public address systems the factor γ does not exceed 7 or 8 per cent.

Frequency distortion is present in an amplifier when it does not amplify signal voltages of all frequencies within the specified frequency range by the same amount. Such non-uniform amplification is caused by the presence of frequency-dependent reactive components in the circuit.

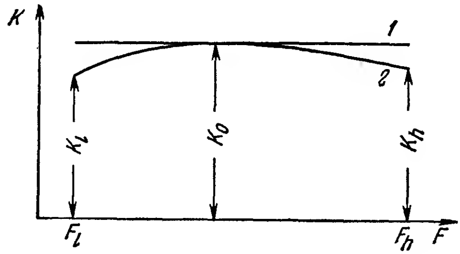


Fig. 2.9. Frequency response curves

1 — ideal amplifier; 2 — practical amplifier

The dependence of K on frequency is usually represented by a frequency characteristic, also called the *frequency response curve.*

Figure 2.9 shows the frequency response of an ideal amplifier *1*. Its response is said to be "flat" over the whole frequency range. The frequency response of a practical amplifier *2* shows that its gain falls off at low and high frequencies.

The values of K are laid off as ordinate, and the frequency as abscissa. F_l and F_h are the limiting or cut-off (low and high) frequencies of the band. K_0 is the *midband* (or *mid-frequency*) *gain*, as it is called. K_l is the *low-frequency gain*, and K_h is the *high-frequency gain* of the amplifier.

Sometimes, frequency-response curves are plotted on a logarithmic scale, with K in decibels as ordinate and the logarithms of frequency as abscissa (Fig. 2.10).

Quantitatively, frequency distortion is expressed as the distortion factor M which is equal to the ratio of the midband gain to the gain at the limiting frequencies

$$M=\frac{K_0}{K} \tag{2.9}$$

Since the drop in amplification in the low-frequency range is different from that in the high-frequency range, frequency dis-

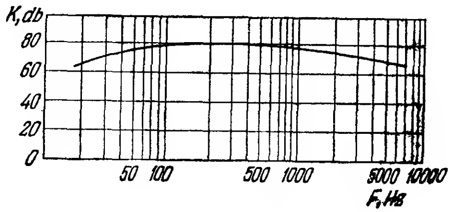


Fig. 2.10. Logarithmic plot of frequency response

tortion is evaluated separately for each:

$$M_l = \frac{K_0}{K_l} \tag{2.10}$$

$$M_h = \frac{K_0}{K_h} \tag{2.11}$$

The total frequency distortion of a multi-stage amplifier is equal to the product of the frequency distortions of the individual stages

$$M_t = M_1 M_2 M_3 \ldots \tag{2.12}$$

Frequency distortion just as gain may be expressed in decibels

$$M_{db} = 20 \log_{10} M$$

The limits of frequency distortion depend on the application of the amplifier.

The human ear does not practically notice frequency distortion if it is less than 25 to 30 per cent at the limiting frequencies of the audio range. Therefore, for audio amplifiers, frequency distortion may be $M_l = M_h = 1.25$ to 1.3, which corresponds to a frequency response flat to within ± 2 or 3 db.

Phase Distortion. When an electric circuit contains capacitive and inductive reactances, its voltages and currents will undergo a phase shift from input to output, and this phase shift will be different at different frequencies.

This is also true of all a. c. amplifiers. For one thing, they contain reactances. For another, their signal consists of the fundamental and higher harmonics.

Because of this lack of direct proportionality between phase shift and frequency, the waveform of the output signal will be distorted. This is phase distortion.

The curve showing phase shift as a function of frequency is called the *phase response characteristic* (Fig. 2.11).

No phase distortion occurs when the phase relations between the components of a complex signal at the output of the amplifier remain the same as at the input.

This is also true of cases where the phase shift is directly proportional to the frequency of higher harmonics (Fig. 2.11, curve *1*).

The departure of the phase-response characteristic from linearity is an indication that phase distortion is present. In audio amplifiers phase distortion is usually disregarded.

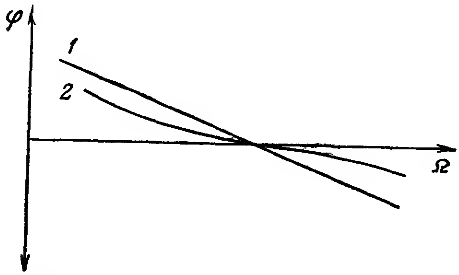


Fig. 2.11. Phase response characteristics
1—practical amplifier; 2—amplifier introducing no phase distortion

In television and radar amplifiers phase distortion may considerably impair the quality of the picture. Therefore the phase shift must not exceed 5° to 20° at the limiting frequencies of the amplification channel.

Review Questions

1. When can a radio receiver operate without an a. f. amplifier?
2. Name the essential properties of an amplifying element.
3. The input and output voltage of a device are the same. Can this device be an amplifier?
4. Name the factors affecting voltage gain.
5. Name the factors affecting power gain.
6. Is non-linear distortion dependent on the signal amplitude?

7. Principles of Amplifier Design

The circuit configuration of an amplifier is to a large extent decided by its intended application, which may be voltage or power amplification.

A *voltage amplifier* is one designed to amplify voltage waveforms in circuits where almost no power is taken from the load. A *power amplifier* is one designed to increase the power of a signal and to supply power to some circuit other than the grid of another valve, such as a loudspeaker, a transmission line, a relay, etc.

When the amplification provided by a single-stage amplifier is insufficient, resort is made to a multi-stage, or cascade, arrangement. In audio-frequency multi-stage or cascade amplifiers, the early stages amplify voltages, and the final (or output) stage

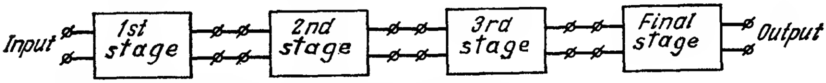


Fig. 2.12. Block-diagram of a multi-stage amplifier

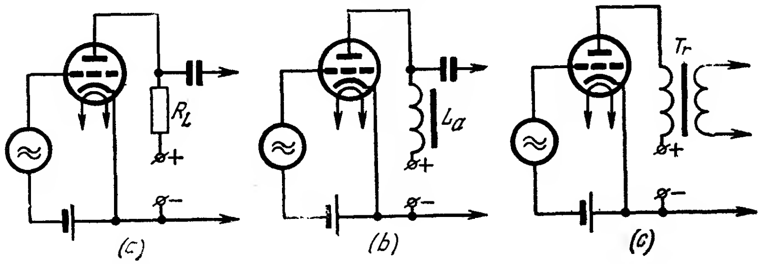


Fig. 2.13. Voltage amplifiers
(a) RC-coupled; (b) LC-coupled; (c) transformer-coupled

amplifies power. A generalized block-diagram of a multi-stage amplifier is shown in Fig. 2.12.

A further classification of amplifiers is according to interstage or load coupling, namely:

(a) resistance-capacitance-coupled amplifiers (Fig. 2.13*a*);
(b) choke-capacitance-coupled amplifiers (Fig. 2.13*b*);
(c) transformer-coupled amplifiers (Fig. 2.13*c*).

Still another basis for amplifier classification is according to output-circuit configuration, namely:

(a) earthed- (common-) cathode amplifiers (Fig. 2.14*a*);
(b) split-load amplifiers (Fig. 2.14*b*);
(c) earthed- (common-) anode amplifiers (cathode followers) (Fig. 2.14*c*).

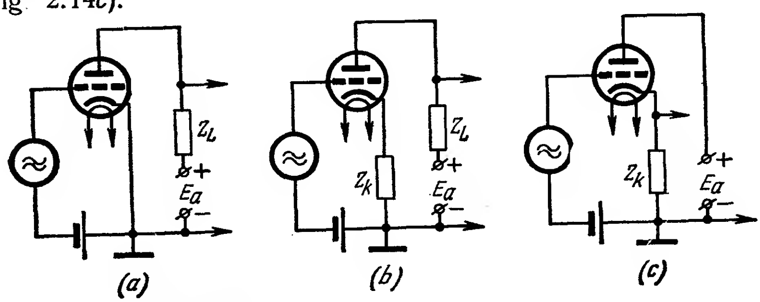


Fig. 2.14. Audio-amplifiers
(a) earthed-cathode; (b) split-load; (c) earthed-anode (cathode follower)

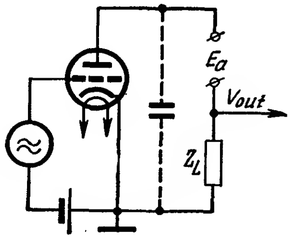


Fig. 2.15. Wrong connection of a load in an earthed-cathode amplifier

In an earthed (common)-cathode amplifier, the cathode is at earth potential at the signal frequency, the input is applied between the control grid and earth, and the output load is placed between the anode and earth.

In an earthed-anode amplifier (cathode follower), the input is applied between the control grid and earth, and the output is derived from the impedance placed between the cathode and earth which provides feedback.

Fig. 2.15 shows a wrong connection of the load in the anode circuit. The anode supply source has a marked capacitance to earth. This stray capacitance (shown by the dashed line in the circuit diagram) will shunt the load impedance and affect the physical properties of the amplifier.

The most commonly used circuit configurations for transistor amplifiers are:

(a) common-base amplifiers;
(b) common-emitter amplifiers;
(c) common-collector amplifiers;
(d) split-load amplifiers.

These types are shown in Fig. 2.16.

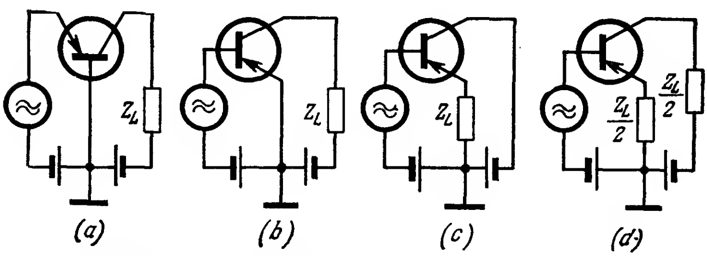


Fig. 2.16. Transistor amplifiers
(a) common-base; (b) common-emitter; (c) common-collector; (d) split-load

In a common-base amplifier, the load impedance is placed between the collector and the negative side of the collector supply source (if the transistor is a *P-N-P* type). The base and the positive side of the collector supply source are connected to earth.

In a common-emitter amplifier, the load impedance is disposed in exactly the same way as in a common-base amplifier, with the emitter, and not the base, connected to earth.

In a common-collector amplifier, the load impedance is connected between the emitter and earth. The collector supply source is placed between earth and the collector.

In a split-load amplifier, part of the load impedance is placed in the collector circuit, and part in the emitter circuit.

In all circuit configurations, the collector supply source is connected so that its stray capacitance will not shunt the load impedance.

Review Questions

1. Name the circuit configurations that can be realised in a two-stage valve amplifier.
2. Which of these circuit configurations may be of practical value in a two-stage valve amplifier?
3. Name the circuit configurations that can be realised in a two-stage transistor amplifier.
4. Which of these circuit configurations may be of practical value in a two-stage transistor amplifier?
5. May the anode circuit of the valve in an amplifier be fed with an alternating supply voltage?

8. Resistance-capacitance-coupled Amplifier

A resistance-capacitance (*RC*)-coupled amplifier is one in which the a. c. voltage coupled to the next stage develops across a resistor placed in the anode circuit.

A simplified circuit diagram of an *RC*-coupled amplifier is shown in Fig. 2.17.

The instantaneous grid voltage e_g is the sum of bias voltage E_g and signal voltage v_g

$$e_g = E_g + V_g \sin \Omega t$$

where V_g is the signal voltage amplitude.

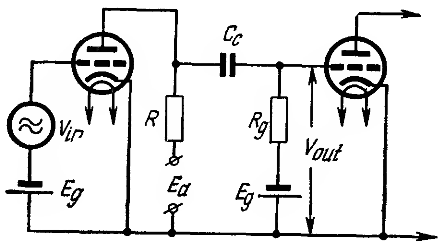


Fig. 2.17. RC-coupled amplifier

In no-signal operation, bias voltage E_g alone is applied to the grid, and a d. c. current I_{a0} flows in the anode circuit. This current, on passing through the anode load resistor R produces a constant voltage drop across this resistance

$$V_R = I_{a0}R$$

The value of d. c. voltage between the cathode and anode is less than the supply voltage E_a by the value of the voltage drop across the anode load resistor

$$V_{a0} = E_a - V_R$$

The signal voltage applied to the valve grid changes the anode current of the valve in such a way that the maximum value of anode current $I_{a\,max}$ corresponds to the maximum value of grid voltage

$$e_{g\,max} = E_g + V_g$$

and the lowest anode current $i_{a\,min}$ corresponds to the minimum grid voltage

$$e_{g\,min} = E_g - V_g$$

As the anode current increases, the voltage drop across the anode load resistor also increases. This brings about a decrease in the anode voltage. Therefore, when the grid voltage is a maximum, the anode voltage will be a minimum and vice versa. Thus the a. c. anode voltage is 180° out of phase with the signal (a. c. grid) voltage.

The a. c. signal voltage applied to the valve grid changes not only the anode current, but also the voltage drop across the anode load resistor and the anode voltage; as a result, the static characteristic no longer reflects the relationship between grid voltage and anode current, and resort is now made to the *dynamic anode-grid* (or *transfer*) characteristic.

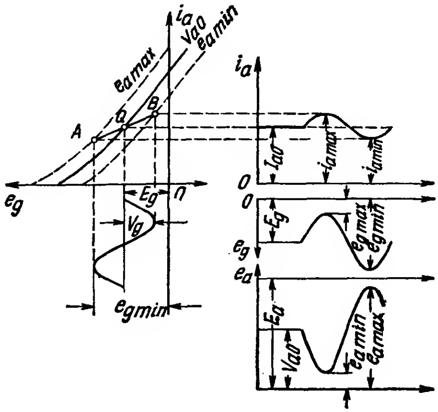


Fig. 2.18. Characteristics of an RC-coupled amplifier

The events discussed above are presented graphically in Fig. 2.18 where Q is the quiescent operating point of the valve fixing the quiescent anode voltage V_{a0}, and AB is called the *path of operation*, which is a segment of the dynamic transfer curve.

Quite frequently the *a. c. (dynamic) load line*, $i_a = \varphi(e_a)$, is used instead of dynamic transfer characteristics $i_a = f(e_g)$.

The a. c. load line (Fig. 2.19) offers a convenient method for determining the a. c. voltage amplitude across the anode load

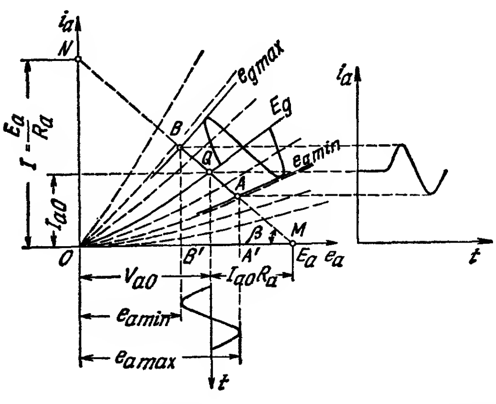


Fig. 2.19. Dynamic load line of an RC-coupled amplifier

resistor (see Fig. 2.17)

$$V_a = \frac{e_{a\,max} - e_{a\,min}}{2} = \frac{A'B'}{2}$$

and the quiescent anode voltage V_{a0}.

The a. c. voltage developed across the anode load resistor is applied to the valve grid of the next stage through a coupling capacitor C_c which "blocks off" the d.c. voltage on the anode of the first valve and prevents it from being applied to the grid of the second valve while allowing the amplified a. c. voltage to drive the next stage. The resistance of C_c to d. c. must be at least several hundred megohms.

R_g is the grid-leak resistor, or one providing a d. c. path to limit the accumulation of charge on, and applying bias voltage to, the grid. If R_g were taken out of the circuit, the grid potential would not be constant. The value of R_g must be from a few hundredths to a few thousandths of the insulation resistance between the grid and cathode. Usually the value of R_g does not exceed 0.5 to 2 megohms. A smaller grid-leak would shunt the anode load.

C_c and R_g form an a. c. voltage divider. To increase the voltage taken from R_g the reactance of C_c must be far less than R_g

$$\frac{1}{\Omega C_c} \ll R_g$$

The stage gain of the circuit shown in Fig. 2.17 is defined as

$$\dot{K} = \frac{\dot{V}_{out}}{\dot{V}_g} \tag{2.13}$$

To derive an expression for the stage gain we shall use the equivalent circuit of Fig. 2.20 where the shunt capacitance C_s, connected in parallel with R_g, is equal to the sum of output valve capacitance C_{out}, the input capacitance C_{in} of the next valve, and the distributed capacitance C_w of the wiring

$$C_s = C_{out} + C_{in} + C_w$$

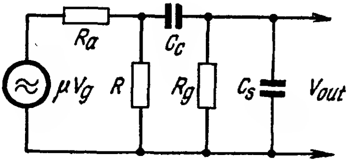


Fig. 2.20. Equivalent circuit of an RC-coupled amplifier

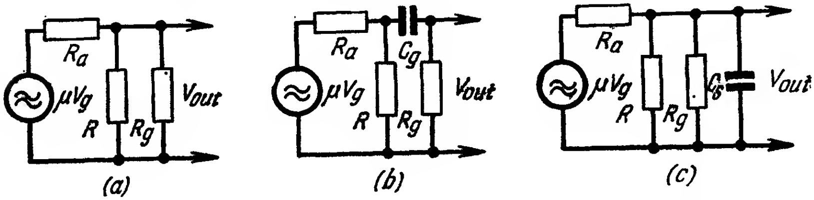


Fig. 2.21. Simplified equivalent circuits of an RC-coupled amplifier
(a) for midband frequencies; (b) for low frequencies; (c) for high frequencies

The value of C_s is from 50 to 200 picofarads.

At the midband frequency, the reactance of C_c is much smaller and that of C_s is much higher than the resistance of R_g

$$\frac{1}{\Omega C_c} \ll R_g, \quad \frac{1}{\Omega C_s} \gg R_g$$

so that both may be disregarded.

Therefore, we may ignore C_s in the low-frequency range, and C_c in the high-frequency range.

Simplified equivalent circuits of the stage are shown in Fig. 2.21.

Let us determine the stage gain K_0 in the midband range. Applying Thévenin's theorem to the circuit of Fig. 2.21*a* produces the voltage-generator equivalent circuit of Fig. 2.22 where

$$V = \mu V_g \frac{R}{R_a + R}$$

$$R_{eq} = \frac{R_a R}{R_a + R}$$

The output voltage of the amplifier

$$V_{out} = V \frac{R_g}{R_{eq} + R_g} \tag{2.14}$$

Substituting the expressions for V and R_{eq} into Eq. (2.14) gives

$$V_{out} = \mu V_g \frac{R}{R_a + R} \frac{R_g}{\frac{R_a R}{R_a + R} + R_g} = \mu V_g \frac{1}{1 + \frac{R_a}{R} + \frac{R_a}{R_g}}$$

$$K_0 = \frac{V_{out}}{V_{in}} = \frac{V_{out}}{V_g} = \mu \frac{1}{1 + \frac{R_a}{R} + \frac{R_a}{R_g}} \tag{2.15}$$

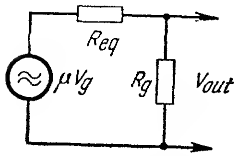


Fig. 2.22. Transformed equivalent circuit for the midband range

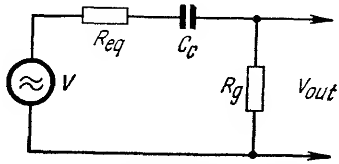


Fig. 2.23. Transformed equivalent circuit for the low-frequency range

Again using the voltage-generator equivalent circuit, (Fig. 2.23), the gain in the low frequency range is

$$\dot{V}_{out} = \dot{V}\frac{R_g}{R_{eg} + R_g + \frac{1}{j\Omega_l C_c}}$$

Substituting the expressions for V and R_{eq} gives

$$\dot{V}_{out} = \mu\dot{V}_g \frac{R}{R_a + R}\,\frac{R_g}{\frac{R_a R}{R_a + R} + R_g + \frac{1}{j\Omega_l C_c}} =$$

$$= \mu\dot{V}_g \frac{1}{\left(1 + \frac{R_a}{R} + \frac{R_a}{R_g}\right)\left[1 + \frac{1}{\left(R_g + \frac{R_a R}{R_a + R}\right) j\Omega_l C_c}\right]}$$

Since, however,

$$R_g \gg \frac{R_a R}{R_a + R}$$

it follows that

$$\dot{V}_{out} = \mu\dot{V}_g \frac{1}{\left(1 + \frac{R_a}{R} + \frac{R_a}{R_g}\right)\left(1 + \frac{1}{j\Omega_l C_c R_g}\right)}$$

and

$$\dot{K}_l = \frac{V_{out}}{V_g} = \dot{K}_0 \frac{1}{1 + \frac{1}{j\Omega_l C_c R_g}} \qquad (2.16)$$

The modulus of the low-frequency gain is

$$K_l = K_0 \frac{1}{\sqrt{1+\left(\frac{1}{\Omega_l C_c R_g}\right)^2}} \tag{2.17}$$

$$K_l = K_0 \frac{1}{\sqrt{1+\left(\frac{1}{\Omega_l \tau_l}\right)^2}} \tag{2.18}$$

where $\tau_l = C_c R_g$ is the time constant of the interstage coupling network.

Equations (2.16) and (2.17) can be used to determine both the frequency distortion M_l and the phase response of the stage

$$M_l = \frac{K_0}{K_l} \sqrt{1+\left(\frac{1}{\Omega_l C_c R_g}\right)^2} \tag{2.19}$$

$$M_l = \sqrt{1+\left(\frac{1}{\Omega_l \tau_l}\right)^2} \tag{2.20}$$

$$\tan\varphi_l = \frac{1}{\Omega_l C_c R_g} \tag{2.21}$$

$$\tan\varphi_l = \frac{1}{\Omega_l \tau_l} \tag{2.22}$$

The frequency distortion M_l and the phase shift φ are related thus

$$M_l = \frac{1}{\cos\varphi_l} \tag{2.23}$$

Now let us determine the stage gain in the high-frequency range

$$\dot{K}_h = \frac{\dot{V}_{out}}{\dot{V}_g} \tag{2.24}$$

Applying Thévenin's theorem to the circuit of Fig. 2.21c produces the voltage-generator equivalent circuit of Fig. 2.24

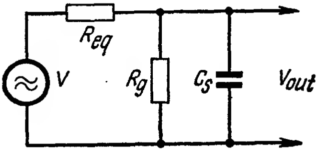


Fig. 2.24. Transformed equivalent circuit for the high-frequency range

in which

$$\dot{V}_{out} = \dot{V}\frac{\dot{Z}_{RC}}{R_{eq} + \dot{Z}_{RC}}.$$

where

$$\dot{Z}_{RC} = \frac{R_g \frac{1}{j\Omega_h C_s}}{R_g + \frac{1}{j\Omega_h C_s}} = \frac{R_g}{1 + j\Omega_h C_s R}$$

Substituting in Eq. (2.24) the expressions for V_{out}, and simplifying gives

$$\dot{K}_h = K_0 \frac{1}{1 + j\Omega_h C_s R_{eq}} \tag{2.25}$$

where

$$R_{eq} = \frac{R_a R}{R_a + R}$$

The modulus of the gain is

$$K_h = K_0 \frac{1}{\sqrt{1 + (\Omega_h C_s R_{eq})^2}} \tag{2.26}$$

Putting

$$C_s R_{eq} = \tau_h$$

we have

$$K_h = K_0 \frac{1}{\sqrt{1 + (\Omega_h \tau_h)^2}} \tag{2.27}$$

The modulus of the frequency distortion is:

$$M_h = \frac{K_0}{K_h} = \sqrt{1 + (\Omega_h C_s R_{eq})^2} \tag{2.28}$$

or

$$M_h = \sqrt{1 + (\Omega_h \tau_h)^2} \tag{2.29}$$

The phase-shift tangent is determined from Eq. (2.25)

$$\tan \varphi_h = -\Omega_h C_s R_{eq} \tag{2.30}$$

or

$$\tan \varphi_h = -\Omega_h \tau_h \tag{2.31}$$

The frequency distortion M_h and the phase shift φ may also be found from the same equation as used for the low-frequency range

$$M_h = \frac{1}{\cos \varphi_h} \tag{2.32}$$

From the foregoing we may conclude that the gain is a maximum in the midband range. As the frequency is decreased, the reactance of C_c increases, and the voltage drop across it also increases, bringing down the output voltage across R_g so that the gain K_l is considerably smaller than K_0.

As the frequency increases, the reactance of C_s decreases, and the shunting effect of C_s on the anode resistor is reduced. Lowering the anode resistor results in lowering K_h. Thus, the frequency-response characteristic of the amplifier in the low- and high-frequency ranges falls off.

As is seen from Eq. (2.21) and (2.30), the phase shift caused by C_c in the low-frequency range is positive ($\varphi_l > 0$), while the phase shift caused by C_s in the high-frequency range is negative ($\varphi_l < 0$). At the midband frequency Ω_0 the phase shift is equal to the algebraic sum of phase shifts φ_l and φ_h and is equal to zero

$$\frac{1}{\Omega_0 C_c R_g} + (-\Omega_0 C_s R_{eq}) = 0 \tag{2.33}$$

or

$$\frac{1}{\Omega_0 \tau_l} - \Omega_0 \tau_h = 0 \tag{2.34}$$

Solving Eq. (2.33) for the frequency, we obtain

$$\Omega_0 = \frac{1}{\sqrt{C_c R_g C_s R_{eq}}} \tag{2.35}$$

or

$$\Omega_0 = \frac{1}{\sqrt{\tau_l \tau_h}} \tag{2.36}$$

Amplifier analysis and experiments show that the midband range, that is, the region where the gain is very near to K_0, covers a considerable spectrum of frequencies (Fig. 2.25). This is also true of the phase response of the amplifier (Fig. 2.26).

Non-linear distortion in an RC-coupled amplifier, provided the proper operating conditions have been selected, is negligibly small. However, if an amplifier accepts large signals, it may be driven into operation within the lower curved region of the characteristic where noticeable non-linear distortion often occurs.

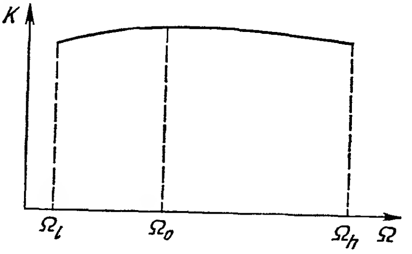


Fig. 2.25. Frequency response characteristic of an RC-coupled amplifier

As shown above, the gain is controlled by the anode resistor; the greater R, the less the difference between the stage gain and the amplification factor of the valve.

This relationship is also influenced by the a.c. anode resistance of the valve itself. Therefore, it is frequently more convenient to relate the gain to the ratio

$$\frac{R}{R_a}=\alpha$$

and not the anode load resistor. A plot of this relationship is shown in Fig. 2.27.

As is seen, R has a marked effect on K_0 only when the ratio $\frac{R}{R_a}=\alpha$ does not exceed three or four. Further increase in the anode load resistor has but a very slight effect on the gain.

There are some more reasons why the anode load resistance should not be made too large. Firstly, as R is made higher, the voltage drop across it increases, and the d.c. anode voltage is reduced. This increases the a.c. anode resistance of the valve and decreases its mutual conductance.

Secondly, an increase in the anode load resistance would cause an increase in the equivalent resistance R_{eq} and, consequently.

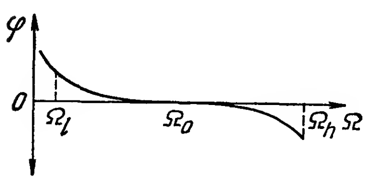


Fig. 2.26. Phase-response characteristic of an RC-coupled amplifier

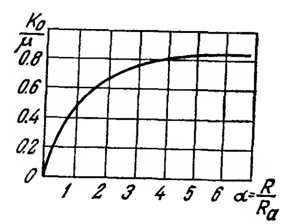


Fig. 2.27. Plot of the gain as a function of the anode resistance

in the frequency distortion M_h and the phase shift φ_h in the high-frequency range. Therefore, the value of R is usually chosen to be two to five times the a.c. anode resistance of the valve, so that $\alpha = 2$ to 5.

The value of the grid-leak resistor R_g in audio amplifiers is usually 5 to 10 times that of the anode resistance. At such values of R and R_g, the gain changes but slightly over a considerable frequency range.

Among the advantages of RC-coupled amplifiers are simplicity and low cost of components. This is the reason why such amplifiers are widely used as voltage amplifiers in different types of radio equipment.

Review Questions

1. Can the anode load resistor be smaller than the a.c. anode resistance of the valve?
2. How large should the ratio of R to R_a be to give a gain less than unity?
3. How large should C_c be to affect the frequency response in the high-frequency range?
4. Why is it that the instantaneous anode voltage of an RC-coupled amplifier is always lower than the anode supply voltage?
5. Can amplifiers use self-bias due to grid current?

9. Design of an *RC*-coupled Amplifier

Given:

1. The input and output voltages of the amplifier, i.e. the gain that it must provide.
2. The bandwidth, F_l-F_h.
3. The limits of low- and high-frequency distortion, M_l and M_h.

To find:

1. Valve type and its operating conditions.
2. Circuit parameters R, R_g and C_c.
3. Frequency distortion M and gain K at the cut-off frequencies.

In some cases, it is desired to calculate circuit parameters which will secure a maximum stage gain with the valve type selected.

Design Procedure:

1. The choice of the valve is governed by the required gain and frequency range.

Narrow-band amplifiers and stages with $K_0 < 100$ usually employ triodes with a high amplification factor (50 to 100). Broad-band amplifiers and amplifiers with $K_0 = 100$ to 200 employ pentodes.

For the purpose of valve selection it may be assumed that the gain of the triode stage is

$$K_0 \cong (0.7 \text{ to } 0.8)\,\mu$$

2. Having selected the type of valve, find the equivalent resistance from Eq. (2.28)

$$R_{eq} = \frac{\sqrt{M_h^2 - 1}}{\Omega_h C_s}$$

3. Once R_{eq} is found, determine the anode load resistor

$$R = \frac{R_{eq} R_a}{R_a - R_{eq}}$$

The a.c. anode resistance of a pentode valve is usually much higher than the equivalent resistance R_{eq}.

Therefore the anode load resistance is

$$R \cong R_{eq}$$

4. Now determine the value of the grid-leak resistor R_g and find the capacitance of C_c using Eq. (2.19)

$$R_g = (5 \text{ to } 10)\ R$$

$$C_c \geqslant \frac{1}{\Omega_l R_g \sqrt{M_h^2 - 1}}$$

5. Find the midband frequency Ω_0 and the midband gain K_0

$$\Omega_0 = \frac{1}{\sqrt{C_c R_g C_s R_{eq}}}$$

$$K_0 = \mu \frac{1}{1 + \frac{R_a}{R} + \frac{R_a}{R_g}}$$

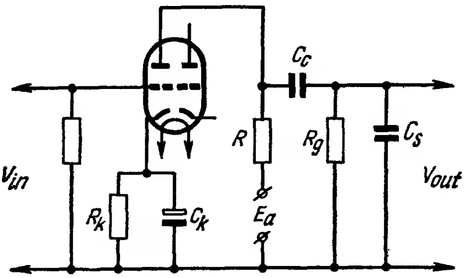


Fig. 2.28. An amplifier employing a bantam dual triode

6. Using the value of R, plot the a.c. load line of the valve and find I_{a0} and V_{a0} graphically.

7. Calculate the frequency distortion and gain at the cut-off frequencies

$$M_l = \frac{K_0}{K_l} = \sqrt{1 + \left(\frac{1}{\Omega_l C_c R_g}\right)^2}$$

$$M_h = \frac{K_0}{K_h} = \sqrt{1 + (\Omega_h C_s R_{eq})^2}$$

$$K_l = \frac{K_0}{M_l}$$

$$K_h = \frac{K_0}{M_h}$$

Example 2.1. Design an RC-coupled amplifier using one half of a bantam dual triode valve (Fig. 2.28) for $V_g = 0.5$ volt; frequency range, 50 to 6,000 hertz; limits of frequency distortion: $M_l = M_h = 1.05$; shunt capacitance $C_s = 200$ picofarads; anode supply voltage $E_a = 250$ volts.

The parameters of the valve are $g_m = 2$ mA/V, $\mu = 97.5$, $R_a = 49$ kilohms, $P_{a\,max} = 1$ watt, $V_L = 6.3$ volts, $I_L = 0.345$ ampere.

To find: the anode load resistor R, grid-leak resistor R_g, coupling capacitor C_c, the gain at the low, midband and high frequencies, self-bias resistor R_k and capacitance of the bypass capacitor C_k in the cathode circuit.

Solution:

1. The equivalent resistance is

$$R_{eq} = \frac{\sqrt{M_h^2 - 1}}{\Omega_h C_s} = \frac{\sqrt{1.05^2 - 1}}{6.28 \times 6 \times 10^3 \times 200 \times 10^{-12}} \simeq 42 \text{ kilohms}$$

2. The anode load resistance is

$$R=\frac{R_{eq}R_a}{R_a-R_{eq}}=\frac{42\times 49}{49-42}=294 \text{ kilohms}$$

or taking the nearest standard value, $R=300$ kilohms.

3. The grid-leak resistance is

$$R_g=(5 \text{ to } 10)\,R=5\times 300=1.5 \text{ megohms}$$

4. The capacitance of the coupling capacitor

$$C_c\geqslant\frac{1}{R_g\Omega_l\sqrt{M_l^2-1}}$$

$$=\frac{1}{1.5\times 10^6\times 6.28\times 50\sqrt{1.05^2-1}}=6{,}600 \text{ picofarads, or taking the}$$

nearest standard value, $C_c=6{,}800$ picofarads.

5. The midband gain is

$$K_0=\mu\frac{1}{1+\frac{R_a}{R}+\frac{R_a}{R_g}}=97.5\frac{1}{1+\frac{49}{300}+\frac{49}{1{,}500}}\cong 81.5$$

6. The gain at 50 and 6,000 hertz is

$$K_{50}=K_{6{,}000}=\frac{K_0}{M}=\frac{81.5}{1.05}=77.5$$

7. From the characteristics of Fig. 2.29 $E_g=-1.5$ volts.

8. Plot the a.c. (dynamic) load line. To do this, divide E_a into R to obtain a point on the current axis:

$$E_a\div R=250\div 300\times 10^3=0.83 \text{ mA}$$

which gives point N in the plot of Fig. 2.29. The other point is on the voltage axis; it is located by the specified anode supply voltage $E_a=250$ volts. The a.c. load line is thus drawn through points N and M.

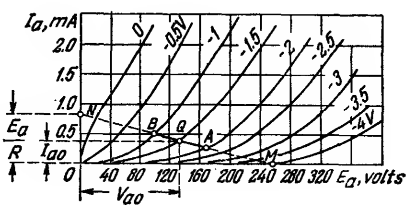


Fig. 2.29. Characteristics of the bantam dual triode

9. The Q-point (the intersection of the load line and the grid-bias curve) defines the quiescent anode current I_{a0} and the quiescent anode voltage V_{a0}

$$I_{a0}=0.4 \text{ mA}; \quad V_{a0}=132 \text{ volts}$$

10. The self-bias resistor is

$$R_k=\frac{|E_g|}{I_{a0}}=\frac{1.5}{0.4\times 10^3}\cong 3{,}800 \text{ ohms}$$

11. The capacitance of the bypass capacitor in the cathode circuit is

$$C_k=10 \text{ microfarads}$$

Review Questions

1. What effect has the a.c. anode resistance of a valve on the frequency response of the associated amplifier stage?
2. Name the methods for reducing distortion in the low-frequency range.
3. May a valve amplifier use an electrolytic as the coupling capacitor?
4. What should be the working voltage for the coupling capacitor in a valve amplifier?

10. Transformer-coupled Amplifiers

Transformer coupling is employed in voltage amplifiers and is the basic one for power amplifiers.

The circuit diagram of a transformer-coupled amplifier is shown in Fig. 2.30. The amplifier functions as follows.

The audio-frequency signal applied to the grid of the first valve produces an a.c. component in the anode circuit. This

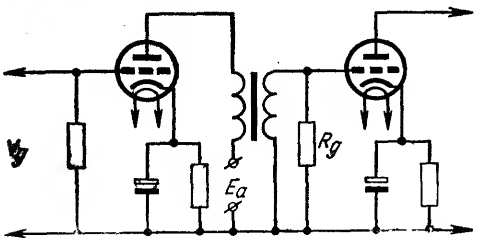


Fig. 2.30. Transformer-coupled amplifier

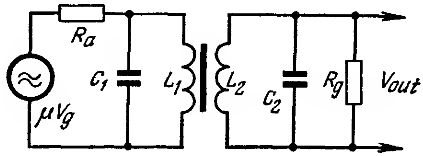


Fig. 2.31. Equivalent circuit of a transformer-coupled amplifier

component, passing through the primary winding of the transformer, sets up an alternating magnetic field around it. The magnetic field of the primary induces an alternating emf in the secondary.

If the number of turns in the secondary is greater than that of the primary, $N_2 > N_1$, the secondary voltage will be higher than the one across the primary terminals. Thus, a transformer-coupled amplifier both amplifies and steps up the voltage.

The a.c. resistance of the transformer primary is comparatively low, being usually from several hundred to several thousand ohms. Therefore the anode may be considered to be at full or nearly full potential of the anode supply source. This is why in a transformer-coupled amplifier the anode supply voltage may be considerably lower than in an RC-coupled amplifier employing a similar type of valve.

The primary and secondary windings are electrically isolated from each other so that no coupling (or blocking) capacitor C_c is needed in a transformer-coupled amplifier.

R_g across the secondary winding is provided to improve the stability of amplifier operation and frequency response.

The equivalent circuit of a transformer-coupled amplifier is shown in Fig. 2.31. However, it is not convenient for analysis and is replaced with another circuit, in which the components of the secondary circuit are referred to the primary side (Fig. 2.32). The following designations are used in Figs. 2.31 and 2.32:

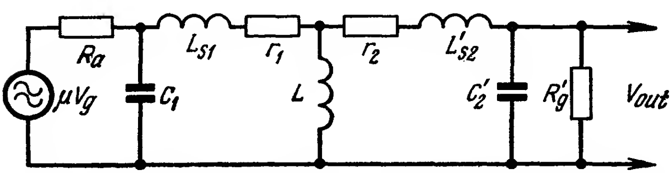


Fig. 2.32. Equivalent circuit of a transformer-coupled amplifier referred to the primary circuit

$V'_{out}=\frac{V_{out}}{n}$ = amplifier output voltage referred to the primary

$n=\frac{N_2}{N_1}$ = turns ratio

N_1 = number of turns in the primary

N_2 = number of turns in the secondary

r_1 = resistance of the primary

$r'_2=r_2/n^2$ = resistance of the secondary referred to the primary

L = total inductance

L_1 = primary inductance

L_2 = secondary inductance

L_{s1} = primary leakage inductance

$L'_{s2}=\frac{L_{s2}}{n^2}$ = secondary leakage inductance referred to the primary

C_1 = total capacitance of the primary circuit, consisting of the valve output capacitance C_{out}, capacitance C_{w1} of the wiring in the primary, and distributed capacitance C_{tr1} of the transformer primary

C_2 = total capacitance of the secondary circuit consisting of the input capacitance C_{in} of the valve in the next stage, the distributed capacitance C_{w2} of the secondary wiring, and distributed capacitance C_{tr2} of the transformer secondary

$C'_2=n^2c_2$ = total capacitance of the secondary circuit, referred to the primary circuit

$R'_g=\frac{R_g}{n^2}$ = shunting resistance referred to the primary.

The value of the leakage inductance L_s depends on the transformer design. The greater the distance between the primary and secondary windings, the higher is L_s.

To reduce the leakage inductance, the transformer windings are made in sections. In such windings, sections of the secondary are located between sections of the primary.

In modern transformers L_s is such that

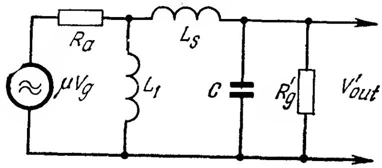
$$\sigma=\frac{L_s}{L_1}=0.007 \text{ to } 0.05$$

where σ is the leakage factor.

For the purpose of analysis, the circuit of Fig. 2.32 may be safely replaced by a simpler circuit of Fig. 2.33. The following considerations justify the replacement:

(1) L may be replaced by L_1 because usually $L_{s1} \ll L_1$ in a transformer.

Fig. 2.33. Simplified equivalent circuit of a transformer-coupled amplifier referred to the primary circuit



(2) Resistances r_1 and r_2' are usually much smaller than R_a; their effect on circuit operation may be allowed for by a slight increase in R_a to R_a'. At the high frequencies, the current branching off into L is small, and it may be considered that

$$R_a' = R_a + r_1 + r_2'$$

At the low frequencies

$$R_a' = R_a + r_1$$

(3) In exactly the same way, the effect of C_1 may be allowed for by adding a small capacitance C_2' to it

$$C = C_1 + C_2'$$

(4) L_{s1} and L_{s2}' may be considered to be connected in series and may be replaced by a single inductance L_s

$$L_s = L_{s1} + L_{s2}'$$

Let us see how the various elements in the circuit of Fig. 2.33 function at different frequencies. In the low-frequency range the reactance of L_s is so small as to be neglected. The low-frequency equivalent circuit is shown in Fig. 2.34*a*.

In the midband range, on the other hand, not a single circuit element of Fig. 2.33 may be neglected.

In the high-frequency range, L_1 may be disregarded because its reactance $\Omega_h L_1$ is very high. The key factors in this range are L_s and C. The high-frequency equivalent circuit is shown in Fig. 2.34*b*.

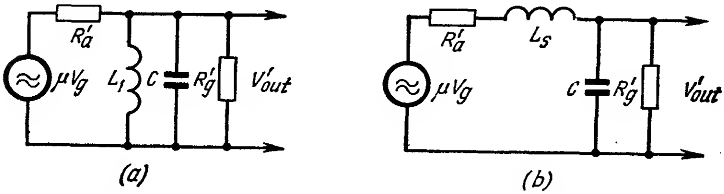


Fig. 2.34. Equivalent circuits of a transformer-coupled amplifier
(a) for low frequencies; (b) for high frequencies

In the circuit of Fig. 2.34a, L_1 and C form a parallel resonant circuit whose impedance is very high at a certain lower resonant frequency Ω_{0l}. At this frequency, the stage can be represented by a simple equivalent circuit (Fig. 2.35).

The resonant frequency is given by

$$\Omega_{0l}=\frac{1}{\sqrt{L_1C}}$$

The output voltage referred to the primary is

$$V'_{out}=\frac{V_{out}}{n}=\mu V_g\frac{R'_g}{R'_a+R'_g}$$

and the actual output voltage is

$$V_{out}=V_g\frac{n\mu R'_g}{R'_a+R'_g} \tag{2.37}$$

The stage gain K_0 at the resonant frequency Ω_{0l} is

$$K_{0l}=\frac{V_{out}}{V_g}=\frac{n\mu R'_g}{R'_a+R'_g}=\frac{n\mu}{1+\frac{R'_a}{R'_g}} \tag{2.38}$$

If $R_g=\infty$, then $R'_g=\infty$, and $K_{0l}=n\mu$.

The gain at lower frequencies other than the resonant frequency Ω_{0l} may be determined by applying Thévenin's theorem. This will give the circuit of Fig. 2.34a, which is then transformed into that of Fig. 2.36 where

$$R_{eq}=\frac{R'_aR'_g}{R'_a+R'_g}$$

Applying the usual stage analysis to the circuit of Fig. 2.36, we obtain the following expressions for the gain and frequency

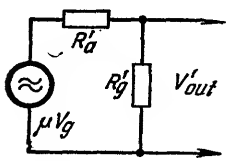


Fig. 2.35. Equivalent circuit of an amplifier at resonant frequency

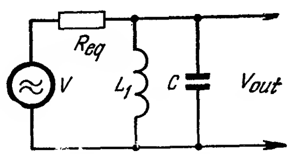


Fig. 2.36. Equivalent reduced circuit

distortion in the low-frequency range:

$$\left.\begin{aligned} K_l &= \frac{n\mu}{\sqrt{\left(1+\frac{R_a'}{R_g'}\right)^2+\left(\frac{R_a'}{\Omega_l L_1}\right)^2}} \\ &\text{and} \\ M_l &= \sqrt{1+\left(\frac{R_a' R_g'}{R_a'+R_g'}\right)^2\left(\frac{1}{\Omega_l L_1}\right)^2} \end{aligned}\right\} \tag{2.39}$$

Once the frequency distortion M_l is found, it is an easy matter to determine the necessary inductance L_1 of the primary

$$L_1 = \frac{R_a' R_g'}{R_a'+R_g'}\,\frac{1}{\Omega_l\sqrt{M_l^2-1}} = \frac{R_{eq}}{\Omega_l\sqrt{M_l^2-1}} \tag{2.40}$$

If R_g is very high or is not provided, the equation for L_1 becomes more convenient for computation

$$L_1 = \frac{R_a'}{\Omega_l\sqrt{M_l^2-1}} \tag{2.41}$$

In the high-frequency range the stage equivalent circuit (see Fig. 2.34*b*) is a series resonant circuit whose capacitance is shunted by R_g'.

At $\Omega = \Omega_{0h}$, or the upper resonant frequency, voltage resonance takes place in the resonant circuit. This causes a sharp increase in the voltage across the resonant circuit and in the output voltage. The frequency-response characteristic for this frequency range will have a sharp peak (Fig. 2.37).

Neglecting the effect of R_g, an approximate value of the resonant frequency may be found from the usual equation

$$\Omega_{0h} \cong \frac{1}{\sqrt{L_s C}} \tag{2.42}$$

At frequencies higher than the resonant frequency, the gain falls off sharply. The rapid fall-off of the gain at these frequen-

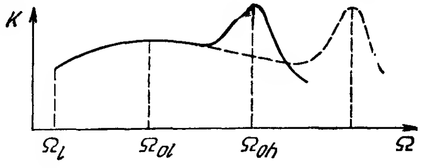


Fig. 2.37. Frequency-response characteristic of a transformer-coupled amplifier

cies is caused by the fact that the current in the circuit and the reactance of C are reduced simultaneously. Therefore, the bandwidth of a transformer-coupled amplifier in the high-frequency range is actually limited to the upper resonant frequency Ω_{0h}. In approximate form, the frequency-response characteristic is shown in Fig. 2.37. The sharp peak on the frequency-response characteristic may be flattened by decreasing the value of R_g. However, it should be kept in mind that reducing R_g will not only improve the frequency response, but will also lower the stage gain K. Because of this, R_g is selected to have a value of 0.1 to 1 megohm. When a reduction in gain is undesirable, the circuit parameters are selected in such a way that the resonant frequency Ω_{0h} will lie outside the operating range of frequencies (the dashed line in Fig. 2.37).

Review Questions

1. How will the shunting of the transformer primary with a resistance affect the gain?
2. How will a resistive shunt across the transformer primary affect the frequency response of the amplifier?
3. Can the instantaneous anode voltage be higher than the anode supply voltage?
4. When can the stage gain be greater than the amplification factor of the valve?
5. How will the frequency response of an amplifier be affected by the replacement of its triode with a pentode?

SUMMARY

1. An audio-frequency amplifier operates in the range from 16 to 20,000 hertz.

There are two basic types of audio-frequency amplifiers, voltage and power.

2. Amplification distorts the waveform of the signal. Distortion is caused by the non-linear and reactive components present in the amplifier.

3. The RC-coupled amplifier is most commonly used in voltage amplification. It is simple and inexpensive and has high performance.

4. The gain of an RC-coupled amplifier depends on the valve parameters and anode load resistor.

5. The gain of a transformer-coupled triode amplifier may be higher than the amplification factor of the valve. The frequency-response characteristic of such an amplifier sharply falls off in the low- and high-frequency ranges.

Problems

2.1. The input voltage of an amplifier is 1 millivolt, and its output voltage is 15 volts Determine the gain K and express it in decibels.

Answer: $K = 15{,}000$, $K_{db} = 83.5$ db.

2.2. Express the gain $K = 15$; 200; 450; 3,000; 15,000 in decibels.

Answer: $K_{db} = 23.56$; 46; 53; 69.5; 83.5 db.

2.3. Find the total gain of a three-stage amplifier and express it in decibels, if $K_1 = 50$, $K_2 = 50$ and $K_3 = 20$.

Answer: $K_t = 50{,}000$; $K_{db} = 94$ db.

2.4. What will be K_1 of the first stage of a two-stage amplifier, having a gain of $K_{db} = 66$ db, if the gain of the second stage is $K_2 = 40$?

Answer: $K_1 = 50$.

2.5. A four-stage amplifier has a gain of $K_{db} = 80$ db. The input voltage is 2 millivolts.

What will be the output voltage of the amplifier and the output voltage of each stage if the stages have the same gain?

Answer: $V_{out} = 20$; $K_t = 10{,}000$; $V_{out\,1} = 20$ millivolts; $V_{out\,2} = 200$ millivolts; $V_{out\,3} = 2$ volts.

2.6. The midband gain is $K_0 = 50$, while $K_l = 42$ and $K_h = 45$. Determine the frequency distortion M_l and M_h and express it in decibels.

Answer: $M_l = 1.19$; $M_{l\,db} = 1.5$ db; $M_h = 1.11$; $M_{h\,db} = 0.82$ db.

CHAPTER III
AUDIO-FREQUENCY POWER AMPLIFIERS

11. General

A *power amplifier* is one designed chiefly to supply the signal to a load at the required power level.

Power amplifiers are usually employed as output stages in radio receivers, rediffusion amplifiers, motion-picture sound amplifiers, test amplifiers, radar, aircraft radio equipment, etc.

The power output of such amplifiers varies from a few tenths of a watt to tens of kilowatts.

The levels of voltages and currents in power amplifiers considerably exceed those encountered in voltage amplifiers. Voltage amplifiers commonly use only a small part of the valve characteristic. In power amplifiers the general trend is to use the valve characteristic as fully as possible. This, of course, leads to a considerable increase in non-linear distortion.

It often happens that a power amplifier operates within the region of positive grid-voltage values. In this case the grid draws current and the grid circuit dissipates power. This means that not only the final amplifier but also the driver (penultimate) stage has to be a power amplifier. The driver may account for 3 to 5 per cent of the power output of the final stage.

The operating conditions of a power amplifier depend on the nature and magnitude of the load. In the general case, the load is an impedance. At the high frequencies, it will be reactive in its effect. At the midband frequencies, the load is mainly resistive. In the subsequent discussion, the load will be assumed to be purely resistive.

Whatever the effect of the load, it is possible to select the operating voltages (grid bias and signal-voltage amplitude) so that anode current may be caused to flow in the valve during the complete cycle of the input voltage or during any fraction of it.

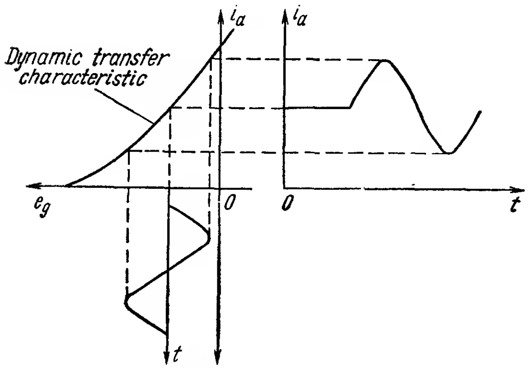


Fig. 3.1. Graphic representation of Class A_1 operation

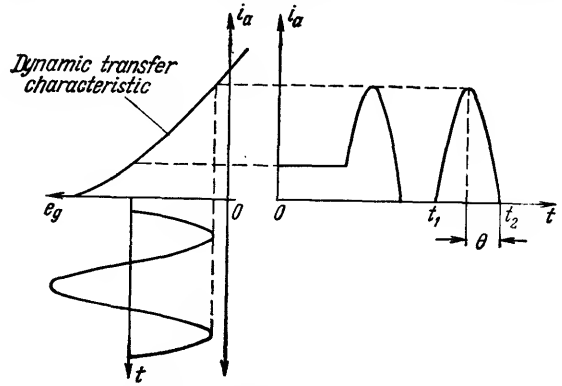


Fig. 3.2. Graphic representation of Class AB_1 operation

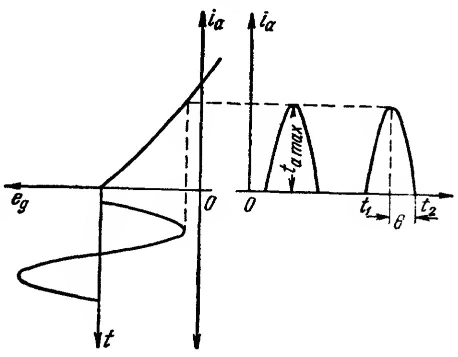


Fig. 3.3. Graphic representation of Class B_1 operation

The period, measured in electrical degrees, during which the valve conducts is called the *operating angle*. Very often it is measured from current maximum to cut-off, and is called the *cut-off angle*, defined as

$$\theta = \Omega\,(t_2 - t_1)/2$$

In both cases, the grid may or may not draw current or, which is the same, the grid may or may not be driven positive.

For convenience, all these forms of operation have been assigned letter symbols with numerical subscripts as follows.

(1) **Class A.** The anode current flows during the complete grid-voltage cycle. Fig. 3.1 shows Class A_1 operation, where the subscript "1" means that the grid does not go positive and draws no current. This happens because the valve is biased into the linear portion of its transfer characteristic. If the bias were not so large, the grid would go positive, grid current would flow, and the operation would be Class A_2.

(2) **Class B** (Figs. 3.3 and 3.5). The grid bias is set at the current cut-off point. Very often the grid-input voltage is necessarily quite large, and the grid is driven slightly positive so that grid current is flowing, producing Class B_2 operation (Fig. 3.5). Obviously, the valve partly operates in the non-linear portion of its characteristic. The operating angle is 180°, and the cut-off angle, 90°.

(3) **Class AB**. The grid bias is higher than for Class A operation and less than the cut-off bias required for Class B. The

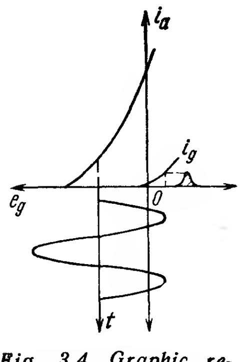


Fig. 3.4 Graphic representation of Class AB_2 operation

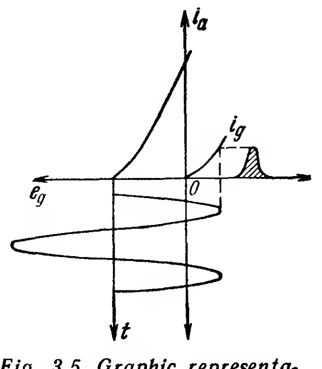


Fig. 3.5. Graphic representation of Class B_2 operation

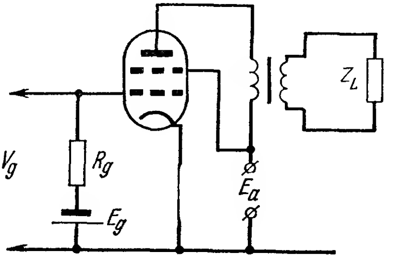


Fig. 3.6. Transformer-coupled power amplifier

cut-off angle is from 120° to 130°. If the grid does not go positive, the operation is Class AB_1 (Fig. 3.2). If it does the operation is Class AB_2 (Fig. 3.4).

It should be noted that low-power amplifiers are nearly always Class A_1 circuits.

Class B operation is confined mostly to push-pull (two-valve) amplifiers where it provides for high efficiency and high power output, which fact makes it especially suitable for high-level and battery-operated equipments. It might be used in single-ended (single-valve) amplifiers, but non-linear distortion would be prohibitive.

Power amplifiers may also be classed according to the method of coupling the anode circuit to the load. The most popular method is transformer coupling (Fig. 3.6).

Review Questions

1 Why is it that in Class A operation non-linear distortion is lower than in Class B operation?

2. Can readings of a milliammeter in the anode circuit give a clue as to the class of operation?

3. In which class of operation is the anode dissipation in the no-signal state greater?

12. Analysis of a Single-ended Triode Power Amplifier

As will be recalled, triode valves have fairly linear characteristics. This linearity makes the triode characteristics a convenient tool for analytical calculation of optimum operating conditions for a power amplifier, and simplifies stage analysis.

The key factor governing the operating conditions of a power amplifier stage is the load resistance. It affects the power output

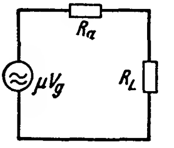


Fig. 3.7. Simplified equivalent circuit o, a power amplifier

of the stage, efficiency, non-linear distortion, and other characteristics.

Let us see how the load resistance affects the power output of an amplifier by reference to the midband equivalent circuit of Fig. 3.7.

The power P in the load R_L is given by

$$P=\frac{1}{2} I_1^2 R_L \tag{3.1}$$

where I_1 is the amplitude of the first-harmonic anode current.

As follows from the equivalent circuit, the amplitude of the first-harmonic anode current I_1 is given by

$$I_1=\frac{\mu V_g}{R_a+R_L} \tag{3.2}$$

Substituting Eq. (3.2) in Eq. (3.1) gives

$$P=\frac{1}{2}\left(\frac{\mu V_g}{R_a+R_L}\right)^2 R_L=\frac{1}{2}\,\mu V_g^2 \frac{R_L}{(R_a+R_L)^2}$$

Placing R_a outside the brackets in the denominator and putting $\frac{R_L}{R_a}=\alpha$, we have

$$P=\frac{1}{2}\,\frac{\mu^2 V_g^2}{R_a}\,\frac{\alpha}{(1+\alpha)^2} \tag{3.3}$$

Now we determine the value of α for maximum power transfer to the load.

Assume that the amplitude of the grid voltage is constant and independent of α.

Writing out the derivative $\frac{dP}{d\alpha}$ and equating it to zero gives

$$\frac{dP}{d\alpha}=P'=\left[\frac{1}{2}\frac{\mu^2V_g^2}{R_a}\frac{\alpha}{(1+\alpha)^2}\right]'=0$$

$$\frac{1}{2}\frac{\mu^2V_g^2}{R_a}\left[\frac{\alpha}{(1+\alpha)^2}\right]'=0$$

$$\left[\frac{\alpha}{(1+\alpha)^2}\right]'=\frac{(1+\alpha)^2-2\alpha(1+\alpha)}{(1+\alpha)^4}=0$$

$$1-\alpha^2=0$$

$$\alpha=\pm 1$$

The value $\alpha=-1$ is meaningless. Therefore, we drop the "—" sign and conclude that maximum power transfer will occur when $\alpha=1$, that is, when $R_L=R_a$. This fully checks with the maximum power transfer theorem which states that in a circuit where a generator, with internal resistance R_G, feeds power to a load, with impedance R_L (no net reactance in either), maximum power will be supplied to the load when R_L is equal to R_G.

Substituting the optimum value of α into the power equation of the amplifier, we have

$$P_{max}=\frac{\mu^2V_g^2}{8R_a} \tag{3.4}$$

In determining the optimum value of α, we assumed that the grid voltage amplitude is constant and the voltage is independent of α. Such an assumption is, of course, arbitrary.

The load resistance also governs the maximum signal voltage that may be applied to the grid. Referring to the dynamic transfer characteristic which represents the "with-signal" operation of the valve, we note that the linear portion of the curve grows longer as the load resistance is increased. The longer linear portion of the curve means, however, a higher maximum input (grid) voltage.

The last point is particularly important in Class A_1 operation, because the operating region of the curve is then fully within the area of negative grid-bias values.

However, there is a set-back to the advantage. As the load resistance is increased, the amplitude of the first-harmonic current decreases. With a limited supply voltage E_a, this may reduce the power output of the amplifier. This suggests that in cases where the amplitude of grid voltage is likely to vary, the conditions for maximum power transfer to the anode load are different.

Detailed analysis would show that for maximum power transfer the anode load resistance should be

$$R_L = 2R_a$$

or

$$\alpha = R_L/R_a = 2$$

The efficiency of an amplifier operating under optimum conditions is

$$\eta = P_s/P_0 = \frac{V_a I_{a1}/2}{E_a I_0} \qquad (3.5)$$

where P_s = signal power applied to the input (grid)
P_0 = power furnished by the supply source
V_a = output (anode) voltage
I_a = output (anode) current
E_a = anode supply voltage
I_0 = direct component of anode current.

In terms of the anode supply voltage and the anode efficiency, the amplitude of the a.f. voltage is given by

$$V_a = \xi E_a$$

where ξ is the anode efficiency whose value for a.f. amplifiers is 0.3 to 0.5.

In terms of the alternating anode current and the minimum value $I_{a\,min}$ the direct component of the anode current is defined as

$$I_0 = I_{a1} + I_{a\,min} = I_{a1}(1+\beta)$$

where

$$\beta = I_{a\,min}/I_{a1} = \text{(approx.) } 0.15 \text{ to } 0.2$$

Substituting the expressions for V_a and I_0 in Eq. (3.5) gives

$$\eta = P_s/P_0 = \frac{\xi E_a I_{a1}/2}{E_a I_{a1}(1+\beta)} = 0.5 \div 2 \div (1+0.2) = \text{(approx.) } 0.2$$

Under no-signal (quiescent) conditions, all power furnished by the anode supply source is dissipated by the anode as heat

$$P_a = P_0 = P_s/\eta = 5P_s \qquad (3.6)$$

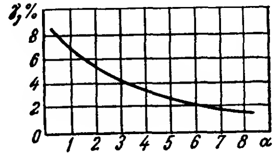


Fig. 3.8. Plot of non-linear distortion as a function of α in a triode

This relation between power output and anode dissipation is ordinarily used when selecting valves.

In conclusion, let us see how the load resistance affects non-linear distortion.

It has been pointed out that increasing the load resistance extends the linear part of the dynamic transfer characteristic of the valve which, in turn, reduces non-linear distortion.

An approximate plot of non-linear distortion as a function of α appears in Fig. 3.8.

Maximum power transfer is obtained with $\alpha=2$. Yet, in modern triode amplifiers, α is selected to be 3 or 4, although this entails a decrease of 10 to 12 per cent in power output. This reduction in power output is more than offset by reduction in non-linear distortion.

However, it should be kept in mind that an increase in the load resistance should be accompanied by an increase in the signal voltage, negative grid bias, and anode supply voltage. Therefore, when E_a is limited, preference has to be given to $\alpha=2$.

The load impedance Z_L of the amplifier may differ considerably from the optimum value of the anode load resistance R_L. To secure an impedance match between Z_L and the anode circuit, an output matching transformer is provided at the amplifier output. The load impedance, connected in the secondary winding may be reflected into the primary winding in such a way that

$$Z'_L=\frac{Z_L}{n^2}=R_{L\,opt}$$

This is why transformer coupling (Fig. 3.6) is the most commonly used arrangement in power amplifiers.

Review Questions

1. Can a dynamic speaker be directly connected into the anode circuit of a valve?

2. Why is the efficiency of a Class A_1 amplifier always less than 50 per cent?
3. Name the methods for reducing non-linear distortion.

13. Frequency Response of the Final Stage

Consider the final stage of Fig. 3.6. Its equivalent circuit referred to the primary circuit is exactly the same as shown in Fig. 2.32.

Simplified equivalent circuits of the stage for the various frequencies of the range are shown in Fig. 3.9, where

$$R'_a = R_a + r_1 + r'_2 = R_a + 2r_1 \tag{3.7}$$

$$Z'_L = \frac{Z_L}{n^2} \tag{3.8}$$

$$L_s = L_{s1} + \frac{L_{s2}}{n^2} \tag{3.9}$$

$$V'_{out} = \frac{V_{out}}{n} \tag{3.10}$$

$$n = \frac{N_2}{N_1} \tag{3.11}$$

To determine frequency distortion, let us find the stage gain in the various frequency ranges.

The midband gain is

$$K_0 = \frac{V_{out}}{V_g} = \frac{nV'_{out}}{V_g}$$

where V'_{out} (see Fig. 3.9*b*) is

$$V'_{out} = \mu V_g \frac{Z'_L}{R'_a + Z'_L}$$

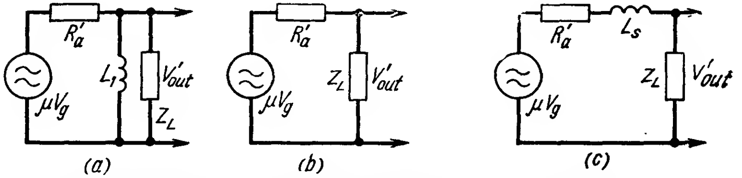


Fig. 3.9. Simplified equivalent circuits of the output stage
(a) for low frequencies; (b) for midband frequencies; (c) for high frequencies

Substituting the expression for V'_{out} in the equation for the gain gives

$$K_0 = n\mu \frac{Z'_L}{R'_a + Z'_L} \tag{3.12}$$

Next, we apply Thévenin's theorem to the equivalent circuits of Fig. 3.9*a* and *b*, and find the stage gain at the low and high frequencies of the range. The modulus of the gain at the low frequencies is given by

$$K_l = K_0 \frac{1}{\sqrt{1+\left(\frac{R_{eq}}{\Omega_l L_1}\right)^2}} \tag{3.13}$$

The modulus of the frequency distortion in the low-frequency range is

$$M_l = \sqrt{1+\left(\frac{R_{eq}}{\Omega_l L_1}\right)^2} \tag{3.14}$$

where

$$R_{eq} = \frac{R'_a Z'_L}{R'_a + Z'_L} \tag{3.15}$$

The modulus of the gain in the high-frequency range is

$$K_h = K_0 \frac{1}{\sqrt{1+\left(\frac{\Omega_h L_s}{R'_a + Z'_L}\right)^2}} \tag{3.16}$$

The modulus of the frequency distortion in the high-frequency range is

$$M_h = \sqrt{1+\left(\frac{\Omega_h L_s}{R'_a + Z'_L}\right)^2} \tag{3.17}$$

As seen from Equation (3.14), the frequency distortion at the low frequencies depends on L_1 and R_{eq}. Hence, if the frequency distortion is specified in advance, it is easy to find the necessary inductance of the transformer primary from the known load

$$L_1 = \frac{R_{eq}}{\Omega_l \sqrt{M_l^2 - 1}} \tag{3.18}$$

However, it should be borne in mind that a reduction in the load, although this brings down frequency distortion, may decrease the output voltage and power output of the stage.

The frequency distortion at the high frequencies depends on the leakage inductance L_s and the load impedance Z_L. If Z_L is specified in advance, the frequency distortion may be reduced only by lowering the leakage inductance.

In conclusion, we shall discuss the electrical properties of the stage output transformer.

The power P, due to the alternating component of the current, is dissipated as heat by the anode load resistance R_L which is the sum of the output load impedance referred to the primary circuit, Z'_L, and the resistances of the transformer windings:

$$R_L = Z'_L + r_1 + r'_2 = Z'_L + 2r_1 = \frac{Z_L}{n^2} + 2r_1 \qquad (3.19)$$

$$P = \frac{1}{2} I_1^2 R_L = \frac{1}{2} I_1^2 (Z'_L + 2r_1) \qquad (3.20)$$

where I_1 is the amplitude of the first-harmonic anode current.

The power dissipated in Z'_L is the power output of the stage

$$P_{out} = \frac{1}{2} I_1^2 Z'_L. \qquad (3.21)$$

The power P_t dissipated in the resistance of the windings is wasted as heat by the transformer

$$P_t = \frac{1}{2} I_1^2 (r_1 + r'_2) \qquad (3.22)$$

The power dissipation of the transformer is evaluated in terms of η_t, the efficiency of the transformer,

$$\eta_t = \frac{P_{out}}{P} = \frac{\frac{1}{2} I_1^2 Z'_L}{\frac{1}{2} I_1^2 (Z'_L + 2r_1)} = \frac{Z'_L}{Z'_L + 2r_1} \qquad (3.23)$$

Noting that

$$2r_1 + Z'_L = R_L$$

Equation (3.23) may be rewritten as follows:

$$\eta_t = \frac{R_L - 2r_1}{R_L} \qquad (3.24)$$

The efficiency of a transformer depends on its power rating. Table 3.1 gives approximate values of efficiency for transformers of different power ratings.

TABLE 3.1

Power output, P_{out}, watts	<5	5 to 100	> 100
Transformer efficiency, η_t	0.7 to 0.8	0.8 to 0.9	0.9 to 0.95

Equation (3.24) and Table 3.1 may be used for determining the permissible resistance r_1 of the transformer primary winding and the turns ratio n as follows.

Solving Equation (3.24) for r_1 gives

$$r_1 = \frac{R_L}{2}(1-\eta_t) \tag{3 25}$$

Substituting Eq. (3.25) in Equation (3.19) and solving it for n yields

$$R_L = \frac{Z_L}{n^2} + 2r_1 = \frac{Z_L}{n^2} + 2\,\frac{R_L}{2}(1-\eta_t)$$

Hence,

$$n = \sqrt{\frac{Z_L}{R_l \eta_t}} \tag{3.26}$$

If Z_L is lower than R_L the turns ratio n will be less than unity, i.e. the output transformer will be of the step-down type

$$n = \frac{N}{N_1} < 1; \quad N_2 < N_1$$

A step-down transformer can match a low load impedance to the anode circuit of a valve having a considerable a.c. anode resistance R_a.

Review Questions

1. Does the a.c. anode resistance of the valve affect the frequency response of the final stage?
2 How can high-frequency distortion be reduced?
3. How can the efficiency of a transformer be raised?
4. When can the output transformer be a step-up one?

14. Design of a Triode Power Amplifier

Using the stage analysis presented above, we can readily select the type of valve and operating conditions for an amplifier answering a particular specification.

Given:

1. Power output, P_{out}.
2. Output load impedance, Z_L.
3. Frequency range, F_l-F_h.
4. Limits of frequency distortion M_l and M_h at the limiting frequencies of the range.
5. Limits of non-linear distortion γ.

To Find:

1. Type of valve.
2. Anode load resistance, R_L.
3. Amplitude of the alternating component I_1 of the anode current.
4. Maximum (I_{max}) and minimum (I_{min}) anode current.
5. Grid bias voltage E_g and signal-voltage amplitude V_g.
6. Direct component $I_{a(dc)}$ of the anode current.
7. Anode supply voltage E_a and power P_0.
8. Anode dissipation, P_a.
9. Resistance of the transformer primary winding, r_1.
10. Cathode resistor R_k.
11. Inductance of the transformer primary winding, L_1.
12. Leakage inductance, L_s.
13. Turns ratio, n.

Design Procedure:

1. Referring to Table 3.1, find the transformer efficiency and determine the available power of the stage

$$P = \frac{P_{out}}{\eta_t}$$

2. Determine the expected anode dissipation

$$P_a \cong (5 \text{ to } 6)\, P$$

and select a valve from a valve manual such that

$$P_a \leqslant P_{a\,(safe)}$$

where $P_{a\,(safe)}$ is the safe maximum anode dissipation.

3. Find the anode load resistance for the expected value of α

$$R_L = \alpha R_a$$

4. Determine the amplitude of the alternating anode current

$$I_1 = \sqrt{\frac{2P}{R_L}}$$

5. Calculate the maximum and minimum values of the anode current

$$I_{max} = I_0 + I_1 = I_1(1+\beta) + I_1 = I_1(2+\beta)$$
$$I_{min} = I_0 - I_1 = \beta I_1$$

where $\beta = 0.15$ to 0.2.

6. Plot the a.c. load line. For this purpose, lay off I_{max} on the current axis of the anode characteristic (Fig. 3.10) and draw a straight line parallel to the voltage axis until it cuts the curve $E_g = 0$. The intersection (A) is one point of the a.c. load line. Drop a perpendicular from point A to the voltage axis. The segment OA' represents the minimum anode voltage $e_{a\,min}$.

From point A' lay off the segment $A'B'$

$$A'B' = 2V_a = 2I_1R_L.$$

The segment OB' represents the maximum anode voltage $e_{a\,max}$.

From point B', erect a perpendicular until it cuts the line I_{min} at point B. Connecting points A and B will give the a.c. load line AB of the valve.

Find $E_{g\,min}$ from the valve characteristics.

7. Find the bias voltage and the signal voltage amplitude

$$|E_g| = V_g = \frac{|E_{g\,min}|}{2}$$

Laying off E_g on the load line, locate the operating point Q.

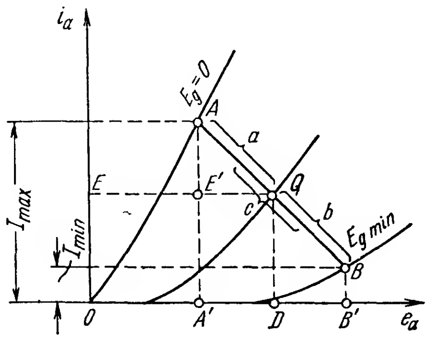


Fig. 3.10. Graphic calculation of a triode power amplifier

8. Determine the direct component of the anode current (this component is represented by the intercept OE on the current axis) and find the direct anode voltage $E_{a\,(dc)}$ (represented by OD).

9. Find the anode dissipation for the quiescent (no-signal) state

$$P_a = E_{a\,(dc)} I_{a\,(dc)} \leqslant P_{a\,(safe)}$$

which is represented by the area of rectangle $OEQD$. The power output is represented by the area of triangle $AE'Q$.

10. Determine non-linear distortion γ:

$$\gamma_2 = \frac{1}{2}\frac{a-b}{a+b}$$

$$\gamma_3 = \frac{1}{2}\frac{(a+b)-2c}{a+b+c}$$

$$\gamma = \sqrt{\gamma_2^2+\gamma_3^2} < \gamma_{lim}$$

If the non-linear distortion γ turns out to be greater than the specified limit, select a higher value of β or increase the anode load resistance R_L and carry out the calculation anew.

11. Find the resistance of the transformer primary

$$r_1 = \alpha R_a \frac{1-\eta_t}{2}$$

12. Find the anode supply voltage

$$E_a = E_{a\,(dc)} + I_{a\,(dc)} r_1 + |E_g|$$

If the bias is supplied from an external source,

$$E_a = E_{a\,(dc)} + I_{a\,(dc)} r_1$$

13. Calculate the bias resistor R_k and the power it will dissipate; the resistor type is selected according to this power.

14. Find the inductance of the transformer primary

$$L_1 = \frac{R_{eq}}{\Omega_l \sqrt{M_l^2 - 1}}$$

where

$$R_{eq} = \frac{R_L (R_a + 2r_1)}{R_L + R_a + 2r_1}$$

15 Find the maximum leakage inductance

$$L_s = \frac{R_a + Z_L'}{\Omega_h} \sqrt{M_h^2 - 1}$$

16. Calculate the turns ratio

$$n = \sqrt{\frac{Z_L}{R_L \eta_t}}$$

15. Single-ended Pentode Power Amplifier

Low-power amplifiers and the output stages of radio receivers frequently employ audio-frequency pentodes and beam-power tetrodes having the properties of pentode valves.

Pentodes and beam-power tetrodes are more sensitive than triodes. In pentode amplifiers the a.c. grid voltage necessary to obtain the required power in the anode circuit is one-half to one-third of what it is in amplifiers employing triodes. Therefore, pentode amplifiers may have fewer stages.

The high sensitivity of pentodes is attributed to the higher mutual conductance g_m and the higher amplification factor μ, as compared with triodes.

The provision of a screen grid held at constant potential considerably improves the anode voltage efficiency $\xi = \frac{V_a}{E_a}$, which in pentodes is 0.65-0.75 as against 0.3 to 0.5 in triodes. The anode current efficiency β of pentodes is about the same as that of triodes. This is why the efficiency of a pentode stage is considerably higher than that of a stage employing a triode

$$\eta = \frac{P}{P_0} = \frac{\frac{1}{2} V_a I_1}{E_a I_{a\,(dc)}} = \frac{1}{2} \xi (1 - \beta)$$

In a pentode amplifier, $\xi = 0.65$ to 0.75; $\beta = 0.2$ to 0.3; and $\eta = 0.22$ to 0.28.

The overall efficiency of a pentode amplifier is about 0.26.

The higher efficiency of an amplifier reduces the power drawn from the anode supply source and anode dissipation in "no-signal" operation

$$P_{a\,max} = P_0 = \frac{P}{\eta}$$

Assuming $\eta = 0.25$, then

$$P_a = 4P \tag{3.27}$$

In triode amplifiers, at $\alpha = 2$ to 4, the anode dissipation is considerably greater:

$$P_a = 3P \frac{2 + \alpha}{\alpha} = (4.5 \text{ to } 6)\, P$$

Thus, a pentode power amplifier is more economical than a triode amplifier.

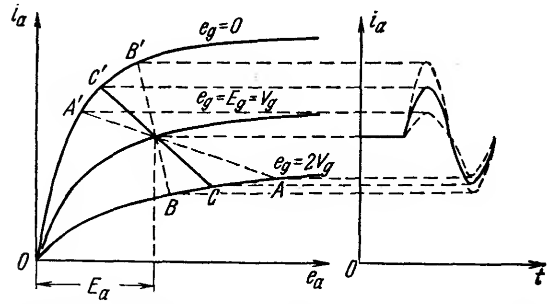


Fig. 3.11. Diagram for valve operation at different values of anode load resistance

Pentodes have a smaller transfer capacitance, which fact provides for better stability of operation. However, pentodes have certain disadvantages of their own. For one thing, their anode characteristics are extremely non-linear, and this leads to greater non-linear distortion.

In triode amplifiers, non-linear distortion usually decreases as the anode load resistance is increased. In a pentode, an increase in the anode load resistance is accompanied by a considerable increase in non-linear distortion. This is attributed to the fact that, as R_L is made higher, the slope of the dynamic load line is reduced, and the pentode amplifier operates within the non-linear portion of the characteristics. In Fig. 3.11, this condition is represented by the line AA'.

A low anode load resistance also considerably increases non-linear distortion because the slope of the dynamic load line increases too much (line BB'). With this type of load line, the positive half-cycle of the grid voltage causes considerably greater changes in the anode current than the negative half-cycle. As a result, the sinusoidal voltage applied to the grid brings about non-sinusoidal changes in the anode current.

For each type of pentode and beam-power tetrode, it is possible to select a value of anode load resistance, $R_L = R_{L\,opt}$, such that changes in the anode current will be symmetrical (line CC'). The optimum anode load resistance $R_{L\,opt}$ is given in the valve manual.

Another important disadvantage of pentode amplifiers is increased frequency and non-linear distortion at the high frequencies.

In triode amplifiers, the frequency response curve usually falls off sharply (representing a loss of gain) at the high frequencies due to the leakage inductance. In pentode amplifiers this does not happen because the a.c. (dynamic) anode resistance is hundreds of thousands of ohms

$$M_h = \sqrt{1 + \left(\frac{\Omega_h L_s}{R'_a + Z'_L}\right)^2} \cong 1$$

where

$$R_a \gg \Omega_h L_s$$

More specifically the frequency and non-linear distortion in a pentode amplifier is attributed to the reactive component of the load. As already noted, at the high frequencies the load is an impedance such that

$$\dot{Z} = R_Z + j\Omega_h L_Z$$

where $\Omega_h L_Z$ is the inductive component of the load impedance due to earphones, a loudspeaker, etc.

Figure 3.12 shows the high-frequency equivalent circuit of the final stage where R_L and L_L are such that

$$R_L = Z'_L + 2r_1 = \frac{Z_L}{n^2} + r_1 + r'_2$$

$$L_L = \frac{L_Z}{n^2}$$

As the signal frequency increases, the anode load impedance Z_L increases, too

$$\dot{Z}_L = R_L + j\Omega_h L_L$$

and the output voltage of the stage also increases. This is another way of saying that at the high frequencies, the frequency response curve goes up and not down.

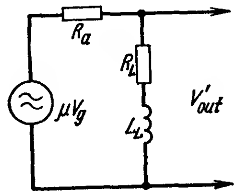


Fig. 3.12. High-frequency equivalent circuit of the output stage

In triode amplifiers, too, the anode load increases at the high frequencies, but we have disregarded this because (1) the leakage inductance has a greater effect, and (2) a triode amplifier may be regarded as a current source with a low internal resistance.

If the load impedance is much greater than the a.c. anode resistance of the valve (in our case, $\alpha = 2$ to 4), changes in the load impedance will affect the output voltage very slightly.

A pentode amplifier is a current source with a high internal resistance. Therefore, the anode current is chiefly determined by the a. c. anode resistance and much less by the anode load impedance

$$I_1 = \frac{\mu \dot{V}_g}{R_a + \dot{Z}_L} = \frac{\mu \dot{V}_g}{R_a} \frac{1}{1 + \frac{\dot{Z}_L}{R_a}} \cong \frac{\mu \dot{V}_g}{R_a}$$

since

$$\frac{\dot{Z}_L}{R_a} \ll 1$$

In other words, an increase in the anode load impedance at the high frequencies will not change the value of I_1.

It is readily seen that a higher output voltage will correspond to a larger anode load impedance

$$\dot{V}'_{out} = \dot{I}_1 \dot{Z}_L$$

Non-linear distortion at the high frequencies is also attributed to an increase in the equivalent impedance of the anode load. The modulus of the equivalent impedance becomes greater than the optimum value R_{opt} and non-linear distortion rises as explained above. As a counter-measure, pentode amplifiers incorporate a compensating RC network which keeps the load impedance constant (i.e. independent of the signal frequency) at the high frequencies.

Figure 3.13 shows an amplifier with a compensating network and also an equivalent circuit of the amplifier. According to alternating-current theory, if in this circuit

$$R = R_L = \sqrt{\frac{L_L}{C}}$$

the equivalent impedance will be constant at any frequency and numerically equal to R_L. This is called "perpetual resonance". However, when $R_c = R_L$, the resistor will dissipate a considerable part of the amplifier power output.

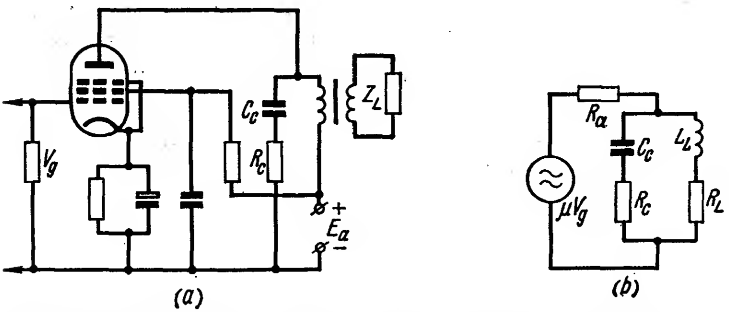


Fig. 3.13. Pentode power amplifier with a compensating network (a) and its equivalent circuit (b)

To decrease the losses in R_c, it is usually made slightly larger than R_L

$$R_c = (1.5 \text{ to } 2)\, R_L$$

The effect of a compensating network may be explained as follows.

At the low frequencies, the reactance of C_c is high, and the RC network does not shunt the load impedance. At the high frequencies, the reactance of C_c decreases, and R_c shunts the load, thereby compensating the increase in Z_a.

The components for a compensating network are given by

$$C_c = \frac{L_L}{R^2} \tag{3.28}$$

where

$$L_L = L_s + \frac{L_Z}{n^2}$$

and

$$R_c = (1.5 \text{ to } 2)\, R_L$$

Review Questions

1. Why do the final stages use beam tetrodes and pentodes?
2. Explain why a pentode amplifier has a higher efficiency.
3. Why is the leakage inductance of the output transformer greater in a pentode than in a triode amplifier?
4. Why is it that $R_L < R_a$ in a pentode amplifier?
5. Explain why the gain of the early stages is lower when the final stage uses a pentode or a beam tetrode.

16. Design of Class A Pentode and Beam Tetrode Power Amplifiers

An important point about the design of a pentode amplifier is that one has to allow for the non-linearity of the valve characteristic and the effect of operating voltages (grid bias and signal-voltage amplitude) on its parameters. Because of this, it is difficult to idealise the characteristics and calculate the stage analytically. Instead, a pentode amplifier is calculated graphically.

The graphic method, just as any graphic method, is not precise. Yet, the results are sufficiently accurate for most practical applications.

Given:

1. Power output, P_{out}.
2. Load resistance, Z_L.
3. Load inductance, L_Z.
4. Frequency range, F_l-F_h.
5. Frequency distortion M_l and M_h at the limiting frequencies of the range.
6. Non-linear distortion γ.

To Find:

1. Type of valve and anode supply voltage, E_{a0}.
2. Signal voltage, V_g.
3. Grid bias voltage, E_g.
4. The direct component of anode current, $I_{a(dc)}$.
5. Anode dissipation, P_a.
6. Resistance r_1 of the transformer primary.
7. Self-bias (cathode) resistor, R_k.
8. Inductance of the transformer primary L_1.
9. Turns ratio, n.
10. Parameters of the compensating network, R_c and C_c.

Design Procedure:

1. Find the power output of the valve

$$P = \frac{P_{out}}{\eta_t}$$

The value of η_t is selected from Table 3.1.

2. Find the anode dissipation

$$P_a = 4P$$

3. Referring to a valve manual, choose the type of valve satisfying the condition

$$P_{a\,max} \geqslant P_a$$

4. Find i_{max} and calculate the approximate value of the d.c. anode voltage.

For a pentode in Class A operation, the total instantaneous anode current is a maximum, i_{max}, when the instantaneous grid voltage, e_g, is zero. This current may be easily found from the pentode characteristics of Fig. 3.14.

In terms of peak anode current and peak anode voltage, the power output of the amplifier is

$$P = \frac{1}{2} V_a I_1 \tag{3.29}$$

As seen from Fig. 3.14,

$$i_{max} = I_{a\,(dc)} + I_1$$

where

$$I_{a(dc)} = \frac{I_1}{1-\beta}$$

Substituting $I_{a(dc)}$ and solving the equation for I_1 gives

$$I_1 = \frac{1-\beta}{2-\beta} i_{max}$$

In pentodes, $\beta = (0.2 \text{ to } 0.3)$, and so

$$I_1 \cong 0.4 i_{max}$$

As is known,

$$V_a = \xi E_a$$

where $\xi = 0.65$ to 0.75.

Substituting the values of I_1 and V_a into Eq. (3.29) gives

$$P = \frac{1}{2} V_a I_1 = \frac{1}{2} 0.65 E_a \times 0.4 i_{max}$$

Finally, in terms of i_{max} and E_a,

$$P \cong \frac{E_a i_{max}}{8} \tag{3.30}$$

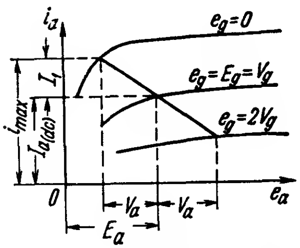


Fig. 3.14. Graphical method for determining the maximum value of total instantaneous anode current

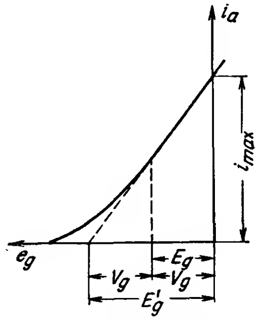


Fig. 3.15. Graphical method for determining grid voltage

The values of $E_{a\,max}$, $P_{a\,max}$ and i_{max} and the approximate values of anode load resistances R_L for the most commonly used Soviet pentodes and tetrodes are given in Table 3.2.

TABLE 3.2

Valve type	Valve	$E_{a\,max}$, volts	$P_{a\,max}$, watts	i_{max}, mA	R_L, ohms
2П2M	Pentode	200	2	24	20,000
2П1П	Beam tetrode	100	1.5	20	10,000
6П3C	Beam tetrode	400	20	150-180	2,500
6П6C	Beam tetrode	300	13	100	6,500
6П9	Pentode	300	9	60	10,000
6П1П	Beam tetrode	300	12	100	5,500
30П1C	Beam tetrode	120	6	100	1,800
6П13C	Beam tetrode	200	14	150	5,000
6П14П	Pentode	250	12	130	5,200
6П15П	Pentode	300	12	60	8,000
6П18П	Pentode	300	12	160	2,500

Once the total instantaneous anode current is known, it is possible to determine the required total anode voltage

$$E_a = \frac{8P}{i_{max}}$$

5. Find the bias voltage E_g and the signal-voltage amplitude V_g graphically, employing an idealised transfer characteristic $i_a = f(e_g)$ of the valve (Fig. 3.15).

Place the Q-point in the middle of the idealised transfer characteristic and determine E'_g. Then

$$E_g = \frac{E'_g}{2}$$

Assume that the signal-voltage amplitude is equal to the bias voltage or is slightly below it

$$V_g \leqslant |E_g|$$

The idealised transfer characteristic coincides with the linear part of the static transfer characteristic. Therefore, E'_g may be defined as

$$|E'_g| = \frac{i_{max}}{g_m}$$

Hence,

$$|E_g| = V_g = \frac{i_{max}}{2g_m} \tag{3.31}$$

6. Determine $I_{a(dc)}$ from the characteristic of Fig. 3.14 and check to see that

$$P_a = E_a I_{a(dc)} < P_{a\,max}$$

7. Plot the a.c. (dynamic) load line of the valve and determine the anode load resistance R_L. If the calculated operating conditions are close to those recommended in the valve certificate, the anode load resistance R_L may be looked up in the manual. If the calculated operating voltages considerably differ from the recommended ones (which is usually the case when the valve is underrated), the optimum anode load resistance R_L should be determined graphically. For this purpose, place a calibrated straight-edge on the family of anode characteristics and rotate it about the selected operating point (Fig. 3.16) until the segments a and b are equal. This indicates that anode-current

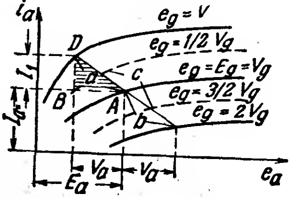


Fig. 3.16. Diagram for graphical determination of the operating point

changes are symmetrical, i. e. the second-harmonic anode current is at its lowest. Marking this position of the straight-edge by points on the characteristic, find graphically the amplitude of the first-harmonic current, the amplitude of V_a, an approximate value of the power output, and the anode load resistance

$$P=\frac{1}{2}I_1V_a$$

$$R_L=\frac{V_a}{I_1}$$

The power output is approximately represented by the area of triangle ABD

$$P=\frac{1}{2}I_1V_a=\frac{AB\times BD}{2}$$

If the obtained power output is less than is required, slightly increase the anode supply voltage E_a, and, sometimes, the grid voltage (V_g and E_g), and determine the new values of currents, voltages and power out ut.

8. Find the non-linear distortion

$$\gamma=\gamma_3=\frac{1}{2}\,\frac{2c-(a+b)}{a+b+c}$$

If the obtained value of γ exceeds the specified one, slightly reduce the a.c. grid voltage amplitude.

9. Calculate the anode load resistance

$$R_L=\frac{V_a}{I_1}$$

10. Find the resistance of the transformer primary

$$r_1=\frac{R_L}{2}(1-\eta_t)$$

11. Determine the value of the self-bias resistor

$$R_k=\frac{|E_g|}{I_{a(\mathrm{dc})}+I_{g2}}$$

The screen-grid current I_{g2} should be looked up in the manual or determined from the valve characteristic.

12. Find the anode supply voltage

$$E_{a\,sup}=E_a+I_{a(\mathrm{dc})}r_1+|E_g|$$

13. Determine the equivalent resistance

$$R_{eq}=\frac{R_L(R_a+2r_1)}{R_L+R_a+2r_1}\cong R_L$$

because $R_a \gg R_L$.

14. Calculate the inductance of the transformer primary

$$L_1=\frac{R_{eq}}{\Omega_l\sqrt{M_l^2-1}}$$

15. Find the leakage inductance

$$L_s=\frac{R_a+R_L}{\Omega_h\sqrt{M_h^2-1}}$$

16. Determine the turns ratio

$$n=\sqrt{\frac{Z_L}{R_L\eta_t}}$$

17. Calculate the parameters of the compensating network

$$R_c=(1 \text{ to } 2)R_L$$

$$C_c=\frac{L_s+\frac{L_Z}{n^2}}{R^2}$$

Example 3.1. Design a power amplifier to meet the following specifications: power output $P_{out}=2.25$ watts; amplifier load $Z_L=5$ ohms; load inductance $L_Z=0.003$ henry; frequency range $F_l=80$ Hz, $F_h=6{,}000$ Hz; limits of frequency distortion $M_l=M_h=1.25$; non-linear distortion $\gamma\leqslant 8\%$.

Solution: 1. Assume that $\eta_t=0.75$ and find the power to be supplied by the valve:

$$P=\frac{P_{out}}{\eta_t}=\frac{2.25}{0.75}=3 \text{ watts}$$

2. Determine anode dissipation in the no-signal state

$$P_a=4P=4\times 3=12 \text{ watts}$$

3. Select a type 6П14П pentode. Pentode specifications: $V_f=6.3$ volts; $I_f=0.76$ ampere; $I_{g2}=5$ milliamperes; $E_{g2}=250$ volts; $g_m=11.3$ milliamperes/volt, $\mu=226$; $R_a=20$ kilohms; $P_{a\,max}=12$ watts.

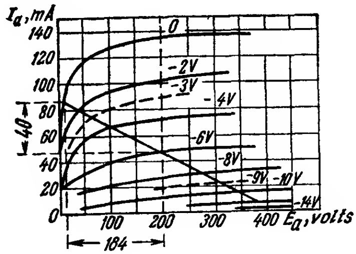


Fig. 3.17. An example of graphical computation of an amplifier employing a type 6П14П valve

4. Assume that $i_{max} = 140$ milliamperes and determine the anode supply voltage

$$E_a = \frac{8P}{i_{max}} = \frac{8\times 3}{0.14} = 172 \text{ volts}$$

or, taking the nearest standard value, 200 volts.

5. Find the grid bias voltage and the amplitude of the alternating grid voltage

$$|E_g| = \frac{i_{max}}{2g_m} = \frac{140}{2\times 11.3} \cong 6.18 \text{ volts}$$

Choose $V_g = |E_g| = 6$ volts.

6. Referring to the valve characteristic (Fig. 3.17) find the quiescent current, $I_{a\,(dc)} = 48$ milliamperes.

7. Plot the load line. From the line find $I_1 = 40$ milliamperes and $V_a = 184$ volts.

Now find the power output of the valve

$$P = \frac{1}{2} V_a I_1 = \frac{1}{2}\, 184 \times 0.04 = 3.68 \text{ watts} > P_{total}$$

8. Find the non-linear distortion. To do this, determine a, b, and c from the dynamic load line: $a = 22$, $b = 22$, $c = 27.5$

$$\gamma_3 = \frac{1}{2}\,\frac{2c-(a+b)}{a+b+c} = \frac{1}{2}\,\frac{2\times 27.5-(22+22)}{22+22+27.5} = 0.077$$

$$\gamma_3 = 7.7\% < 8\%$$

9. Find the anode load resistance

$$R_L = \frac{V_a}{I_1} = \frac{184}{0.04} = 4{,}600 \text{ ohms}$$

10. Determine the resistance of the transformer primary

$$r_1 = R_L \frac{1-\eta_t}{2} = 4{,}600\,\frac{1-0.75}{2} = 575 \text{ ohms}$$

11. Find the resistance of the self-bias resistor

$$R_k=\frac{|E_g|}{I_{a(dc)}+I_{g2}}=\frac{6}{(48+5)\times 10^{-3}}=113 \text{ ohms}$$

12. Find the voltage of the anode supply source

$$E_{a\,sup}=E_a+I_{a\,(dc)}r_1+|F_g|=200+0.048\times 575+6\cong 234 \text{ volts}$$

13. Calculate the inductance of the transformer primary

$$R_{eq}=\frac{R_L(R_a+2r_1)}{R_L+R_a+2r_1}=\frac{4{,}600\,(20{,}000+2\times 575)}{4{,}600+20{,}000+2\times 575}\cong 3{,}800 \text{ ohms}$$

$$L_1=\frac{R_{eq}}{\Omega_l\sqrt{M_l^2-1}}=\frac{3{,}800}{6.28\times 80\sqrt{1.25^2-1}}=10 \text{ henrys}$$

14. Find the leakage inductance

$$L_s=\frac{R_a+R_L}{\Omega_h}\sqrt{M_h^2-1}=\frac{20{,}000+4{,}600}{6.28\times 6{,}000}\sqrt{1.25^2-1}=0.49 \text{ henry}$$

15. Determine the turns ratio

$$n=\sqrt{\frac{Z_L}{R_L\eta_t}}=\sqrt{\frac{5}{4{,}600\times 0.75}}=0.038$$

16. Calculate the parameters of the compensating network

$$R_c=2R_L=2\times 4{,}600=9{,}200 \text{ ohms}$$

$$C_c=\frac{L_s+\frac{L_Z}{n^2}}{R^2}=\frac{0.49+\frac{0.003}{(0.038)^2}}{(9{,}200)^2}=3.04\times 10^{-8} \text{ farad}$$

or, taking the nearest standard value, $C=0.03$ microfarad.

Review Questions

1. Name methods for reducing non-linear distortion in a pentode amplifier.
2. Can the screen grid of a pentode or tetrode be connected directly to the anode?
3. Why does an increase in the anode load resistance result in greater non-linear distortion?
4. Can the power output of an amplifier be evaluated approximately without detailed circuit calculation?

17. Push-pull Power Amplifier

Amplifiers with a power output of 3 to 5 watts do well with a single triode, tetrode, or pentode arranged into a single-ended or single-valve circuit. Most a. f. amplifying valves are designed for such comparatively low power.

In some applications, the power delivered by a single valve may be insufficient. In the absence of a valve capable of handling a greater power, two or more valves may be used in parallel, that is, with the like electrodes connected together. Then two similar valves will have an anode current twice that of a single valve, while the anode resistance will be half of one valve. Thus the power output will be doubled. However, the direct anode current increases in the same proportion, posing additional difficulties in transformer design and introducing additional non-linear distortion.

As an alternative, valves may be used in *push-pull*. Figure 3.18 shows two valves connected in this way. The valves are fed from the previous stage by means of a transformer which has the centre tap of its secondary connected to earth ($-E_g$).

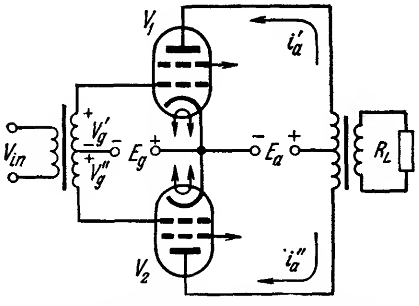


Fig. 3.18. Push-pull amplifier circuit

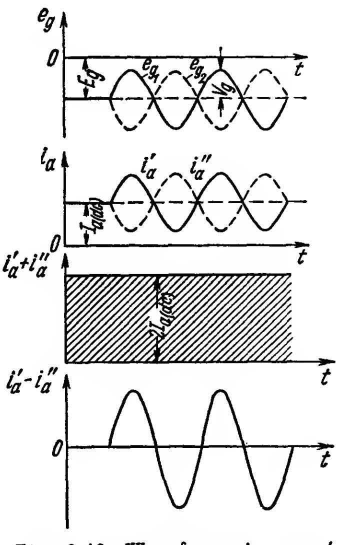


Fig. 3.19. Waveforms in a push-pull amplifier

At any instant when the top terminal of the input transformer secondary is positive with respect to the centre tap, the bottom terminal of the secondary winding will be negative.

The centre tap is connected to the valve cathodes via the grid bias source. Therefore, the signal voltages applied to the grids of V_1 and V_2 are separated by a half-period in time and by 180° in phase or, as is commonly said, they are in phase opposition.

Owing to this, an increase in the anode current in, say, V_1 will be accompanied by a decrease in the current in the other valve (V_2). One half-cycle later the situation reverses.

Figure 3.19 shows the curves relating grid voltages and anode currents.

Let us determine how the anode currents of V_1 and V_2 affect the output transformer primary. We shall express the instantaneous current values in terms of a power series. The current through V_1 is then

$$i'_a = I'_{a\,(dc)} + I'_1 \cos \Omega t + I'_2 \cos 2\Omega t + I'_3 \cos 3\Omega t + \dots \quad (3.32)$$

and the current through V_2

$$i''_a = I''_{a\,(dc)} + I''_1 \cos \Omega \left(t + \frac{T_1}{2}\right) + I''_2 \cos 2\Omega \left(t + \frac{T_1}{2}\right) + I''_3 \cos 3\Omega \left(t + \frac{T_1}{2}\right) + \dots \quad (3.33)$$

where T_1 is the period of the first-harmonic current. In terms of the first-harmonic period, the angular frequency of the signal is:

$$\Omega = \frac{2\pi}{T_1}$$

Substituting it into Eq. (3.33) gives

$$i''_a = I''_{a\,(dc)} + I''_1 \cos \frac{2\pi}{T_1} \left(t + \frac{T_1}{2}\right) + I''_2 \cos 2 \frac{2\pi}{T_1} \times \left(t + \frac{T_1}{2}\right) + I''_3 \cos 3 \frac{2\pi}{T_1} \left(t + \frac{T_1}{2}\right) + \dots$$

$$i''_a = I''_{a\,(dc)} + I''_1 \cos \left(\frac{2\pi}{T_1} t + \pi\right) + I''_2 \cos \left(2 \frac{2\pi}{T_1} t + 2\pi\right) + I''_3 \cos \left(3 \frac{2\pi}{T_1} t + 3\pi\right) + \dots$$

Reverse substitution of $\frac{2\pi}{T_1}=\Omega$ and the usual trigonometric relations give

$$i''_a = I''_{a\,(dc)} + I''_1 \cos(\Omega t + \pi) + I''_2 \cos(2\Omega t + 2\pi) + I''_3 \cos(3\Omega t + 3\pi)$$

$$i''_a = I''_{a\,(dc)} - I''_1 \cos \Omega t + I''_2 \cos 2\Omega t - I''_3 \cos 3\Omega t + \ldots \quad (3.34)$$

Comparing Equations 3.32 and 3.34, the following conclusions can be drawn:

(1) the first and third harmonics of the anode currents are 180° out of phase;

(2) the second and all other even harmonics of the anode currents are in phase.

In the common anode supply wire connecting the positive terminal of the supply source with the centre-tap of the output transformer the instantaneous current is equal to the sum of the instantaneous anode currents through valves V_1 and V_2

$$i_a = i'_a + i''_a$$

Substituting the expressions for i'_a and i''_a from Equations (3.32) and (3.34) gives

$$i_a = I'_{a\,(dc)} + I''_{a\,(dc)} + (I'_1 - I''_1) \cos \Omega t + (I'_2 + I''_2) \cos 2\Omega t + (I'_3 - I''_3) \cos 3\Omega t + \ldots \quad (3.35)$$

If the valves have similar characteristics, i. e. if the stage arms are balanced, the like current components are equal

$$I'_{a\,(dc)} = I''_{a\,(dc)}, \quad I'_1 = I''_1, \quad I'_2 = I''_2, \quad I'_3 = I''_3 \ldots$$

Noting this equality, we finally obtain

$$i_a = 2I_{a\,(dc)} + 2I_2 \cos 2\Omega t$$

Thus, when the amplifier arms are balanced, only the d. c. components and even harmonics of the anode currents flow in the common anode supply wire.

Since no first-harmonic current appears in the anode supply circuit, the parasitic feedback usually taking place in a multistage amplifier through the internal resistance of the common power supply is reduced to a marked degree. This improves the stability of operation of the multistage circuit.

The currents through the halves of the primary winding flow in opposite directions. Therefore, the magnetic flux due to the currents in valves V_1 and V_2 is proportional to the difference between instantaneous currents i'_a and i''_a

$$\Phi = A\,(i'_a - i''_a) \tag{3.36}$$

where A is a proportionality factor allowing for transformer design and the number of primary turns.

Substituting the expressions for the currents from Equations (3.32) and (3.34) into Eq. (3.36) we get

$$\Phi = A\,(I'_{a\,(dc)} - I''_{a\,(dc)}) + A\,(I'_1 + I''_1)\cos\Omega t$$
$$+ A\,(I'_2 - I''_2)\cos 2\Omega t + A\,(I'_3 + I''_3)\cos 3\Omega t + \ldots \tag{3.37}$$

For a stage with similar valves and a balanced primary winding Equation (3.37) reduces to

$$\Phi = 2A\,(I_1 \cos\Omega t + I_3 \cos 3\Omega t + \ldots)$$

Thus, when the amplifier arms are balanced the magnetic flux, and, consequently, the secondary voltage are solely controlled by the odd harmonics of the anode current.

The direct components of the anode currents flowing in opposite directions establish magnetic fluxes around the primary winding, equal in magnitude and opposite in direction.

In a precisely balanced circuit, the resultant d.c. flux is zero. The absence of d. c. magnetisation makes it possible to reduce the size and weight of the output transformer to a considerable degree. The resultant magnetic flux due to the even-harmonic currents which have opposite directions will also be equal to zero. Thus the voltages across the secondary terminals of the output transformer will be due solely to odd-harmonic currents.

In the general case, according to Eq. (3.37), the magnetic flux due to even-harmonic currents is proportional to the difference in amplitude between the harmonics. Therefore, considerably wider limits may be specified for the non-linear distortion caused by the second-harmonic anode current, the strongest harmonic after the fundamental. As a result, the anode characteristic may be utilized better and the direct component, $I_{a(dc)}$, of the anode current may be increased.

The anode current efficiency β may be close to zero in the push-pull circuit, and the relation

$$\frac{I_1}{I_{a(dc)}} = 1 - \beta$$

may be close to unity.

A reduction in the direct component of the anode current causes a reduction in the power drawn from the anode supply source, and, consequently, improves the efficiency of the stage.

A push-pull circuit may be Class A, Class AB, or even Class B, which cannot be with a single-ended circuit because of prohibitively high non-linear distortion ($\gamma = 30\%$).

Because the arms of a push-pull amplifier are balanced in respect to the anode supply source, simultaneous changes in i'_a and i''_a due to the ripple in the supply voltage cause the magnetic flux around the primary to vary by equal amounts in opposite directions, so that these increments cancel each other. This is why a. c. hum is not heard at the output of a push-pull amplifier. This is also true of the a. c. hum due to the filament voltage, usually observed in single-ended amplifiers. This is why a push-pull amplifier needs only very little filtering.

Now we are in a position to formulate the advantages of the push-pull amplifier as follows:

(1) the power output is doubled;

(2) the second-harmonic non-linear distortion is reduced;

(3) the valves may be more fully utilized in Class A operation. Class AB and Class B operation is possible;

(4) the efficiency of the stage is increased;

(5) the amplifier is less sensitive to the ripple in the anode supply voltage;

(6) the first harmonic of the signal does not find its way into the anode supply circuit; this decreases parasitic interstage coupling and makes the amplifier more stable.

The disadvantages of the push-pull circuit are as follows:

(a) the preceding stage must have a balanced output. This creates difficulties in coupling a single-ended stage to a push-pull one;

(b) the push-pull circuit requires at least two valves. However, in a low-power amplifier this is not a problem since one double valve may be used instead of two single ones.

Design of a Push-Pull Stage. Since the arms of a push-pull amplifier are balanced, its equivalent circuit diagram for the mid-frequency range may be presented as in Fig. 3.20. The dotted line indicates the anode supply circuit through which odd harmonics do not pass. The circuit with two a. c. sources may be replaced by a circuit with a single source having twice the voltage and internal resistance (Fig. 3.21). Such a circuit does not differ in any way from the equivalent circuit of a single-

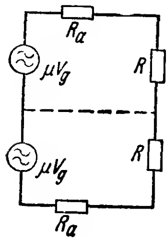


Fig. 3.20. Mid-band equivalent circuit of a push-pull amplifier

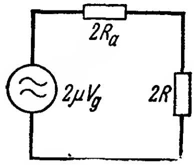


Fig. 3.21. Transformed equivalent circuit of a push-pull amplifier

ended stage. Therefore, the design equations given for the single-ended circuits may be also used in the design of a push-pull stage.

Practical calculations of a Class A push-pull amplifier are usually performed in the following manner.

Determine the power output to be supplied by one arm

$$P' = \frac{P_{out}}{2\eta_t}$$

Then calculate the arm, using the procedure given for a single-ended stage. From the data thus obtained calculate the characteristics and parameters of the whole stage.

When calculating one arm of a push-pull amplifier, it is important to consider the following. As already mentioned, the valves of a push-pull amplifier have better anode current efficiency. If we assume that $\beta = 0$, then the no-signal maximum anode dissipation of a triode will be

$$P_{a\ max} = 2P' \frac{2+\alpha}{\alpha}$$

Hence, the number of valves in one arm will be

$$m = \frac{P_{a\ max}}{P_{a\ safe}}$$

With a higher anode current efficiency, a valve operates partly within the lower curved part of the characteristic. In this case, the a. c. anode resistance R_a is 10-15 per cent greater than the valve given in the valve manual. The increased a. c. anode resistance will inevitably affect $E_{a\ sup}$ and V_g.

The design equation for non-linear distortion is also slightly modified. While in a single-ended stage the anode load resistance is selected to obtain a symmetrical change in the anode current, in a push-pull amplifier the load resistance is selected for maximum power output from the valve. The second harmonic, which inevitably appears with such a load resistance, will be minimized by the symmetry of the circuit configuration. In this case, non-linear distortion is

$$\gamma=\gamma_3=\frac{I_3}{I_1}$$

However, the absolute symmetry of the arms is a rare occurrence. Experience shows that the valves used in the amplifier arms cannot be made fully identical.

Inequality of the valves used in the amplifier arms is expressed by the asymmetry factor δ

$$i''_a=i'_a\,(1+\delta)$$

If the valve asymmetry cannot be disregarded, the non-linear distortion should be calculated from the following equation:

$$\gamma=\sqrt{\left(\frac{\delta}{2-\delta}\gamma_2\right)^2+\gamma_3^2} \tag{3.38}$$

The equation for the turns ratio n is also slightly modified in that it uses the doubled value of the anode load resistance determined for one arm

$$n=\sqrt{\frac{Z_L}{2R'_L\eta_t}}$$

Review Questions

1. How are the valves connected in the push-pull circuit? What is the advantage of this connection?
2. Why should the grid voltages be 180° cut of phase in a push-pull circuit?
3. How can the arm of a push-pull stage be balanced for d. c.?
4. Why has a push-pull amplifier a greater efficiency than a single-ended one?
5. What will be the changes in readings of a meter measuring the anode current of a Class B amplifier when the signal voltage is applied to the grid?

18. Negative Feedback in Audio-frequency Amplifiers

A feedback amplifier may be defined as one in which a part of the output signal is fed back to its input. When the feedback is used to decrease the effective input, this is *negative* feedback.

The feedback signal may be proportional to changes in either the output voltage, or the load current, or both. Accordingly, one has:

(1) *voltage* feedback;
(2) *current* feedback;
(3) *bridge* feedback.

Block diagrams of amplifiers employing negative feedback are shown in Figs. 3.22, 3.23, and 3.24.

In the circuit of Fig. 3.22, the feedback voltage, V_{fb}, proportional to the output voltage and applied through the feedback circuit to the amplifier input along with the signal voltage V_{in}, is given by

$$\dot{V}_{fb} = \dot{\beta}\dot{V}_{out} \tag{3.39}$$

The quantity β is known as the *feedback-path gain*, although some authors call it the *feedback factor*.

In the circuit of Fig. 3.23 the negative feedback voltage V_{fb} is proportional to the load current I_L and the coupling impedance Z_c

$$\dot{V}_{fb} = \dot{\beta} Z_c \dot{I}_L$$

The value of the negative feedback voltage in the circuit of Fig. 3.24 is proportional to both the output voltage and the load current

$$\dot{V}_{fb} = \dot{\beta}(\dot{V}_{out} + Z_c\dot{I}_L) \tag{3.40}$$

In the above circuits the negative feedback voltage is returned to the input circuit in series with the signal. This is *series* negative feedback.

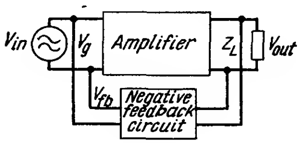


Fig. 3.22. Block diagram of amplifier with voltage negative feedback

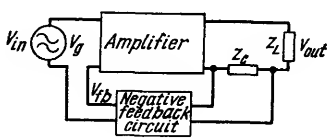


Fig. 3.23. Block diagram of amplifier with current negative feedback

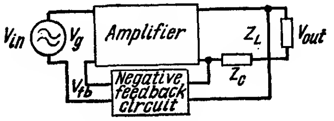


Fig. 3.24. Block diagram of amplifier with negative bridge feedback

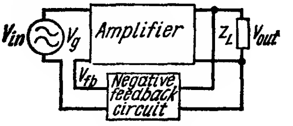


Fig. 3.25. Block diagram of amplifier with parallel negative voltage feedback

In some amplifiers the negative feedback voltage is applied across to the input terminals (Fig. 3.25). This is *parallel* or *shunt* feedback.

Consider the physical properties of an amplifier employing negative voltage feedback.

In the circuit of Fig. 3.22, V_g is the sum of the input voltage V_{in} and the negative feedback voltage V_{fb}

$$\dot{V}_g = \dot{V}_{in} + \dot{V}_{fb} \tag{3.41}$$

Let the gain of an amplifier without negative feedback, or the *open-loop gain* of the amplifier, be K, such that

$$\dot{K} = \frac{\dot{V}_{out}}{\dot{V}_g} \tag{3.42}$$

and the gain of the same amplifier with negative feedback, or the *closed-loop gain* of the amplifier, be K_{fb}, such that

$$\dot{K}_{fb} = \frac{\dot{V}_{out}}{\dot{V}_{in}} \tag{3.43}$$

Substituting the expression for V_{in} from Eq. (3.41) into Eq. (3.43) gives

$$\dot{K}_{fb} = \frac{\dot{V}_{out}}{\dot{V}_g - \dot{V}_{fb}}$$

Substituting the expression for V_{fb} from (3.39) yields

$$\dot{K}_{fb} = \frac{\dot{V}_{out}}{\dot{V}_g - \beta\dot{V}_{out}} = \frac{\dot{V}_{out}}{\dot{V}_g\left(1 - \beta\frac{\dot{V}_{out}}{\dot{V}_g}\right)} \tag{3.44}$$

Noting Eq. (3.42), we finally have

$$\dot{K}_{fb} = \frac{\dot{K}}{1 - \beta\dot{K}} \tag{3.45}$$

The term βK in Eq. (3.45) is frequently called the *feedback factor*, or *return ratio*, or *loop gain*, since the input signal is multiplied by K and then by β as it takes a closed path around the complete feedback amplifier. It should be noted, however, that in some texts the quantity $(1-\beta K)$ is called the feedback factor.

In the general case, K and βK are complex quantities.

When βK is negative (negative feedback), K_{fb}, the closed-loop gain, is less than K, the open-loop gain

$$K_{fb}=\frac{K}{1+\beta K}<K$$

This is explained by the fact that, with negative feedback, the grid voltage is lower than the input signal voltage by the value of feedback. The reduction of the voltage applied to the grid leads to a reduced voltage at the amplifier output. As a result, the stage gain also decreases.

When βK is positive (positive feedback), the modulus of K_{fb} may be greater than K. This is explained by the fact that the feedback voltage is in phase with the signal voltage and boosts the voltage at the valve grid. As a consequence, the voltage at the output of the amplifier also increases.

When $\beta K=1$, the denominator in Eq. (3.45) is zero, and K_{fb} is infinity. Then even a negligible input voltage of any waveform will cause considerable voltages at the output. The equality $\beta K=1$ is the condition for the self-oscillation of an amplifier, that is, the amplifier becomes an oscillator generating spurious oscillations. This phenomenon is ruinous to amplifier performance.

Negative feedback has its effect not only on the gain, but also on distortion, gain stability and stability of output voltage.

As will be recalled, any distortion and noise finally turn up as an impairment in the signal waveform. In an amplifier with feedback, any distortion of the signal waveform gives rise to a voltage which is returned to the input in anti-phase with the signal. This voltage, amplified by the amplifier, balances out the distortion.

Consider the effect of negative feedback on distortion in the mid-band range where phase shifts in the amplifier and feedback circuit are small, so that both K and β are real and not complex quantities.

Let V_h be the harmonic voltage at the output of an amplifier without feedback, and $V_{h\,fb}$ at the output of one with feedback.

$V_{h\,fb}$ may be defined as

$$V_{h\,fb} = V_h - K_0\beta_0 V_{h\,fb} \tag{3.46}$$

where K_0 is the mid-band gain and β_0 is the mid-band feedback-path gain.

Solving Eq. (3.46) for $V_{h\,fb}$ yields

$$V_{h\,fb} = V_h/(1+\beta_0 K_0) \tag{3.47}$$

As is seen, the harmonic voltage at the amplifier output is reduced in the ratio $1/(1+\beta_0 K_0)$.

If the voltage at the fundamental frequency is held constant by increasing, say, V_{in}, the non-linear (harmonic) distortion will be reduced in the same ratio

$$\gamma_{fb} = \gamma/(1+\beta_0 K_0) \tag{3.48}$$

Reduction in non-linear distortion is especially important in power amplifiers. In addition, negative feedback improves the utilization of the valve characteristics and the electrical efficiency.

The relation derived for the harmonic output voltage also holds for other sources of interference and noise in amplifiers, such as valve noise, mains hum, etc.:

$$V_{n\,fb} = V_n/(1+\beta_0 K_0) \tag{3.49}$$

where V_n is the noise voltage in an amplifier without feedback and $V_{n\,fb}$ is the noise voltage in one with feedback.

It should be noted that in the low- and high-frequency ranges of the bandwidth, phase shifts appear both in the amplifier and the feedback circuit. This inevitably modifies the relations derived for the effect of negative feedback in the mid-band range.

Negative feedback in an amplifier considerably reduces frequency and phase distortion. The flattening of the frequency response may be pictured as follows.

As the gain at the limiting frequencies of the range decreases, the amplifier output voltage goes down and so does the negative feedback voltage fed in anti-phase to the amplifier input. As a result, the total grid voltage goes down less at these frequencies than at the mid-band frequencies, and there is a relative increase in the output voltage. In other words, negative feedback affects the mid-band stage gain much more than the gain at the low and high frequencies.

The reduction in frequency distortion is accompanied by a reduction in phase distortion. In Chapter II we have defined the

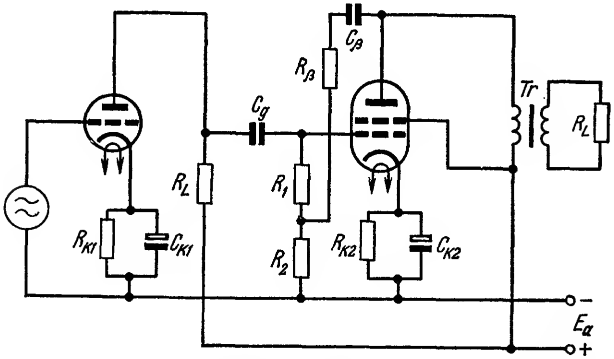


Fig. 3.26. Final stage of an amplifier with negative feedback

relation between the phase shift and the frequency distortion as follows

$$\cos\varphi = \frac{1}{M}$$

The closer M to unity, the smaller the phase shift φ.

Consider practical amplifier circuits employing negative feedback.

Figure 3.26 shows a commonly used final stage of an amplifier with negative feedback. In this arrangement, the negative feedback circuit consists of a blocking capacitor C_β, resistor R_β and resistor R_2 connected in series with the valve grid-leak resistor R_1. The reactance of C_β at the lowest frequency of the range must be one-third to one-fifth of the total resistance of R_β and R_2.

Approximately, the feedback-path gain is given by

$$\beta \simeq \frac{R_2}{R_2 + R_\beta} \tag{3.50}$$

If the anode load resistance R_L and the a. c. anode resistance of the valve are comparatively low, then a more accurate equation should be used for β

$$\beta = \frac{R_2}{R_2 + R_\beta}\,\frac{R_{eq}}{R_{eq} + R_1} \tag{3.51}$$

where R_{eq} is defined as

$$\frac{1}{R_{eq}} = \frac{1}{R_a} + \frac{1}{R_L} + \frac{1}{R_1}$$

Figure 3.27 shows an amplifier in which negative feedback is carried across two stages. In this circuit, β depends upon the

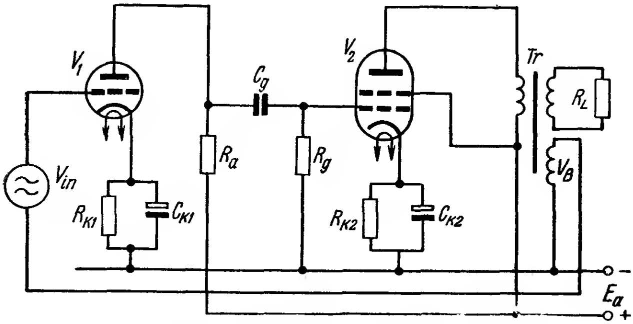


Fig. 3.27. Two-stage amplifier with negative feedback

ratio of the turns N_β in the feedback winding and the turns N_1 in the transformer primary

$$\beta = \frac{N_\beta}{N_1}$$

The connection for the leads of N_β to the circuit components is found experimentally.

Figure 3.28 shows the circuit of an amplifier with parallel negative voltage feedback. The negative feedback voltage is fed from the anode circuit of V_2 to its input through R_β.

The feedback factor for this circuit is given by

$$\beta_0 K_0 = (1 + K_0) \frac{R_{eq}}{R_\beta} \tag{3.52}$$

where K_0 is the open-loop gain of the second stage.

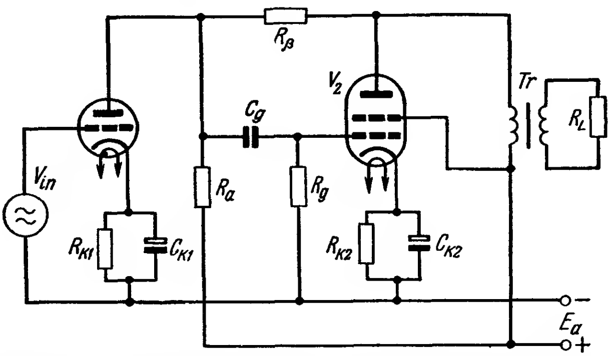


Fig. 3.28. Circuit with negative parallel voltage feedback

Parallel negative feedback is best adapted for an amplifier using a pentode in the first stage.

Review Questions

1. Why should the signal voltage be increased in an amplifier with negative feedback?
2. Is it possible to obtain large power output from a stage with negative feedback?
3. What factors control the amount of feedback?
4. Name the physical phenomena minimizing non-linear distortion in a stage with negative feedback.
5. Quote the value of the product βK at which the voltage gain will be close to unity.

19. Phase-inversion Circuits

Two voltages, equal in magnitude and opposite in phase, must be applied to the grids of the valves connected in push-pull. The simplest way of doing this would be to use an input transformer with a centre-tapped secondary (Fig. 3.18). Such a transformer can match any signal source to the balanced input of the push-pull amplifier. Transformer coupling is especially convenient when the push-pull stage draws grid current.

However, an input transformer increases the cost and weight of the equipment as well as frequency and phase distortion in it. This is why special circuits are used to couple a signal source to the balanced input of a push-pull amplifier operating without grid current.

Amplifier circuits which develop two voltages equal in magnitude and opposite in phase are called *phase-inversion circuits*.

Phase-inversion circuits are resistance-coupled audio-frequency amplifiers with a balanced output. The simplest phase-inversion circuit is shown in Fig. 3.29. In this circuit, the anode load consists of two parts. One part, $R_L = \frac{R_L}{2}$, is connected between the anode and the positive terminal of the anode supply source. The other part, $R_L^{''} = \frac{R_L}{2}$, is connected between the negative terminal of the anode supply source and cathode through a cathode resistor R_{k1} and a bypass capacitor C_{k1}. Both R_{k1} and C_{k1} present a low impedance at the signal frequency.

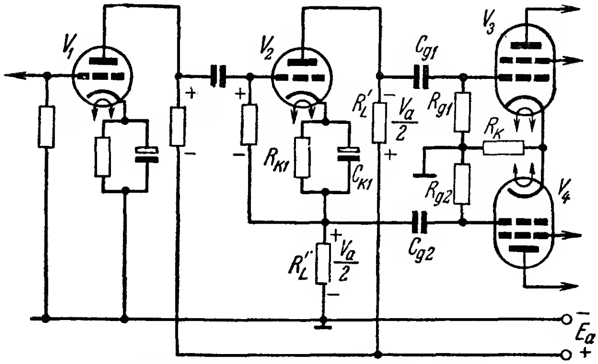


Fig. 3.29. Single-ended split-load phase inverter

The top of R'_L is connected to the grid of the valve V_3 via C_{g1}, while the top of R''_L is connected to the grid of the valve V_4 via C_{g2}. When at any moment the grid of V_1 goes positive, the amplified instantaneous alternating voltage across R'_L will be negative and that across R''_L, positive.

Thus, the split anode load furnishes alternating voltages which are 180° out of phase. These voltages will be equal if the equivalent resistances of the split load are equal. With the equality of R'_L and R''_L, the circuit is sufficiently balanced at the mid-band frequencies.

The gain of this type of phase inverter is always smaller than 2

$$K_0 = \frac{V_a}{V_{in}} < 2$$

owing to tight negative feedback with $\beta = 0.5$, because $R_\beta = 0.5 R_L$.

$$K_{fb} = \frac{K_0}{1+\beta K_0} = \frac{K_0}{1+0.5\,K_0} = \frac{K_0}{K_0\left(\frac{1}{K_0}+0.5\right)} < 2$$

Higher gain is obtainable with push-pull circuits. One of the valves shifts the signal through 180°. Application of a twin triode instead of two separate valves in such a circuit is logical and keeps the cost down.

Figure 3.30 shows a push-pull self-balancing phase inverter. The signal voltage applied to the grid of V_1 is amplified and

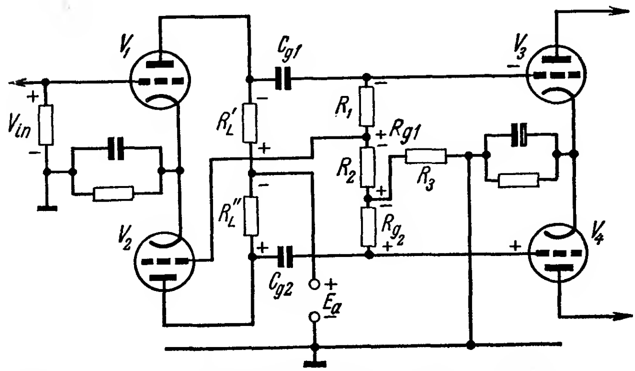


Fig. 3.30. Push-pull self-balancing inversing circuit

applied to the grid of V_3 through C_{g1}. This voltage is 180° out of phase with the signal voltage. The grid of V_2 is actuated by an a. c. voltage from R_2 which, together with R_1, forms a voltage divider. This voltage is also 180° out of phase with the input voltage. The voltage applied to the grid of V_3 is higher than the input voltage K_0 times. Therefore, in order to supply a voltage equal to input voltage V_{in} to the grid of valve V_2, R_2 must be $1/K_0$th fraction of R_{g1}

$$R_2 = \frac{R_{g1}}{K_0} = \frac{R_1 + R_2}{K_0} \tag{3.53}$$

The voltage amplified by V_2 is applied to the grid of V_4 via C_{g2}. If Eq. (3.53) is satisfied and the circuit parameters of valves V_1 and V_2 are similar, equal and opposite voltages will be applied to the grids of valves V_3 and V_4.

The polarity of the a. c. voltages acting in the circuit is indicated by "+" and "—".

Let us see how the self-balancing phase inverter operates.

Let the voltage V_g' be greater than V_g'', so that the circuit is unbalanced. The increase in V_g' will cause a voltage increase across R_1 and R_2 and, consequently, a voltage increase at the grid of valve V_2. Thus, an increase in V_g' will be accompanied by an increase in the output voltage of valve V_2. As a result, an additional voltage will appear across R_3, opposite in phase to V_g'. Hence any change in V_g' is accompanied by an additional balancing voltage across R_3, such that the unbalance of the circuit will be minimized to a fraction of its original value.

Review Questions

1. Can a transformer without a centre tap be used to drive a push-pull stage?
2. Why is the gain of a split-load phase-inverter less than two?
3. Can a push-pull amplifier be driven by a single-ended stage?
4. Which of the phase-inversion circuits is most economical?
5. How can the grid voltage be checked for symmetry?

SUMMARY

1. The basic type of power amplifier is one using a transformer to match the load to the a.c. anode resistance of the valve.
2. The frequency response of the output stage falls off at the higher and lower frequencies of the band.
3. Frequency distortion in the low-frequency range is caused by the low inductance of the transformer primary.
4. Frequency distortion in the high-frequency range is caused by an increased leakage inductance.
5. The condition for maximum power output in a stage with a constant driving voltage is the equality of the load resistance and the a.c. anode resistance of the valve.
6. For maximum power output in the anode circuit with a constant anode supply voltage, the anode load resistance must be twice the a.c. anode resistance of the valve.
7. Final stages of radio receivers and low-power amplifiers use pentodes and beam-power tetrodes because these valves have higher sensitivity and better efficiency than triodes.
8. Medium- and high-power amplifiers employ push-pull circuits. The efficiency of a push-pull amplifier is higher than that of a single-ended amplifier.
9. In a push-pull amplifier, non-linear distortion due to even harmonics of anode current is lower than it is in a single-ended amplifier.
10. The push-pull circuit is less sensitive to the ripple in the anode supply voltage.
11. A single-ended stage may be coupled to a push-pull stage via an input transformer or a phase-inversion circuit.
12. Phase inverters furnish voltages that are equal but opposite in phase.
13. Audio-frequency amplifiers often use negative feedback which decreases all types of distortion appearing during amplification.

CHAPTER IV

AUDIO-FREQUENCY TRANSISTOR AMPLIFIERS

20. Brief Outline of Transistors

As compared with valves, semiconductor devices are smaller in size and weight, more economical in power consumption, more robust mechanically, and durable.

Two main types of transistors are the *point-contact type* and the *junction type*. Although historically the older of the two, point-contact transistors are not widely used, and the basic commercial type is the junction transistor.

Transistors may be further classed into *P-N-P* and *N-P-N*. In a *P-N-P* transistor, usually fabricated from germanium, conduction in the emitter and collector regions is by holes, or positive charge carriers, while in the base region conduction is by electrons, or negative charge carriers. In an *N-P-N* transistor, usually made of silicon, conduction in the emitter and collector regions is by electrons, and in the base region by holes.

Still another classification of transistors is according to the mechanism by which charge carriers are transported across the junctions. In some, charge transfer is by diffusion, and the transistors are called *diffusion transistors* (not to be confused with *diffused* transistors, so called because of the manufacturing process involved). In others, charge transfer is due to the *drift* caused by an internal electric field, and the transistors are called *drift transistors*. In each class, units may be fabricated by a large variety of processes varying from manufacturer to manufacturer. Thus, alloy transistors transfer carriers from emitter to collector by diffusion. There is little or no electric field in the base region to accelerate motion of the carriers to the collector. The reason for this is the constant resistivity of the base region. In diffused transistors, on the other hand, there is a gradual variation of resistivity in the base, and there may be an assisting, or sweeping, electric field considerably reducing transit time in the base.

Although the transistor is often likened to the valve, the physical processes taking place in the two are markedly different. For example, the valve deals with only one type of current carrier, electrons moving through a vacuum. The finite transit time of electrons begins to be felt only in the microwave region. The parameters of the valve depend on the operating voltages and currents and are almost independent of the environments. The interelectrode capacitances of the valve, affecting operation of a valve amplifier at the higher frequencies, are mainly determined by valve construction.

In contrast, the transistor has two types of current carrier, electrons and holes. The flow of current in the collector circuit is produced by injection of minority carriers from the emitter into the base and their motion into the collector region. This takes place in a solid material, and not in a vacuum. Therefore, the finite transit time of carriers in transistors affects their operation already in the HF (high-frequency) band. As a result, the *short-circuit gain*, or the *current amplification factor*, α, for the common-base configuration, a major parameter of the transistor, is frequency-dependent:

$$\alpha = F(f)$$

At the higher frequencies, the current amplification alpha becomes a complex quantity:

$$\dot{\alpha} = \alpha_0/(1 + jf/f_\alpha)$$

and its modulus is

$$\alpha = \alpha_0/\sqrt{1 + (f/f_\alpha)^2}$$

where

α_0 = amplification factor at zero or very low frequency

f = operating frequency

f_α = *alpha cut-off frequency* at which the current amplification in the common-base connection drops to 0.707 of its low-frequency value (3 db down).

As the width of the base increases, the amplification factor alpha becomes more and more frequency-dependent.

The electric properties of the transistor are strongly affected by ambient temperature and heating in operation.

The apparent capacitances formed in the transistor on each side of the depletion region, known as the *barrier* or *junction* *capacitances*, vary with the operating voltages. The so-called *diffusion* or *storage capacitances*, formed in a forward-biased transistor, may be many times the input capacitance of the valve.

Since there is always a current flowing in the input circuit of the transistor, its input resistance is only a fraction of the valve.

Accordingly, equivalent circuits for the transistor, especially for operation at the high frequencies, turn out to be far more complicated than they are for the valve.

Moreover, while the behaviour of the valve operating with a negative grid (zero grid current) may be described by a single family of characteristics, such as

$$i_a = F(e_g) \text{ with } E_a \text{ held constant}$$

or

$$i_a = F(e_a) \text{ with } E_g \text{ held constant}$$

at least two families of characteristics, *input* and *output*, are necessary to describe the behaviour of the transistor.

The input and output characteristics quoted in transistor data sheets are usually plotted for typical, or "bogey", transistors and do not faithfully reflect the properties of actual units because the spread between them is still considerable.

Review Questions

1. What is the difference and the similarity between the valve and the transistor?
2. Why are the transistor parameters dependent on temperature?
3. Explain the dependence of transistor parameters on frequency.
4. Does the open-emitter collector current I_{CBO} affect the amplifying properties of the transistor?
5. Can an r.f. transistor operate in an a.f. amplifier?

21. Analysis of Transistor Amplifiers

Analysis of a single-stage transistor amplifier involves the following parameters:

— the current gain

$$K_i = I_{out}/I_{in}$$

— the voltage gain

$$K_v = V_{out}/V_{in}$$

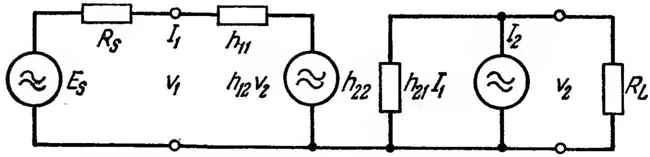


Fig. 4.1. Equivalent circuit for a transistor amplifier

— the power gain

$$K_p = P_{out}/P_{in}$$

— the input resistance

$$Z_{in} = V_{in}/I_{in}$$

— the output resistance

$$Z_{out} = V_{out}/I_{out}$$

It is usual to think of the transistor as a four-terminal or two-port network having the simple equivalent circuit of Fig. 4.1 which is drawn up in terms of the h-parameters, usually quoted in transistor data sheets.

It has been noted that in the general case the transistor parameters are complex quantities. At the low frequencies, for which the equivalent circuit of Fig. 4.1 is valid, it will be safe to treat the h-parameters as real quantities. Then by Kirchhoff's second law we may write for the input circuit

$$E_s = I_1 R_s + h_{11} I_1 - h_{12} V_2 \tag{4.1}$$

By Kirchhoff's first law for the output circuit we have

$$h_{21} I_1 = I_{h22} + I_2 \tag{4.2}$$

where I_{h22} is the current in branch h_{22} and I_2 is the output current of the two-port.

As will be recalled,

$$I_{h22}/I_2 = h_{22}/(1/R_L) = R_L h_{22}$$

Therefore,

$$I_{h22} = R_L h_{22} I_2$$

Substituting the expression for I_{h22} in Eq. (4.2) gives

$$h_{21} I_1 = R_L h_{22} I_2 + I_2$$

Solving for I_2 gives

$$I_2 = I_1 h_{21}/(1 + R_L h_{22}) \tag{4.3}$$

Current Gain. Numerically, the stage current gain is equal to the ratio of the output and input currents. Hence,

$$K_i = I_2/I_1 = I_1h_{21}/(1 + R_Lh_{22}) \div I_1 = h_{21}/(1 + R_Lh_{22}) \quad (4.4)$$

Input Resistance. It is equal to the ratio of V_1 to I_1:

$$R_{in} = V_1/I_1$$

From Eq. (4.1),

$$V_1 = E_s - R_sI_1 = h_{11}I_1 - h_{12}V_2 \quad (4.5)$$

In terms of the load resistance and the output current

$$V_2 = I_2R_L$$

Substituting the expression for I_2 from Eq. (4.3) gives

$$V_2 = R_LI_1h_{21}/(1 + R_Lh_{22})$$

Therefore,

$$V_1 = h_{11}I_1 - h_{12}V_2 = h_{11}I_1 - \frac{R_Lh_{12}h_{21}}{1 + R_Lh_{22}}I_1 = I_1\left(h_{11} - \frac{R_Lh_{12}h_{21}}{1 + R_Lh_{22}}\right)$$

whence

$$R_{in} = V_1/I_1 = \frac{I_1\left(h_{11} - \frac{R_Lh_{12}h_{21}}{1 + R_Lh_{22}}\right)}{I_1} = h_{11} - \frac{R_Lh_{12}h_{21}}{1 + R_Lh_{22}} \quad (4.6)$$

or

$$R_{in} = \frac{h_{11} + R_Lh_{11}h_{22} - R_Lh_{12}h_{21}}{1 + R_Lh_{22}} = \frac{h_{11} + R_L\Delta_h}{1 + R_Lh_{22}} \quad (4.7)$$

where

$$\Delta_h = h_{11}h_{22} - h_{12}h_{21}$$

Voltage Gain. As already defined,

$$K_v = V_2/V_1$$

In terms of output and input currents, we have

$$K_v = V_2/V_1 = I_2R_L/I_1R_{in}$$

Noting that

$$I_2/I_1 = K_i$$

we may write

$$K_v = K_i(R_L/R_{in})$$

Substituting the expressions for K_i and R_{in} from Eqs. (4.4) and (4.7) gives

$$K_v = K_i(R_L/R_{in}) = \frac{h_{21}}{1 + h_{22}R_L}\,\frac{R_L(1 + h_{22}R_L)}{h_{11} + \Delta_hR_L} = \frac{h_{21}R_L}{h_{11} + \Delta_hR_L} \quad (4.8)$$

By definition, the voltage gain is the ratio of V_2 to V_1. More accurately, however, the output voltage ought to be related to the signal voltage E_s

$$K_v' = V_2/E_s \tag{4.9}$$

Writing E_s in terms of V_1, we have

$$E_s = V_1 + I_1 R_s = V_1 + R_s(V_1/R_{in}) = V_1(1 + R_s/R_{in})$$

Substituting the expression for E_s in Eq. (4.9) gives

$$K_v = V_2/E_s = \frac{V_2}{V_1\left(1+\frac{R_s}{R_{in}}\right)} = K_v/(1 + R_s/R_{in})$$

Substituting the expressions for K_v and R_{in}, we may re-write the equation for K_v' thus

$$K_v' = \frac{h_{21}R_L}{(h_{11} + \Delta_h R_L) + R_s(1 + h_{22}R_L)} \tag{4.10}$$

Output Resistance. Numerically, it is equal to the ratio of the output voltage V_2 at R_L = infinity to the output current I_2 at $E_s = 0$

$$R_{out} = V_2/I_2$$

The respective equivalent circuit is shown in Fig. 4.2. By Kirchhoff's law, for the input circuit

$$0 = I_1 R_s + I_1 h_{11} + h_{12} V_2$$

Hence, the input current is

$$I_1 = -h_{12}V_2/(R_s + h_{11}) \tag{4.11}$$

The equation for the output current is

$$I_2 = h_{21} I_1 + h_{22} V_2 \tag{4.12}$$

Substituting the expression for I_1 from Eq. (4.11) in Eq. (4.12) gives

$$I_2 = -\frac{h_{12}h_{21}}{R_s + h_{11}} V_2 + h_{22}V_2 = V_2\left(h_{22} - \frac{h_{12}h_{21}}{R_s + h_{11}}\right)$$

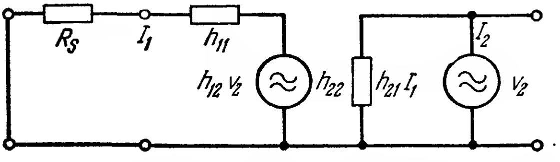


Fig. 4.2. Equivalent circuit for determining the output resistance of a transistor amplifier

Simplifying,

$$I_2 = V_2(h_{22}R_s + \Delta_h)/(R_s + h_{11})$$

whence,

$$V_2/I_2 = R_{out} = (R_s + h_{11})/(h_{22}R_s + \Delta_h) \tag{4.13}$$

Power Gain. By definition,

$$K_p = P_2/P_1$$

or in terms of currents and voltages

$$K_p = P_2/P_1 = I_2V_2/I_1V_1 = K_iK_v$$

Substituting the expressions for K_i and K_v from Eqs. (4.4) and (4.8) gives

$$K_p = K_iK_v = \frac{h_{21}}{1+h_{22}R_L}\,\frac{h_{21}R_L}{h_{11}+\Delta_hR_L} = \frac{h_{21}^2R_L}{(1+h_{22}R_L)\,(h_{11}+\Delta_hR_L)} \tag{4.14}$$

When R_s is perfectly matched to R_{in}, and R_{out} to R_L, power output is a maximum, and the power gain is given by

$$K_{p\,max} = \frac{h_{21}^2}{(\sqrt{\Delta_h} + \sqrt{h_{11}h_{22}})^2}$$

Instead of the h-parameter equivalent circuit, a single-stage transistor amplifier may be analysed by use of equivalent circuits drawn up in terms of the z- and the y-parameters. The relationships between the three sets of parameters are summarized in Table 4.1

TABLE 4.1

	h-parameters	z-parameters	y-parameters
Input resistance	$\frac{h_{11}+R_L\Delta_h}{1+h_{22}R_L}$	$\frac{\Delta_R+R_{11}R_L}{R_{22}+R_L}$	$\frac{y_{22}+y_L}{\Delta_y+y_{11}y_L}$
Output resistance	$\frac{R_s+h_{11}}{h_{22}R_s+\Delta_h}$	$\frac{\Delta_R+R_{22}R_s}{R_{11}+R_s}$	$\frac{y_{11}+y_s}{\Delta_y+y_{22}y_s}$
Current gain	$\frac{h_{21}}{1+h_{22}R_L}$	$\frac{R_{21}}{R_{22}+R_L}$	$\frac{y_{21}y_L}{\Delta_y+y_{11}y_L}$
Voltage gain	$\frac{h_{21}R_L}{h_{11}+\Delta_hR_L}$	$\frac{R_{21}R_L}{\Delta_R+R_LR_{11}}$	$\frac{y_{21}}{y_{22}+y_L}$

where

$$\Delta_h = h_{11}h_{22} - h_{12}h_{21}$$
$$\Delta_R = R_{11}R_{22} - R_{12}R_{21}$$
$$\Delta_y = y_{11}y_{22} - y_{12}y_{21}$$
$$y_s = 1/R_s$$
$$y_L = 1/R_L$$

Review Questions

1. Which of the four-terminal parameters of the transistor are the easiest to measure?
2. How can the h-parameters be expressed in terms of the z-parameters?
3. How can the z-parameters be expressed in terms of the h-parameters?
4. Explain why the input resistance of the transistor depends on the load resistance.
5. Explain why the output resistance of the transistor depends on the source resistance.

22. Transistor Arrangements

The numerical values of the h- and other parameters of a transistor depend on the way it is connected into the circuit. The three transistor arrangements of practical importance are *common-base*, *common-emitter*, and *common-collector*.

Common-base Circuit. A simplified schematic and an equivalent circuit of a transistor arranged into the common-base configuration appear in Fig. 4.3.

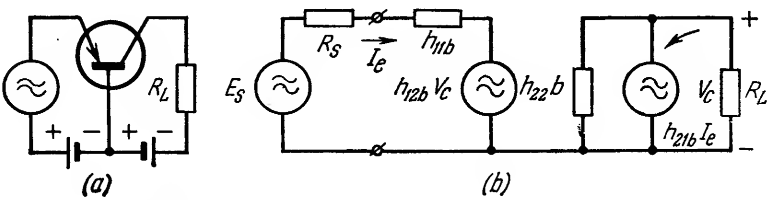


Fig. 4.3. Common-base amplifier
(a) schematic; (b) equivalent circuit

As is seen, the equivalent circuit of Fig. 4.3*b* is similar to that of Fig. 4.1. Therefore, noting that

$$I_1 = I_e$$
$$I_2 = I_c$$
$$V_2 = V_c$$
$$h_{11} = h_{11b}$$
$$h_{22} = h_{22b}$$
$$h_{12} = h_{12b}$$
$$h_{21} = h_{21b} = -\alpha$$

we may use the relationships already derived. The subscript "*b*" in the *h*-parameters indicates that they apply to the common-base circuit. The quantity $h_{21b} = -\alpha$ is the static (d.c.) amplification factor for the emitter-to-collector current. The stage current gain then is

$$K_i = I_c/I_e = h_{21b}/(1 + R_L h_{22b}) \tag{4.15}$$

or

$$K_i = -\alpha/(1 + R_L h_{22b}) \tag{4.16}$$

The "—" sign indicates that the output current is in anti-phase with the input current.

The input resistance of the stage is given by

$$R_{in} = \frac{h_{11b} + R_L \Delta_{hb}}{1 + R_L h_{22b}} \tag{4.17}$$

The voltage gain is

$$K_v = \frac{h_{21b} R_L}{h_{11b} + \Delta_{hb} R_L} \tag{4.18}$$

or

$$K_v = -\frac{\alpha R_L}{h_{11b} + \Delta_{hb} R_L} \tag{4.19}$$

The output resistance is

$$R_{out} = \frac{R_s + h_{11b}}{h_{22b} R_s + \Delta_{hb}} \tag{4.20}$$

The power gain is given by

$$K_p = \frac{h_{21b}^2 R_L}{(1 + h_{22b} R_L)(h_{11b} + \Delta_{hb} R_L)} \tag{4.21}$$

or

$$K_p = \frac{\alpha^2 R_L}{(1 + h_{22b} R_L)(h_{11b} + \Delta_{hb} R_L)} \tag{4.22}$$

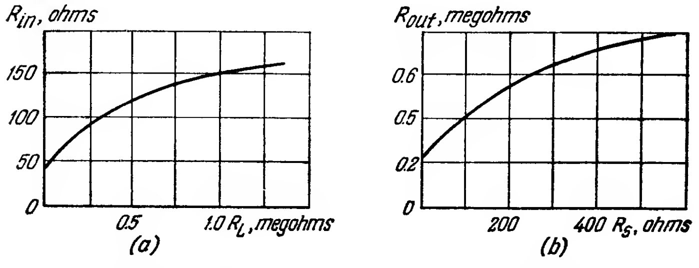


Fig. 4.4. Input resistance as a function of load resistance (a), and output resistance as a function of source resistance (b) in common-base circuit

The input resistance of the common-base circuit is low, but it increases with increase of the load resistance in the collector circuit. The output resistance is usually several hundred kilohms. Plots of R_{in} and R_{out} as functions of R_L and R_s, respectively, are shown in Fig. 4.4.

The current gain of the common-base configuration is always less than unity. The voltage gain may be several hundreds, if the load resistance (the one in the collector lead) is sufficiently high.

Common-emitter Circuit. A simplified schematic and an equivalent circuit of the common-emitter arrangement are shown in Fig. 4.5.

The equivalent circuit of Fig. 4.5*b* is similar to that of Fig. 4.1. Therefore, we may safely use the relationships already established. However, the numerical values are different, and an additional subscript, *e*, is therefore added to the parameter symbols.

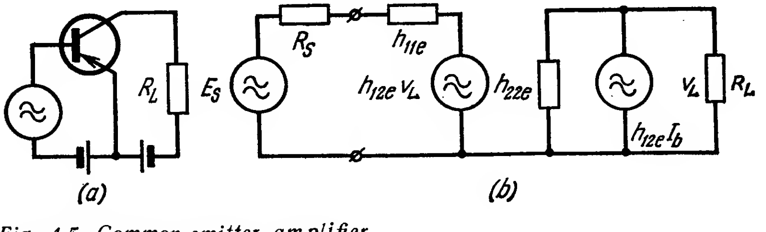


Fig. 4.5. Common-emitter amplifier
(a) schematic; (b) equivalent circuit

If the h_b-parameters of a transistor are quoted in the data sheet, its h_e-parameters can be derived by use of the following approximate equations:

$$h_{11e} \cong h_{11b}/(1-\alpha)$$
$$h_{12e} \cong (\Delta_{hb} - h_{12b})/(1-\alpha)$$
$$h_{21e} \cong \alpha/(1-\alpha) = \beta$$
$$h_{22e} \cong h_{22b}/(1-\alpha)$$

where β is the static (d.c.) short-circuit current gain or the current amplification factor for the common-emitter circuit.

The design relationships applying to the common-emitter configuration are summarized in Table 4.2.

TABLE 4.2

Current gain	$K_i = \dfrac{h_{21e}}{1 + R_L h_{22e}}$
Input resistance	$R_{in} = \dfrac{h_{11e} + R_L \Delta_{he}}{1 + R_L h_{22e}}$
Voltage gain	$K_v = \dfrac{h_{21e} R_L}{h_{11e} + \Delta_{he} R_L}$
Output resistance	$R_{out} = \dfrac{R_s + h_{11e}}{h_{22} R_s + \Delta_{he}}$
Power gain	$K_p = \dfrac{h_{21e}^2 R_L}{(1 + h_{22e} R_L)(h_{11e} + \Delta_{he} R_L)}$

The amplification factor beta is always much greater than unity. The input resistance of the common-emitter stage is markedly higher than that of the common-base arrangement, but the output resistance is lower. Plots of R_{in} and R_{out} as functions of R_L and R_s, respectively, appear in Fig. 4.6.

The power gain of the common-emitter configuration is noticeably greater than that of the common-base arrangement.

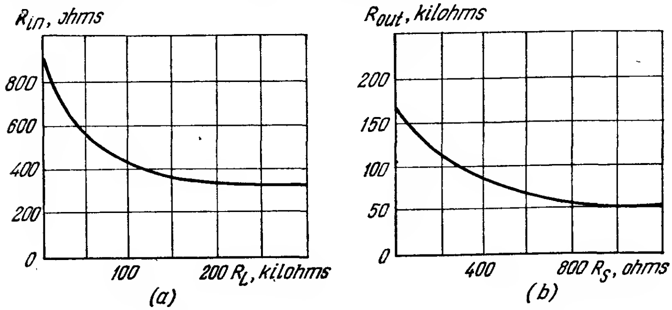


Fig. 4.6. Plots of R_{in} vs. R_L (a), and of R_{out} vs. R_s (b) in common-emitter circuit

Common-collector Circuit. The simplest arrangement of a transistor used as an amplifier with a resistive load is shown in Fig. 4.7 along with its equivalent circuit. As is seen, the equivalent circuit is developed in terms of the h_c-parameters which are related to the h_b-parameters as follows:

$$h_{11c} \cong h_{11b}/(1-\alpha)$$
$$h_{12c} \cong 1$$
$$h_{21c} \cong -1/(1-\alpha)$$
$$h_{22c} = h_{22b}/(1-\alpha)$$

The design relationships describing the common-collector arrangement are tabulated in Table 4.3.

The current gain K_i of the stage differs little from that of the common-emitter configuration. The voltage gain is less than unity. The input resistance of the common-collector circuit is very high and may be hundreds of kilohms (Fig. 4.8*a*). The output resistance is very low, being anywhere between tens and hundreds of ohms (Fig. 4.8*b*).

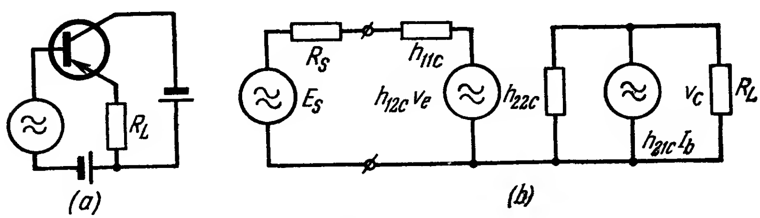


Fig. 4.7. Common-collector amplifier
(a) schematic; (b) equivalent circuit

TABLE 4.3

Current gain	$K_i = \dfrac{h_{21c}}{1 + R_L h_{22c}}$
Input resistance	$R_{in} = \dfrac{h_{11c} + R_L \Delta_{hc}}{1 + R_L h_{22c}}$
Voltage gain	$K_v = \dfrac{h_{21c} R_L}{h_{11c} + \Delta_{hc} R_L}$
Output resistance	$R_{out} = \dfrac{R_s + h_{11c}}{h_{22c} R_s + \Delta_{hc}}$
Power gain	$K_p = \dfrac{h_{21c}^2 R_L}{(1 + h_{22c} R_L)(h_{11c} + \Delta_{hc} R_L)}$

The approximate values of the input and output resistance for a single-stage amplifier using a transistor connected into the common-collector circuit are given by

$$R_{in} = R_L (1 + \beta) \tag{4.23}$$

$$R_{out} = R_s/(1 + \beta) \tag{4.24}$$

The common-collector circuit has the lowest power gain of all.

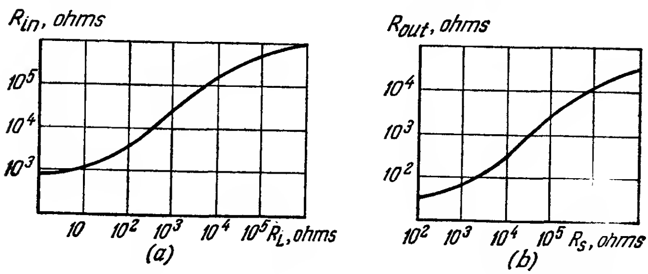


Fig. 4.8. Plots of R_{in} vs. R_L (a), and of R_{out} vs. R_s (b) in common-collector circuit

To sum up, the comparative advantages and disadvantages of the three transistor arrangements are as follows:

1. The lowest input and the highest output resistance is offered by the common-base circuit.
2. The highest input and the lowest output resistance is offered by the common-collector configuration.
3. The highest power gain is secured by the common-emitter arrangement.

Review Questions

1. Explain why the current gain in the common-base circuit is smaller than unity.
2. Explain why the connection of the transistor into the common-base circuit can raise the upper limiting frequency of the amplifier.
3. Why is the common-emitter configuration used more widely than the common-base circuit?
4. Why is it that in the common-collector circuit the voltage gain is smaller than unity?
5. How do you explain that the common-collector circuit has a considerable input resistance?

23. Multi-stage Transistor Amplifiers

When a single-stage amplifier is inadequate, use is made of multi-stage amplifier circuits.

While in valve amplifiers the number ot stages is determined by the relative magnitudes of the voltage necessary to drive the valves in the final stage and the signal voltage, in transistor circuits the number of stages is decided by the relative magnitudes of power output, P_{out}, and the power delivered by the signal source to the input of the first amplifying stage, P_{in1}. The requisite power gain, in decibels, is then given by

$$K_p = 10 \log_{10} (P_{out}/P_{in1})$$

The power gains of the individual stages may range between 15 and 30 db, according to transistor type and stage configuration.

Approximately, the number of amplifying stages in preliminary design may be obtained from

$$N = 10 \log_{10} (P_{out}/P_{in1})/(15 \text{ to } 30)$$

For a two-stage amplifier, the three transistor arrangements may theoretically lead to nine amplifier configurations, but only some of them are used in practice. As an example, with both stages arranged into the common-base circuit, it is difficult to match the high output resistance of the preceding stage to the low input resistance of the succeeding stage. Inter-stage matching is most conveniently obtained with the common-emitter and common-collector configurations. The common-emitter circuit has the additional advantage of high gain. This is the reason why the common-emitter circuit is used most. The common-base arrangement is employed mainly when it is desired to extend the bandwidth in the direction of the higher frequencies. The common-collector circuit is resorted to when a signal source of high output resistance has to be matched to a load of low input resistance.

The stages may be either *RC* or transformer coupled. *RC*-coupling is mainly used in preamplifiers; transformer coupling is most often found between the penultimate and final stages and also between the final stage and the load.

RC-coupled Amplifiers. In *RC*-coupled amplifiers, the transistor is most often connected into the common-emitter circuit because this arrangement provides for the highest signal amplification in power.

Unfortunately, a common-emitter amplifier suffers from temperature instability. This is mainly due to the fact that the collector-junction leakage current I_{CBO} and the current amplification factor are extremely temperature-sensitive. If not adequately controlled, temperature variations in these parameters might cause a shift in the *Q*-(operating) point of the circuit, with the result that it would not be able to amplify and reproduce the signal with a minimum of distortion.

There are several methods for improving the *Q*-point (bias) stability by use of negative direct-current feedback and suitable biasing arrangements. One very popular method of improving bias stability is to add some external emitter resistor as shown in Fig. 4.9*a* in conjunction with a bias-voltage divider R_1R_2.

In this circuit, the *Q*-point will be positioned by the base-to-emitter voltage v_{be} which is numerically equal to the difference in voltage drop between R_2 and R_e

$$v_{be} = vR_2 - vR_e$$

Should variations in temperature cause an increase in the base current, a greater current will flow through R_1 and R_e.

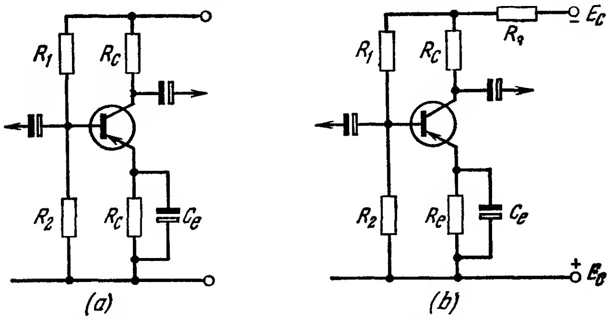


Fig. 4.9. RC-coupled transistor amplifiers
(a) with bias network and negative series feedback; (b) with bias network and series-parallel negative feedback

An increase in the current through the bias-voltage divider will bring about an increase in the voltage drop across R_1, and the d. c. voltage applied to the base will be brought down, thereby decreasing the base current. An increase in the voltage drop across R_e will bring down v_{be}, and this, too, will lead to a decreased base current.

Figure 4.9*b* shows a stabilization arrangement using series and parallel negative feedback. As the base current and, consequently, the collector current increases, the voltage drop across R_3 also increases, and the voltage drop across the bias-voltage divider R_1R_2 is brought down along with v_{be}.

The stability factor S of the circuit shown in Fig. 4.9*a* is given by

$$S = 1 + \frac{\alpha R_2}{(1-\alpha) R_2 + R_e'} \tag{4.25}$$

Since the amplification factor alpha for existing transistors used in amplifiers ranges between 0.96 and 0.98, Eq. (4.25) may be re-written thus:

$$S = (\text{approx.})\ 1 + R_2/R_e'$$

where

$$R_e' = (R_e/R_1)(R_1 + R_2)$$

As is seen, the value of S decreases as R_2 decreases and R_e increases.

It should be remembered, however, that R_1R_2 acts as an additional load on the collector circuit of the preceding stage. With decrease of R_1 and R_2, the effective load resistance of the preceding stage will be brought down, and the gain of that stage will be impaired.

The parameters of the common-emitter circuit are more frequency-dependent than they are in the common-base configuration. Among other things, the *beta cut-off frequency* f_T, defined as the frequency at which the common-emitter current amplification becomes equal to unity, decreases while the capacitance C_{ce} shunting the load increases.

The beta cut-off frequency is given by

$$f_T = f_\alpha/(1+\beta)$$

and the output capacitance,

$$C_{ce} = C_{cb}(1+\beta)$$

The decrease in f_T and the increase in C_{ce} result in that the current gain in the high-frequency range of the band is impaired, and the frequency response falls off

$$\beta_f = \beta_0/\sqrt{1+(f/f_T)^2}$$

Consider the effect of the various circuit elements in the two-stage RC-coupled amplifier of Fig. 4.10 on its gain. The equivalent circuit of the first stage is shown in Fig. 4.11*a* where:

$I_{b1}K_i$ = constant-current generator equivalent to the transistor T_1
R_{out} = output resistance of T_1
R_{c1} = resistance in the collector lead of T_1
R_{in2} = input resistance of T_2
R_{12} = equivalent resistance of the bias network

$$R_{12} = R_1R_2/(R_1+R_2)$$

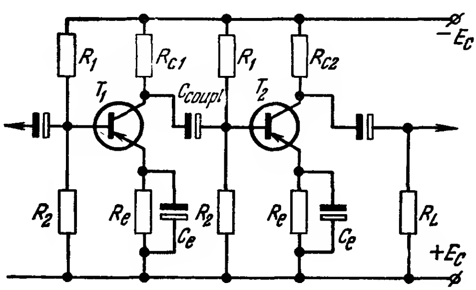


Fig. 4.10. Two-stage RC-coupled transistor amplifier

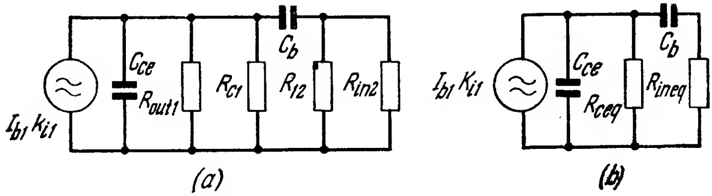


Fig. 4.11. Equivalent circuit for the first stage
(a) full; (b) reduced

C_{coupl} = coupling capacitor
C_{ce} = collector-emitter (shunt) capacitance.

The equivalent circuit of Fig. 4.11*a* can be reduced to that of Fig. 4.11*b* where

$$R_{ceq} = R_{out}R_c/(R_{out} + R_c)$$
$$R_{in2eq} = R_{in2}R_{12}/(R_{in2} + R_{12})$$

The equivalent circuit of Fig. 4 11*b* is similar to that of a valve single-stage amplifier. Therefore, we may use the same method of analysis. Figure 4.12 shows equivalent circuits for the low, middle, and high frequencies of the band.

Consider the parameters of the stage at the mid-band frequencies.

The current gain is given by

$$K_{i0} = h_{21e}/(1 + R_{Leq}h_{22e}) \tag{4.26}$$

The input resistance is

$$R_{in0} = (h_{11e} + R_{Leq}\Delta_{he})/(1 + R_{Leq}h_{22e}) \tag{4.27}$$

The voltage gain is

$$K_v = (h_{21e}R_{Leq})/(h_{11e} + \Delta_{he}R_{Leq}) \tag{4.28}$$

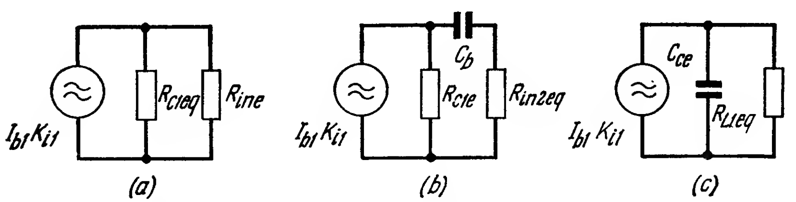


Fig. 4.12. Equivalent circuits for the first stage
(a) mid-band; (b) low-frequency; (c) high-frequency

The output resistance is

$$R_{out}=(R_s+h_{11e})/(h_{22e}R_s+\Delta_{he}) \tag{4.29}$$

The power gain is

$$K_{p0}=\frac{h_{21e}^2 R_{Leq}}{(1+h_{22e}R_{Leq})(h_{11e}+\Delta_{he}R_{Leq})} \tag{4.30}$$

where R_{Leq} is the equivalent load resistance of the first stage, defined as

$$R_{Leq}=R_{c1eq}R_{in2eq}/(R_{c1eq}+R_{in2eq})$$

The frequency distortion in the low-frequency range is

$$M_l=\sqrt{1+\left[\frac{1}{\Omega_l C_{coupl}(R_{c1eq}+R_{in2eq})}\right]^2} \tag{4.31}$$

The high-frequency distortion due to C_{ce} is given by

$$M_h=\sqrt{1+(\Omega_h R_{Leq}C_{ce})^2} \tag{4.32}$$

The frequency distortion caused by variations in the properties of the transistor with frequency is allowed for by use of an adjusted value of h_{21e}

$$h_{21e}=\beta_f=\beta_0/\sqrt{1+(\Omega_h/\Omega_T)^2}$$

Design of an *RC*-coupled Transistor Amplifier (Fig. 4.9).

Given:

1. Supply voltage, v_{sup}.
2. Peak value of the load voltage equal to that of the input voltage to the next stage, $V_L=V_{in}$.
3. Peak value of the load current equal to that of the input current adjusted for the current in the bias network of the next stage, $I_L=I'_{in}$.
4. Frequency range, F_l-F_h.
5. Limits of frequency distortion at the low and high frequencies of the band, M_l and M_h.
6. Signal voltage E_s and signal-source resistance R_s.

To Find:

1. Transistor type.
2. D. c. (no-signal) operating conditions.
3. Parameters of the bias stabilization network.

4. Stage parameters, such as current gain, input resistance, voltage gain, and output resistance.
5. Coupling and bypass capacitors, C_{coupl} and C_e.

Design Procedure:

1. Select the transistor type so that the maximum rated collector-to-emitter voltage will be higher than the supply voltage

$$v_{ce\ rated} > v_{sup}$$

The beta cut-off frequency should be higher than the upper desired frequency

$$f_T > F_h$$

2. Select the minimum collector current such that

$$I_{c\ min} = \text{(approx.) } 5 \text{ to } 10\ I_{co\ (rated)}$$

and such that

$$I_c \geqslant I_L + I_{c\ min}$$

If the peak value of the load current is small, the quiescent collector current should be chosen close to the value given in the data sheet.

3. Select the minimum collector-to-emitter voltage

$$v_{ce\ min} \geqslant 0.8 \text{ to } 1 \text{ volt}$$

4. Determine the collector-to-emitter voltage

$$v_{ce} = v_{ce\ min} + V_L$$

If the load voltage is low, v_{ce} should be taken close to that specified in the data sheet.

5. From the transistor characteristic, find the base current I_b (Fig. 4.13). If the transistor characteristics are not available, I_b can be found from

$$I_b = \text{(approx.) } I_{c\ (dc)}/\beta_0$$

6. Select the voltage across R_e:

$$v_{R_e} = (0.15 \text{ to } 0.2)\, v_{sup}$$

7. Determine the load resistance in the collector lead:

$$R_c = (v_{sup} - v_{ce} - v_{R_e})/I_{c\ (dc)}$$

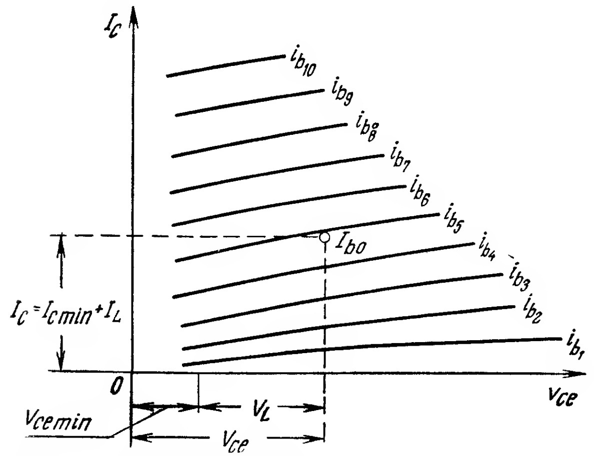


Fig. 4.13. To calculation of base current from transistor characteristics

8. To allow for the effect of R_c on the peak alternating current at the stage output, find the adjusted value of the collector current:

$$I_{c\,(\mathrm{dc})} = I_{c\,min} + \frac{V_L (R_c + R_L)}{R_c R_L}$$

9. Determine the resistance in the emitter lead:

$$R_e = v_{R_e}/I_c$$

10. Select the bias current:

$$I_{12} = (3 \text{ to } 5)\, I_b$$

11. Determine the resistance of R_2 in the stabilization network:

$$R_2 = (v_{R_e} + v_{eb})/I_{12} = (\text{approx.})\,(v_{R_e} + 0.2)/I_{12}$$

12. Calculate R_1 in the stabilization network:

$$R_1 = (v_{sup} - v_{R_e} - v_{eb})/(I_{12} + I_b)$$

13. Calculate the stability factor:

$$S = (\text{approx.})\,(1 + R_2)/R'_e \leqslant 5 \text{ to } 8$$

where

$$R'_e = (R_e/R_1)\,(R_1 + R_2)$$

If S is too high, specify a larger current for the stabilization network.

14. Calculate the equivalent load resistance in the collector circuit:

$$R_{Leq} = R_c R_L (R_c + R_L)$$

where

$$R_L = V_L / I_L$$

15. Using design Equations (4.26), (4.27), (4.28) and (4.29), find the current gain K_i, the input resistance R_{in}, the voltage gain K_v, and the output resistance R_{out}.

16. Calculate the capacitance of the coupling capacitor:

$$C_{coupl} = \frac{1}{\Omega_l (R_{ceq} + R_L) \sqrt{M_l^2 - 1}}$$

17. Find the frequency distortion in the high-frequency range of the band:

$$M_h = \sqrt{1 + (\Omega_h R_{Leq} C_{ce})^2}$$

where

$$R_{Leq} = R_{ceq} R_L / (R_{ceq} + R_L)$$

18. Determine the capacitance of the bypass capacitor C_e:

$$C_e \geqslant \frac{1}{\Omega_l R_\Sigma \sqrt{M_{le}^2 - 1}}$$

where M_{le} is the limit of frequency distortion due to the emitter circuit ($M_{le} = 1.01$ to 1.02)

$$R_\Sigma \cong R_{in} + R_s$$

Transformer-coupled Amplifier. Transformer coupling is used with a view to securing high power gain.

The point is that for maximum power output it is essential that R_s, the source resistance, be matched to the input resistance of the stage, and the output resistance of the stage be matched to the load resistance, that is

$$R_s = R_{in}$$
$$R_{out} = R_L$$

On the other hand, the output resistance of a common-base or a common-emitter stage is ordinarily greater than the input resistance. This is where a step-down coupling transformer comes in.

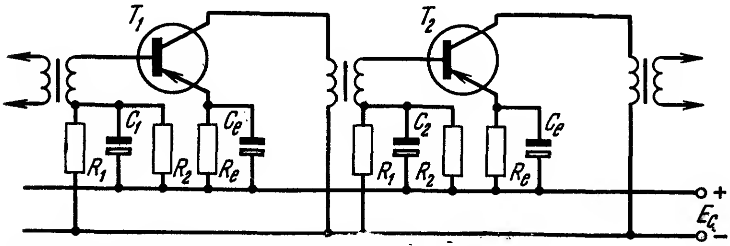


Fig. 4.14. Transformer-coupled transistor amplifier

Transformer coupling may be used in both common-base and common-emitter circuits. Practically, use is mostly made of the common-emitter configuration, such as shown in Fig. 4.14.

In practice, transformer coupling is also used to match the penultimate stage of the a. f. amplifier into a single-ended or push-pull final stage.

Consider operation of a transformer-coupled two-stage amplifier. The resistance of the transformer primary to direct current is low, about several hundred ohms. Therefore, the voltage drop across the primary is a few tenths of a volt, and the collector circuit may operate on a lower supply voltage.

With a reduced d. c. voltage drop across the collector load, it is possible to use a greater value for the stabilizing resistor R_e in the emitter lead. The increase in R_e improves the stability of the stage and the reliability of the amplifier as a whole.

Among the disadvantages of a transformer-coupled amplifier are higher cost, larger size and heavier weight. The size of the transformer cannot often be reduced because it might lead to core saturation and associated non-linear distortion.

An equivalent circuit of the output stage of a transformer-coupled amplifier, referred to the primary, is shown in Fig. 4.15 where

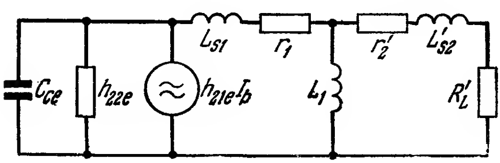


Fig. 4.15. Equivalent circuit of a transformer-coupled transistor amplifier

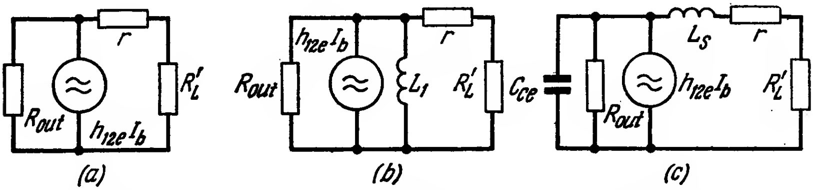


Fig. 4.16. Equivalent circuits for transformer-coupled transistor amplifier (a) at mid-band frequencies; (b) at low frequencies; (c) at high frequencies

L_1 = primary inductance
r = primary and secondary series loss resistance, defined as

$$r = r_1 + r_2/n^2$$

L_s = primary and secondary leakage inductances referred to the primary, defined as

$$L_s = L_{s1} + L_{s2}/n^2$$

R'_L = load resistance referred to the primary, such that

$$R'_L = R_L/n^2$$

At the mid-band frequencies, the effect of the reactive elements may be neglected. At the low frequencies of the band, the leakage inductance and the shunt capacitance C_{ce} may also be neglected, while the inductive reactance has a marked effect. In the high-frequency range, operation of the stage will be markedly affected by the shunt capacitance C_{ce} and the leakage inductances L_s.

Equivalent circuits for the various frequencies of the band are shown in Fig. 4.16. The one for the mid-band frequencies is similar to the equivalent circuit of an RC-coupled amplifier. Therefore, the parameters may be calculated from the equations already derived.

In the low-frequency range of the band, some of the current from the constant-current generator will branch off through L_1, so that the load current and voltage will be reduced, thereby decreasing current and voltage gain. Because of this, the frequency response will fall off at the low frequencies. The frequency distortion occurring at those frequencies is given by

$$M_l = \sqrt{1 + R_{ceq}/\Omega_l L_1)^2} \tag{4.33}$$

where

$$R_{ceq} = R_{out}R'_L/(R_{out} + R'_L)$$

In the high-frequency range of the band, the leakage inductances begin to be felt. As the frequency increases, the voltage drop across the reactance ΩL_s goes up, and the load voltage goes down. The frequency distortion occurring at the higher frequencies due to L_s is given by

$$M'_h \cong \sqrt{1 + \left(\frac{\Omega_h L_s}{R_{out} + R'_L}\right)^2} \qquad (4.34)$$

At the high frequencies of the band, the shunt capacitance C_{ce} may also introduce frequency distortion. With increase of frequency, part of the current from the constant-current generator will branch off through the reactance of C_{ce}, decreasing the load current. The frequency distortion due to this cause is given by

$$M''_h = \sqrt{1 + (\Omega_h C_{ce} R_{ceq})^2} \qquad (4.35)$$

If M''_h is too high, the value of R'_L must be decreased. The overall frequency distortion in the high-frequency region is given by

$$M_h = M'_h M''_h$$

Review Questions

1. Why is *RC* coupling used in practical amplifiers most often?
2. How can the input resistance of a transistor amplifier be raised?
3. How can the polarity of the coupling electrolytic capacitor be determined in the common-emitter circuit?
4. Why cannot two adjacent stages in an amplifier be both common-collector circuits?
5. Why is transformer coupling used between the penultimate and final stages most of all?

24. Transistor Power Amplifiers

The final stages of receivers may be connected either single-ended or push-pull.

Single-ended configurations are mainly used in small-size receivers with a power output of 50 to 100 milliwatts. If the

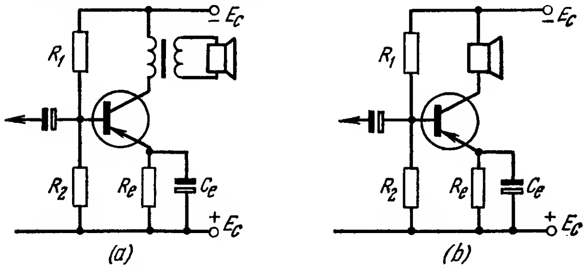


Fig. 4.17. Power amplifiers (final stages)
(a) transformer-coupled output; (b) direct-coupled speaker in the collector lead

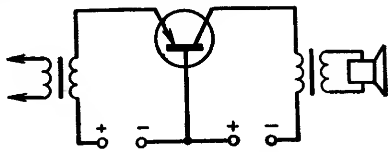


Fig. 4.18. Common-base power amplifier

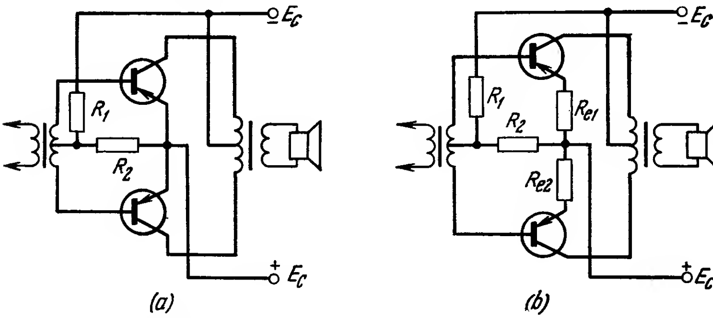


Fig. 4.19. Push-pull power amplifiers
(a) with bias network; (b) with bias network and series negative feedback

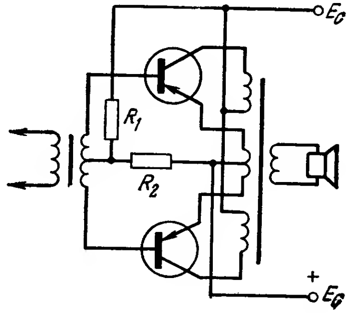


Fig. 4.20. Push-pull power amplifier with the emitter circuit connected to a tapping on the output transformer

impedance of the speaker differs markedly from the optimum load impedance, a matching transformer is provided at the output (Fig. 4.17). Speakers of a sufficiently high impedance may be directly connected to the collector circuit.

Figure 4.18 shows a common-base power amplifier.

Transistor push-pull power amplifiers come in a great variety of circuit configurations. Two of them are shown in Fig. 4.19. The one in (*b*) uses negative current feedback to reduce distortion.

In the amplifier of Fig. 4.20, the emitter circuit is connected to a tapping on the output transformer. The negative current feedback effected in this arrangement minimizes both frequency and non-linear distortion.

Figure 4.21 shows a common-collector push-pull amplifier. This configuration is often used in cases where the output characteristics of the transistors are rather non-linear.

As with single-ended circuits, the trend is likewise towards the use of circuits with direct-connected loads. Omission of the matching transformer minimizes frequency distortion in the low-frequency range. In one such circuit (Fig. 4.22) the power supply has a centre tap (or there may be two power supplies). Figure 4.23 shows a push-pull amplifier with a single supply source and a speaker connected through a high-capacitance capacitor.

Bias stabilization is effected by provision of a voltage divider in the base circuit and of an external resistance in the emitter lead. Sometimes the divider is extended to include a thermistor (Fig. 4.24) to provide temperature compensation.

In the circuit of Fig. 4.25 temperature compensation is accomplished by crystal diodes. Another distinction of this circuit is that there is no centre tap on the input matching transformer. When the upper diode (in the diagram) is conducting, the lower one is cut off, and the whole of the secondary voltage appears across the upper arm.

As an example, consider the procedure for the calculation of a common-emitter power amplifier (Fig. 4.17*a*).

Given:

1. Power output, P_{out}.
2. Load resistance, R_L.
3. Lowest desired frequency, F_l.
4. Limit of frequency distortion, M_l.

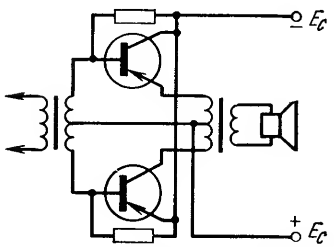


Fig. 4.21. Common-collector push-pull amplifier

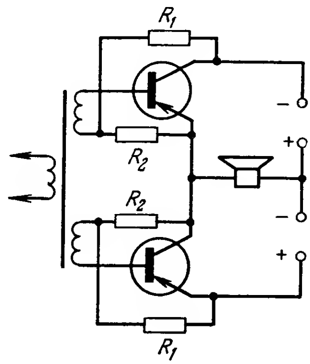


Fig. 4.22. Push-pull amplifier with direct-connected load and two supply sources

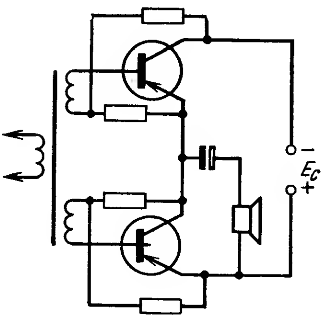


Fig. 4.23. Push-pull power amplifier with direct-connected load and one supply source

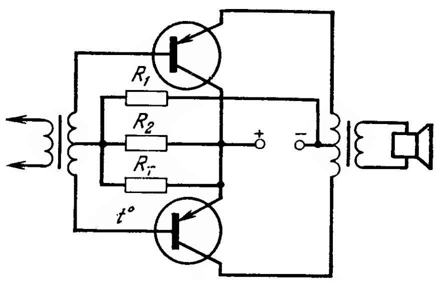


Fig. 4.24. Push-pull power amplifier with a thermistor in the bias network

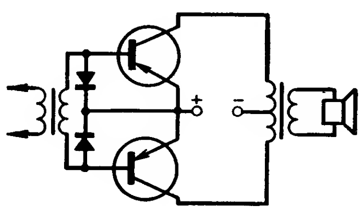


Fig. 4.25. Push-pull power amplifier with temperature-compensating diodes

To Find:

1. Transistor type.
2. Operating conditions in the collector circuit.
3. Collector load resistance, R_c.
4. Base current and input voltage.
5. Input resistance of the stage and the power dissipated by the coupling circuit.
6. Non-linear distortion, γ.
7. Primary inductance, L_1.
8. Transformation ratio of the output transformer, n.

Design Procedure:

1. The available power, that is, one to be delivered by the stage

$$P_{st} = \frac{P_{out}}{\eta_T}$$

2. From considerations of the power handling capability of the stage, select the transistor type

$$P_{rated} \geqslant P_{av}$$

and

$$v_{c\,rated} \geqslant 2E_{sup},$$

if the supply voltage is specified in advance.

3. Select the quiescent-point collector voltage

$$v_{c0} < v_{c\,max}/2$$

4. The output characteristics of junction transistors are linear. Therefore the collector voltage efficiency may be raised to about unity and the collector current efficiency β may be brought down to 0.01-0.02. Then, the efficiency of the stage will be very near to 0.5.

The efficiency of the stage is usually specified between 0.45 and 0.48. As the supply voltage is reduced, the efficiency goes down to 0.3-0.4. Once the efficiency has been determined, the power dissipated by the collector circuit is found:

$$P_{diss} = P_{st}/\eta$$

which is a maximum in the no-signal (quiescent) state.

5. Determine the quiescent collector current

$$I_{c0} = P_{diss}/v_{c0}$$

6. Determine the approximate value of the collector load resistance

$$R_c = v_{c0}/I_{c0} = 0.4v_{c0}^2/P \text{ (approx.)}$$

7. Plot the dynamic load line of the transistor from its static characteristics. The load line should pass through points with coordinates (v_{c0}, I_{c0}) and ($0.1v_{c0}$, $I_{c0}+I_c$).

8. Determine the base current in the quiescent state, note it for several values of the collector voltage, transfer the values thus found onto the static input characteristics and plot the dynamic input characteristic. If instead of a family of input characteristics the data sheet gives only a typical curve for a specified collector voltage, use may in a first approximation be made of one static input characteristic.

9. Determine E_{b0}, $2I_b$, and $2V_{be}$ graphically from the dynamic input characteristic.

10. Calculate the average input resistance

$$R_{in} = 2V_{be}/2I_b$$

11. Select the current for the bias stabilization network

$$I_{12} = (2 \text{ or } 3)\, I_{b0}$$

12. Calculate R_2 and R_1

$$R_2 = E_{b0}/I_{12}$$

$$R_1 = (v_{c0} - E_{b0})/(I_{12} + I_{b0})$$

If the emitter lead contains an external feedback resistor R_e, calculate R_1 and R_2 from the equations

$$R_2 = (E_{b0} + v_e)/I_{12}$$

$$R_1 = \text{(approx.)}\,(v_{c0} - E_{b0} - v_e)/(I_{12} + I_{b0})$$

13. Determine the equivalent input resistance

$$R_{in\,eq} = R_{in}R_{12}/(R_{in} + R_{12})$$

where

$$R_{12} = R_1R_2/(R_1 + R_2)$$

14. Determine the power dissipated by the coupling circuit:

$$P_{in} = (2V_{be})^2/8R_{in\,eq}$$

15. To determine the non-linear distortion due to the coupling circuit, plot a transfer (mutual) characteristic, that is, one relating the collector current to the signal voltage E_s. The need for the transfer characteristic arises because some of the

signal voltage will inevitably be dropped across the internal resistance R_s of the signal source. Hence,

$$E_s = V_{be} + I_b R_s$$

and the input voltage will be dependent on the base current so that the dynamic (a.c.) characteristic will be curved (non-linear).

The transfer characteristic $I_c = F(E_s)$ is calculated as follows. The source resistance is specified in advance to be anywhere between 50 and 300 ohms, and the signal source voltage is calculated for various design values of I_b and I_c

$$E_{s1} = V_{be1} + I_{b1} R_s$$
$$E_{s2} = V_{be2} + I_{b2} R_s$$
$$\cdots\cdots\cdots$$
$$\cdots\cdots\cdots$$

Once the $I_c = F(E_s)$ curve is plotted, determine the collector-current harmonics by any of the graphical methods and find non-linear distortion (harmonic content):

$$\gamma = \frac{\sqrt{I_{c2}^2 + I_{c3}^2 + I_{c4}^2}}{I_{c1}}$$

16. Calculate the inductance of the transformer primary

$$L_1 = \frac{R_{eq}}{2\pi F_l \sqrt{M_l^2 - 1}} = \frac{R_c}{2\pi F_l \sqrt{M_l^2 - 1}} \text{ (approx.)}$$

Since $R_{out} \gg R_c$, the equivalent resistance may be taken equal to the load resistance

$$R_{eq} = \text{(approx.) } R_c$$

17. Calculate the transformation ratio of the output transformer

$$n = \sqrt{R_L / R_c \eta_T}$$

Example 4.1. Calculate a transistor power amplifier (the final stage), if $P_{out} = 1$ watt, $R_L = 5$ ohms, $F_l = 100$ hertz, and $M_l = 1.1$.

Solution.

1. Putting $\eta_T = 0.7$, the power delivered by the stage should be

$$P_{st} = P_{out} / \eta_T = 1 \div 0.7 = 1.41 \text{ watts}$$

2. From a transistor data manual we select a П201А power transistor whose parameters are: $v_{c\,max} = 30$ volts, $I_{c\,max} = 1.5$ amperes, rated collector dissipation without a heat-sink $P_{c\,rated} = 1$ watt,

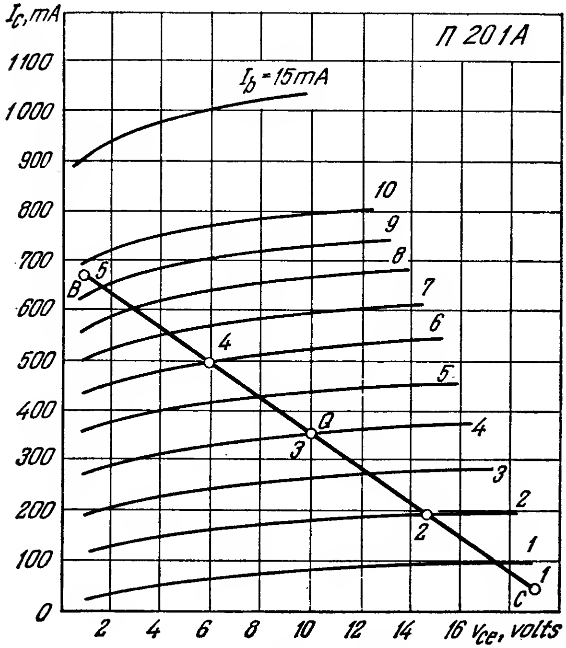


Fig 4.26. Output characteristics of Soviet-made П201А transistor

and with a heat-sink = 10 watts. The output and input characteristics are shown in Figs. 4.26 and 4.27.

3. Select the quiescent collector voltage

$$v_{c0} = 10 \text{ volts} < \frac{v_{c\,max}}{2}$$

4. Put $\eta = 0.4$ and find P_{diss}

$$P_{diss} = P_{st}/\eta = 1.41 \div 0.4 = 3.52 \text{ watts}$$

5. Calculate the quiescent collector current

$$I_{c0} = P_{diss}/v_{c0} = 3.52 \div 10 = 0.352 \text{ ampere}$$

6. Calculate the collector load resistance

$$R_c = v_{c0}/I_{c0} = 10 \div 0.352 = 28.4 \text{ ohms}$$

7. Plot the load line. The load line in our example passes through the Q-point with the coordinates $I_{c0} = 0.352$ ampere, $v_{c0} = 10$ volts and has the terminal points B and C with the

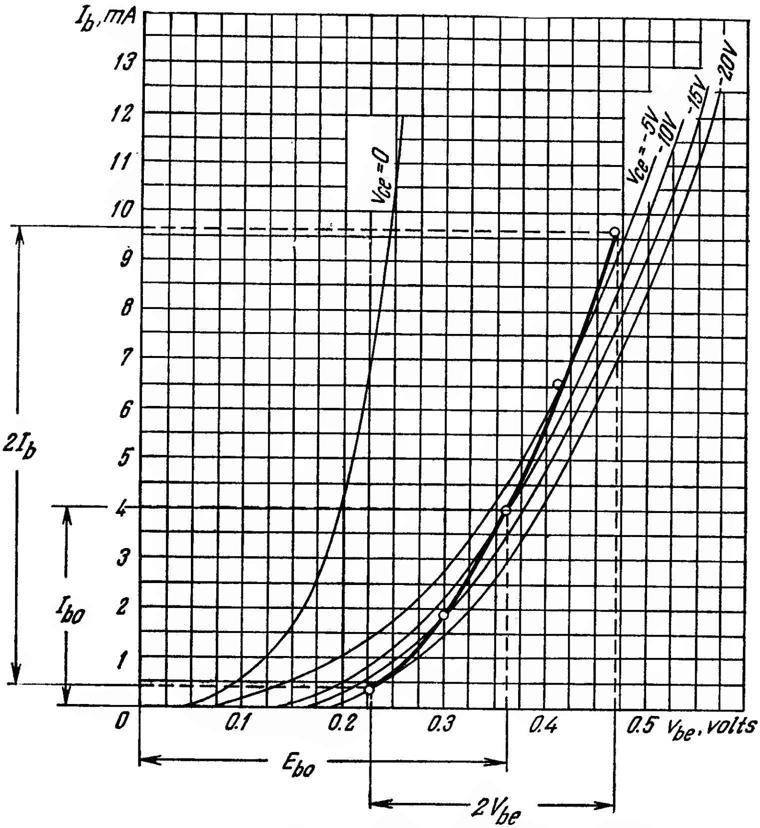


Fig 4.27 Input characteristics of Soviet-made П201А transistor

coordinates:

$$\text{point } B\text{: } v_{c\,min} = \text{(approx.) } 0.1v_{c0} = 1 \text{ volt}$$
$$I_{c\,max} = I_{c0} + (0.9v_{c0}/R_c) = 0.67 \text{ ampere}$$
$$\text{point } C\text{: } v_{c\,max} = 1.9v_{c0} = 19 \text{ volts}$$
$$I_{c\,min} = I_{c0} - (0.9v_{c0}/R_c) = 0.035 \text{ ampere}$$

From the load line, find the quiescent base current $I_{b0} = 4$ milliamperes, the maximum base current $I_{b\,max} = 9.6$ milliamperes, and the minimum base current $I_{b\,min} = 0.4$ milliampere.

8. From the data thus obtained, plot the dynamic input characteristic.

9. From the dynamic input characteristic, find the quiescent base voltage $E_{b0}=0.365$ volt, the minimum base voltage $v_{be\,min}=$ $=0.23$ volt, the maximum base voltage $v_{be\,max}=0.47$ volt

$$2I_b=I_{b\,max}-I_{b\,min}=9.6-0.4=9.2 \text{ milliamperes}$$
$$2V_{be}=v_{be\,max}-v_{be\,min}=0.47-0.23=0.24 \text{ volt}$$

10. Determine the average input resistance

$$R_{in}=2V_{be}/2I_b=0.24\div 9.2\times 10^{-3}=26.1 \text{ ohms}$$

11. Select the current for the stabilization network

$$I_{12}=(2 \text{ or } 3)\, I_{b0}=2\times 4=8 \text{ milliamperes}$$

12. Calculate R_1, R_2, and R_e. Putting $v_e=1$ volt:

$$R_2=(E_{b0}+v_e)/I_{12}=(0.365+1)\div 8\times 10^{-3}=171 \text{ ohms (approx.)}$$
$$R_1=\frac{v_{c0}-E_{b0}-v_e}{I_{12}+I_{b0}}=\frac{10-0.365-1}{(8+4)\times 10^{-3}}=720 \text{ ohms}$$
$$R_e=v_e/I_{c0}=1\div 0.352=2.84 \text{ ohms}$$

13. Calculate the equivalent input resistance

$$R_{12}=R_1R_2/(R_1+R_2)=720\times 171\div(720+171)=138 \text{ ohms}$$
$$R_{in\,e}=R_{in}R_{12}/(R_{in}+R_{12})=26.1\times 138\div(26.1+138)=21.9 \text{ ohms}$$

14. Find the power dissipated by the input circuit

$$P_{in}=(2V_{be})^2/8R_{in\,e}=(0.24)^2\div(8\times 21.9)=0.33 \text{ milliwatt}$$

15. Plot the dynamic transfer characteristic

$$I_c=F(E_s)$$

Put $R_s=150$ ohms and locate the points of the dynamic transfer characteristic by reference to the output characteristic:

Point *1*: $I_{c(1)}=35$ milliamperes; $I_{b1}=0.4$ milliampere; $v_{be1}=$ $=0.23$ volt

$$E_{s1}=v_{be1}+R_sI_{b1}=0.23+150\times 0.4\times 10^{-3}=0.29 \text{ volt}$$

Point *2*: $I_{c(2)}=190$ milliamperes; $I_{b2}=2$ milliamperes; $v_{be2}=$ $=0.32$ volt

$$E_{s2}=0.32+150\times 2\times 10^{-3}=0.62 \text{ volt}$$

Point *3*: $I_{c(3)} = 352$ milliamperes; $I_{b3} = 4$ milliamperes; $v_{be3} = 0.365$ volt

$$E_{s3} = 0.365 + 150 \times 4 \times 10^{-3} = 0.965 \text{ volt}$$

Point *4*: $I_{c(4)} = 500$ milliamperes; $I_{b4} = 6$ milliamperes; $v_{be4} = 0.415$ volt

$$E_{s4} = 0.415 + 150 \times 6 \times 10^{-6} = 1.315 \text{ volts}$$

Point *5*: $I_{c(5)} = 670$ milliamperes; $I_{b5} = 9.6$ milliamperes; $v_{be5} = 0.47$ volt

$$E_{s5} = 0.47 + 150 \times 9.6 \times 10^{-3} = 1.91 \text{ volts}$$

The dynamic transfer characteristic plotted by these points is shown in Fig. 4.28.

In order to find the collector-current harmonics, calculate the instantaneous voltages:

$$2V_{in} = E_{s5} - E_{s1} = 1.91 - 0.29 = 1.62 \text{ volts}$$
$$V_{in} = 0.81 \text{ volt}$$
$$V_{in}/2 = 0.81 \div 2 = 0.405 \text{ volt}$$

and the instantaneous currents:

$$I_{c\,min} = 35 \text{ milliampere at } E_{c\,min}$$
$$I_{c\,max} = 670 \text{ milliamperes at } E_{c\,max}$$
$$I_{c0} = 395 \text{ milliamperes at } E_{s\,min} + V_{in}$$
$$I_{c1} = 565 \text{ milliamperes at } E_{s\,min} + 3V_{in}/2$$
$$I_{c2} = 222 \text{ milliamperes at } E_{s\,min} + V_{in}/2$$

Then, the collector-current harmonics are:

$$I_{c1} = \frac{(I_{c\,max} - I_{c\,min}) + (I_{c1} - I_{c2})}{3}$$
$$= \frac{(670 - 35) + (565 - 222)}{3} = 326 \text{ milliamperes}$$

$$I_{c2} = \frac{0.5\,(I_{c\,max} + I_{c\,min}) - I_{c0}}{2} = -21 \text{ milliamperes}$$

$$I_{c3} = \frac{(I_{c\,max} - I_{c\,min}) - 2\,(I_{c1} - I_{c2})}{6} = 8.5 \text{ milliamperes}$$

$$I_{c4} = \frac{(I_{c\,max} + I_{c\,min}) - 4\,(I_{c1} + I_{c2}) + 6I_{c0}}{12} = 0.58 \text{ milliampere}$$

$$I_{c\,av} = \frac{(I_{c\,max} + I_{c\,min}) + 2\,(I_{c1} + I_{c2})}{6} = 373 \text{ milliamperes}$$

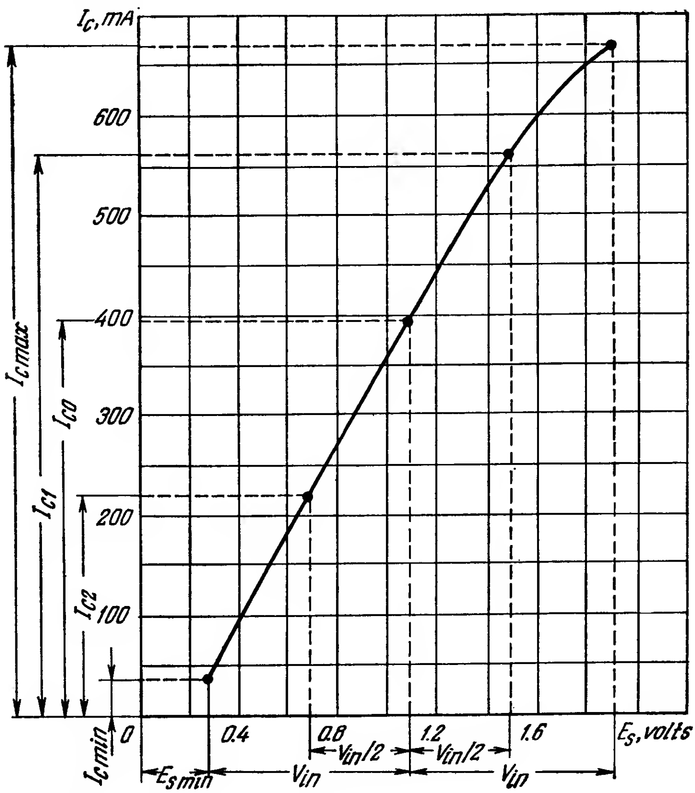


Fig. 4.28 Dynamic transfer characteristic

Hence, the non-linear distortion is

$$\gamma = \frac{\sqrt{I_{c2}^2 + I_{c3}^2 + I_{c4}^2}}{I_{c1}} = 0.07 \text{ (approx.)}$$

16. Calculate the inductance of the transformer primary:

$$L_1 = \frac{R_c}{2\pi F_l \sqrt{M_l^2 - 1}} = 0.0985 \text{ henry (approx.)}$$

17. Determine the transformation ratio of the output transformer

$$n = \sqrt{R_L / R_c \eta_T} = 0.5$$

Review Questions

1. Why is the efficiency of single-ended transistor power amplifiers greater than that of valve amplifiers?
2. Why is non-linear distortion greater in a transistor amplifier than in a valve amplifier?
3. Why will the split load minimize non-linear distortion?
4. How can a push-pull amplifier be built without a phase-inversion stage?
5. Develop the circuit diagram of a push-pull amplifier without a phase inverter.

CHAPTER V

AERIAL-INPUT CIRCUITS

25. General

An *aerial-input* (or *aerial-coupling*) *circuit* is one which couples the aerial to the receiver and partly filters out unwanted frequencies. Most receivers use resonant circuits for the purpose.

Aerial-input circuits may be classed according to (1) the type of resonant circuit or circuits used, and (2) the method of coupling between the aerial and the receiver tuned circuit.

As to type of resonant circuit, it may be either single or coupled. The most commonly used type is a single resonant circuit.

As to method of coupling between the aerial and the receiver, the choice is mainly decided by the type of aerial system used. On long, medium, and short waves, untuned aerials are commonly used. They are usually earthed vertical wires, although inverted-L and T-aerials are also common. With them, use is made of capacitative coupling (Fig. 5.1*a*), inductive coupling (Fig. 5.1*b*), and capacitative-inductive coupling (Fig. 5.1*c*).

Communication receivers for short and ultra-short wavelengths, as well as TV receivers, use tuned balanced aerials from which the r. f. energy is conveyed to the receiver by a transmission line, and this should be coupled to the receiver through a matching transformer or tapped coil (autotransformer).

Broadcast receivers, especially portable ones use built-in *ferrite-rod* (or *loopstick*) aerials.

In analysis, the aerial-coupling circuit along with the tuned circuit of the receiver are represented by a common equivalent circuit, in which the aerial is approximated by a constant-voltage generator of emf E_A and of internal impedance Z_A, which has components r_A, C_A, and L_A whose effect is felt on long, medium and partly short waves. An equivalent circuit for the aerial on these wavelengths is shown in Fig. 5.2. Other factors controlling

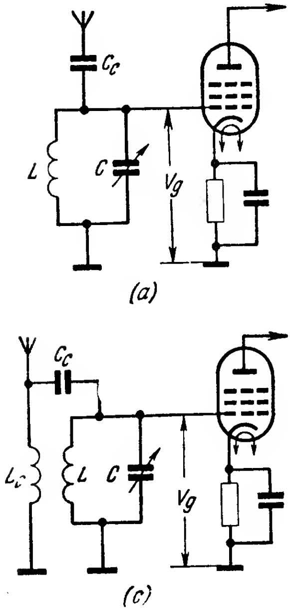


Fig. 5.1. Single-tuned aerial-input circuits

(a) capacitative coupling; (b) inductive coupling; (c) capacitative-inductive coupling

the aerial parameters are its geometry, location, height above terrain, etc. For a standard broadcast aerial with an effective height of 4 metres, $r_A = 25$ ohms, $L_A = 20$ microhenrys, and $C_A = 150$ to 300 picofarads. These are average values which may vary considerably in specific cases.

Analysis of aerial-input circuits deals with:

(1) The voltage gain defined as the ratio of the voltage at the first stage of the receiver to that fed by the aerial.

(2) Variations in the voltage gain with frequency over the operating band.

(3) The selectivity of the aerial-input circuit.

(4) The effect of the aerial on the associated tuned circuit. The point is that the aerial and the tuned circuit make up a

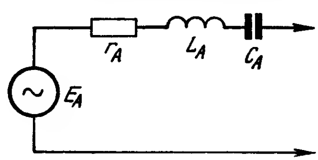


Fig. 5.2. Equivalent circuit of an aerial

coupled circuit, in which the primary may affect the secondary and vice versa. This aspect materially affects the choice of coupling between the two.

Review Questions

1. Name the factors governing the choice of aerial coupling.
2. Why is it important to reckon with the effect of the aerial on the associated tuned circuit?
3. Which parameter of the receiver is affected by variations in the voltage gain of the aerial-coupling circuit with frequency?

26. Capacitatively-coupled Aerial-input Circuit

Let the capacitatively-coupled aerial-input circuit of Fig. 5.1*a* be replaced with its equivalent circuit, shown in Fig. 5.3*a*. With the coupling capacitance C_c series-connected with the aerial lead, the total capacitance

$$C'_A = C_A C_c/(C_A + C_c)$$

decreases, and the capacitative reactance of the circuit becomes much higher than its inductive reactance, so that the latter may be dropped from the equivalent circuit. For the same reason, r_A (at *b*) may be omitted from the circuit without sacrificing the accuracy to any considerable extent.

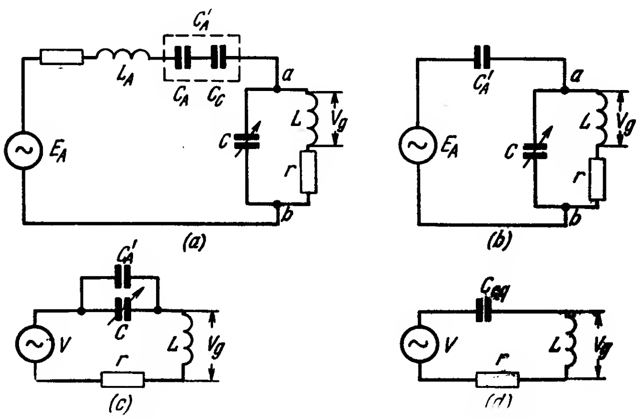


Fig. 5.3. Conversion of an aerial-input circuit into an equivalent series-resonant circuit

To simplify the equivalent circuit still more, we apply Thévenin's theorem.

Taking points ab as the output terminals of a two-port and its impedance with the input port short-circuited as the internal impedance, we obtain the equivalent circuit of Fig. 5.3c.

The final equivalent circuit is given in Fig. 5.3d. In this circuit, C'_A and C are lumped into an equivalent capacitance $C_{eq}=C'_A+C$. This is a series resonant circuit having the following parameters:

$$\left.\begin{array}{c} L=L \\ C_{eq}=C+C'_A \\ r=r \end{array}\right\} \qquad (5.1)$$

Using the equivalent circuit, determine the basic parameters of the aerial-input circuit.

The voltage gain at resonance is given by

$$K_0=\frac{V_g}{E_A}$$

The resonance voltage V_g across the inductance (see Fig. 5.3d) may be written in terms of the Q-factor of the tuned circuit

$$V_g=VQ \qquad (5.2)$$

where V is the voltage of the equivalent generator and is determined from the circuit of Fig. 5.3b, with the load disconnected from points ab. From circuit (b) it is seen that E_A is applied to a capacitative voltage divider consisting of series-connected capacitances C'_A and C. V across C may be found from

$$\frac{E_A}{V}=\frac{C}{C_{tot}}$$

where

$$C_{tot}=\frac{C'_A C}{C'_A+C}$$

Therefore,

$$V=E_A\frac{C'_A}{C'_A+C}=E_A\frac{C'_A}{C_{eq}}$$

Substituting this expression in (5.2) gives

$$V_g=VQ=E_AQ\frac{C'_A}{C_{eq}}$$

Hence, the voltage gain at resonance is

$$K_0 = \frac{V_g}{E_A} = Q\frac{C'_A}{C_{eq}} \tag{5.3}$$

From Eq. (5.3) it is possible to evaluate the voltage gain K_0 by comparing it with the Q-factor of the unloaded tuned circuit (that is, taken separately). A comparison will show that the voltage gain of the aerial-input circuit at resonance is considerably smaller than the Q-factor of the unloaded tuned circuit, which is due to the coupling capacitance C_c in the aerial-input circuit.

With comparatively high values of the Q-factor of the tuned circuit, K_0 does not exceed 10.

Example 5.1. The effective Q-factor (that is, that of a loaded tuned circuit) is 100. The tuned circuit is coupled to an aerial through $C_c = 15$ picofarads. The tuned-circuit capacitance is $C = 250$ picofarads. Determine the voltage gain of the aerial-input circuit.

Solution: Putting $C_A = 200$ picofarads, find the capacitance C'_A coupled into the tuned circuit by the aerial:

$$C'_A = C_A C_c/(C_A + C_c) = 14 \text{ picofarads}$$

Hence, the equivalent capacitance of the aerial-input circuit is

$$C_{eq} = C'_A + C = 14 + 250 = 264 \text{ picofarads}$$

The voltage gain of the aerial-input circuit is

$$K_0 = Q_{ef} C'_A / C_{eq} = 100 \times 14 \div 264 = 5.3$$

Variations in the voltage gain with frequency may be found from

$$K_0 = QC'_A/C_{eq}$$

The value of $Q = \omega_0 L/r$ remains almost unchanged over the frequency range because an increase in $\omega_0 L$ is accompanied by an increase in the series loss resistance r. The equivalent capacitance C_{eq} of a tuned circuit with a variable tuning capacitor is approximately inversely proportional to the square of the frequency. Consequently, the voltage gain of the aerial-input circuit is likewise a quadratic function of the tuning frequency. A plot of the function $K_0 = \varphi(f)$ is shown in Fig. 5.4. As is seen, the voltage gain of the aerial-input circuit is low at the low frequencies of the tuning range and high at the high frequencies. This is a disadvantage of capacitative coupling between the aerial and the tuned circuit of the receiver.

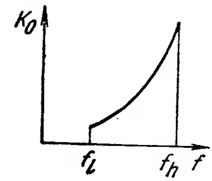


Fig. 5.4. Voltage gain of an input circuit, capacitatively coupled to an aerial, as a function of frequency

In the case just examined we have seen that the aerial couples a capacitance C'_A into the tuned circuit. A more rigorous analysis would show that the aerial also couples an additional loss resistance into the tuned circuit. In other words, the aerial detunes the tuned circuit owing to an increase in the tuned-circuit capacitance and impairs its Q-factor. Therefore, for practical purposes, the Q-factor of the loaded tuned circuit should be taken equal to 0.8 to 0.9 of its Q-factor when unloaded. That is, in Eq. (5.3), the Q must be replaced with the Q_{ef}.

The effect of the aerial on the tuned circuit is of practical importance in alignment of the aerial-input circuit. If the resultant detuning were a fixed quantity, it could be easily compensated by adjustment of a tuned-circuit parameter. Unfortunately, C_A may, due to some factors, vary from a half to twice its middle value, and the value of C'_A coupled into the tuned circuit will also be a varying quantity. C'_A may be made more or less fixed through loose coupling between the aerial and the tuned circuit. This is why the coupling capacitor is usually 10 to 30 picofarads.

The selectivity of the aerial-input circuit is determined from its series-resonant equivalent circuit, using the resonance-curve equation:

$$Y = 1/\sqrt{1+x^2} \tag{5.4}$$

where

$$x = (\omega/\omega_0 - \omega_0/\omega)\, Q_{ef} = \text{approx. } 2\Delta f Q_{ef}/f_0$$

and Δf is the separation from the resonant frequency. The resonance curve described by Eq. (5.4) is shown in Fig. 5.5. The separation from the resonant frequency, $\Delta f = f_0 - f$, is laid off as abscissa, f_0 is the resonant frequency of the aerial-input circuit, and f is the signal frequency. When $f_0 = f$, the circuit is at resonance.

The bandwidth of the aerial-input circuit can be derived from Eq. (5.4). Taking Δf as a half-bandwidth, the ordinate

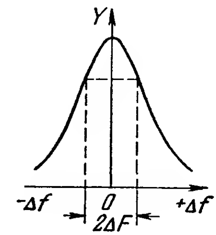


Fig. 5.5. Resonance curve of the aerial-input circuit

Y corresponding to a specified harmonic content (frequency distortion) M will locate the bandwidth, $2\Delta F$.

Substituting M for Y and $2\Delta F$ for $2\Delta f$ in Eq. (5.4), and writing out x in full gives

$$M = \frac{1}{\sqrt{1+\left(\frac{2\Delta F}{f_0} Q_{ef}\right)^2}} \tag{5.5}$$

Solving Eq. (5.5) for $2\Delta F$ gives the bandwidth

$$2\Delta F = \frac{f_0}{Q_{ef}} \sqrt{\frac{1}{M^2} - 1} \tag{5.6}$$

Capacitatively-coupled aerial-input circuits are mainly used in fixed-tuned or narrow-band receivers.

Review Questions

1. Why is it that the voltage gain of an aerial-input circuit is less than its Q-factor?
2. What effect has the coupling capacitance on the voltage gain of the aerial-input circuit?
3. Why is it that the coupling capacitance is chosen to be small?
4. What factors control the selectivity of the aerial-input circuit?

27. An Inductively-coupled Aerial-input Circuit

An inductively-coupled aerial-input circuit is shown in Fig. 5.1*b*.

Its equivalent circuit appears in Fig. 5.6*a*. Neglecting the aerial inductance L_A, which is a small fraction (20 microhenrys) of that of the coupling coil L_c, we obtain two resonant circuits

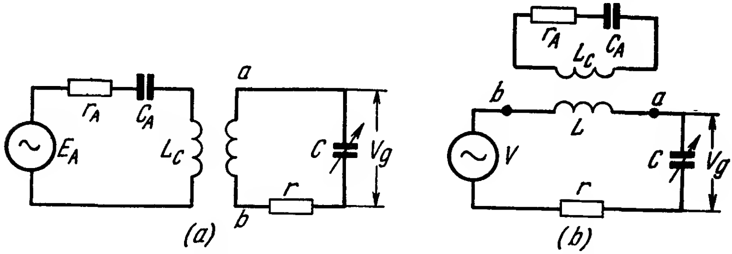


Fig 5.6. Equivalent circuits of an inductively-coupled aerial-input circuit

tuned to different frequencies (stagger tuning). The frequency of the aerial circuit is

$$f_A = \frac{1}{2\pi\sqrt{L_c C_A}}$$

The frequency of the tuned circuit

$$f = \frac{1}{2\pi\sqrt{LC}}$$

The aerial circuit has a fixed frequency f_A, which is chosen when the receiver is being designed. The frequency of the tuned circuit can be changed within a definite range by a variable capacitor.

To decrease the effect of the aerial on the tuned circuit, the coupling between them is set to be less than optimum.

The voltage gain of the aerial-input circuit may be determined from the circuit of Fig. 5.6*b*, obtained with the aid of Thévenin's theorem. By definition, the voltage gain at resonance is

$$K_0 = \frac{V_g}{E_A} \tag{5.7}$$

The output voltage of the tuned circuit is

$$V_g = VQ$$

where Q is the figure-of-merit of the tunable circuit.

The equivalent-generator voltage V is found from Fig. 5.6*a* across points *a*-*b*, with the load disconnected

$$V = I_A \omega M$$

The current I_A in the aerial circuit is

$$I_A = \frac{E_A}{Z_A}$$

Hence

$$V = E_A \frac{\omega M}{Z_A}$$

where:

$$Z_A = \sqrt{r_A^2 + \left(\omega L_c - \frac{1}{\omega C_A}\right)^2} = \omega L_c - \frac{1}{\omega C_A}$$

The resistance r_A of the aerial circuit is considerably less than its reactance, and so the contribution of r_A to Z_A may be disregarded.

After substitutions, we obtain

$$V = E_A \frac{\omega M}{\omega L_c - \frac{1}{\omega C_A}} = E_A \frac{\omega M}{\omega L_c \left(1 - \frac{1}{\omega^2 L_c C_A}\right)} \qquad \text{(a)}$$

Since

$$\frac{1}{L_c C_A} = \omega_A^2$$

then

$$V = E_A \frac{M}{L_c \left(1 - \frac{\omega_A^2}{\omega^2}\right)}$$

Writing M in (a) in terms of the coefficient of coupling k,

$$M = k\sqrt{L_c L}$$

we get

$$V = E_A \frac{k}{1 - \frac{f_A^2}{f^2}} \sqrt{\frac{L}{L_c}}$$

V_g at the output of the equivalent circuit is Q times V:

$$V_g = QE_A \left(\frac{k}{1 - f_A^2/f^2}\right) \sqrt{\frac{L}{L_c}} \qquad (5.8)$$

Substituting Eq. (5.8) in Eq. (5.7), we obtain

$$K_0 = \frac{V_g}{E_A} = Q \frac{k}{1 - \frac{f_A^2}{f^2}} \sqrt{\frac{L}{L_c}} \qquad (5.9)$$

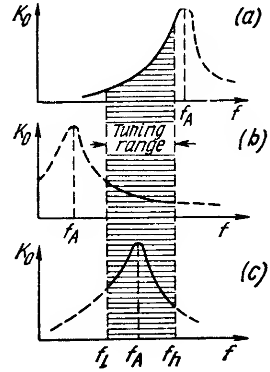


Fig. 5.7. Variations in the voltage gain of the aerial-circuit with inductive coupling

The voltage gain of an inductively and capacitatively coupled aerial-input circuit is considerably less than the Q of the tuned circuit due to incomplete transfer of the emf from the aerial.

The selectivity of the aerial-input circuit is determined as that of an equivalent series circuit, because two coupled circuits may be represented by an equivalent series circuit which allows for the effect of one circuit on the other.

The selectivity and bandwidth can be found from Eqs. (5.4) and (5.6), assuming that $Q_{eq}=0.8$ to $0.9\ Q$ (approximately).

The voltage gain as a function of frequency may be found as follows. The transmission of the signal from the aerial to the first valve may be thought of as transmission of the signal voltage from the aerial into the tuned circuit where this voltage is amplified Q times. If the Q of the input tuned circuit is deemed constant over the entire frequency range, then variations in gain with frequency will only be due to the resonant properties of the aerial circuit and voltage transfer from the aerial to the tuned circuit. Equation (5.9) does not give an exact picture of the resonant properties of the aerial circuit. The resonance curve of the aerial circuit is unsymmetrical, due to the loss resistance coupled in from the secondary circuit (Fig. 5.7).

The frequency range of the aerial-input circuit may use different portions of the resonance curve of the aerial, namely below, above or near the resonant frequency of the aerial circuit.

When $f_A > f_0$ ($\lambda_A < \lambda_0$), the aerial circuit (case a) is called short. When $f_A < f_0$ ($\lambda_A > \lambda_0$), the aerial circuit (case b) is called long.

With a short aerial circuit (see Fig. 5.7*a*), the voltage gain in the tuning range is represented by the rising portion of the resonance curve (the full line), reflecting a considerable change in gain with frequency.

With a long aerial circuit (see Fig. 5.7*b*), the voltage gain is represented by the descending portion of the curve, reflecting the fact that the gain remains almost unchanged over the range. When the resonant frequency of the aerial circuit is within the tuning range (see Fig. 5.7*c*), the gain over the range is not uniform, and this case is not used in practice.

Most modern receivers employ a long aerial circuit.

The effect of the aerial on the tuned circuit is insignificant.

Since, however, the aerial parameters are varying quantities, the amount of detuning caused by the aerial will also be varying.

To make the relative detuning as small as it is with capacitative coupling, the degree of coupling within the aerial-input circuit should not exceed a definite value.

Review Questions

1. What defines the aerial circuit as short or long?
2. Which of the circuit elements makes the aerial circuit short or long?
3. How can the effect of the aerial on the tuned circuit be minimized?
4. What governs the bandwidth of the aerial-input circuit?

28. Aerial-input Circuit for a Ferrite-rod Aerial

With a ferrite-rod aerial (also called a loopstick aerial), the aerial-input circuit is a single resonant circuit composed of the tuned-circuit capacitor and the tuned-circuit inductor inside which there is a ferrite rod.

Ferrite rods used for aerials are usually circular or square in cross-section. Circular rods are usually 8 to 10 mm in diameter and 70 to 150 mm long.

Like a frame aerial, a ferrite-rod aerial is a directional one, securing an amount of spatial selectivity for the receiver. In table receivers, the ferrite-rod aerial can be oriented for a maximum signal with a knob; in portable receivers this is done by moving about the receiver itself.

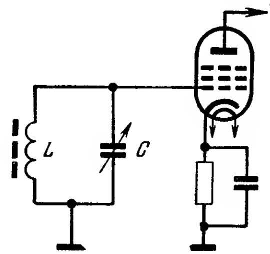


Fig. 5.8. Ferrite-aerial coupling circuit connected directly in the grid lead

In broadcast receivers, built-in ferrite-rod aerials are usually used in the long and medium wavebands The initial relative permeability of the ferrite rods used for the purpose is from 600 to 1000 (in Soviet-made types). The effective height is low, being 0.005 to 0.015 m. The Q-factor of aerial-input circuits with ferrite rods is 100 to 200.

The joint action of a ferrite-rod aerial of effective height h_{ef} and the associated resonant circuit of quality Q is equivalent to an increase in the effective height to

$$h'_{ef} = h_{ef}Q \tag{5.10}$$

where h'_{ef} is the effective height of a loaded ferrite-rod aerial. If receiver sensitivity is known (which for receivers with built-in aerials is defined in terms of the electric intensity E at the point of reception), the voltage V_{crt} across the input resonant circuit will be given by

$$V_{crt} = Eh'_{ef} \tag{5.11}$$

It defines the efficiency of the ferrite-rod aerial.

Ferrite-rod aerial-input circuits may be connected to the first stage of the receiver in several ways. In valve receivers, the usual method is direct connection as shown in Fig. 5.8. In transistor receivers, use is made of either transformer coupling (Fig. 5.9*a*) or tapped-coil (autotransformer) coupling (Fig. 5.9*b*). Tapped-coil connection reduces the shunting effect of the low input resistance of the transistor on the tuned coupling circuit.

The selectivity and bandwidth of the ferrite-aerial coupling circuit depend on the loaded Q-factor and can be determined from Eqs. (5.4) and (5.6). It should be remembered, however,

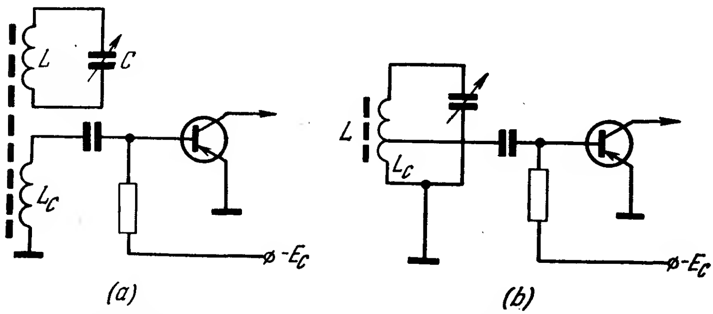


Fig. 5.9. Connection of a ferrite-aerial coupling circuit to a transistor receiver: (a) transformer coupling; (b) tapped-coil coupling

that in transistor receivers using the arrangements of Fig. 5.9, the condition for maximum power output (perfect impedance matching) will halve the loaded Q-factor:

$$Q_{ef} = Q/2$$

Review Questions

1. Define the spatial selectivity of the receiver.
2. How is the efficiency of a ferrite-rod aerial evaluated?
3. Why is it that in transistor receivers the aerial coupling circuit is not directly connected to the receiver?

29. Calculation of Aerial-input Circuits for Long, Medium and Short Waves

Calculation of a Capacitatively-coupled Aerial-input Circuit

Given:

1. Frequency band, f_l-f_h.
2. Maximum capacitance C_{max} of the variable capacitor.
3. Loaded Q-factor, Q_{ef}.
4. Aerial parameters, C_A, r_A.

To Find:

1. Coupling capacitance C_c.
2. Voltage gain K_0 at three points of the band.

Design Procedure:

1. Assume that C_c is 10 to 30 picofarads.
2. Determine the tuned-circuit capacitance (with allowance for the aerial effect) at three points of the band:

(a) f_l

$$C_{eq\,(max)} = C_{max} + C'_A = C_{max} + \frac{C_A C_c}{C_A + C_c}$$

(b) $f_m = \frac{f_h + f_l}{2}$

$$C_{eq\,(av)} = C_{eq\,(max)} \left(\frac{f_l}{f_m}\right)^2$$

(c) f_h

$$C_{eq\,(min)} = C_{eq\,(max)} \left(\frac{f_l}{f_h}\right)^2$$

3. Determine the voltage gain for the three points of the band

$$K_0 = Q_{ef} \frac{C_A}{C_{eq}}$$

4. Plot K_0 as a function of frequency.

Calculation of a Long Inductively-coupled Aerial-input Circuit

Given:

1. Frequency band, f_l-f_h.
2. Loaded Q-factor, Q_{ef}.
3. Tuned-circuit inductance, L.
4. Aerial capacitance, $C_{A\,min}$-$C_{A\,max}$.

To Find:

1. Coefficient of coupling, k.
2. Coupling inductance, L_c.
3. Voltage gain, K_0, at three points of the band.

Design Procedure:

1. Determine the maximum (f_h) and minimum (f_l) frequencies of the equivalent aerial circuit.

For long and medium wavelengths, the maximum frequency is:

$$f_{Ah} = (0.5 \text{ to } 0.8) f_l$$

For the short-wave range:

$$f_{Ah}=(0.25 \text{ to } 0.3)\,f_l$$

The minimum frequency, regardless of the range:

$$f_{Al}=\frac{1}{\sqrt{\frac{C_{A\,max}}{C_{A\,min}}}}\,f_{Ah}$$

2. Find the optimum coefficient of coupling k_{opt} which will limit the detuning due to the aerial to β:

$$k_{opt}=2\sqrt{\frac{\beta(1-A)(1-B)}{B-A}}$$

where

$$A=\left(\frac{f_{Al}}{f_h}\right)^2$$

$$B=\left(\frac{f_{Ah}}{f_l}\right)^2$$

$$\beta=0.5\,\frac{1}{Q_{ef}}$$

The coefficient of coupling for the subsequent computation should be so selected that $k<k_{opt}$ and does not exceed its realisable value.

With a universal winding

$$k=0.5\text{-}0.6$$

With a single-layer cylindrical winding

$$k=0.4\text{-}0.5$$

3. Determine the inductance of the coupling coil

$$L_c=\frac{2.53\times10^{10}}{C_{A\,min}\,f^2_{Ah}}$$

where L is in microhenrys, C in picofarads, and f_A in kilohertz.

4. Find the voltage gain of the aerial-input circuit at three points of the band

$$K_0=Q_{ef}\frac{k}{1-\left(\frac{f_{Al}}{f}\right)^2}\sqrt{\frac{L}{L_c}}$$

Calculation is carried out for $f_A=f_{Al}$ because we obtain a minimum value of K_0 in this case.

SUMMARY

1. Any aerial-input circuit is characterised by voltage gain which is smaller than the Q of the tuned circuit used.

2. The selectivity of the aerial-input circuit is chiefly determined by the resonant properties of the tunable circuit.

3. From the circuits studied above the voltage gain changes least with frequency in a long inductively-coupled aerial-input circuit.

4. The aerial affects the tunable circuit introducing resistance and reactance into it. With loose coupling, this effect is negligible.

5. The most commonly used arrangement is a single-tuned, inductively-coupled aerial-input circuit.

Capacitative coupling is mainly used on narrow bands or when a receiver operates at a fixed frequency.

6. Aerial-input circuits with built-in ferrite-rod aerials are mainly used on long and medium waves, in both valve and transistor receivers.

Problems

5.1. Determine the voltage gain of an inductively-coupled aerial-input circuit, if $C=200$ picofarads; $f=900$ kilohertz; $r=15$ ohms; $C_c=20$ picofarads; $C_A=150$ picofarads; $Q_{ef}=0.9Q$.

Answer: $K_0=4.3$.

5.2. Determine the voltage gain of the aerial-input circuit in Problem 5.1 when the tuned-circuit capacitance C is halved.

Answer: $K_0=17.7$.

5.3. From the data of Problem 5.1, find the percentage change of the tuned-circuit capacitance as the aerial capacitance C_A changes from 150 to 300 picofarads.

Answer: $\frac{\Delta C'_A}{C_{eq}}=0.51\%$.

5.4. Solve Problem 5.3 when the coupling capacitance C_c is 60 picofarads.

Answer: $\frac{\Delta C'_A}{C_{eq}}=2.8\%$.

5.5. Determine the voltage gain of an inductively-coupled aerial-input circuit at two points of the range, if $L=300$ microhenrys; $Q=67$; $L_c=2{,}100$ microhenrys; $f_h=1{,}300$ kilohertz; $f_l=400$ kilohertz; $k=0.2$.

The aerial capacitance C_A changes from 150 to 300 picofarads.

Answer: $\left.\begin{array}{l}\text{For } f_l = 400 \text{ kilohertz, } K_0 = 6.8 \\ \text{For } f_h = 1{,}300 \text{ kilohertz, } K_0 = 5.2\end{array}\right\} (f_{A\,h} = 0.7 f_l)$

5.6. Calculate inductive coupling for an aerial-input circuit if:

(a) Frequency band: $f_l = 150$ kilohertz, $f_h = 400$ kilohertz; tuned-circuit inductance $L = 2{,}080$ microhenrys; the Q of the tuned circuit is 50; $C_A = 150$ to 300 picofarads.

Answer: $K_0 = 4.7$ ($f = 150$ kilohertz; $f_{A\,h} = 0.8 f_l$)
$K_0 = 3.3$ ($f = 400$ kilohertz)

(b) $f_l = 6 \times 10^3$ kilohertz; $f_h = 20 \times 10^3$ kilohertz; $L = 1.3$ microhenrys; $Q = 67$; $C_A = 150$ to 300 picofarads.

Answer: $\left.\begin{array}{l}K_0 = 5.5 \ (f = 6 \text{ megahertz}) \\ K_0 = 5.3 \ (f = 20 \text{ megahertz})\end{array}\right\}$ at $f_{A\,h} = 0.25 f_l$, $k = 0.6$

CHAPTER VI

RADIO-FREQUENCY AMPLIFIERS

30. General

The functions of a radio-frequency (r.f.) amplifier are to increase the voltage of the radio-frequency signal and to secure the required selectivity of the receiver.

The voltage applied to the input of an r.f. amplifier is from units to hundreds of microvolts, depending on the sensitivity of the receiver. Since normal operation of a detector is obtained only when the radio-frequency signal applied to its input is several volts, it follows that before the signal reaches the detector it should be amplified a million times or more. Such voltage gain may be obtained only with the aid of several amplifier stages.

An r.f. amplifier stage, just as an a.f. one, contains a valve or a transistor and a load. The load is a resonant circuit tuned to the frequency of the signal applied to the input of the stage. This resonant circuit may be a single tuned circuit or a band-pass filter.

R.f. amplifiers in which single tuned circuits serve as load are known as *tuned amplifiers*. R.f. amplifiers employing band-pass filters for load are called *band-pass* or *filter amplifiers*.

Band-pass amplifiers have a nearly rectangular resonance curve, are usually fixed-tuned and are chiefly used as i.f. (intermediate-frequency) amplifiers in superheterodyne receivers. In some receivers one of r.f. amplifier stages is *RC*-coupled. It does not possess resonance properties and is called an *untuned stage*.

The present chapter deals with tuned and band-pass amplifiers for long, medium and short waves. Amplification in the microwave region is treated in Chapter XIV.

Amplifier circuits are evaluated in terms of stage gain and selectivity, and — in the case of wideband amplifiers — also the gain/frequency response.

Stage gain is defined as the ratio of input to output voltage:

$$\dot{K} = \frac{\dot{V}_{out}}{\dot{V}_{in}}$$

If an amplifier has several stages with gains $\dot{K}_1$, $\dot{K}_2$, $\dot{K}_3 \ldots \dot{K}_n$, the total gain is the product of the gains of the individual stages

$$\dot{K} = \dot{K}_1 \dot{K}_2 \dot{K}_3 \ldots \dot{K}_n \quad (6.1)$$

Stage selectivity is evaluated from the resonance curve and is determined by the ordinate for a given am unt off resonance, Δf. The selectivity of a multi-stage amplifier is the product of the selectivities Y of the individual stages:

$$Y = Y_1 Y_2 Y_3 \ldots Y_n \quad (6.2)$$

Review Questions

1. What is the shape of the resonance curve of an untuned amplifier?
2. Define the gain and selectivity of a multi-stage amplifier.

31. Tuned Amplifiers

Tuned amplifiers are usually of the wideband variety because they are designed to amplify a definite band of frequencies. Such amplifiers may be tuned by varying either the tuned-circuit capacitance or, occasionally, the tuned-circuit inductance.

Single-stage Tuned-anode Amplifier. A relatively weak r.f. signal V_g applied to the input of the stage (Fig. 6.1) gives rise to an alternating current I_a in the anode circuit of the valve.

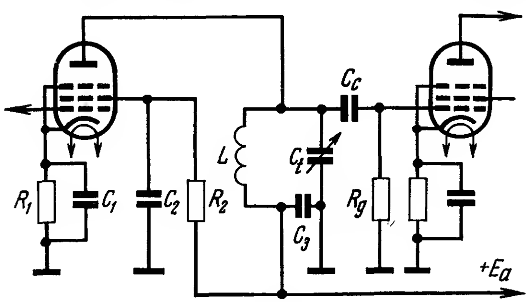


Fig. 6.1. Tuned-anode amplifier

The parallel tuned circuit, connected into the anode circuit, is tuned to the frequency of the incoming signal and presents a high resistance to I_a

$$R_{0e}=\frac{\omega_0^2L^2}{R}$$

such that the output voltage of the stage is

$$I_aR_{0e}=V_a$$

Other frequencies will not produce the same effect, because the impedance of the tuned circuit will be less than the *equivalent resonant resistance* R_{0e} at these frequencies. The output voltage from the tuned circuit is fed to the grid of the next stage by a coupling network C_cR_g.

A few words should be said about connection of the variable tuning capacitor C_t. When the operator rotates the capacitor, he introduces an additional capacitance into the circuit, thus changing its tuning. This effect is known as the "hand capacitance". The hand capacitance may be eliminated by connecting the rotor to the receiver chassis which is at earth (zero) potential. The bypass capacitor C_3 does the job without short-circuiting the anode supply source.

If the bypass capacitor is not to change the total capacitance of the tuned circuit, the capacitance of the circuit must be many times the maximum capacitance of the tuning capacitor C_t.

Figure 6.2 shows an r.f. equivalent circuit of the stage. The valve is replaced by a generator of emf μV_g and of internal resistance R_a. The anode load is represented by an RLC resonant circuit along with the circuits parallel to it.

C is the capacitance of the tuned circuit, L is the tuned-circuit inductance, R is the resistance of the tuned circuit, R_g is the grid-leak resistor, R_{in} is the input resistance of the next stage (from the grid to cathode), and C_s is the total shunt capacitance of the circuit, i.e. the sum of the valve output capaci-

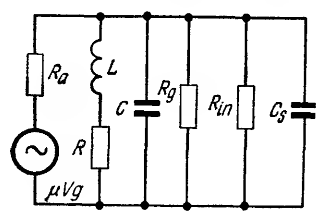


Fig. 6.2. Equivalent circuit of a tuned-anode amplifier

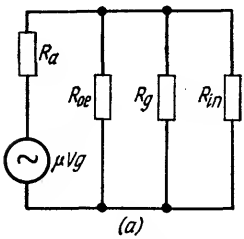


Fig. 6.3. Equivalent circuits of a tuned amplifier at resonance

tance C_{out}, the input capacitance C_{in} of the next stage, and the distributed-capacitance C_w of the wiring

$$C_s = C_{out} + C_{in} + C_w$$

The coupling capacitor C_c whose reactance to r.f. currents is negligible is not shown in the equivalent circuit.

The stage gain may be found from the equivalent circuits of Fig. 6.3 which hold at resonance. In circuit *a*, the tuned circuit $LC + C_sR$ is represented by its resonant resistance R_{0e} and by R_g and R_{in}, which are connected in parallel with it. Replacing the three parallel resistances by R'_{0e} we obtain circuit *b*.

To determine the stage gain from Eq. (1.9), we put

$$Z_a = R'_{0e}$$

Then

$$I_a = \frac{\mu V_g}{R_a + R'_{0e}}$$

Multiplying both sides by R'_{0e} gives

$$I_a R'_{0e} = \frac{\mu V_g}{R_a + R'_{0e}} R'_{0e}$$

Noting that $I_a R'_{0e} = V_a$, we have

$$V_a = \frac{\mu V_g}{R_a + R'_{0e}} R'_{0e}$$

Dividing both sides by V_g gives the stage gain

$$K_0 = \frac{V_a}{V_g} = \frac{\mu R'_{0e}}{R_a + R'_{0e}} \tag{6.3}$$

As a rule, r.f. amplifiers employ pentodes. In most r.f. pentodes, $R_a > 0.5$ megohm, while the resonant resistance R_{0e} of the tuned circuit is about 0.2 megohm only on long waves; on medium and short waves R_{0e} does not exceed a few tens of kilohms. Hence, R'_{0e} may be disregarded as compared with R_a in the denominator of Eq. (6.3).

Then

$$K_0 = \frac{\mu}{R_a} R'_{0e}$$

Replacing $\frac{\mu}{R_a}$ by the mutual conductance g_m gives

$$K_0 = g_m R'_{0e} \tag{6.4}$$

Or, in words, the stage gain of a tuned anode amplifier is determined only by the mutual conductance g_m of the valve and by the equivalent anode load resistance R'_{0e}.

The value of R'_{0e} and its relation to R_g and R_{in} which shunt the tuned circuit may be established from the curcuit of Fig. 6.3*a*

$$\frac{1}{R'_{0e}} = \frac{1}{R_{0e}} + \frac{1}{R_g} + \frac{1}{R_{in}}$$

It follows that

$$R'_{0e} = \frac{R_{0e} R_g R_{in}}{R_{0e} R_g + R_{0e} R_{in} + R_g R_{in}}$$

Dividing the numerator and denominator by $R_g R_{in}$, we obtain

$$R'_{0e} = \frac{R_{0e}}{\frac{R_{0e}}{R_{in}} + \frac{R_{0e}}{R_g} + 1} \tag{6.5}$$

The shunting effect of R_g and R_{in} on the tuned circuit will be minimized when

$$R_g \gg R_{0e}$$
$$R_{in} \gg R_{0e}$$

Example 6.1. Determine the stage gain of a tuned amplifier employing a 6K1П pentode for $R_{0e} = 40$ kilohms, $R_g = 1$ megohm. R_{in} is so high as to be neglected.

Solution. Referring to the valve manual, we find that a 6K1П pentode normally has $g_m = 1.85$ milliamperes per volt and $R_a = 0.45$ megohm.

Then

$$R'_{0e}=\frac{R_{0e}}{\frac{R_{0e}}{R_{in}}+\frac{R_{0e}}{R_g}+1}=\frac{40}{\frac{40}{1,000}+1}=39 \text{ kilohms}$$

The gain is

$$K_0=g_m R'_{0e}=1.85\times10^{-3}\times39\times10^3=72$$

The change of the equivalent resistance from R_{0e} to R'_{0e} may be regarded as a result of the shunting effect of R_g and R_{in} on the resistive component R of the tuned circuit

$$R'_{0e}=\frac{\omega_0^2L^2}{R'} \tag{6.6}$$

where R' is the new value of the tuned-circuit resistance equal to

$$R'=R+\Delta R$$

In Fig. 6.4*a*, the anode load R'_{0e} is represented by an equivalent resonant circuit of resistance R'.

The resistance R' may be obtained from Eq. (6.6):

$$R'=\frac{\omega_0^2L^2}{R'_{0e}} \tag{6.7}$$

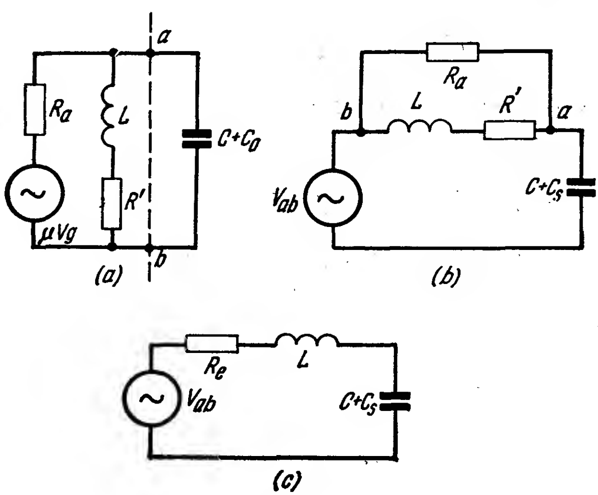


Fig. 6.4. Reducing the equivalent circuit of a tuned amplifier to a series resonant circuit

Replacing R'_{0e} by its expression from Eq. (6.5), and noting that

$$\frac{\omega_0^2 L^2}{R_{0e}} = R$$

we obtain

$$R' = R\left(1 + \frac{R_{0e}}{R_g} + \frac{R_{0e}}{R_{in}}\right) \tag{6.8}$$

From Eq. (6.8) it is seen that the incremental resistance of the tuned circuit depends upon the ratios R_{0e}/R_{in} and R_{0e}/R_g.

Now we shall transform the circuit of Fig. 6.4*a* by Thévenin's theorem, assuming that *ab* is the output port of the four-terminal network. Then $C+C_s$ and the internal resistance of the network seen looking into *ab* will be connected in series with the equivalent generator V_{ab} (Fig. 6.4*b*). The internal resistance seen looking into *ab* is determined with the network input short-circuited.

In Fig. 6.4*b*, the parallel connection at points *ab* is replaced by an equivalent series combination consisting of R_e, L and $C+C_s$, where R_e is the equivalent resistance which may be obtained from Eq. 1.11

$$R_e = R' + \frac{\omega_0^2 L^2}{R_a}$$

Multiplying the second term of the right-hand side by $\frac{R}{R}$, we obtain

$$R_e = R' + R\frac{\omega_0^2 L^2}{R R_a} = R' + R\frac{R_{0e}}{R_a}$$

The final equivalent circuit of a tuned amplifier (Fig. 6.4*c*) is a series resonant circuit with parameters L, $C+C_s$ and R_e.

Noting Eq. (6.8), the total resistance of the series equivalent circuit is

$$R_e = R\left(1 + \frac{R_{0e}}{R_g} + \frac{R_{0e}}{R_{in}}\right) + R\frac{R_{0e}}{R_a}$$

After simplification,

$$R_e = R\left(1 + \frac{R_{0e}}{R_g} + \frac{R_{0e}}{R_{in}} + \frac{R_{0e}}{R_a}\right) \tag{6.9}$$

The selectivity and bandwidth of a single-stage amplifier are determined by the Q-factor of the loaded tuned circuit

$$Q_{ef}=\frac{\omega_0 L}{R_e}=\frac{\omega_0 L}{R\left(1+\frac{R_{0e}}{R_g}+\frac{R_{0e}}{R_{in}}+\frac{R_{0e}}{R_a}\right)}$$

$$Q_{ef}=\frac{Q}{1+\frac{R_{0e}}{R_g}+\frac{R_{0e}}{R_{in}}+\frac{R_{0e}}{R_a}} \tag{6.10}$$

From Eq. (6.10) it follows that $Q_{ef}<Q$. This is an indication that the resonance curve of a single-stage amplifier is more flattened than that of the unloaded tuned circuit.

The resonance curve of an amplifier is described by a simple expression

$$Y=\left(\frac{1}{\sqrt{1+x^2}}\right)^n \tag{6.11}$$

where: x = generalized separation from resonance, defined as

$$x=\frac{2(f_0-f)}{f_0}Q_{ef},$$

f_0 = signal (resonant) frequency
f = unwanted frequency
n = number of tuned circuits (including the aerial-input tuned circuit).

In calculation, the quantity $d=\frac{1}{Y}$ is often used, showing the attenuation of the unwanted signal at $\Delta f=f-f_0$ from resonance.

The bandwidth of a tuned amplifier can be derived from Eq. (6.11). Taking Δf as a half-bandwidth, the ordinate Y corresponding to a specified harmonic content (frequency distortion) M will define the bandwidth, $2\Delta F$. Substituting M for Y and $2\Delta F$ for $2\Delta f$, re-arranging, and solving for $2\Delta F$ will give the bandwidth of a tuned-anode amplifier

$$2\Delta F=\frac{f_0}{Q_{ef}}\sqrt{\frac{1}{\sqrt[n]{M^2}}-1} \tag{6.12}$$

If $n=1$ and the frequency distortion $M=0.707$, the bandwidth $2\Delta F=\frac{f_0}{Q_{ef}}$ is that of the tuned circuit taken separately.

Example 6.2. Find the Q_{ef} and the selectivity of the tuned anode amplifier in Example 6.1. It is also known that $Q=60$; $f_0=500$ kilohertz; $f=510$ kilohertz.

Solution. The loaded Q-factor is

$$Q_{ef} = \frac{Q}{1+\frac{R_{0e}}{R_g}+\frac{R_{0e}}{R_{in}}+\frac{R_{0e}}{R_a}} = \frac{60}{1+\frac{40}{1,000}+\frac{40}{450}} \cong 53$$

The generalized detuning

$$x = \frac{2(f-f_0)}{f_0} Q_{ef} = \frac{2\times 10}{500} 53 = 2.1$$

Using $d = 1/Y$, the selectivity is

$$d = \sqrt{1+x^2} = \sqrt{1+2.1^2} = 2.32$$

A tuned amplifier incorporated into a broadcast or communication receiver is expected to cover a considerable frequency range. Such a range is usually broken up into a number of bands. Within each band, the amplifier is tuned with a variable capacitor, while the bands are changed by inserting a different coil into the tuned circuit for each band (Fig. 6.5).

As the frequency is increased within a band, the stage gain increases due to the higher resonant resistance of the tuned circuit

$$R_{0e} = \rho Q = \omega_0 LQ$$

where ρ is the characteristic impedance of the tuned circuit.

As we move from the low through the medium to the high frequencies the inductance of the tuned circuit decreases and so

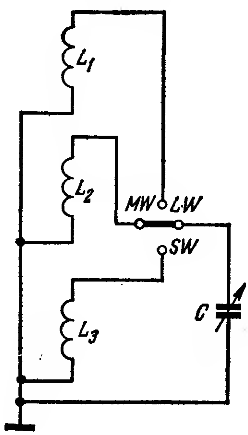


Fig. 6.5. Circuit for band-changing

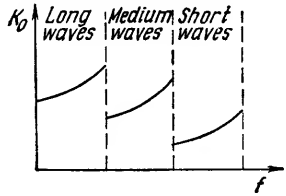


Fig. 6.6. Variations in gain with frequency in a tuned-anode amplifier

does the characteristic impedance ρ of the circuit. This is accompanied by a reduction in R_{0e} and, as a result, the stage gain drops. Figure 6.6 shows the stage gain as a function of frequency over the entire range.

Example 6.3. Determine the stage gain of a tuned amplifier using a type 1K1П valve at the extreme points of the range, if the frequency range is from 520 to 1,500 kilohertz, the tuned circuit inductance L is 167 microhenrys and $Q=40$.

Solution. Find the resonant resistance of the tuned circuit

(a) at $f_l = 520$ kilohertz

$$R_{0e\,min} = \rho Q = 2\pi \times 167 \times 10^{-6} \times 520 \times 10^3 \times 40 \cong 22 \text{ kilohms}$$

(b) at $f_h = 1{,}500$ kilohertz

$$R_{0e\,max} = 2\pi \times 167 \times 10^{-6} \times 1{,}500 \times 10^3 \times 40 \cong 63 \text{ kilohms}$$

Now find the gain. The mutual conductance of the 1K1П valve is $g_m = 0.75$ milliampere per volt

(a) at f_l, $K_{0\,min} = g_m R_{0e\,min} = 1.75 \times 22 = 16.5$

(b) at f_h, $K_{0\,max} = g_m R_{0e\,max} = 0.75 \times 63 = 47$

The Single-stage Amplifier with Transformer and Tapped-coil Coupling. In a transformer-coupled amplifier (Fig. 6.7*a*), the alternating component I_a of anode current flows through L_a and induces in the tuned-circuit coil L an emf of mutual induction at the signal frequency. The current in the tuned circuit and the voltage at the stage output are a maximum when the tuned circuit is tuned to the signal frequency.

Physically, L_a is a coil wound on the same former with L. There is no anode supply voltage applied to the tuned circuit, so that neither a bypass capacitor C_3 nor a coupling capacitor C_c are necessary. Grid bias to the next valve is supplied through L, because of which grid-leak resistor R_g may be eliminated from the circuit.

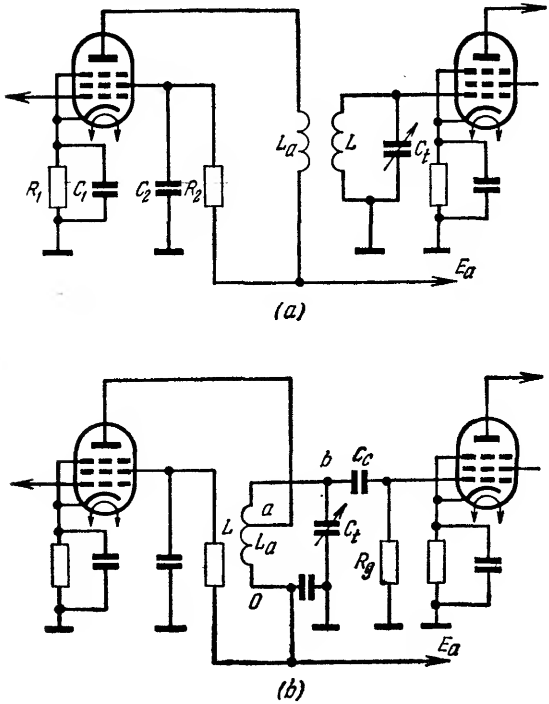


Fig. 6.7. Tuned amplifier circuits
(a) transformer-coupled amplifier;
(b) autotransformer-coupled amplifier connected to the anode

Figure 6.7*b* shows a single-stage amplifier with a tapped parallel tuned circuit. As to operating principle, it is similar to the arrangement of Fig. 6.1. The reason for its inclusion here is the role that is additionally played by the autotransformer action of the tapped coil.

In the tapped-coil arrangement (Fig. 6.7*b*), the anode circuit is loaded into a parallel tuned circuit whose resistance is $R'_{0e\,II} = R'_{0e}p^2$, where p is the coupling parameter defined here as the ratio of the inductance included between the taps to the inductance of the whole coil, or the tapping-down factor

$$p = \frac{L_a}{L}$$

In this case, the load voltage (across terminals *oa*) is given by

$$V_{oa} = I_a R'_{0e\,II} = I_a R'_{0e} p^2$$

Noting that

$$Z_a = R'_{0e\,II}$$

and using Eq. (1.9), we get

$$I_a = \frac{\mu V_g}{R_a + R'_{0e\,II}}$$

Since for pentodes, $R_a \gg R'_{0e\,II}$, we have

$$I_a = g_m V_g$$

Substituting this expression into the expression for V_{oa}, gives:

$$V_{oa} = g_m V_g R'_{0e} p^2$$

The output voltage (across terminals *ob*) is $L/L_a = 1/p$ times V_{oa}.
Or

$$V_{out} = V_{oa} \frac{1}{p} = g_m V_g R'_{0e} p$$

Hence, with a tapped parallel tuned circuit, the stage gain is

$$K_0 = \frac{V_{out}}{V_g} = g_m R'_{0e} p \tag{6.13}$$

In the transformer-coupled amplifier (see Fig. 6.7*a*) the stage gain depends not only on the valve and tuned circuit parameters but also on the degree of coupling between the tuned circuit and the anode circuit of the valve. At what is known as the *optimum coupling* the stage gain is a maximum which may exceed the gain of a stage in which no tapping-down for tuned-circuit connection is used. However, optimum coupling is seldom used. The degree of coupling between the valve and the tuned circuit with transformer coupling is given by the tapping-down factor or coupling parameter $p = \frac{M}{L}$.

It can be shown that the gain of a transformer-coupled stage is given by Eq. (6.13). At $p = 1$, this equation reduces to one giving the stage gain of a tuned-anode amplifier

$$K_0 = g_m R'_{0e}$$

For the transformer-coupled circuit

$$R'_{oe}=\frac{R_{oe}}{\frac{R_{oe}}{R_{in}}+1} \tag{6.14}$$

For tapped-coil connection, Eq. (6.5) holds.

The selectivity of both arrangements is expressed in terms of the effective or loaded figure-of-merit, Q_{ef}, as found from Eq. (6.10). If the coupling between the tuned circuit and the valve is loose and the value of R_a is large, the term $\frac{R_{oe}}{R_a}$ in the denominator may be disregarded. Then the Q_{ef} for the transformer-coupled circuit will be

$$Q_{ef}=\frac{Q}{1+\frac{R_{oe}}{R_{in}}} \tag{6.15}$$

and for the tapped-coil circuit

$$Q_{ef}=\frac{Q}{1+\frac{R_{oe}}{R_g}+\frac{R_{oe}}{R_{in}}} \tag{6.16}$$

It should be noted that tapping-down and transformer coupling reduce the loading of the circuit by the valve, i.e. the Q_{ef} is increased. As a result, the resonance curve of the amplifier becomes sharper.

The gain within each band increases as the frequency is raised for the same reasons as in the arrangement with a directly-connected tuned anode circuit.

Plots relating the gain to frequency appear in Fig. 6.8. This gain-frequency response is obtained by choosing the appropriate values of p for each band.

The Parallel-feed Single-stage Amplifier. The name *parallel-feed* applies to an arrangement in which the anode supply source,

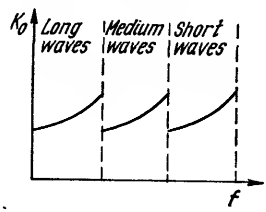


Fig. 6.8. Variations in gain with frequency in transformer-coupled and tapped-coil amplifiers

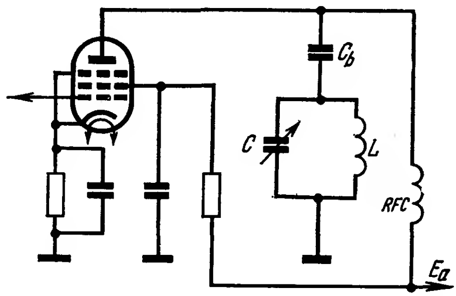


Fig. 6.9 Parallel-feed tuned amplifier

the valve and the tuned circuit are connected in parallel (Fig. 6.9). The circuits discussed earlier may be called *series-feed*, because in them the valve, tuned circuit, and d.c. supply are all connected in series.

In the parallel-feed circuit, the direct current from the anode supply source does not flow through the r.f. tuned circuit, but goes to the valve through a second coil *RFC* (r.f. choke). Direct current cannot flow through L because a blocking capacitor C_b is placed in the circuit to prevent it. On the other hand, the r.f. current generated by the valve can easily flow through C_b to the tuned circuit because the value of C_b is intentionally chosen to have a low reactance at the radio frequency. The r.f. current cannot flow through the d.c. supply because the inductance of *RFC* is intentionally made so large that it has a very high reactance at the radio frequency.

The output signal of the tuned circuit is applied directly to the grid of the next valve, so that there need be no grid resistor R_g.

Review Questions

1. Explain why the anode load of an amplifier is a parallel tuned circuit?
2. What would be the consequences of applying the output voltage of the anode tuned circuit to the grid of the next valve without a blocking capacitor?
3. How would the absence of a grid-leak resistor affect operation of the amplifier circuit?
4. Which circuit parameters affect the Q-factor of a tuned circuit?

5. Which circuit parameters outside the tuned circuit affect the bandwidth of the amplifier?
6. Why does the stage gain go up within a frequency band?
7. Why is it that in an amplifier with a directly-connected tuned circuit the gain on short waves is lower than on long waves?
8. How can the gain be maintained constant in the various bands of an amplifier with a transformer-coupled or a tapped-coil anode tuned circuit?
9. Explain the names "series-feed" and "parallel-feed" circuits.

32. Band-pass Amplifiers

As distinct from tuned amplifiers, band-pass amplifiers are mostly fixed-frequency amplifiers. That is, their tuned circuits do not have to be retuned when the receiver is in operation. Band-pass amplifiers are widely used as intermediate-frequency amplifiers in superheterodyne receivers.

In a band-pass amplifier, the anode load is a band-pass filter. Band-pass filters may have widely differing circuit configurations and may be connected to the anode of the amplifier valve in many ways. Therefore there exists a great variety of band-pass amplifier types.

Figure 6.10 shows the arrangement employed in most communication and broadcast radio receivers. The band-pass circuit is an r.f. transformer with tuned primary and tuned secondary, in which the primary circuit is directly connected to the anode.

The gain of a single-stage band-pass amplifier may be determined from its equivalent circuit (Fig. 6.11). Applying Théve-

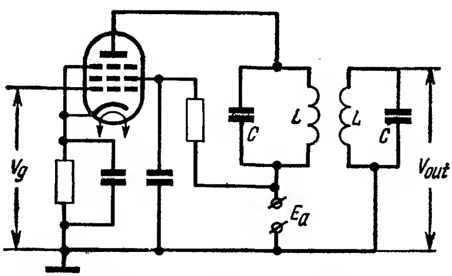


Fig. 6.10. Circuit of a single-stage band-pass amplifier

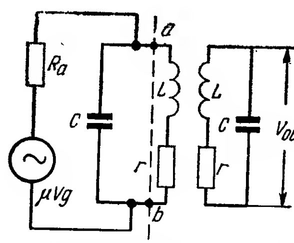


Fig. 6.11. Equivalent circuit of a single-stage band-pass amplifier

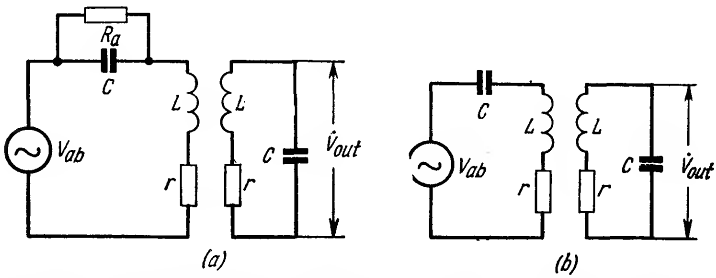


Fig. 6.12. Reduction of the equivalent circuit of a band-pass filter amplifier to that of a double-tuned circuit

nin's theorem, we can transform it into a simpler circuit (Fig. 6.12*a*). The emf of the equivalent generator (V_{ab}) is measured across the capacitor C of the primary (terminals *ab* in Fig. 6.11) with the right-hand part of the circuit being disconnected:

$$V_{ab} = I_a \frac{1}{\omega_0 C} = I_a \rho$$

The anode current is chiefly determined by the a.c. anode resistance R_a of the valve. Since $R_a \gg \frac{1}{\omega C}$, then

$$I_{ab} = g_m \dot{V}_g$$

or, after substitution

$$V_{ab} = g_m V_g \rho \tag{6.17}$$

With the value of R_a thus allowed for, the equivalent circuit becomes as shown in Fig. 6.12*b*. Since the reflected, or coupled-in resistance $\frac{1}{\omega^2 C^2 R}$ is small in comparison with r, the two circuits may be considered identical. For a band-pass filter

$$Q_f = \frac{V_{out}}{V_{ab}}$$

For a filter having identical primary and secondary circuits,

$$Q_f = Q \frac{\eta}{\sqrt{(\eta^2 - x^2 + 1)^2 + 4x^2}} \tag{6.18}$$

where $\eta = Qk$ = parameter defining overall coupling between the primary and secondary

k = coefficient of coupling

Q = Q-factor of the filter tuned circuits
x = generalized frequency shift from the resonant value.

At resonance, that is, when $x=\frac{2\Delta f}{f_0}Q=0$, we have the resonant transfer function of the filter

$$Q_{f0}=Q\frac{\eta}{\eta^2+1} \tag{6.19}$$

By definition, the output voltage of the filter at resonance is

$$V_{out\ 0}=Q_{f0}V_{ab}$$

Using Eq. (6.17), we obtain

$$V_{out\ 0}=Q_{f0}g_mV_g\rho$$

At resonance, the stage gain is

$$K_0=\frac{V_{out\ 0}}{V_g}=\frac{Q_{f0}V_{ab}}{V_g}$$

Replacing Q_{f0} and V_{ab} by their expressions, we have:

$$K_0=Q\frac{\eta}{\eta^2+1}\frac{g_mV_g\rho}{V_g}$$

Noting that $Q\rho=R_{0e}$, we finally obtain

$$K_0=\frac{\eta}{\eta^2+1}g_mR_{0e} \tag{6.20}$$

Comparing the gains of a single-stage band-pass amplifier and a single-stage tuned amplifier [Eq. (6.4)], it may be seen that they differ by a factor $\frac{\eta}{\eta^2+1}$. This factor has a maximum value 0.5 at critical coupling between the tuned circuits ($\eta=1$). Consequently, with similar valves and tuned circuits, the stage gain of a band-pass amplifier is only a half of the gain of a stage with a single-tuned circuit. This is because some of the energy is lost as it is transferred from the valve through the primary to the secondary side of the double-tuned circuit. A more correct picture is obtained when a comparison is made for the same bandwidth and the same valves. In this case, R_{0e} will not be the same for both circuits. It can be shown that a single-tuned circuit amplifier stage will have a gain which is 1.4 and not 2 times that of a stage in a band-pass amplifier. Comparison of four-stage amplifiers having the same bandwidth will show that a band-pass amplifier has a gain approximately 35 per cent higher than that of the amplifier employing single-tuned circuits.

The selectivity of a band-pass amplifier, as follows from the equivalent circuit of Fig. 6.12*b*, is decided by the frequency response of the band-pass filter. The equation of the relative response may be derived from the ratio of the Q_f, the transfer function of the filter, to Q_{f0}, the transfer function at resonance [Equations (6.18) and (6.19)]. The selectivity is then

$$Y = \frac{Q_f}{Q_{f0}} = \frac{\eta^2 + 1}{\sqrt{(\eta^2 - x^2 + 1)^2 + 4x^2}} \tag{6.21}$$

As will be recalled, the resonance curve of a band-pass filter may have one or two humps. To determine the conditions under which a single-hump curve changes into a double-hump one, it is necessary to find the values of generalized detuning x at which the ordinate of the curve is a maximum, $Y = Y_{max}$. This is done by analysing the roots of Eq. (6.21). First, we deem η constant, and equate the derivative of the discriminant to zero:

$$\frac{d\,[(\eta^2 - x^2 + 1)^2 + 4x^2]}{dx} = 0$$

$$\frac{d\,[(\eta^2 + 1)^2 - 2x^2(\eta^2 + 1) + 4x^2 + x^4]}{dx} = 0$$

$$-4x(\eta^2 + 1) + 4x^3 + 8x = 0$$

$$4x(-\eta^2 + x^2 + 1) = 0$$

Solving the last equation gives three roots. The first, $x_1 = 0$, corresponds to resonance, when $Y = 1$. The remaining roots, $x_{2,3} = \pm\sqrt{\eta^2 - 1}$, are located symmetrically about the point of resonance and characterize the peaks of the double-hump curve.

The roots may be real or imaginary, according to the value of η. Consider three cases of coupling:

1. Critical coupling ($\eta = 1$). The roots are equal: $x_1 = x_2 = x_3 = 0$, and there is only one peak, or a single-humped curve.

2. The coupling is less than critical ($\eta < 1$). The roots x_2 and x_3 are imaginary. The curve will also be single-humped, but the peak will be located below $Y = 1$.

3. $\eta > 1$. The roots are real and the curve has two humps.

Figure 6.13 illustrates the cases discussed above. As η is increased, the bandwidth becomes wider, and the ratio of the maximum to minimum response increases.

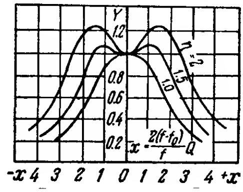


Fig. 6.13. Composite resonance curves of a double-tuned circuit

It is interesting to compare the bandwidth ratios $K_{bw\ 20db}$ for single- and double-tuned circuits. For a single-tuned circuit, $K_{bw\ 20db}=10$. For a double tuned circuit, when $\eta=0.5$, $K_{bw\ 20db}=4.1$; when $\eta=1$, $K_{bw\ 20db}=3.2$, and at η_{max}, $K_{bw\ 20db}=2.32$.

With any coupling, the resonance curve of a double-tuned circuit comes nearer to the ideal resonance curve than that of a single-tuned circuit. The best approximation is obtained with a maximum value of η. However, such coupling deepens the trough between the humps and results in that the gain varies within the bandwidth. This is why critical or nearly critical coupling should be maintained.

In some communication receivers with particularly stringent requirements for adjacent-channel rejection, multi-section band-pass filters are used as the load of the mixer or of the early stage of the intermediate-frequency amplifier. Figure 6.14 shows a filter consisting of four tuned circuits. As may be seen, the parameters of each tuned circuit have different values. The output of the filter is terminated in a resistance R equal to the image impedance of the filter. In addition to multi-section filters, i.f. amplifiers may use crystal filters for bandwidth control.

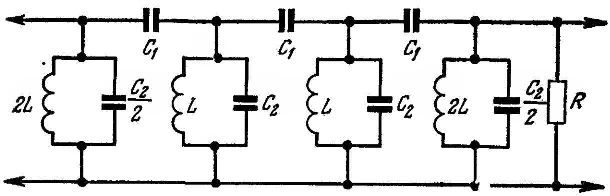


Fig. 6.14. Multiple-section filter circuit

Review Questions

1. Name the factors affecting the shape of the resonance curve of a band-pass amplifier.
2. What sets apart the resonance curve of a band-pass amplifier at critical coupling from that of a tuned amplifier with the same bandwidth?
3. How will the shape of the resonance curve and frequency distortion be affected in a band-pass amplifier by $\eta = 1.5$ and more?

33. Input Impedance of the Amplifier Valve

The input impedance of the amplifier valve in the next stage has a considerable effect on operation of a radio-frequency amplifier. The input impedance Z_{in} is defined as the opposition that the input circuit of the valve offers to alternating (signal) current:

$$Z_{in} = \frac{\dot{V}_g}{\dot{I}_g} \tag{6.22}$$

The input impedance Z_{in} is made up of two components, the resistance R_{in} and the reactance X_{in}. The value and nature of the input impedance are directly related to the impedance of the anode circuit, or, more specifically, the impedance of the anode tuned circuit, which is purely resistive at resonance, but becomes inductive or capacitative off resonance.

Of practical interest are cases where the r.f. amplifier load is resistive and inductive.

Figure 6.15 shows a triode amplifier circuit and its equivalent circuit for long, medium and, partly, short waves. In the equivalent circuit the valve is shown as a generator of emf μV_g and of internal resistance R_a. C_{gk} and C_{ag} are valve capacitances

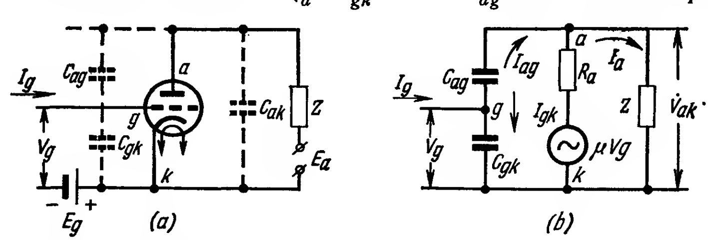


Fig. 6.15. Explaining the valve input impedance

coupling the input and output circuits of the valve, that is, anode and cathode (terminals a and k). The presence of a negative bias source $|E_g| > |V_g|$ in the grid circuit implies that there is no resistive (convection) current in the grid circuit; the input current of the valve is a displacement current due to interelectrode capacitances.

An idea about the effect of the input impedance can be gleaned from a vector diagram. Let us plot a vector diagram for the equivalent circuit when the anode load impedance is resistive, $Z=R$ (Fig. 6.16). To begin with, we lay off the vector of the input voltage V_g. μV_g is μ times V_g and is in phase with it. The anode current I_a passes through two resistances, R_a and R, and is in phase with μV_g. The output voltage V_{ak} appears across R and is in phase with I_a.

In the equivalent circuit all the components are connected at points a, g, k. The voltages between points gk and ak are known. Consequently, the unknown voltage between points ag is the sum of two known values

$$\dot{V}_{ag} = \dot{V}_g + \dot{V}_{ak}$$

Capacitative currents I_{ag} and I_{gk} flowing, respectively, through C_{ag} and C_{gk} lead V_{ag} and V_g by 90°.

The grid current I_g, as seen in the equivalent circuit, is a vector sum

$$\dot{I}_g = \dot{I}_{gk} + \dot{I}_{ag}$$

That the grid current I_g leads the output voltage V_g by 90° is an indication that the input impedance is capacitative in its effect. The resistive component of the grid current, represented by a projection of $\dot{I}_g$ upon $\dot{V}_g$, is zero. Consequently, the input resistance $R_{in} = \infty$.

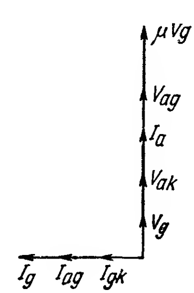


Fig. 6.16. Vector diagram for determining valve input resistance when the load is resistive

Thus, the input impedance of an amplifier valve with a resistive anode load in the long-, medium- and, partly, short-wave ranges, is composed of a resistive component R_{in}, which tends to infinity, and a reactive component which is determined by the input capacitance C_{in} of the valve.

For a triode, the input capacitance may be found from the equation:

$$C_{in}=C_{gk}+C_{ag}(1+K_0) \tag{6.23}$$

For pentodes,

$$C_{in}\cong C_{gk}$$

A resistive anode load is most favourable for amplifier operation because R_{in} is so high that its shunting effect may be neglected in calculating the gain and selectivity of the amplifier.

A vector diagram for an inductive load, $Z=R+jX$, is shown in Fig. 6.17. As before, the datum vector is V_g, such that μV_g is in phase with it. Because of the inductive properties of the anode load, the anode current lags behind μV_g by an angle φ such that

$$\tan\varphi=\frac{X}{R_a+R}$$

The load voltage V_{ak} leads I_a by an angle ψ such that

$$\tan\psi=\frac{X}{R}$$

where $\tan\psi>\tan\varphi$, and, consequently, $\psi>\varphi$.

The anode-to-grid voltage is

$$\dot{V}_{ag}=\dot{V}_g+\dot{V}_{ak}$$

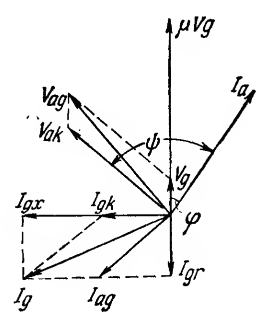


Fig. 6.17. Vector diagram for determining valve input impedance when the load is inductive

Capacitative currents I_{gk} and I_{ag}, respectively, lead V_g and V_{ag} by 90°. The resulting grid current is

$$\dot{I}_g = \dot{I}_{gk} + \dot{I}_{ag}$$

On resolving the vector $\dot{I}_g$ in two rectangular components, representing resistive and reactive components, I_{gr} and I_{gx}, we can see that the reactive component I_{gx} leads V_g by 90°. This component is capacitative in its effect and is, therefore, identified with the input capacitance C_{in} of the valve. The resistive component I_{gr} is in anti-phase with V_g, and not in phase, as usually is the case in ordinary electric circuits. Such a case is possible only in circuits containing valves and similar non-linear components. The input resistance R_{in} is negative. The concept of negative resistance has a definite physical meaning. While a positive resistance is associated with irrevocable loss of energy as heat, a negative resistance should be regarded as a source of electric energy. The presence of a negative input resistance is an evidence that a radio-frequency energy appears at the input of the valve, delivered from the anode circuit through C_{ag}. In other words, when the input resistance is negative, positive feedback is established from anode to grid through C_{ag} owing to the inductive properties of the anode load. This is undesirable feedback, because it may lead to self-oscillation of the amplifier.

Review Questions

1. Describe the effect that the resistive component of the input impedance has on the preceding circuit.
2. Describe the same for the reactive component.
3. Why can the impedance of the tuned anode circuit vary?
4. Explain the physical meaning of negative resistance.

34. Stability of Radio-Frequency Amplifiers

Apart from its theoretical implications, the stability of radio-frequency amplifiers has a very important practical significance. Any system, including an amplifier, may be considered stable if, when brought out of balance, it becomes stable after the disturbing force has been removed. Conversely, a system is unstable if it remains unbalanced after the external force is no longer applied. For instance, an amplifier is stable when an accidental electrical pulse produces in it a damped oscillatory process, that is, one dying away when the force is no longer applied. An amplifier is unstable if an oscillatory process changes into an undamped one and self-oscillation appears in the stage. In this case, the amplifier acts as an oscillator, or is self-excited.

It should be noted that an amplifier may become self-excited with or without a signal voltage applied to its input. One cause of self-oscillation in amplifiers is parasitic positive feedback.

Parasitic feedback in radio-frequency amplifiers may be due to anode-grid capacitance. In multi-stage amplifiers it may also be due to the common supply source. A further cause of feedback in the individual stages or in an amplifier as a whole may be capacitative or inductive interaction between wires, coils, and other amplifier components. This undesirable feedback may be prevented by optimum component layout, proper wiring, coil screening, particularly screening of the wires running to the control grid of the valve.

Parasitic Feedback in an Amplifier Through Anode-Grid Capacitance. Let us analyse a simplified circuit of a single-stage amplifier in which the tuned circuit of the previous stage is connected to its grid and has the same parameters as the tuned anode circuit (Fig. 6.18). Although their parameters are similar, the tuned circuits may be tuned to different frequencies because the input and output capacitances of the valve may affect them differently. As a result, the anode load may become inductive, the input resistance of the valve may go negative, and positive feedback may take place in the amplifier.

To determine the stability conditions, let us represent the grid circuit of the amplifier by an equivalent circuit (Fig. 6.19*a*). In this circuit, the valve on the grid side is represented by a negative input resistance, $-R_{in}$, connected in parallel with the tuned circuit. Fig. 6.19*b* shows another equivalent circuit in which the

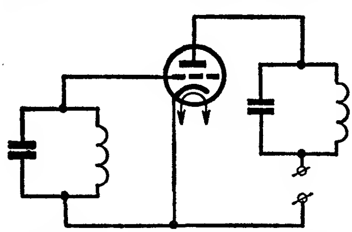


Fig. 6.18. Circuit diagram concerning self-oscillation of r. f. amplifiers

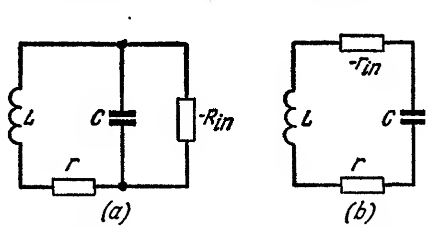


Fig. 6.19. Equivalent circuits of the grid circuit

parallel negative resistance $-R_{in}$ is replaced by its equivalent series negative resistance, $-r_{in}$. The negative resistance connected in series with the tuned circuit acts as a source of radio-frequency oscillations, reducing energy losses or, which is the same, building up energy in the tuned circuit until parasitic oscillation occurs. From this observation, we can intuitively conclude that for stable operation of the amplifier, that is, one without parasitic oscillation, the energy supplied to the tuned circuit to make up for the losses should be insufficient to lead to its build-up, that is, to an undamped oscillatory process. Another way of stating this is that the resultant resistance of tuned circuit must be positive. Therefore, the condition of stability may be represented in the following form: $|r| > |r_{in}|$, i. e. the total series loss resistance r must exceed the reflected or coupled-in resistance r_{in} in absolute value.

According to a theory of self-oscillation and stability of tuned amplifiers advanced by the Soviet scientist Siforov in 1931, the condition of stability for any multi-stage circuit is:

$$\omega C_{ag} g_m R_{0e}^2 p^2 \leqslant (0.18 \text{ to } 0.32) \qquad (6.24)$$

This expression holds for all the three circuit configurations discussed in Sec. 31. Without tapping-down or when the tuned circuit is conductively connected, the parameter $p = 1$. Multiplying both sides by g_m and noting that $g_m^2 R_{0e}^2 p^2 = K_0^2$, we obtain

$$\omega C_{ag} K_0^2 \leqslant (0.18 \text{ to } 0.32)\, g_m \qquad (6.25)$$

As is seen, the stability condition is most easily met at the low frequencies of the range, with a small capacitance C_{ag} and a small value of K_0.

From Eq. (6.25), we have

$$K_0 \leqslant \sqrt{\frac{(0.18 \text{ to } 0.32)\, g_m}{\omega C_{ag}}} \qquad (6.26)$$

The right-hand side of this expression may be called the *stable gain*

$$K_{st} \cong 0.42 \sqrt{\frac{g_m}{\omega C_{ag}}} \qquad (6.27)$$

Thus, the *stage gain K_0 of any amplifier must always be less than the stable gain as determined by Eq. (6.27)*, which suggests optimum connection of the tuned circuit to the anode. If K_0 for a stage with a conductively connected tuned circuit or one without tapping down exceeds K_{st}, the condition of stability is not met, and use should be made of transformer coupling or tapped-coil coupling.

The parameter p is given by

$$p \leqslant \frac{K_{st}}{K_0} \tag{6.28}$$

Example 6.4. Select the circuit configuration for a single-stage tuned amplifier using a 6K1П pentode and operating at $f=1{,}500$ kilohertz; the tuned circuit has a resistance of $R_{0e}=80$ kilohms.

Solution. Referring to the valve manual, we find that the 6K1П pentode has $C_{ag}=0.01$ picofarad and $g_m=1.85$ milliamperes per volt.

The gain is

$$K_0=g_m R'_{0e}$$

Putting $R_{0e} \cong R'_{0e}$, we obtain

$$K_0=1.85\times10^{-3}\times80\times10^{3}=148$$

The stable gain is

$$K_{st}=0.42\sqrt{\frac{g_m}{\omega C_{ag}}}=0.42\sqrt{\frac{1.85\times10^{-3}}{6.28\times1{,}500\times10^{3}\times0.01\times10^{-12}}}=59$$

Since $K_{st}<K$, the tuned circuit cannot be directly connected. The coupling parameter p for both transformer and tapped-coil coupling is

$$p=\frac{K_{st}}{K_0}=\frac{59}{148}=0.4$$

TABLE 6.1

Valve type	g_m, mA/V	$\frac{g_m}{C_{ag}}$, A/V-F	$K_{st}=0.42\sqrt{\frac{g_m}{C_{ag}}}$		
			150 kHz	1.5 MHz	15 MHz
6K1П	1.85	1.85×10^{11}	185	49	15
6K3	2.0	6.7×10^{11}	353	111	35
6K4	4.7	9.4×10^{11}	420	136	42
6K4П	4.4	12.5×10^{11}	486	157	48
1K1П	0.75	0.75×10^{11}	118	37	12

As follows from Eq. (6.27), a maximum and stable gain depends on the ratio $\frac{g_m}{C_{ag}}$. The maximum value of this ratio is obtained with low capacitance C_{ag} and high mutual conductance g_m. This condition is satisfied by r. f. pentodes because of which they are most commonly used valves in r. f. amplifiers.

Table 6.1 gives the values of g_m and the stable gain for the most common types of Soviet-made radio-frequency pentodes.

Parasitic Feedback in an Amplifier Through a Common Supply Source. An important cause of parasitic oscillation in radio and audio frequency amplifiers, especially multi-stage ones, is feedback via a common supply source, particularly the anode supply source. In the VHF-UHF bands feedback can also take place through the common filament supply. The coupling element in these cases is the internal impedance of the supply.

For an insight into the causes of parasitic oscillation in the case of a common supply source consider a three-stage tuned amplifier shown in diagrammatic form in Fig. 6.20. It may be added that the analysis that follows fully applies also to audio-frequency amplifiers.

We choose a spontaneous voltage change at the input of the 1st stage as the point of departure. We assume that the polarity of this instantaneous voltage is such that the grid goes positive. Noting that the valve inverts the voltage phase by 180°, and treating the valve as an a. c. generator, the polarities of instan-

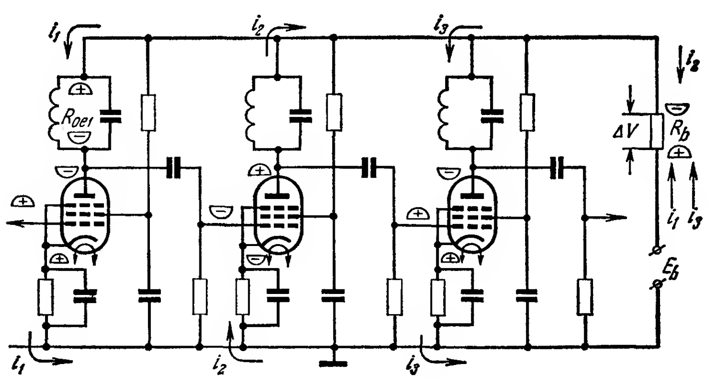


Fig. 6.20. Parasitic feedback through the common anode power source in a tuned milti-stage amplifier

taneous voltages for each valve and the direction of alternating anode current components i_1, i_2 and i_3 will be as shown in the circuit diagram of Fig. 6.20. As is seen, the currents i_1 and i_3 are in the same direction through the internal resistance R_b of the battery, and i_2 is in the opposite direction. In view of the gain of each stage, it should be assumed that $i_1+i_3>i_2$. At the same time, the instantaneous voltage ΔV across R_b is in a polarity corresponding to the direction of the total current. ΔV across R_b is in the same polarity as the voltage across the anode load resistance R_{0e1} of the first valve, and the two add together. Consequently, the grid voltage of the second valve will increase, and this is positive feedback which is likely to bring about parasitic oscillation. In two-stage amplifiers, the conditions for positive feedback do not exist. This may be proved by similar reasoning.

To prevent self-oscillation through the common supply source, so-called L-section decoupling filters, consisting of resistors R_f and capacitors C_f, are used. Such filters are connected into the anode circuits of multi-stage amplifiers as shown in Fig. 6.21. R_f has a value of several thousand ohms. The value of the capacitor depends on the frequency at which the amplifier operates and may be from several tenths to several hundredths of a microfarad.

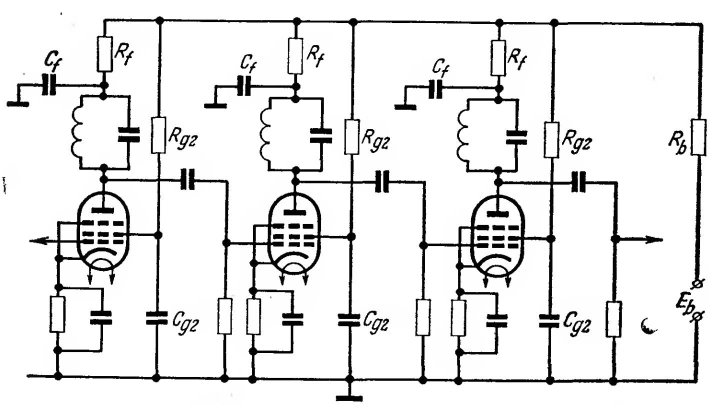


Fig. 6.21. Multi-stage circuit of a tuned amplifier with decoupling filters

Review Questions

1. What happens in an ustable amplifier?
2. Define positive feedback.
3. Define the condition necessary for self-oscillation of an amplifier.
4. Define the condition sufficient for self-oscillation of an amplifier.
5. What valve makes a stable amplifier?
6. What networks are used to neutralize self-oscillation through the common anode supply source?

35. Transistor Radio-Frequency Amplifiers

Like valve amplifiers, a transistor r. f. amplifier may be of the tuned or of the band-pass variety.

On long, medium and short waves, the transistor is usually connected into a common-emitter circuit, while in the VHF and UHF bands use is sometimes made of the common-base arrangement.

Transistor amplifiers differ from valve amplifiers chiefly in interstage coupling. The low input resistance of a transistor might have a considerable shunting effect on the tuned circuit of the preceding stage. This would sharply decrease the gain and selectivity of the stage. Operation of a transistor amplifier is also affected by the output resistance of the transistor, which is much lower than the output (a. c. anode) resistance of an amplifier valve. This is why transformer and tapped-coil coupling is used extensively in r. f. transistor amplifiers.

Figure 6.22 shows the circuit diagram of a common-emitter r. f. tuned amplifier.

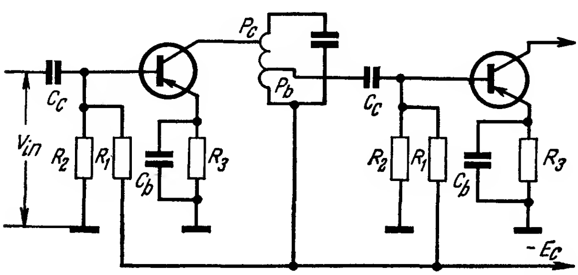


Fig. 6.22. Transistor tuned amplifier with tapped-coil coupling

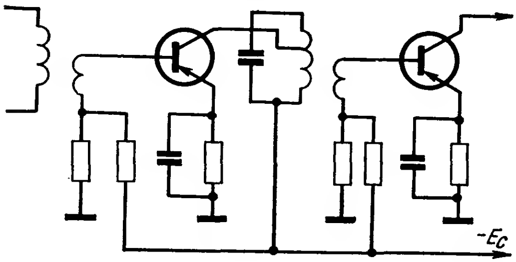


Fig. 6.23. Transistor tuned amplifier with transformer coupling

Interstage coupling is by a tapped-coil. The tapped-down voltage of the first stage is fed through a coupling capacitor C_c to the base of the second-stage transistor. The d. c. voltage is fed from a common source, E_c, through the tuned circuit to the collector, and through a divider R_1 and R_2 to the base.

The stability of operation of a transistor amplifier largely depends on the position of the quiescent operating point which is apt to shift as the transistor is heated. To stabilise the position of the operating point, the circuit employs negative direct-current feedback. This feedback is provided by R_3 connected in the emitter circuit. Such an arrangement is similar to the current feedback arrangement in valve circuits.

To eliminate a. c. feedback, R_3 is bypassed to earth by C_b. Should the operating point shift due to temperature changes, it will be restored by the feedback voltage built up across R_3 and applied to the transistor base.

In an alternative arrangement shown in Fig. 6.23, interstage transformer coupling is augmented by tapped-coil coupling from the tuned circuit to the collector. It should be noted, however, that in both configurations the tuned circuit may be connected to the collector circuit directly, provided the output resistance of the transistor is sufficiently high.

The circuit diagram of a band-pass transistor amplifier appears in Fig. 6.24. The load is an r. f. transformer with tuned primary and tuned secondary. The shunting effect of the collector circuit on the primary, and of the input of the next transistor on the secondary is minimized by tapping the coils.

As with valve amplifiers, the main method of transistor-amplifier analysis is the equivalent-circuit technique. In this technique, the transistor is treated as a linear two-port network.

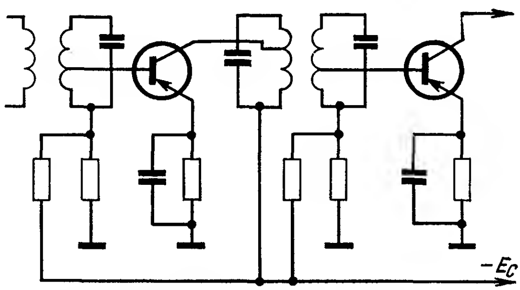


Fig. 6.24. Transistor band-pass amplifier

Analysis of r. f. circuits is based on the Y-(short-circuit admittance) parameters, derived from a pi-equivalent circuit, such as shown in Fig. 6.25*a*. The equivalent circuit shows the admittances Y_{11}, Y_{12}, Y_{22} and a constant-current generator simulating the amplifying action of the transistor.

The current of the constant-current generator is given by the product $Y_{21}V_{be}$ in which Y_{21}, defined as the ratio of the collector current to the emitter voltage with the input short-circuited, is in effect the mutual conductance, g_m, of the transistor, so that

$$Y_{21}V_{be}=\dot{g}_m V_{be}$$

It should be noted that $\dot{g}_m$ is a complex quantity defined as

$$\dot{g}_m = Y_{21} = g_{m0}/(1+jf_0/f_\alpha) \tag{6.29}$$

where $g_{m0}=y_{21}$ = low-frequency mutual conductance
f_0 = operating frequency
f_α = alpha cut-off frequency defined as one at which the absolute value of g_m is $1/\sqrt{2}$ times its low-frequency value (3 db down).

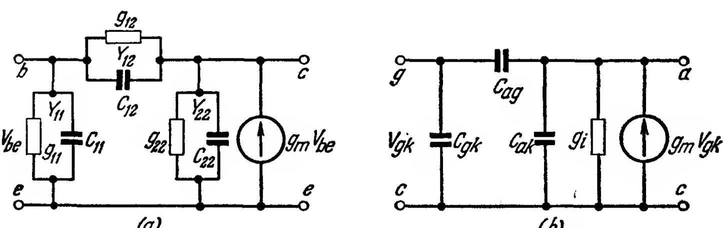


Fig. 6.25. Pi-equivalent circuits of a transistor and a valve

The equivalent circuit of the transistor is similar to that for a valve (Fig. 6.25*b*). In each circuit, Y_{21} describes feedback from the output to the input of the amplifying element. In the valve, this feedback is due to the grid-anode capacitance C_{ag}; in the transistor it is caused by the conductance and capacitance in parallel. It should be noted that in transistor circuits the feedback admittance is greater than it is in valve circuits, for which reason the former are less stable than the latter.

In analysis of a transistor amplifier it is important to allow for the complex character of $\dot{g}_m$, which implies its dependence on frequency.

All elements of the equivalent circuit of a transistor may be calculated from the known low-frequency admittances y_{11}, y_{12}, $y_{21}=g_{m0}$, y_{22}, and from the specified r.f. parameters, namely, the collector-junction capacitance, the spreading (bulk) resistance of the base r_b', and the alpha cut-off frequency, f_α.

The appropriate design equations have the form:

$$\left.\begin{aligned} g_{11} &= \frac{y_{11}+a^2/r_b'}{1+a^2} \\ g_{12} &= \frac{y_{12}+a\omega_0 C_c}{1+a^2} \\ g_{22} &= y_{22}+\frac{a\omega_0 C_c(1+r_b' y_{21})-a^2 y_{12}}{1+a^2} \end{aligned}\right\} \tag{6.30}$$

$$\left.\begin{aligned} C_{11} &= \frac{1-y_{11}r_b'}{\omega_\alpha r_b'(1+a^2)} \\ C_{12} &= \frac{C_c-y_{12}/\omega_\alpha}{1+a^2} \\ C_{22} &= \frac{C_c(1+r_b' y_{21})-y_{12}/\omega_\alpha}{1+a^2} \end{aligned}\right\} \tag{6.31}$$

where $a=f_0/f_\alpha$ is the relative frequency.

Tuned Transistor Amplifier. Consider the amplifier of Fig. 6.22. Its equivalent circuit is shown in Fig. 6.26. For simplicity, we shall ignore the effect of feedback through Y_{12}.

In the equivalent circuit of Fig. 6.26, the collector circuit is approximated by a constant-current generator $g_m V_{be}$ and the output admittance Y_{22} composed of g_{22} and C_{22}. The tuned circuit coupled to the collector through a tapped coil consists of L, C, and g_c which represents the losses within the tuned circuit.

The effect of the input of the next stage is taken care of by Y_{11} made up of g_{11} and C_{11}, also connected to the tuned circuit

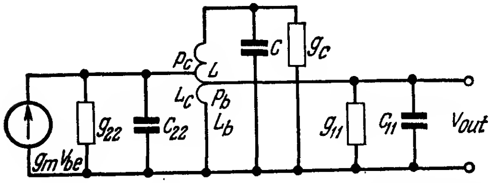


Fig. 6.26. Equivalent circuit of a single-stage transistor amplifier with tapped-coil coupling

through a tap on the coil. The parameters p_c and p_b give the amount of tapping-down on the collector and base sides and are defined as

$$p_c = L_c/L \qquad p_b = L_b/L \tag{6.32}$$

For simplicity, the equivalent circuit of Fig. 6.26 may be reduced to that of Fig. 6.27. Now, on lumping together all conductances and all capacitances, we obtain the final equivalent circuit of Fig. 6.28, where the total capacitance of the tuned circuit is

$$C_\Sigma = C + C_{22}p_c + C_{11}p_b \tag{6.33}$$

and the total conductance is

$$G_{0e} = g_c + g_{22}p_c + g_{11}p_b \tag{6.34}$$

Referring to the equivalent circuit of Fig. 6.28, the stage gain at resonance is

$$\dot{K}_0 = \dot{V}_{out\,0}/\dot{V}_{be}$$

The output voltage at resonance, $\dot{V}_{out\,0}$, is $1/p_b$ times the tuned-circuit voltage at resonance, $\dot{V}_{LC0}$, that is

$$\dot{V}_{out\,0} = \dot{V}_{LC0}p$$

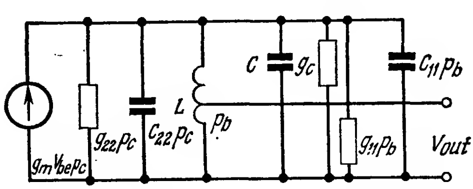


Fig. 6.27. Reduced equivalent circuit of a transistor stage

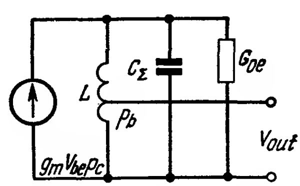


Fig. 6.28. Final equivalent circuit of a transistor stage

where

$$\dot{V}_{LC0} = g_m V_{be} p_c R_{0e}$$

After appropriate manipulations,

$$\dot{K}_0 = g_m p_c p_b R_{0e}$$

The magnitude of this expression gives the design equation of the gain at resonance

$$K_0 = |g_m| p_c p_b R_{0e} = |g_m| p_c p_b (1/G_{0e}) \qquad (6.35)$$

where $|g_m| = g_{m0}/\sqrt{1+a^2}$ is the magnitude of the mutual conductance.

At an operating frequency f_0 far from the cut-off frequency f_α, we may deem

$$g_m = (\text{approx.})\, g_{m0} = y_{21}$$

A similar expression has been derived in analysis of the valve tuned amplifier. However, Eq. (6.13) contains only one value of p, corresponding to tapped-coil coupling on one side of the circuit.

Example 6.5. Calculate the gain of a single-stage amplifier using a П-401 transistor, if $f_0 = 465$ kilohertz, $p_c = 0.4$, $p_b = 0.05$, $L = 290$ microhenrys and $Q_{ef} = 50$.

Solution. For a П-401 transistor, $f_\alpha = 10$ megahertz, $y_{21} = g_{m0} = 65{,}000$ micro-mhos (or micro-siemens).

Determine the effective resistance of the loaded tuned circuit at resonance

$$R_{0e} = \rho Q_{ef} = 2\pi f_0 L Q_{ef} = 6.28 \times 465 \times 10^3 \times 290 \times 10^{-6} \times 50 = 42{,}000 \text{ ohms}$$

Since $f_0 \ll f_\alpha$, put $|g_m| = g_{m0}$

The gain is

$$K_0 = |g_m| p_c p_b R_{0e} = 65{,}000 \times 10^{-6} \times 0.05 \times 0.4 \times 42 \times 10^3 = 55$$

For maximum gain, the coupling parameters on the collector and base sides should be as follows

$$p_c = (\text{approx.}) \sqrt{\frac{G_{0e} - g_c}{2g_{22}}} \qquad (6.36)$$

$$p_b = (\text{approx.}) \sqrt{\frac{G_{0e} - g_c}{2g_{11}}} \qquad (6.37)$$

The selectivity of the transistor stage is found as that approximated in terms of the effective (loaded) Q-factor of the tuned circuit given by

$$Q_{ef} = R_{0e}/\rho = 1/G_{0e}\rho \tag{6.38}$$

The gain-frequency response is described by Eq. (6.11).

All that has been derived for a tapped-coil-coupled transistor amplifier mainly applies to a transformer-coupled transistor amplifier as well.

Band-pass Transistor Amplifier. The band-pass transistor amplifier of Fig. 6.24 may be analysed by a similar technique. In the general case, the gain of a band-pass transistor amplifier is given by

$$K = |g_m| \rho_1 p_c p_b Q_f \tag{6.39}$$

where $|g_m|$ = magnitude of the mutual conductance
$\rho_1 = 1/\omega_0 C_1$ = characteristic impedance of the primary
p_c and p_b = coupling parameters (tapping-down factors) on the collector and base side, respectively
Q_f = transfer function of the band-pass circuit.

If the band-pass circuit is an r.f. transformer made up of two identical resonant circuits (Fig. 6.24), Eq. (6.39) should include the transfer function of the r.f. transformer at resonance [see Eq. (6.19)]

$$Q_{f0} = Q\eta/(\eta^2 + 1)$$

Then the gain of the band-pass transistor amplifier at resonance will be

$$K_0 = |g_m| R_{0e} \frac{\eta}{\eta^2+1} p_c p_b \tag{6.40}$$

Eq. (6.40) checks well with the expression for the gain of the valve band-pass amplifier. It may be added that Eq. (6.39) holds for any band-pass amlpifier using a multi-section band-pass circuit, if Q_f is known.

The coupling parameters p_c and p_b may be found from

$$p_c = \text{(approx.)} \sqrt{\frac{G_{0e} - g_{c1}}{2g_{22}}} \tag{6.41}$$

$$p_b = \text{(approx.)} \sqrt{\frac{G_{0e} - g_{c2}}{2g_{11}}} \tag{6.42}$$

where $G_{0e} = 1/R_{0e}$ = conductance of the tuned circuit at resonance
g_{c1} and g_{c2} = loss conductances of the tuned circuits.

The gain-frequency response for a band-pass transistor amplifier is that of the tuned-primary, tuned-secondary r.f. transformer used and is described by Eq. (6.21).

Stability of Transistor Amplifiers. In deriving equivalent circuits for transistor amplifiers and in their analysis we have ignored the feedback admittance Y_{12} shown in Fig. 6.25.

A more rigorous analysis shows that this admittance has a direct bearing on the magnitude and polarity of the input conductance of the transistor stage, so that it becomes negative at a frequency close to the resonant value. In practical amplifiers, it is hardly possible to tune all the stages precisely to the same frequency. Therefore, it may so happen that as one of the stages is slightly detuned from resonance, the input conductance might go negative. Then, as in a valve amplifier, the conditions would be created for self-oscillation.

Mathematical analysis shows that, as with valve amplifiers, the stable gain is decided by $\sqrt{g_m/\omega_0 C_{fb}}$, where for the transistor C_{fb} is C_{12} (Fig. 6.25), and for valve, C_{ag}.

The C_{12} of transistors is hundreds of times as great as the C_{ag} of valves, while g_{m0} for transistors is 10 to 20 times the g_m value for valves. Therefore, all other things being equal, the stable gain of a transistor stage is one-fifth to one-sixth of that of a comparable valve stage.

The limitation imposed on the transistor amplifier by the stability criterion may be removed by eliminating the internal feedback.

One way to do this is by applying an external feedback, equal in magnitude and opposite in phase to the internal feedback. This technique is called *unilateralisation*, and the circuit is said to be *unilateralised*. A unilateralisation circuit should consist of

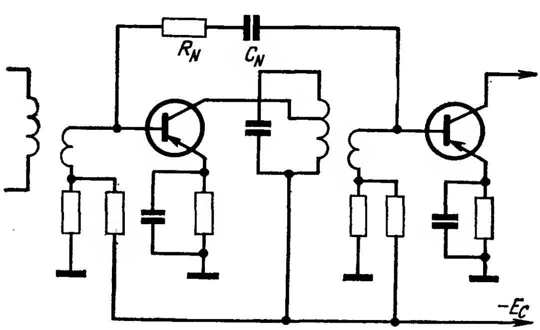


Fig. 6.29. Unilateralised transformer-coupled tuned transistor amplifier

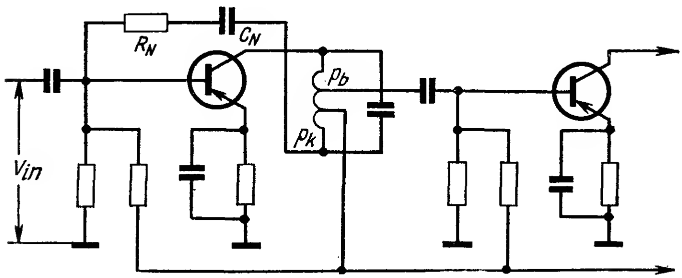


Fig. 6.30. Unilateralised tapped-coil-coupled tuned transistor amplifier

the same elements as the internal feedback path Y_{12} shown in Fig. 6.25. They are usually symbolized C_N and g_N.

A parallel unilateralising network produces an undesirable d.c. coupling between the points of connection. Therefore, preference should be given to a series combination of C_N and R_N. Their values can be calculated as follows:

$$C_N = (\text{approx.})\, 0.9C_c/m \tag{6.43}$$

$$R_N = (\text{approx.})\, 1.1m/\omega_0 C_c \tag{6.44}$$

where $m = p_b/p_c$.

The circuit diagrams of unilateralised tuned transistor amplifiers are shown in Figs. 6.29 and 6.30. With transformer coupling, the unilateralising network is connected to the coupling coil, so that the phase in the unilateralising circuit is inverted. With tapped-coil coupling, the same effect is obtained by connecting the unilateralising circuit to the bottom terminal of the

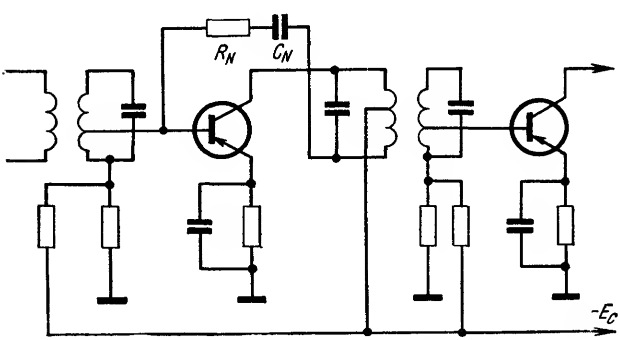


Fig. 6.31. Unilateralised band-pass transistor amplifier

output tuned circuit. The same arrangement is used in the band-pass amplifier (Fig. 6.31).

Satisfactory results can also be obtained by an external capacitor alone. In this case the process is referred to as *neutralisation.*

Review Questions

1. Why do the gain and selectivity of a transistor stage depend on the input resistance of the succeeding stage?
2. What valve circuit is the common-emitter analogous to?
3. What two-port parameters are used in analysis of the r. f. behaviour of transistor circuits?
4. Which of the two-port parameters describes feedback in a transistor circuit?
5. What is the polarity of unilateralising feedback?

36. Calculation of Radio-frequency Amplifiers

This section deals with tuned and band-pass valve amplifiers and also a transistor amplifier.

Calculation of a Wideband Valve Tuned Amplifier

Given.

1. Frequency band, f_l-f_h.
2. Required gain, K_0.
3. Q_{ef} of the tuned circuit, deemed to be constant over the range.
4. Tuned-circuit inductance, L.
5. Number of stages, n.

To Find:

1. Type of valve.
2. Connection of the tuned circuit to the anode.
3. The coupling parameter p and inductance L_a of the coil, for transformer and tapped-coil coupling.
4. Minimum gain, $K_{0\,min}$.

Design Procedure:

1. Select the type of valve and establish the operating voltages from its characteristics.
2. Find the resonant resistance of the tuned circuit for the limiting frequencies of the band

$$R_{0e\ min} = \omega_l L Q_{ef}$$
$$R_{0e\ max} = \omega_h L Q_{ef}$$

3. Choose the type of connection for the tuned circuit from the condition of stability

$$K_{0\ max} = R_{0e\ max} g_m \leqslant 0.42 \sqrt{\frac{g_m}{\omega_h C_{ag}}} = K_{st}$$

In certain cases, even though the stability condition is satisfied by direct connection, preference may be given to transformer or tapped-coil coupling in order, for example, to improve the response on all bands.

4. Determine the coupling parameter (tapping-down factor) for transformer and tapped-coil coupling

$$p \leqslant \frac{K_{st}}{K_{0\ max}}$$

5. Find the inductance of the anode coil:
(a) transformer coupling

$$L_a = \left(\frac{p}{k}\right)^2 L$$

with $k = 0.4$ to 0.6;
(b) tapped-coil coupling

$$L_a = pL$$

6. Determine the minimum gain of the entire amplifier

$$K_{0\ min} = (g_m R_{0e\ min} p)^n$$

If $K_{0\ min}$ happens to be smaller than the required value, use another type of valve with a higher value of $\frac{g_m}{\omega C_{ag}}$ and recalculate the stage.

Calculation of the Band-pass Amplifier

Given:

1. Operating frequency, f_0.
2. Bandwidth, $2\Delta F$.
3. Frequency distortion, M.
4. Gain, K_0.
5. Selectivity $d=\frac{1}{Y}$, for the amount off resonance equal to Δf.
6. Number of coupled tuned circuits m (including the coupled tuned circuits of the mixer).

To Find:

1. Type of valve.
2. Q of the band-pass circuit.
3. C and L parameters.
4. Coefficient of coupling, k, for the coupled tuned circuits.
5. Gain K_0 (as a check).
6. Selectivity, d, (as a check) with Δf off resonance.

Design Procedure:

1. Select the required type of valve and establish its operating voltages from its characteristics.
2. Determine the frequency distortion for a coupled circuit:

$$M' = \sqrt[m]{M}$$

3. Find the Q of the tuned circuits from considerations of the desired bandwidth, using the chart of Fig. 6.32.
Entering the chart with $\eta=0.5$ to 1.5, find the value of the generalised detuning x_1, using the curve M'. For a bandwidth of $2\Delta F=6$ or 7 kilohertz, the value of η should be close to unity. For a broader bandwidth, η should be greater than unity.
Find the Q-factor:

$$Q=\frac{f_0 x_1}{2\Delta F}$$

The tuned circuit possessing the obtained Q-factor must be realizable. The Q-factor seldom exceeds 100-150.

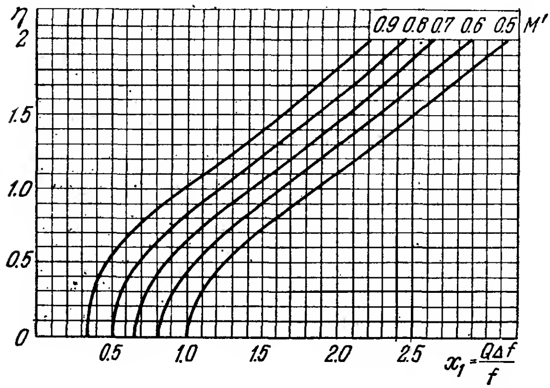


Fig. 6.32. Design curves for determining the Q of the band-pass circuit

4. Check whether the selectivity condition can be met with the chosen value of η

$$d=\left(\frac{\sqrt{(1-x_2^2+\eta^2)^2+4x_2^2}}{2\eta}\right)^m, \text{ with } \eta \geqslant 1 \tag{6.45}$$

$$d=\left(\frac{\sqrt{(1-x_2^2+\eta^2)^2+4x_2^2}}{\eta^2+1}\right)^m, \text{ with } \eta \leqslant 1 \tag{6.46}$$

where $x_2=\frac{2\Delta f}{f_0}Q$, and Δf is a specified amount off resonance (usually 10 kilohertz).

If d happens to be less than specified, change the parameter η and recalculate the Q.

5. Select the tuned-circuit capacitance of the band-pass circuit from the following two conditions:

(a) maximum stable gain, using Eq. (6.24) and putting $\omega_0 C_{ag} g_m R_{0e}^2 = 0.25$. Noting that $R_{0e}=Q/\omega_0 C$ and deeming $C = C_1 = C_2$, we get

$$C \geqslant 2Q\sqrt{\frac{C_{ag}g_m}{\omega_0}} \tag{6.47}$$

(b) minimum shunting effect of the a.c. anode resistance of the valve on the tuned circuit, Eq. (6.31).

This criterion stems from the fact that one side of the band-pass circuit is shunted by the a.c anode resistance of the valve, so that the two sides differ in the Q-factor, and the resonance curve becomes unsymmetrical. To minimize the asymmetry, the change in the Q-factor should not exceed 25 per cent. In other words, choose either $R_{0e} \leqslant 0.25R_a$, or $R_{0e} = Q/\omega_0 C$, where

$$C \geqslant \frac{4Q}{\omega_0 R_a} \tag{6.48}$$

Of the values obtained select the largest.

6. Find the tuned-circuit inductance

$$L = \frac{2.53 \times 10^{10}}{C f_0^2}$$

where L is in microhenrys, C in picofarads, and f_0 in kilohertz.

7. Find the coefficient of coupling between the tuned circuits

$$k = \frac{\eta}{Q}$$

8. Determine the mutual inductance M between the tuned circuits

$$M = kL$$

9. Find the resonant resistance of the tuned circuit

$$R_{0e} = \sqrt{\frac{L}{C}}\, Q$$

It should be noted that the final band-pass circuit of the i.f. amplifier is shunted by the detector stage. Therefore, the value of R_{0e} ought not to exceed 100 kilohms. If $R_{0e} > 100$ kilohms, go through the calculation procedure again, beginning with step 5, increasing the capacitance C of the tuned circuit.

10. Calculate the gain

$$K_0 = \left(\frac{\eta}{\eta^2+1} g_m R_{0e}\right)^{m-1}$$

If the obtained value of K_0 happens to be smaller than the assumed one, select a valve with a higher mutual conductance and recalculate where necessary.

11. Calculate the auxiliary circuit components (the decoupling filters $R_f C_f$ and the screen-grid network $R_{g2} C_{g2}$ as shown in Fig. 6.21):

(a) Calculate the resistances (in kilohms):

$$R_f = \frac{E_a - V_{a0}}{I_{a0}}$$

$$R_{g2} = \frac{E_a - V_{g20}}{I_{g20}}$$

where E_a is the anode supply voltage in amperes; V_{a0} is the desired d. c. voltage at the anode in volts; $V_{g2\,0}$ is the desired d. c. voltage at the screen grid; and I_{a0} and $I_{g2\,0}$ are the direct components of the anode and screen-grid currents in milliamperes.

(b) Calculate the capacitances:

At the low frequencies, C_f should have a reactance which is one-tenth to one-twentieth of R_f:

$$C_f \geqslant \frac{1 \text{ to } 2}{628 f_l R_f}$$

C_f will be in microfarads, if f is in megahertz and R_f in kilohms.

$$C_{g2} \geqslant 2.5 \times 10^3 \frac{C_{gk} C_{ak}}{C_{ag}}$$

C_{g2} will be in picofarads, if C_{gk} and C_{ak} are in picofarads.

Example 6.5. Calculate an intermediate-frequency amplifier for a battery-powered receiver, using the following data: $f_0 = 465$ kilohertz; $K_0 \geqslant 1{,}000$; $2\Delta F = 9$ kilohertz; $d = \frac{1}{Y} \geqslant 20$ at $\Delta f = 10$ kilohertz; $M = 0.76$; $m = 3$.

Solution. 1. Use a high-frequency 1К1П pentode for the amplifier. The pentode parameters for standard operating conditions are: $g_m = 0.75$ milliampere per volt; $R_a = 750$ kilohms; $\mu = 570$; $C_{ag} = 0.01$ picofarad.

2. Determine the frequency distortion for the band-pass circuit (i.f. transformer) taken alone:

$$M' = \sqrt[m]{M} = \sqrt[3]{0.76} = 0.91$$

3. Putting $\eta = 1.15$ and, using the chart of Fig. 6.32, determine $x_1 = 1.2$; then

$$Q = \frac{f_0 x_1}{2\Delta F} = \frac{465 \times 1.2}{9} = 62$$

4. Find the selectivity. First, determine

$$x_2 = \frac{2\Delta f}{f_0} Q = \frac{20 \times 62}{465} = 2.66$$

$$d = \left(\frac{\sqrt{(1 - x_2^2 + \eta^2)^2 + 4x_2^2}}{2\eta} \right)^m$$

$$= \left(\frac{\sqrt{(1 - 2.66^2 + 1.2^2)^2 + 4 \times 2.66^2}}{2 \times 1.2} \right)^3 = 2.93^3 = 25$$

The result corresponds to the specified value.

5. Now, select the tuned-circuit capacitance

$$C \geqslant 2Q\sqrt{\frac{C_{ag}g_m}{\omega_0}} = 2\times 62\sqrt{\frac{0.01\times 10^{-12}\times 0.75\times 10^{-3}}{6.28\times 465\times 10^3}}$$
$$= 198 \text{ picofarads}$$

$$C \geqslant \frac{4Q}{\omega_0 R_a} = \frac{4\times 62\times 10^{12}}{6.28\times 465\times 10^3\times 750\times 10^3} = 113 \text{ picofarads}$$

We shall put $C = 220$ picofarads.

6. Determine the tuned-circuit inductance

$$L = \frac{2.53\times 10^{10}}{Cf_0^2} = \frac{2.53\times 10^{10}}{220\times 465^2} = 533 \text{ microhenrys}$$

7. Find the coefficient of coupling between the tuned circuits

$$k = \frac{\eta}{Q} = \frac{1.15}{62} = 0.0185$$

8. Determine the mutual inductance

$$M = kL = 0.0185\times 533 = 9.8 \text{ microhenrys}$$

9. Determine the resonant resistance of the tuned circuit

$$R_{0e} = \sqrt{\frac{L}{C}}\,Q = \sqrt{\frac{533\times 10^{-6}}{220\times 10^{-12}}}\times 62 = 97 \text{ kilohms}$$

10. Compute the gain

$$K_0 = \left(\frac{\eta}{\eta^2+1}g_m R_{0e}\right)^{m-1}$$
$$= \left(\frac{1.15}{1.15^2+1}0.75\times 10^{-3}\times 97\times 10^3\right)^2 = 39^2 = 1{,}520$$

The value of K_0 considerably exceeds the specified gain. Hence, the calculation has been carried out correctly.

Calculation of the Transistor R. F. Amplifier. Consider the calculation of the fixed-frequency, single-tuned r.f. transistor amplifiers shown in Figs. 6.22 and 6.23.

Given:

1. Operating frequency, f_0.
2. Bandwidth, $2\Delta F$.
3. Frequency distortion, M.
4. Gain.

5. Selectivity.
6. Number of tuned circuits, n (including the one in the frequency changer in the case of an i.f. amplifier).

To Find:

1. Type of transistor.
2. Tuned-circuit parameters L and C.
3. Coupling parameters p_c and p_b.
4. Gain (as a check).
5. Selectivity (as a check).

Design Procedure:

1. Select a transistor such that the operating frequency is much lower than its alpha cut-off frequency, f_α.

2. Using Eqs. (6.30) and (6.31), find the equivalent-circuit parameters g_{11}, g_{12}, g_{22}, C_{11}, C_{12}, and C_{22}.

3. Find the tuned-circuit inductance

$$L = \frac{2.53 \times 10^{10}}{Cf_0^2}$$

where C is to be from 200 to 600 picofarads.

4. Find the loss (leakage) conductance of the tuned circuit

$$g_{LC} = \frac{1}{\rho Q} = \frac{1}{Q}\sqrt{C/L}$$

The Q-factor is set at 100 to 200.

5. Find the bandwidth of the single tuned circuit

$$2\Delta F_{tc} = \frac{2\Delta F}{\sqrt{\frac{1}{\sqrt[n]{M^2}} - 1}}$$

6. Find the effective conductance of the tuned circuit, G_{0e}, that will secure the desired bandwidth $2\Delta F_{tc}$:

$$G_{0e} = \frac{1}{Q_e \rho} = \frac{2\Delta F_{tc}}{f_0 \rho}$$

7. Using Eqs. (6.36) and (6.37), find p_c and p_b.

8. Using Eq. (6.35), find the stage gain and overall gain.

9. Using Eq. (6.33), find the tuned-circuit capacitor:

$$G_{tc} = C - C_{22}p_c - C_{11}p_b$$

10. Using Eqs. (6.43) and (6.44), calculate the unilaterali-sation network parameters, C_N and R_N.

11. Using Eqs. (6.11) and (6.38), find the selectivity of the amplifier.

SUMMARY

1. Tuned amplifiers are used to amplify radio-frequency oscillations in a definite range of frequencies.

2. Any tuned amplifier configuration has a varying gain over the range. Increase of frequency within each band is accompanied by an increase in gain.

3. In arrangements with directly connected tuned circuits the gain changes between bands.

4. Transformer and tapped-coil coupling flattens the gain between bands.

5. Maximum gain is obtained in configurations with directly connected tuned circuit.

6. The selectivity of a tuned amplifier is decided by the properties of the loaded tuned circuit as represented by a series equivalent circuit.

The sharpness of the resonance curve of an amplifier depends on the resistances shunting the tuned circuit. The shunting effect is a maximum in arrangements with direct connection or no tapping-down.

7. Band-pass amplifiers are chiefly employed to amplify intermediate-frequency signals in superheterodyne receivers. The band-pass circuit used as the anode load is usually an r.f. transformer with tuned primary and tuned secondary.

8. Parasitic feedback in radio-frequency amplifiers may result in self-oscillation.

9. Self-oscillation cannot take place in a stable amplifier.

10. The condition of stability is one of criteria for the selection of tuned-circuit connection to the anode. Use of trans-former or tapped-coil coupling instead of direct connection will improve the stage stability but will decrease the gain.

11. Radio-frequency amplifiers widely use transistors. The transistors are connected mainly into the common-emitter circuit.

12. Interstage coupling in transistor amplifiers is by transfor-mers or tapped coils. With this type of coupling the high output resistance of the preceding stage can be readily matched to the low input resistance of the next.

Problems

6.1. Determine the gain of a single-stage tuned amplifier from the following data: valve type 6K4П; tuned-circuit inductance $L = 2{,}040$ microhenrys; tuned-circuit capacitance $C = 420$ picofarads; unloaded tuned-circuit $Q = 40$.

Answer: $K_0 = 386$ (the shunting effect of R_{in} and R_g is neglected).

6.2. Determine the value of R_g at which the gain of the amplifier in Problem 6.1 is halved.

Answer: $R_g = 88$ kilohms.

6.3. Find the adjacent-channel attenuation (the selectivity of the receiver), using the data of Problem 6.1, at 10 kilohertz off resonance.

Answer: $d = 4.75$.

6.4. Check whether the data given in Problem 6.1 ensure stable operation of a single-stage tuned amplifier.

Answer: $K_{st} = 454$. The stability is ensured.

6.5. A tuned amplifier employs a type 6K4 pentode and operates at 6.4 megahertz. Tuned-circuit resonant resistance R_{0e} is 40 kilohms. Select the form of coupling for maximum stability and find, if necessary, the coupling parameter p.

Answer: Direct connection of the tuned circuit cannot be used because of poor stability. With transformer and tapped-coil coupling, $p = 0.34$.

6.6. Find the gain of a band-pass single-stage amplifier using type 6K1П valve; $f_0 = 465$ kilohertz; $L = 350$ microhenrys. The Q of the unloaded tuned circuits is 100. Interstage coupling is critical.

Answer: $K_0 = 95$.

6.7. Check the band-pass single-stage amplifier of Problem 6.6.

Answer: The condition of stability is satisfied.

6.8. Find the adjacent-channel attenuation at 10 kilohertz off resonance in the amplifier of Problem 6.6.

Answer: $d = 9.25$.

CHAPTER VII

DETECTION

37. General

Detection is a process the primary purpose of which is to extract from a modulated r.f wave an a.f. signal which can be reproduced as sound. This is done in a receiver stage known as the *detector*.

According to the type of modulation, there may be amplitude, phase, frequency or other types of detection. In any type, detection utilizes a non-linear circuit element or elements.

The present chapter deals only with amplitude detection. The reception and detection of frequency-modulated signals are discussed in Chapter XI.

Figure 7.1 shows the block diagram of an amplitude detector. When amplitude-modulated r.f. signal is applied to the detector input, audio-frequency oscillations appear at its output.

Previously, radio receivers used mainly valve detectors. For their operation, valve detectors depend on the unidirectional conduction of valves or the non-linearity of their volt-ampere characteristics. The most commonly used forms of valve detection are diode, grid and anode-bend detection.

Since the late 40s semiconductor diodes have come into use as detectors.

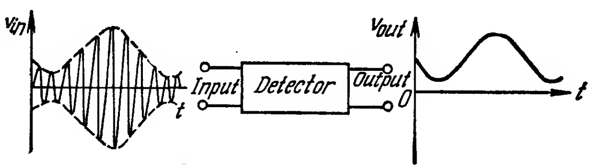


Fig. 7.1. Block diagram of an amplitude detector

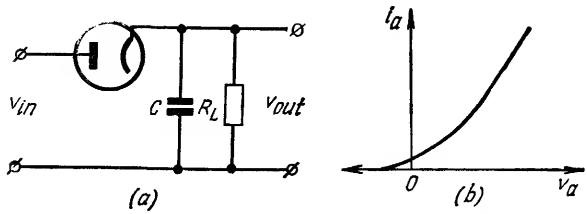


Fig. 7.2. Thermionic diode detector and its static characteristic

As a rule, modern radio receivers use diode detection. The circuit diagram and static characteristic of a thermionic diode detector are shown in Fig. 7.2.

When an r.f. voltage of constant amplitude, such that $v = V \cos \omega t$ (Fig. 7.3*a*), is applied to the detector input, a pulsating current begins to flow through the diode, with a pulse amplitude unvarying with time (Fig. 7.3*b*).

The pulsating current i flowing through the diode may be represented as a sum of the d.c. component and a series of a.c. components:

$$i = I_{ck} + I_1 \cos \omega t + I_2 \cos 2\omega t + I_3 \cos 3\omega t + \ldots$$

where I_{ck} = the direct component, or the average rectified current
I_1 = the amplitude of the first-harmonic current
I_2 and I_3 = the amplitudes of higher harmonics.

The average rectified current is represented in Fig. 7.3*b* by the straight line ($I_{ck} = \Delta I$).

When a modulated radio-frequency wave V_{in} is applied to the detector input (Fig. 7.4*a*), a pulsating current i_a appears in the detector circuit, with the pulse amplitude varying in time exactly as that of the applied voltage (Fig. 7.4*b*). The average rectified current I_{av} in this case will vary in time exactly as the envelope of the modulated wave, that is, as the audio-frequency signal superimposed on the r.f. carrier in modulation

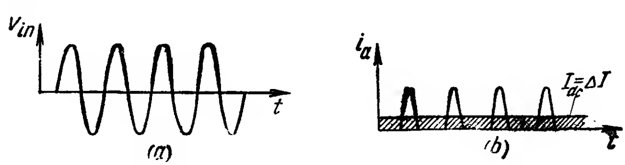


Fig. 7.3. Detector of an unmodulated signal

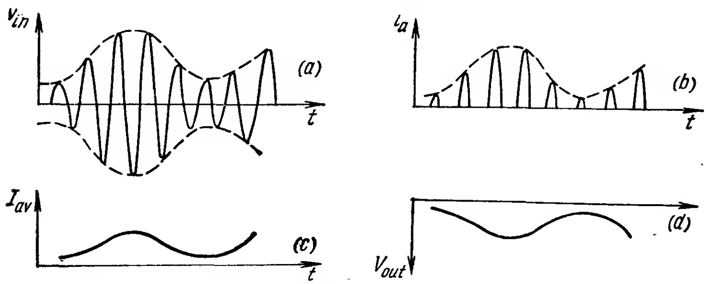


Fig. 7.4. Detection of a modulated wave

(Fig. 7.4*c*). This current may be represented as a sum of a direct current and an alternating current at audio frequency.

The capacitance of the capacitor in the circuit of Fig. 7.2 is such that at the radio frequency the reactance of C is much lower than the detector load resistance R_L, while at the audio frequency its reactance is much higher than R_L

$$\frac{1}{\omega C} \ll R_L, \quad \frac{1}{\Omega C} \gg R_L$$

Since the reactance of C at the radio frequency is low, the radio frequency voltage across the load terminals will also be low. As they pass through R_L, the direct component of the rectified current and the audio frequency current build up a pulsating voltage across the load resistance. This pulsating voltage changes in time so that it faithfully follows variations in the audio frequency signal (Fig. 7.4*d*). This pulsating voltage consists of the d.c. voltage and the audio-frequency voltage. The audio frequency voltage developing across the load resistance is separated by a simple filter, C_gR_g (Fig. 7.5). The reactance of C_g at the audio frequency is much lower than R_g; therefore, the audio-frequency voltage across R_g will only slightly differ in value from the voltage across R_L. The voltage across R_g is fed to an audio frequency amplifier for further amplification.

The electrical properties of a detector are expressed in terms of detector characteristic, voltage gain, frequency distortion, detector linearity, input resistance, and r.f. filtering.

The detector characteristic shows the relative current ΔI as a function of the input voltage V_{in} (Fig. 7.6). The curved part at the foot of the detector characteristic corresponds to

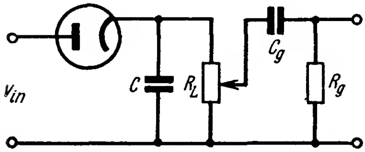


Fig.7.5.Diode detector with a filter to separate the audio-frequency voltage

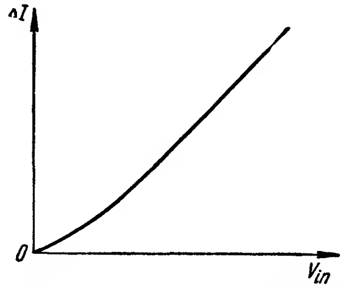


Fig. 7.6. Detector characteristic

relatively small values of input voltage (up to 0.3 volt). When the input voltage exceeds 0.3 to 0.5 volt, the detector characteristic is nearly linear.

The voltage gain of the detector is the ratio

$$K_d = \frac{V_\Omega}{mV} \tag{7.1}$$

where V_Ω = the audio-frequency voltage amplitude at the output of the detector

m = the modulation factor

V = the amplitude of the carrier voltage applied to the circuit.

With the r.f. voltage and modulation factor held constant, an increase in K_d will raise the a.f. voltage at the detector output.

Frequency distortion is evaluated from the frequency response curve of the detector, which relates the gain factor K_d to the intelligence (modulating) frequency F, with the modulation factor m and the carrier voltage amplitude V held constant:

$$K_d = f(F) \tag{7.2}$$

Detector linearity is expressed in terms of the non-linearity factor:

$$\gamma = \frac{\sqrt{V_{\Omega_2}^2 + V_{\Omega_3}^2}}{V_{\Omega_1}} \tag{7.3}$$

The input resistance of the detector is equal to the ratio of the input voltage amplitude to the amplitude of the first-harmonic radio-frequency input current:

$$R_{d\,in}=\frac{V_{in}}{I_1} \qquad (7.4)$$

From the value of the input resistance it is possible to evaluate the shunting effect of the detector on the tuned circuit. As the input resistance is increased, the shunting effect of the detector decreases, and the selectivity of the tuned circuit loaded by the detector remains unchanged.

Radio-frequency filtering is expressed in terms of the ratio of the amplitude of the r.f. voltage at the detector output to the amplitude of the r.f. voltage at the detector input.

The r.f. voltage finding its way into the audio-frequency circuits impairs the stability of the receiver. Therefore, the better the r.f. filtering, the lower the r.f. voltage at the output of the detector and, consequently, at the input of the audio-frequency amplifier.

Square-law Detection. When the maximum amplitude of voltages at the detector input does not exceed 0.3 volt, the lower portion of the volt-ampere characteristic of the detector is ordinarily used. Within this portion the dependence of the current i, flowing through the detector, on the input voltage may, in the general case, be expressed by a quadratic polynomial of the form

$$i=I_0+av+bv^2 \qquad (7.5)$$

It can be shown that with this relation between the detector current and the applied voltage, the incremental current in the detector circuit will be proportional to the square of the amplitude of the applied voltage

$$\Delta I=AV^2 \qquad (7.6)$$

where A is the proportionality factor, and V is the amplitude of the applied voltage. Hence the name, square-law detectors.

Any practical detector, regardless of the circuit configuration, type of valve, etc., acts as a square-law detector at low input voltages. Square-law detectors have disadvantages limiting their application. Firstly, they have a comparatively low input resistance, considerable non-linear distortion and a gain which varies with the amplitude of the applied signal.

The input resistance of a square-law detector, both loaded and unloaded, is determined by the a.c. anode resistance of the diode at the operating point

$$R_{in} = R_{aQ} \tag{7.7}$$

In square-law detection, non-linear distortion is proportional to the depth of modulation

$$\gamma = \frac{m}{4} \tag{7.8}$$

A detailed analysis of a square-law detector would show that the gain is proportional to the amplitude of the signal

$$K_d = A'V \tag{7.9}$$

Review Questions

1. Why is it that detection can be effected only by a non-linear circuit or device?
2. How do the static characteristics of the thermionic and crystal diodes differ?
3. How does the shape of the diode static characteristic affect operation of the detector?
4. What is the function of the capacitor placed across the detector load?
5. Will series or parallel connection of two diodes improve detection?

38. Linear Detection

Detection of Unmodulated Waves. When the minimum amplitude of the input voltage exceeds 0.3 volt, the relation of the incremental current ΔI in the detector circuit to the applied voltage may be described by the equation of a straight line passing through the origin of coordinates

$$\Delta I = bV \tag{7.10}$$

Detectors with the incremental current directly proportional to the applied voltage are called *linear detectors*.

Let us determine the basic electrical properties of a linear detector. In our analysis we shall use an idealised and not a practical diode characteristic (Fig. 7.7*b*). When no r. f. voltage is applied to the detector input (Fig. 7.7*a*), the current is zero. When an r. f. voltage is applied to the detector input, pulses of a rectified current appear in it. The average rectified current I_{av},

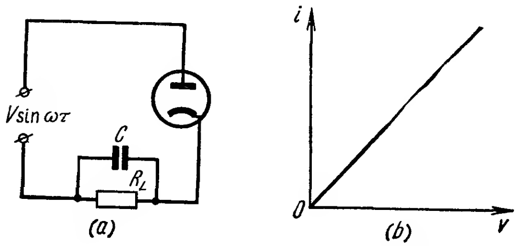


Fig. 7.7. Diode detector circuit and idealised characteristic of the diode

equal to the incremental current ΔI, is represented by the height of a rectangle whose area is equal to the area of the pulse waveform and whose base is equal to the period of oscillation

$$I_{av} = \Delta I$$

The rectified current I, on passing through the load resistance, produces a voltage drop $v_0 = \Delta I R_L$ across it. This voltage drop provides a negative bias at the anode. With negative bias applied to the anode, the instantaneous voltage v between the anode and cathode decreases such that

$$v = V\cos\omega t - v_0 = V\left(\cos\omega t - \frac{v_0}{V}\right) \tag{7.11}$$

Referring to the plot of Fig. 7.8, we may write

$$\frac{v_0}{V} = \cos\theta$$

where θ is the *operating* (or *conduction*) *angle* of the diode. It is defined as the angle, in electrical degrees, from the current maximum to cut-off.

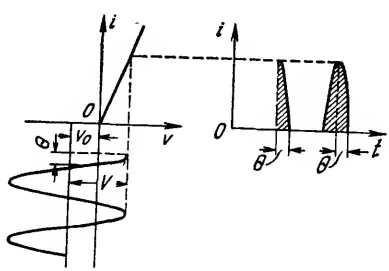


Fig. 7.8. Diagram showing operation of a loaded linear detector

Equation (7.11) may be re-written in the following way:

$$v = V(\cos\omega t - \cos\theta)$$

The instantaneous detector current i may be expressed in terms of the anode voltage v and mutual conductance g_m of the diode

$$i = g_m v = g_m V(\cos\omega t - \cos\theta) \tag{7.12}$$

The average current is numerically equal to the ratio of a half of the area of the current pulse waveform to a half-period in angular units

$$I_{av} = \frac{1}{\pi}\int_0^\theta i\,d(\omega t) \tag{7.13}$$

Substituting the instantaneous current into Eq. (7.13) gives

$$I_{av} = \frac{1}{\pi}\int_0^\theta g_m V(\cos\omega t - \cos\theta)\,d(\omega t) = \frac{1}{\pi}g_m V\int_0^\theta (\cos\omega t$$

$$- \cos\theta)\,d(\omega t) = \frac{1}{\pi}g_m V\left[\int_0^\theta \cos\omega t\,d(\omega t) - \cos\theta\int_0^\theta d(\omega t)\right]$$

Therefore,

$$I_{av} = \frac{g_m V}{\pi}(\sin\theta - \theta\cos\theta) \tag{7.14}$$

Thus, the incremental current through the detector is a function of the amplitude of the applied voltage V and the operating angle θ.

Let us see how the operating angle depends on the circuit parameters

$$\cos\theta = \frac{v_0}{V}$$

but

$$v_0 = \Delta I R_L = R_L\frac{g_m V}{\pi}(\sin\theta - \theta\cos\theta)$$

Then

$$\cos\theta = \frac{v_0}{V} = R_L\frac{g_m V}{\pi V}(\sin\theta - \theta\cos\theta) = \frac{R_L g_m}{\pi}(\sin\theta - \theta\cos\theta) \tag{7.15}$$

Therefore

$$\frac{\sin\theta - \theta\cos\theta}{\cos\theta} = \frac{\pi}{R_L g_m} = \frac{\pi R_a}{R_L}$$

or

$$\tan\theta-\theta=\frac{\pi}{R_L g_m}=\frac{\pi R_a}{R_L} \tag{7.16}$$

Eq. (7.16) shows that the operating angle depends only on the load resistance R_L and the mutual conductance g_m of the diode and does not depend on the amplitude of the applied voltage V. The fact that the operating angle is independent of the applied voltage suggests important conclusions concerning the shape of the detection characteristic.

The incremental current through a linear detector is directly proportional to the amplitude of the applied voltage.

The load resistance of the detector will only affect the gradient of the detector characteristic. The higher the load resistance, the smaller the gradient (Fig. 7.9).

The voltage gain of a linear detector and its dependence on the load resistance can be determined from Eqs. (7.15) and (7.16).

As already noted, the radio-frequency voltage applied to the detector input gives rise to a rectified current through the detector. On passing through R_L, the direct component of the rectified current produces a voltage v_0 across this resistance. Therefore, the voltage gain of the detector will be

$$K_d=\frac{v_0}{V}=\cos\theta \tag{7.17}$$

As follows from Eq. (7.16), the operating angle θ will decrease with increase of the load resistance R_L. Ordinarily the load resistance is tens or even hundreds of times as high as the a. c. anode resistance of the detector diode. Therefore, the operating

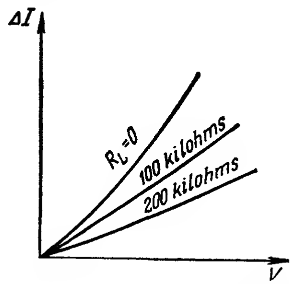


Fig. 7.9. Detection characteristics of a diode detector

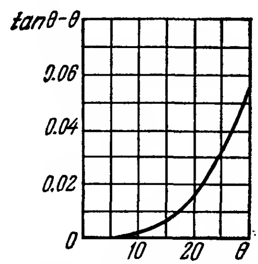


Fig. 7.10. Plot of the function $\tan\theta-\theta=f(\theta)$

angle θ does not exceed 10 to 20°. The cosine of this angle will be close to unity. Consequently, the gain of a linear detector will also be close to unity.

Eq. (7.16) may be solved only graphically. To begin with, a curve relating $\tan\theta-\theta$ to the operating angle is plotted (Fig. 7.10). Then using Eq. (7.16), the difference $\tan\theta-\theta$ is found, then the angle θ, and finally $\cos\theta$.

The output voltage of the detector is found from

$$v_0 = K_d V$$

The rectified current is

$$I = \frac{v_0}{R_L}$$

Determine the relation between the input resistance and load resistance of the detector. As the load resistance increases, the rectified voltage will also increase while the operating angle θ will decrease. The smaller the operating angle, the lower the amplitude of the first-harmonic current I_1 in the detector. Thus, an increase in R_L will also increase the input resistance of the detector:

$$R_{d\,in} = \frac{V}{I_1}$$

The design equation for the input resistance of the detector may be derived from the following relationships.

Since the load resistance considerably exceeds the a.c. anode resistance of the diode, we may assume that the bulk of P_{ac} applied to the detector is dissipated across the load resistance.

Therefore, we may write

$$P_{ac} = P_0$$

but

$$P_{ac} = \frac{V^2}{2R_{d\,in}}$$

and

$$P_0 = \frac{v_0^2}{R_L}$$

Hence

$$\frac{V^2}{2R_{d\,in}} = \frac{v_0^2}{R_L}$$

When the load resistance has a considerable value, the detector output voltage is approximately equal to the amplitude of the applied voltage

$$v_0 \cong V$$

Therefore, the design equation for the input resistance may be simplified thus

$$R_{d\,in} = \frac{R_L}{2} \tag{7.18}$$

Equation (7.18) holds only for detection at frequencies up to several megahertz. As the frequency increases, the shunting effect of the anode-cathode capacitance of the diode begins to tell. At ultra-high frequencies, Eq. (7.18) is not suitable for practical application.

We have seen that in detection of the unmodulated wave, the d. c. voltage appearing across the detector load shifts the operating point to the left from the origin of coordinates by a fixed amount equal to v_0. The situation is somewhat different when the input voltage is a modulated wave.

Detection of the Modulated Wave. In detection of the modulated wave, the average current and, consequently, the voltage drop across the load resistance change in time exactly as the audio-frequency (modulating or intelligence) wave. The position of the operating point depends on the instantaneous voltage drop across the load resistance, so that the applied modulated wave swings about the envelope of the radio-frequency modulated signal, rather than the straight line representing v_0.

The waveforms of the voltages and currents present in the detector circuit are shown in Fig. 7.11.

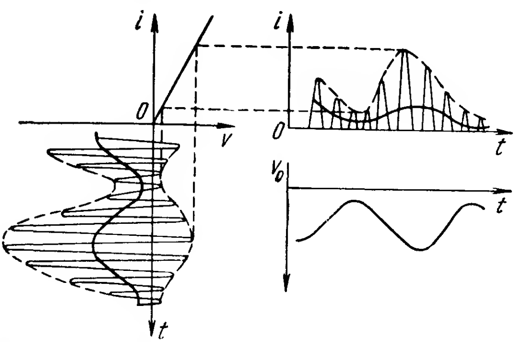


Fig. 7.11. Detection of a modulated wave

Detection of the modulated wave can be analysed, using the relationships already derived. We shall assume that the reactance of C (Fig. 7.5) is

$$\frac{1}{\Omega C} \gg R_L$$

The amplitude of the modulated wave applied to the detector is

$$V_M = V(1 + m\cos\Omega t)$$

Substituting the expression for V_M into Eq. (7.14) gives

$$I_{av} = \frac{g_m V}{\pi}(\sin\theta - \theta\cos\theta)(1 + m\cos\Omega t) = \frac{g_m V}{\pi}(\sin\theta - \theta\cos\theta)$$
$$+ \frac{g_m V}{\pi}(\sin\theta - \theta\cos\theta)\cos\Omega t \qquad (7.19)$$

The first term on the right-hand side of Eq. (7.19) represents the direct component of the rectified current. The second term stands for the audio-frequency current component of amplitude

$$I_{\Omega_1} = \frac{g_m m V}{\pi}(\sin\theta - \theta\cos\theta)$$

The rectified current contains only the first harmonic at the audio frequency. Since there is no second harmonic in the detector current, we may conclude that linear detection is free from second-harmonic distortion.

The amplitude of the audio-frequency voltage across the load resistance is

$$V_\Omega = I_{\Omega_1} R_L$$

Substituting the expression for I_{Ω_1} we obtain

$$V_\Omega = \frac{g_m R_L}{\pi}(\sin\theta - \theta\cos\theta)\, mV$$

By Eq. (7.15)

$$\frac{g_m R_L}{\pi}(\sin\theta - \theta\cos\theta) = \cos\theta$$

Therefore

$$V_\Omega = mV\cos\theta \qquad (7.20)$$

As follows from Eq. (7.20), the voltage gain of a linear detector detecting modulated waves is

$$K_d = \frac{V_\Omega}{mV} = \cos\theta \qquad (7.21)$$

From Eq. (7.21) we may conclude that K_d is independent of the applied voltage amplitude, but depends on the modulating frequency. This can be explained by the fact that the impedance offered by the detector load to audio-frequency currents may be considered equal to R_L only within a limited frequency band where the following condition is satisfied:

$$R_L \ll \frac{1}{\Omega_h C}$$

As the modulating frequency is increased, the reactance of the blocking capacitor may become comparable with the load resistance. At the high modulating frequencies the detector load impedance becomes smaller in magnitude than R_L

$$Z = \frac{R_L \dfrac{1}{j\Omega_h C}}{R_L + \dfrac{1}{j\Omega_h C}} = \frac{R_L}{1 + j\Omega_h C R_L}$$

$$|Z| = \frac{R_L}{\sqrt{1 + (\Omega_h C R_L)^2}}$$

Decreasing the load impedance at the high modulating frequencies leads to a decrease in V_Ω and, consequently, K_d.

A detailed analysis of detector operation would show that frequency distortion at the high modulating frequencies is given by

$$M_h = \sqrt{1 + (\Omega_h C R_{eq})^2} \qquad (7.22)$$

where $R_{eq} = \frac{R_a R_L}{R_a + R_L}$ is the equivalent resistance.

Thus, a linear detector has the following basic properties:

1. The incremental current in a linear detector is directly proportional to the applied voltage amplitude, and the detector characteristic is therefore a straight line.
2. The input resistance of a linear detector depends on the load resistance.
3. The voltage gain of a linear detector does not depend on the amplitude of the applied signal.

Review Questions

1. When will a detector operate as a linear one?
2. How does the load resistance affect the operating angle?
3. Does the detector load resistance affect the voltage?

4. Why is it that non-linear distortion is reduced in linear detection?

5. How can frequency distortion be reduced at the higher modulating frequencies?

39. Diode Detectors

Diode detection is the simplest type of detection and utilizes the unidirectional conduction of diode valves.

There are series and parallel diode detectors.

In a series detector, the signal source, the diode and its load are connected in series (Fig. 7.12).

In a parallel detector, the signal source, the diode and its load are connected in parallel (Fig. 7.13).

Series detection is used in cases where the direct component of the detector current may pass through the signal source.

The equivalent input resistance of the parallel detector ($R_{d\,in\,eq}$) is somewhat less than that of the series detector and it shunts the tuned circuit to a greater degree.

The equivalent input resistance of a parallel detector, arranged as shown in Fig. 7.13 is given by

$$R_{d\,in\,eq} = \frac{R_{d\,in} R_L}{R_{d\,in} + R_L} \quad \text{(a)}$$

However,

$$R_{d\,in} = \frac{R_L}{2} \quad \text{(b)}$$

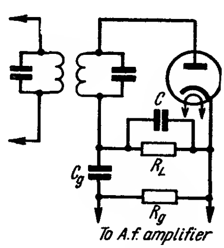


Fig. 7.12. Series detector circuit

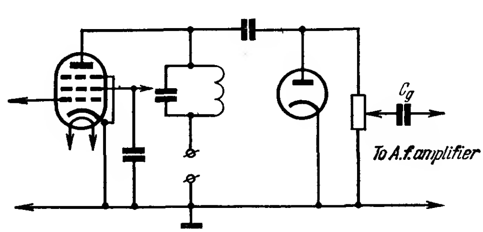


Fig. 7.13. Parallel detector circuit

Therefore, substituting Eq. (b) into Eq. (a), we obtain

$$R_{d\,in\,eq}=\frac{\frac{1}{2}R_LR_L}{\frac{1}{2}R_L+R_L}=\frac{1}{3}R_L \qquad (7.23)$$

Thus, the equivalent input resistance of a parallel detector is one-third smaller than the input resistance of a series detector.

As far as the direct and a.f. components of the detector current are concerned, the diode and the load resistance of a parallel detector are connected in series, just as they are in the series circuit of Fig. 7.12. Therefore, all basic relations derived in the analysis of the series detector apply to the parallel detector.

One of the features of the parallel detector is the presence of a radio-frequency voltage across the load resistance, equal to the voltage across the tuned-circuit terminals. Therefore the detector output should be coupled to a low-pass filter which will not pass the radio-frequency voltage to the input of the next stages.

At present, receivers mainly employ the diode detector shown in Fig. 7.14. The detector load resistance R_L is split into R_1 and R_2. Splitting the load resistance improves r.f. filtering and decreases non-linear distortion.

Non-linear distortion in the diode detector may be due to the indiscriminately chosen values of R_L and C (Fig. 7.12), and also R_g serving as the grid-leak resistor of the audio-frequency amplifier valve.

Non-linear distortion due to C and R_L occurs only at the highest modulating frequencies. If the time constant of the de-

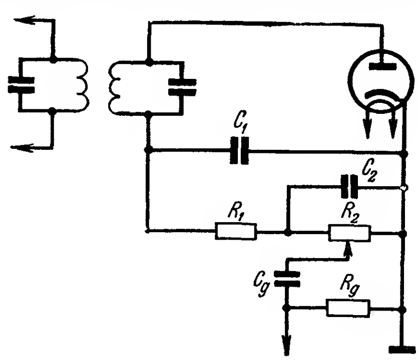


Fig. 7.14. Diode detector circuit

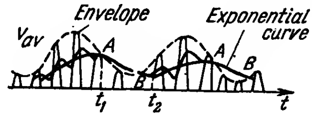


Fig. 7.15. Detector operating into a load impedance

tector load is comparatively large, the discharge of C through R_L will lag behind the change in the amplitude of the r.f. wave caused by modulation. As a result, the output voltage v_{av} of the detector will follow the exponential curve of the capacitor discharge (Fig. 7.15) rather than the envelope of the modulated wave. The result is non-linear distortion.

Referring to the plot of Fig. 7.15, during the time interval from t_1 to t_2 the voltage across the capacitor C is greater than any amplitude of the applied voltage because the capacitor has no time to discharge, and the diode is cut off. As a consequence, the load voltage traces out curve AB rather than the envelope of the modulated wave.

The greater the depth of modulation and the higher the modulation frequency, the larger the non-linear distortion. It may be shown that non-linear distortion will not occur if the selected values of R_L and C are such that

$$R_LC < \frac{\sqrt{1-m^2}}{\Omega_h m} \tag{7.24}$$

If the load resistance R_L is specified in advance, the capacitance of C may be found from the equation

$$C < \frac{\sqrt{1-m^2}}{\Omega_h m R_L} \tag{7.25}$$

In practical calculations it is assumed that the depth of modulation m, as measured at the highest modulating frequency, does not exceed 0.6-0.8. Therefore Eqs. (7.24) and (7.25) may be re-written as follows:

$$R_LC < \frac{0.7 \text{ to } 1.5}{\Omega_h} \tag{7.26}$$

$$C < \frac{0.7 \text{ to } 1.5}{\Omega_h R_L} \tag{7.27}$$

Example 7.1. Determine the capacitance of the blocking capacitor C in a detector whose load resistance R_L is 0.6 megohm and the highest modulating frequency F_h is 5,000 hertz.

Solution. From Eq. (7.27) we have

$$C < \frac{1.5}{\Omega_h R_L} = \frac{1.5}{6.28 \times 5 \times 10^3 \times 0.6 \times 10^6} = 80 \times 10^{-12} \text{ farads} = 80 \text{ picofarads}$$

Non-linear distortion due to the coupling circuit of the audio-frequency amplifier can arise if the equivalent resistance of the detector load at the audio-frequency

$$R_{Leq} = \frac{R_L R_g}{R_L + R_g}$$

is considerably less than R_L. In this case, the audio-frequency current I_{Ω_1} may have a greater amplitude than the direct component I_{dc} (Fig. 7.16), and the lower part of the current half-wave will be clipped, i.e. the waveform of the signal will be distorted.

The detector circuit shown in Fig. 7.14 retains a sufficiently high total load resistance and, consequently, a high input resistance and at the same time minimizes the shunting effect of the grid-leak resistor.

The equivalent circuit of the detector load is shown in Fig. 7.17

$$R_{Leq} = R_1 + \frac{R_2 R_g}{R_2 + R_g}$$

R_g shunts only R_2. In this case, R_g must be 6 to 8 times R_2. The audio-frequency signal voltage is taken off R_2. Since the voltage at the detector output is tapped down, the voltage gain of the detector is reduced. Therefore, R_1 is only a fraction of R_2.

Radio receivers widely use multi-section valves such as diode-triodes, double diode-triodes, diode-pentodes and double diode-pentodes. Their diode part is usually used for detection, and the triode or pentode part for amplification of the detected a.f. signal. A diode detector using a double diode-triode is shown in Fig. 7.18.

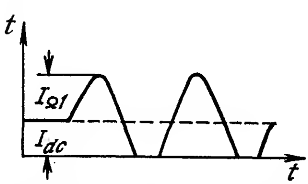


Fig. 7.16. Chart of diode detector operation into a complex circuit

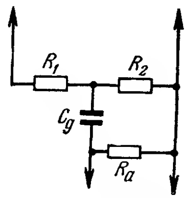


Fig. 7.17. Equivalent load circuit of diode detector

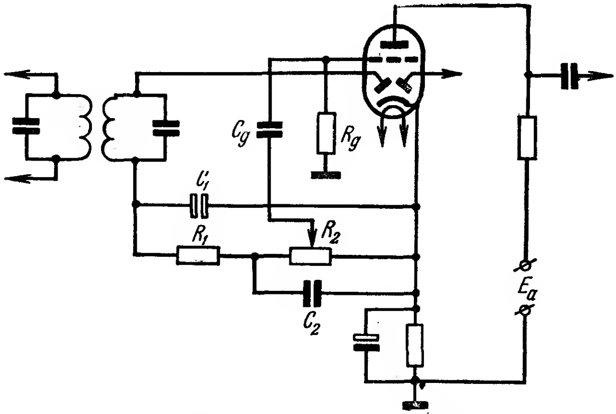


Fig. 7.18. Diode detector employing a double diode-triode

Receivers also use semiconductor diodes. The analysis and configurations of semiconductor-diode detectors somewhat differ from those just discussed.

In analysis of the diode detector it has been assumed that the reverse resistance of the diode is infinitely high. This assumption is, however, valid only for thermionic diodes. In crystal diodes, the reverse resistance is from tens to hundreds of kilohms, that is, it is comparable with the load resistance.

Therefore, other things being equal, the input resistance of a semiconductor diode detector will always be lower than that of a thermionic diode detector.

The input resistance of a semiconductor detector is given by

$$R_{in}=\frac{R_r R_L}{2R_r+3R_L}$$

where R_r is the reverse resistance.

In transistor receivers, the detector load comprises not only R_L, but also the input resistance of the first stage in the a.f. amplifier, which usually is very low. Therefore, R_L for the detector of a transistor receiver is ordinarily chosen to have several kilohms.

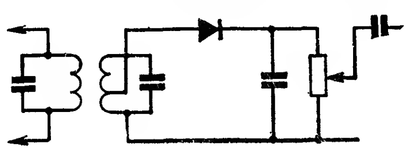


Fig. 7.19. Crystal detector circuit with tapped-down coupling

With this value of R_L, the input resistance of the detector is drastically reduced. To minimize the shunting effect of R_{in} on the amplifier tuned circuit, the latter is coupled through a transformer or a tapped coil (Fig. 7.19).

Review Questions

1. Can a full-wave rectifier be used as a detector?
2. Can a diode detector be arranged as a voltage doubler?
3. Why is it that semiconductor detectors use point-contact diodes?
4. What is the objective of splitting up the load of a diode detector?
5. How can non-linear distortion be reduced in detection of the modulated wave?

40. Calculation of a Diode Detector

The characteristic of a practical diode has a square-law and a linear section. Therefore, the detected signal will vary with the voltage applied to the detector. Rigorous mathematical treatment involves detailed analysis of the actual diode characteristics. For practical purposes it is customary to resort to experimental data obtained for a "bogey", or typical, diode. Table 7.1 gives data on the Soviet-made 6X6 diode when the load resistance R_L is 0.5 megohm. The same data may be used for other types of diodes and values of R_L.

TABLE 7.1

Signal designation	Volts	K_d
Weak	< 0.1	$5V$
Weak-to-medium	0.1-0.5	0.5-0.9
Medium	0.5-2	0.9-0.95
Strong	2	0.95

The general design procedure is as follows:

Given:

1. The equivalent resistance of the anode load tuned circuit (R_{0e}) of an intermediate-frequency amplifier.
2. Modulating frequency range, F_l-F_h.
3. Frequency distortion limits.

To Find:

1. Type of valve.
2. Circuit parameters and the gain of the detector.

Design Procedure:

1. Select the type of valve. In computing the circuit parameters refer to Fig. 7.18.

2. Find the detector input resistance. For the symmetry of the resonance curve, it should shunt the secondary circuit of the i.f. transformer insignificantly

$$R_{d\,in} > (3 \text{ to } 4)\, R_{0e}$$

3. Determine the total detector load resistance

$$R_L = R_1 + R_2 = 2R_{d\,in}$$

4. Put the value of the grid-leak resistor R_g of the next stage in the a.f. amplifier at 0.5 to 3 megohms and find the value of R_2

$$R_2 = \frac{R_g}{6 \text{ to } 8}$$

5. Find the value of R_1

$$R_1 = R_L - R_2$$

6. Find the value of C that will keep non-linear distortion to a tolerable value:

$$C = \frac{1.5 \times 10^{12}}{\Omega_h R_L} \text{ picofarads}$$

7. Assuming that the value of C_1, the filter capacitor, is specified in advance, find the value of C_2

$$C_2 = \frac{C - C_1}{\left(\dfrac{R_2}{R_1 + R_2}\right)^2}$$

8. Find the value of C_g that will keep down frequency distortion in the low-frequency range:

$$C_g \geqslant \frac{1}{\Omega_l R_g \sqrt{M_l^2 - 1}}$$

9. Determine the carrier peak voltage V_{car} assuming that for linear detection the least value of peak voltage (V_{min}) of the modulated wave at the maximum depth of modulation should be

not less than 0.3 to 0.5 volt

$$V_{min}=V_{car}(1-m_{max})$$

$$V_{car}=\frac{V_{min}}{1-m_{max}}$$

Putting $m_{max}=0.9$ volt and $V_{min}=0.3$ volt we have

$$V_{car}\cong 3 \text{ volts}$$

10. Determine the voltage gain of the detector. From Table 7.1 it may be assumed that $K_d\cong 0.9$. Since the audio-frequency signal voltage is taken off R_2, the actual value of K_d will be somewhat smaller

$$K'_d=0.9\frac{R_2}{R_L}$$

Example 7.2. Calculate a diode detector, if $R_{0e}=100$ kilohms; F_l-$F_h=100$ to 5,000 hertz; $M_l=1.02$.

Solution. 1. Select a 6X2П bantam dual diode and use one of its diodes as the detector. Calculate the parameters of the circuit of Fig. 7.14.

2. Determine the necessary detector input resistance

$$R_{d\,in}=4R_{0e}=4\times 100=400 \text{ kilohms}$$

3. Find the total resistance of the detector load

$$R_L=2R_{d\,in}=2\times 400=800 \text{ kilohms}$$

4. Put R_g at 3 megohms, and determine the value of R_2

$$R_2=\frac{R_g}{6 \text{ to } 8}=\frac{3}{6}=0.5 \text{ megohm}$$

5. Find the value of R_1

$$R_1=R_L-R_2=800-500=300 \text{ kilohms}$$

6. Find the capacitance of C

$$C=\frac{1.5\times 10^{12}}{\Omega_h R_L}=\frac{1.5\times 10^{12}}{6.28\times 5{,}000\times 0.8\times 10^{6}}=60 \text{ picofarads}$$

7. Put the value of C_1 at 40 picofarads and determine C_2

$$C_2=\frac{C_{eq}-C_1}{\left(\frac{R_2}{R_L}\right)^2}=\frac{60-40}{\left(\frac{0.5}{0.8}\right)^2}=50 \text{ picofarads}$$

8. Find the capacitance of C_g

$$C_g \geqslant \frac{1}{\Omega_l R_g \sqrt{M_l^2-1}} = \frac{1}{6.28\times 100\times 3\times 10^6 \sqrt{1.02^2-1}}$$
$$\approx 2.7\times 10^{-9} \text{ farads}$$

which may be rounded off to 3,000 picofarads.

9. Determine the amplitude of the signal applied to the detector:

$$V_{min}=0.3 \text{ volt} \quad m_{max}=0.9 \text{ volt}$$
$$V_{car}=\frac{V_{min}}{1-m_{max}}=\frac{0.3}{1-0.9}=3 \text{ volts}$$

10. Find the voltage gain of the detector

$$K_d=0.9\frac{R_2}{R_L}=0.9\frac{0.5}{0.8}=0.565$$

Review Questions

1. List the factors affecting the load resistance of a detector.
2. How can the effect of the detector input resistance on the tuned circuit be minimized?
3. Why is it that in transistor receivers the detector load resistance is lower than it is in valve receivers?
4. Which form of coupling the transistor to the tuned circuit of an a.f. amplifier will increase the detector load resistance?

41. Anode-bend Detection

Anode-bend detection takes advantage of the curvature (or *bend*) at the foot of the anode-grid characteristic $i_a=f(e_g)$.

An anode-bend detector is shown in Fig. 7.20. The control grid accepts both the modulated signal to be detected, V_{in}, and a negative bias voltage, E_g, which is sufficient to shift the quiescent operating point into the lower non-linear part of the

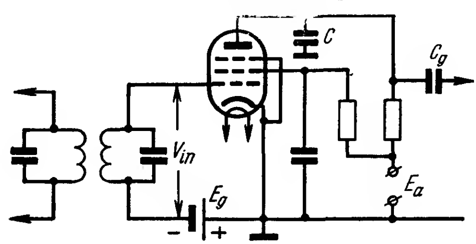


Fig. 7.20. Anode-bend detector

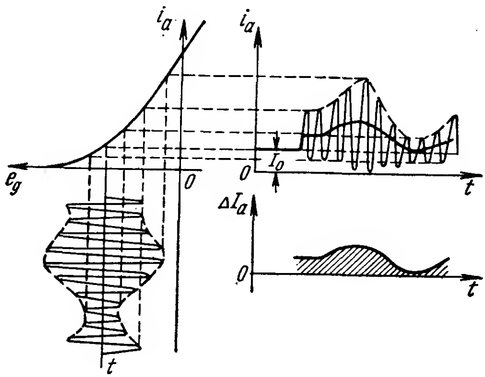


Fig. 7.21. Operation of an anode-bend detector

anode-grid characteristic (Fig. 7.21). As a result, the mutual conductance of the valve will be a maximum during the positive half-cycles, and a minimum during the negative ones. Accordingly, the anode current will increase more in the former case than in the latter, and there will be an increase in the average anode current.

Since the amplitude of the input voltage follows the changes in the modulating (audio-) frequency signal, the average anode current will also change in the same fashion.

The r.f. component of the anode current is bypassed to the cathode by C (Fig. 7.20). The a.f. component flows through the anode load resistance and builds up across it a voltage whose waveform is the same as that of the modulating signal. C_g couples this voltage to the input of the next stage in the audio-frequency amplifier.

The voltage gain of an anode-bend detector is considerably greater than that of a diode detector because in addition to detection it also provides amplification of the signal.

The input resistance of an anode-bend detector depends on the relative values of grid bias voltage and the peak input (modulated signal) voltage. When $E_g > V_{max}$, the instantaneous grid voltage is always negative, there is no grid current flowing, and the input resistance of the anode detector is very high. Under the circumstances, the input resistance is determined only by the dielectric losses in the valve and its associated wiring.

In calculating an anode-bend detector, use may be made of the equation derived for the diode detector. The voltage values, however, should be multiplied by μ.

Review Questions

1. Can an anode-bend detector use self-bias?
2. Why is it that an anode-bend detector affects the tuned circuit insignificantly?
3. Which form of the anode-grid characteristic is preferable for anode-bend detection?

42. Infinite-impedance Detection

Infinite-impedance detection is in effect a modification of anode-bend detection and differs from it in that the load resistance is connected into the cathode circuit.

The circuit of the infinite-impedance detector is shown in Fig. 7.22*a*. The value of the load resistance R_1 is so selected that the operating point of the valve is shifted into the non-linear part of the characteristic. The reactance of C_1 at the radio-frequency must be considerably less than the value of R_1. The audio-frequency voltage appearing across R_1 is fed back into the grid circuit. This voltage is 180° out of phase with the modulated wave envelope, thereby furnishing negative feedback. The voltage gain is therefore considerably less than that of the anode-bend detector.

The input resistance of the infinite-impedance detector may be the same as that of the anode-bend detector. R_2 and C_2 form a filter which eliminates any r.f. that might otherwise appear in the output.

Operation of the infinite-impedance detector differs from that of other detector types in several important aspects. For one thing, the detected output voltage is positive to earth (the chas-

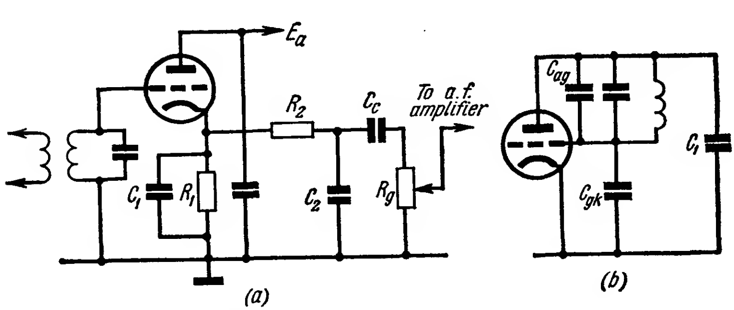


Fig. 7.22. Infinite-impedance detector

sis). For another, this voltage is in phase with the envelope of the modulated wave.

The cathode-load resistance of the infinite-impedance detector and, consequently, its output resistance, is considerably less than that of the diode detector. This markedly minimizes the non-linear distortion caused by the coupling network C_cR_g of the next stage of the a.f. amplifier.

The positive r.f. feedback (regeneration) occurring in the infinite-impedance detector compensates for the losses caused in the tuned circuit by the shunting effect of the input resistance.

Figure 7.22*b* shows the r.f. equivalent circuit of the infinite-impedance detector. This current differs only slightly from the usual Colpitts oscillator. By choosing the proper value of C_1, this circuit may be brought up to a point where it is about to oscillate and the input impedance is close to infinity.

As compared with the diode detector, the infinite-impedance detector has a lower large-signal handling capability. As the amplitude of the incoming signal increases, grid currents may begin to flow, sharply reducing the input impedance.

Review Questions

1. Why is it that the infinite-impedance detector has a lower voltage gain, as compared with the anode-bend detector?
2. Does the infinite-impedance detector affect operation of the tuned circuit?
3. Why is it dangerous to overload the anode-bend and infinite-impedance detectors?

43. Grid-leak Detection

Grid-leak detection utilizes the non-linearity of the grid characteristic.

The grid-leak detector is shown in Fig. 7.23. It provides detection and amplification of the audio-frequency signal. The detection is performed in the grid circuit, and the anode delivers an amplified output.

From Fig. 7.23 it can be seen that the grid-cathode circuit functions in the same manner as the anode-cathode of a diode detector. R_g serves as the detector load. C_c provides a bypass for the r.f. around R_g.

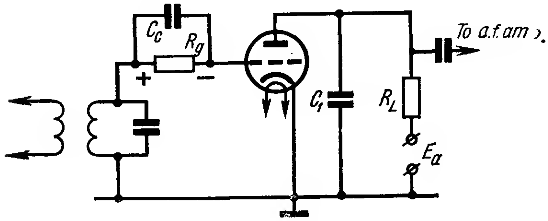


Fig. 7.23. Grid-leak detector circuit

When an unmodulated r.f. voltage is applied between the grid and the cathode, a pulsating current appears in the grid circuit. The direct component $I_{g\,dc}$ of the grid current, passing through R_g, builds up across it a d.c. voltage whose polarity is as shown in Fig. 7.23. The right-hand terminal of R_g is connected to the grid, and the left-hand terminal to the cathode through the tuned-circuit coil. Therefore, the d.c. voltage across R_g, negative to the cathode, decreases the average anode current.

When a modulated r.f. voltage is applied from the tuned circuit to the grid circuit, the average grid current changes in accordance with the envelope of the modulated signal. Now, both the direct and a.f. components of the grid current will flow through R_g. The a.f. component, on passing through R_g, builds up across it, as the diode detector load, an audio-frequency voltage which is also applied to the grid circuit.

The grid potential, changing at the audio-frequency rate, causes audio-frequency variations in the anode current. Applied to the anode load resistor R_L, they give rise to an amplified audio-frequency voltage.

The waveforms of voltages and currents in the grid-leak detector are shown in Fig. 7.24. At (*a*) is a modulated r.f. voltage applied to the input of the detector; at (*b*) are the instantaneous grid current and its average value (the heavy curve); at (*c*) is the voltage drop across R_g (this curve serves as the axis about which the applied modulated voltage swings). Curve (*d*) is the waveform of the alternating anode current, while at (*e*) is the voltage drop across the anode load resistance. The last holds only when the r.f. is removed from the anode current.

The grid-leak detector provides greater amplification than the anode-bend detector. This is explained by the fact that in grid-leak detection the operating point is located on the linear part

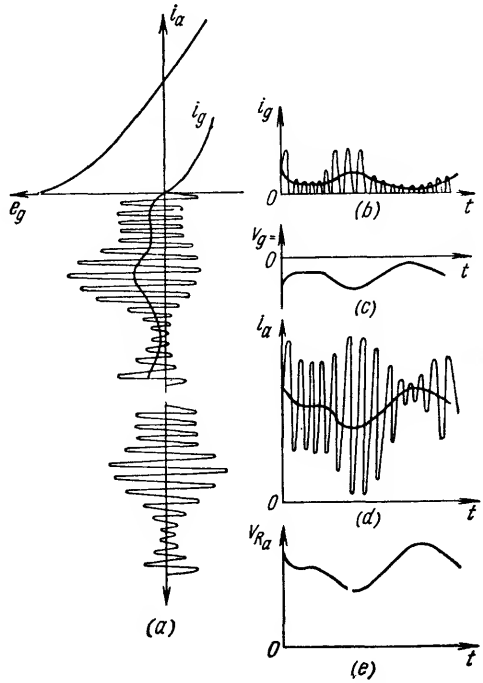


Fig. 7.24. Operation of a grid-leak detector

of the anode-grid characteristic $i_a=f(e_g)$ which has a greater gradient. In anode-bend detection, the operating point is positioned on the lower part of the characteristic which has a smaller gradient. This is why the grid-leak detector is more sensitive, as compared with other types of detector.

It should be noted that the grid-leak detector has some shortcomings, the major one being non-linear distortion in the case of very weak and very strong signals. With a weak signal the distortion is caused by the fact that use is made of the square-law region of the grid characteristic. With a strong signal, the cause lies in the increase of the negative voltage across R_g and in the subsequent shift of the operating point to the curved region of the anode-grid characteristic. If the negative bias is sufficiently high, the operating point may be shifted to the bend of the anode-grid characteristic. In this case, an opposing anode-bend detection, lowering the effect of grid-leak detection, will take place in the anode circuit.

Review Questions

1. What sets the grid-leak detector apart from the diode detector?
2. What is the function of the capacitor placed across the anode load?
3. Why is it that the grid-leak detector is more sensitive than the anode-bend detector?

SUMMARY

1. Detection is a process by which an a.f. signal that can be reproduced as sound is extracted from an r.f. signal in radio-receivers.
2. Detection may be performed only by an electric circuit having a non-linear volt-ampere characteristic.
3. Diodes, triodes, pentodes, multi-section valves and semiconductor diodes may be used as detectors.
4. Electrical properties of a detector depend on the shape of its volt-ampere characteristic.
5. Depending on the volt-ampere characteristic shape, and according to the amplitude of the applied voltage, detectors may perform square-law or linear detection.
6. A square-law detector introduces considerable non-linear distortion proportional to the depth of modulation. The voltage gain of a square-law detector varies with the amplitude of the applied voltage. The input resistance of a square-law detector is equal to its dynamic resistance at the operating point.
7. A linear detector has a high voltage gain, high input resistance, and low non-linear distortion.
8. Modern multi-valve receivers chiefly use linear diode detectors.
9. Anode-bend and grid-leak detectors, employing triodes and pentodes, have a higher sensitivity than diode detectors.

■ Problems

7.1. Determine the voltage gain of a linear detector loaded by $R_l = 300$ kilohms. The mutual conductance of the diode is 1 milliampere per volt.

Answer: $K_d = 0.97$.

7.2. The load resistance of a diode detector is $R_L = 500$ kilohms; the highest modulating frequency is $F_h = 4{,}000$ hertz. Find the capacitance of the capacitor.

Answer: $C \cong 160$ picofarads.

7.3. Determine the dynamic resistance of the detector load. The resistance of the detector load is $R_L = 700$ kilohms; the grid-leak resistor is $R_g = 2.5$ megohms.

Answer: $R_d = 547$ kilohms.

7.4. Find the frequency distortion at $F = 5{,}000$ hertz in a diode detector loaded by $R_L = 0.5$ megohm, shunted by $C = 200$ picofarads.

Answer: $M_h = 1.03$.

CHAPTER VIII
REGENERATIVE AND SUPERREGENERATIVE RECEPTION

44. General

Regenerative reception is based on the application of controllable positive feedback (regeneration) to the grid-leak detector stage.

Feedback is the transfer of part or all of a signal from some stage of an amplifier to a preceding stage or from the anode to the grid or some other circuit within the same stage. When the voltage fed back is in the same phase as the original, there will be an increase in amplification, and the feedback is said to be *positive*.

A simple regenerative receiver is shown in Fig. 8.1. Consider what takes place in its circuitry. The radio-frequency current flowing in the aerial builds up an alternating magnetic field around the aerial coupling coil. This field induces a radio-frequency emf E in the tuned circuit coil L_{tc}, so that a current I_{tc} begins to flow in the tuned circuit (Fig. 8.2).

The tuned circuit is usually adjusted to resonate at the frequency of the signal being received. Therefore, the current in the tuned circuit is in phase with the emf E induced in L_{tc}.

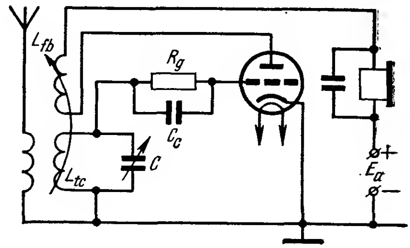


Fig. 8.1. Regenerative receiver circuit

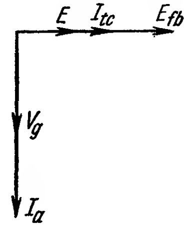


Fig. 8.2. Vector diagram of currents and voltages acting in a regenerative stage

The current I_{tc}, passing through C, builds up a voltage across it

$$\dot{V}_C = \dot{I}_{tc}\frac{1}{j\omega C}$$

The input capacitance C_{in} of the valve is considerably smaller than C_c. Therefore, it may be considered that the grid voltage V_g is equal to the capacitor voltage V_C.

The alternating voltage V_g gives rise to an alternating component, I_a, in the anode current, which is in phase with the grid voltage

$$\dot{I}_a = g_m\dot{V}_g$$

Feedback is provided by the coil L_{fb}, called a "tickler". The current I_a flowing through L_{fb} builds up a magnetic field around the coil. This magnetic field induces an emf of mutual induction in the tuned-circuit coil, which is 90° out of phase with the anode current. This is the feedback emf E_{fb}

$$\dot{E}_{fb} = \pm j\omega M\dot{I}_a$$

where the choice of the plus or minus sign depends on the sign of the mutual inductance M.

Substituting the expressions for I_a and V_g in the equation of E_{fb} gives

$$\dot{E}_{fb} = \pm j\omega Mg_m\dot{I}_{tc}\frac{1}{j\omega C} = \pm\dot{I}_{tc}\frac{Mg_m}{C}$$

or

$$\dot{E}_{fb} = \pm\frac{Mg_m}{C}\dot{I}_{tc}$$

With positive feedback, when M is positive, E_{fb} is in phase with I_{tc} and, consequently, with the incoming signal in the tuned circuit. Then

$$\dot{E}_{fb} = \frac{Mg_m}{C} \dot{I}_{tc} \tag{8.1}$$

Since we shall deal only with positive feedback, M will be considered positive in all the equations. With such phase relation between the incoming signal and the feedback emf, the current flowing through the tuned circuit increases.

The increase in this current is accompanied by an increase in the capacitor voltage and, consequently, in the grid voltage. The increase in the grid voltage due to feedback is equivalent to additional amplification of the r.f. signal in the stage. Because of this, the output voltage of a grid-leak detector employing feedback is considerably higher than that of the usual grid-leak detector. In the final analysis, the increase in the output voltage of the grid-leak detector increases receiver sensitivity and extends the operating range of radio communication. This is the reason why regenerative reception was very popular during the early period of valve receivers.

Review Questions

1. What is the purpose of using positive feedback (regeneration) in a radio receiver?
2. Can a regenerative receiver use transformer or tapped-coil coupling?
3. Draw up the circuit diagram of a regenerative receiver using a transistor.
4. May the tickler coil be connected in the cathode lead?

45. Theory of Regenerative Reception

As already noted, positive feedback augments the current through the tuned circuit of a regenerative stage

$$\dot{I}_{tc} = \frac{\dot{E} + \dot{E}_{fb}}{Z} \tag{8.2}$$

When the tuned circuit is adjusted to resonate at the frequency of the signal being received, the impedance will be a pure resistance ($Z = R$). Substituting the expression for E_{fb} and Z in Eq. (8.2) gives

$$\dot{I}_{tc} = \frac{\dot{E} + \frac{Mg_m}{C} \dot{I}_{tc}}{R}$$

Solving this equation for I_{tc}, we obtain

$$\dot{I}_{tc} = \frac{\dot{E}}{R - \frac{Mg_m}{C}} = \frac{\dot{E}}{R_{eq}} \tag{8.3}$$

where $R_{eq} = R - \frac{Mg_m}{C}$ is the equivalent resistance of the tuned circuit in a stage with positive feedback.

In the preceding section it was stated that the increase in the current through the tuned circuit was due to the feedback emf. Equation (8.3) makes it possible to explain this increase somewhat differently. As is seen, the equivalent resistance of the tuned circuit is brought down to R_{eq}. This improves the Q-factor and, consequently, the selectivity of the tuned circuit, and reduces the bandwidth. A narrow bandwidth is particularly important in short-wave receivers whose tuned circuits usually have a broad bandwidth.

The gain of a regenerative stage is decided by the valve and circuit parameters. If there is no feedback, the tuned-circuit voltage is the signal emf E, and the current is I_{tc}.

Then the voltage across the tuned-circuit capacitor C is

$$\dot{V}_C = \dot{V}_g = \dot{I}_{tc} \frac{1}{j\omega C} = \frac{\dot{E}}{R} \frac{1}{j\omega C} \tag{8.4}$$

With positive feedback, the equivalent resistance of the tuned circuit is given by Eq. (8.3). Substituting it in Eq. (8.4), we have

$$\dot{V}_{gfb} = \frac{\dot{E}}{R_{eq}} \frac{1}{j\omega C} = \frac{\dot{E}}{R - \frac{Mg_m}{C}} \frac{1}{j\omega C} \tag{8.5}$$

The gain due to feedback is numerically equal to the ratio of grid voltage in the stage with feedback to grid voltage in a stage without feedback:

$$\dot{K}_{fb} = \frac{\dot{V}_{gfb}}{\dot{V}_g} \tag{8.6}$$

Substituting the expressions for V_{gfb} and V_g into Eq. (8.6), and simplifying, we obtain

$$\dot{K}_{fb} = \frac{R}{R - \frac{Mg_m}{C}} = \frac{1}{1 - \frac{Mg_m}{RC}} \tag{8.7}$$

Equations (8.6) and (8.7) fully describe the gain due to feedback as a function of the peak value of the incoming signal, valve and circuit parameters. As is seen, the grid voltage of a regenerative stage is limited to $V_{g\,st}$, or the one at which the pulsating anode current attains a steady-state condition. Therefore, no matter how low or high the initial signal voltage may be, the steady-state grid voltage cannot be greater than $V_{g\,st}$.

In other words, when the input signal is small, the gain due to feedback will be much greater than in reception of strong signals. That is, the gain is greatest where it is most needed.

Referring to Eq. (8.7), the gain due to feedback will be at its greatest when the mutual inductance (that is, coupling) is such that the effective resistance of the tuned circuit drops to zero. Now, all energy lost in the grid circuit is fully compensated by the energy inflow from the anode circuit, and the tuned circuit is about to oscillate. The respective value of M is called the *critical mutual inductance*, and the regenerative circuit is said to have attained its *critical point*. Thus, to secure maximum regenerative amplification, positive feedback should be raised to the critical point where the circuit is about to oscillate.

The resonant frequency of the regenerative circuit depends on the tuned-circuit parameters and may differ from the frequency of the incoming signal. Therefore, when the incoming signal and the signal of the regenerative circuit are added together, the result is new frequencies which are the sum and difference of the two signals. These are *beat frequencies*. If the beat frequencies are within the audio range, the earphones connected into the anode circuit of the receiver will reproduce a characteristic whistle interfering with the reception of the desired signal.

Review Questions

1. Why is it that the gain is a maximum with weak signals?
2. How do the parameters of the tuned circuit affect the gain due to feedback?
3. Can a regenerative receiver receive *CW* telegraph signals?

46. Feedback Control Circuits

As stated above, a regenerative circuit uses controllable positive feedback. Feedback may be controlled in the following ways:

(a) by varying the mutual inductance between L_{tc} and L_{fb} (see Fig. 8.1);

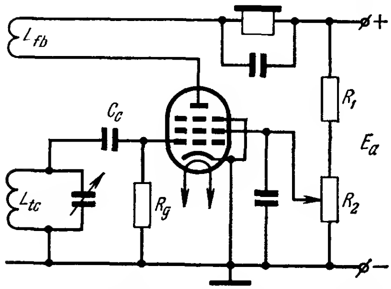


Fig. 8.3. Variable-resistance feedback control

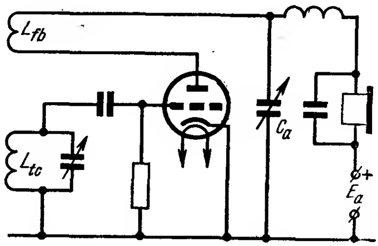


Fig. 8.4. Variable-capacitance feedback control

(b) by varying the mutual conductance of the valve (Fig. 8.3);

(c) by incorporating controllable negative feedback into the circuit (Fig. 8.4).

In the circuit of Fig. 8.1, feedback is controlled by varying the relative position of L_{tc} and L_{fb}. The tickler coil is usually placed inside L_{tc}. This is *variable mutual inductance feedback control.*

In the circuit of Fig. 8.3 the mutual inductance remains fixed, and feedback is controlled by varying the mutual conductance g_m and the a.c. anode resistance R_a of the valve.

Continuous control of g_m and R_a, and, consequently, of feedback can be effected by varying E_{g2} with a potentiometer, R_2. This is *variable-resistance feedback control.*

The circuit of Fig. 8.4 uses *variable-capacitance feedback control.* The mutual inductance between L_{tc} and L_{fb}, and, consequently, positive feedback, remain fixed. Feedback is controlled by C_a. The capacitance of this capacitor is adjusted so that the reactance X_a of the anode circuit is capacitive over the entire frequency range of the radio receiver. Because of this, feedback through the inter-electrode capacitance C_{ag} will be negative.

Changing the setting of C_a, and, consequently, the reactance of the anode circuit at the radio frequency, we can control the a.c. component of the anode voltage V_a, and, hence, the value of negative feedback through C_{ag}. While fixed positive feedback in the circuit brings down the equivalent resistance R_{eq} of the tuned circuit, controllable negative feedback increases this resistance by ΔR:

$$R_{eq} = R - \frac{Mg_m}{C} + \Delta R$$

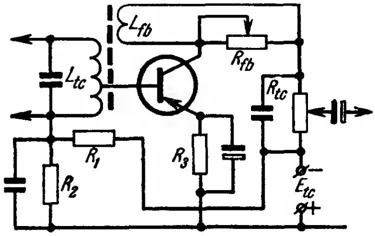


Fig. 8.5. Control of feedback in a transistor receiver

With this method, the resultant feedback can be controlled continuously within wide limits and over a wide frequency range. In transistor receivers, feedback can be controlled by varying the current through the feedback coil (Fig. 8.5). As R_{fb} is decreased, the amount of feedback is also decreased, because the greater part of the alternating collector current will flow through R_{fb}.

Review Questions

1. Can feedback be controlled by placing a variable capacitor across the feedback coil?
2. Can feedback be controlled by varying the potential at the third (suppressor) grid of the pentode?
3. How can feedback be controlled in a transistor receiver?

47. Theory of Superregenerative Reception

One of the main disadvantages of the regenerative stage is its instability at the critical point, when the equivalent resistance of the tuned circuit is nearly zero but still remains positive

$$R_{eq} = R - \frac{Mg_m}{C} > 0$$

Even an insignificant change in the operating condition causing an increase in the mutual conductance of the valve is likely to change this inequality in such a way that the equivalent resistance of the tuned circuit will become negative and the circuit will begin to oscillate, with beats appearing in its output. The reception of the wanted signal will be accompanied by a characteristic whistle.

This disadvantage is non-existent in superregenerative reception which additionally provides an amplification of thousands of times in a single stage.

Briefly, the principle of superregenerative reception is as follows. Feedback in the regenerative detector is increased past the critical point so as to permit self-oscillation with rapid build-up. Obviously, regenerative amplification is a maximum now. In order to preserve intelligibility, it is arranged that the oscillation is stopped very soon after initiation and then allowed to build up again. This process goes on continuously, and the circuit oscillates in bursts.

The frequency of the bursts is usually above the audible range; hence the bursts will not be reproduced at the output of the superregenerative detector. The oscillation is interrupted by a *quenching voltage*. It may be supplied either by the detector itself (a *self-quenching* superregenerative circuit) or by a separate oscillator (a *separately quenched* superregenerative circuit), as shown in Fig. 8.6, operating at the *quench frequency*, f_q.

In the circuit of Fig. 8.6 the regenerative detector operates as an anode-bend detector, that is, the operating point of the valve is located on the lower bend of the characteristic. During the positive half-cycles of the quenching voltage, the operating point of the valve is shifted to that part of the characteristic where mutual conductance is higher. During the negative half-cycles of the quenching voltage the operating point is shifted to, or even beyond, cut-off.

The operating point is positioned so that, when no quenching voltage is applied, the circuit is about to oscillate.

Now the positive half-cycles of the quenching voltage will drive the circuit to self-oscillation, and the negative half-cycles will stop or quench the oscillation. Hence the name "quenching voltage".

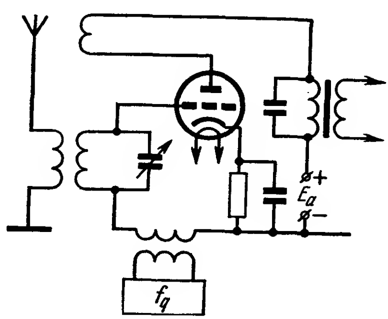


Fig. 8.6. Separately quenched superregenerative receiver

The build-up and stopping of oscillation depends on the magnitude of the tuned-circuit time constant $\tau_{tc}=\frac{2L}{R_{eq}}$ relative to the quench-frequency period $T_q=\frac{1}{f_q}$.

Consider operation of the circuit with and without a signal. In the no-signal state, only a small current is flowing in the tuned circuit, due to thermal agitation of electrons in the tuned-circuit components. This is *thermal noise current*, and the respective voltage is *thermal noise voltage*, V_n, which usually amounts to several microvolts. For simplicity, assume that when the quenching voltage is positive, self-oscillation will take place, i. e.

$$R_{eq}=R-\frac{Mg_m}{C}<0$$

while when this voltage is negative, self-oscillation will not occur ($R_{eq}>0$).

As soon as the equivalent resistance of the tuned circuit goes negative, the amplitude of the tuned-circuit current begins to increase exponentially

$$I_{mtc}=I_{m0}e^{\frac{R_{eq}}{2L}t}$$

If R_{eq} is sufficiently high in absolute value and the time constant of the tuned circuit is $\tau_{tc}<\frac{T_q}{2}$, the oscillation will reach a steady state determined by the parameters and operating voltages of the valve.

During the negative half-cycles of the quenching voltage, the oscillation is stopped and the current amplitude begins to decrease exponentially

$$I_{mtc}=I_{mst}e^{\frac{R_{eq}}{2L}t}$$

reaching its initial value I_{m0} at the end of each negative half-cycle.

Thus, the circuit generates bursts of oscillation and the average anode current increases at their rate (Fig. 8.7).

If the tuned-circuit time constant is greater than the half-period of the quenching voltage, i.e. if $\tau_{tc}>\frac{T_q}{2}$, the oscillation

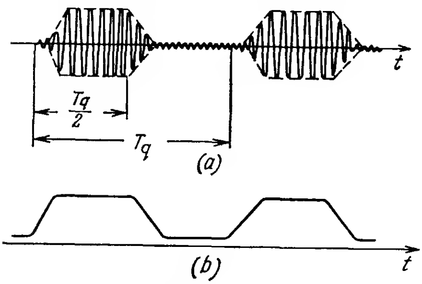


Fig. 8.7. Waveforms of oscillation in the superregenerative circuit when

$$\tau_{tc} < \frac{T_q}{2}:$$

(a) oscillation in the tuned circuit; (b) variations in the average anode current

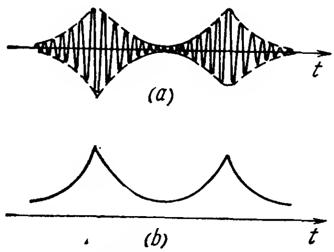


Fig. 8.8. Waveforms of oscillation in the superregenerative circuit when

$$\tau_{tc} > \frac{T_q}{2}:$$

(a) oscillation in the tuned circuit; (b) variations in the average anode current

in the tuned circuit will decrease before it attains a steady state. The respective waveforms of oscillation in the anode circuit are shown in Fig. 8.8.

Since initially the build-up of oscillation is controlled by the noise current, the bursts will differ both in duration and amplitude from one another. Therefore, variations in the average anode current will occur at random and cause so-called *superregenerative hiss* in the earphones (Fig. 8.9).

Now consider operation of the superregenerative receiver with a sinewave signal E_s causing in the tuned circuit a signal current that considerably exceeds the noise current

$$I_{ms} > I_{mn}$$

As soon as the circuit begins to oscillate, the initial amplitude of the tuned-circuit current will be determined by the signal current and will remain unchanged. Therefore, the bursts of

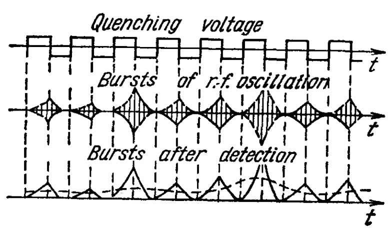


Fig. 8.9. Oscillation in a superregenerative circuit in the no-signal state

oscillation become regular, and the average anode current becomes constant.

Disappearance of superregenerative hiss in the earphones is an indication that the average anode current has ceased to vary at random.

Variations in the average anode current with the amplitude of the incoming signal depend on the mode of operation, that is, the character of oscillation.

Let the time constant of the tuned circuit be greater than the half-period of the quenching voltage. As soon as the circuit begins to oscillate, the amplitude of the current in the tuned circuit will change exponentially. At the end of the positive half-cycle of the quenching voltage the amplitude of the tuned-circuit current will be

$$I_{mtc}=I_{ms}e^{\frac{R_{eq}}{2L}\frac{T_q}{2}}=I_{ms}e^{\frac{R_{eq}}{4L}T_q}$$

The initial amplitude of the signal current I_{ms} is proportional to the signal voltage. Consequently, the final amplitude of the tuned-circuit current, I_{mtc}, and the variations in the average current of the anode-bend detector, ΔI_a, will also be proportional to the signal voltage. If the signal is modulated by an audio frequency, the average anode current will also vary at the audio-frequency rate (Fig. 8.10), or follow the envelope of the modulated r.f. signal.

By definition, a detector whose output is the envelope of the input wave is a *linear detector*. So, the superregenerative circuit discussed is linear.

Now consider variations in the average anode current when the amplitude of the tuned-circuit current is allowed to reach a steady-state value.

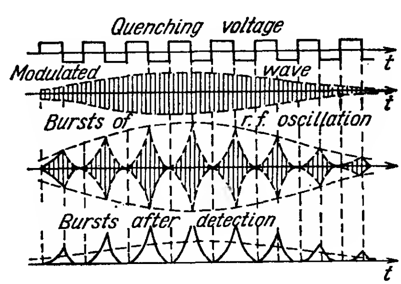


Fig. 8.10. Oscillation in the superregenerative circuit on reception of modulated waves

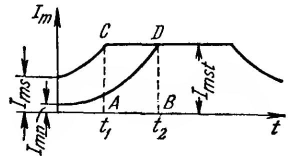


Fig. 8.11. Diagram of incremental anode current

In "with-signal" operation, the steady-state oscillation reaches a maximum amplitude at time t_1. In "no-signal" operation the maximum amplitude of oscillation in the tuned circuit will be reached somewhat later, at time t_2.

The incremental current will be proportional to the area of the figure *ACDB* (Fig. 8.11).

$$\Delta I = I_{m\ st}(t_2 - t_1) = I_{m\ st}\,\Delta t \tag{8.8}$$

$$I_{m\ st} = (I_{ms} + I_{mn})\,e^{\frac{R_{eq}}{2L}t_1} = I_{mn}e^{\frac{R_{eq}}{2L}t_2}$$

The noise current is considerably smaller than the signal current, and its value in the brackets may be neglected

$$I_{ms}e^{\frac{R_{eq}}{2L}t_1} = I_{mn}e^{\frac{R_{eq}}{2L}t_2}$$

Solving this equation for $\Delta t = t_2 - t_1$

$$\frac{I_{ms}}{I_{mn}} = e^{\frac{R_{eq}}{2L}(t_2 - t_1)} = e^{\frac{R_{eq}}{2L}\Delta t}$$

we have

$$\Delta t = \frac{2L}{R_{eq}}\ln\frac{I_{ms}}{I_{mn}}$$

Substituting the expression for Δt in Eq. (8.8) gives

$$\Delta I = I_{m\ st}\frac{2L}{R_{eq}}\ln\frac{I_{ms}}{I_{mn}}$$

Thus, the incremental current is proportional to the logarithm of the incoming signal. Therefore, this mode of operation is *logarithmic* or *non-linear* operation. As the amplitude of the incoming signal increases, the bursts of oscillation grow longer. Consequently, the average current in the anode circuit will also increase.

A superregenerative stage may provide an additional gain of several hundred thousand, and its value is practically independent of the type of valve used. The high gain of the superre-

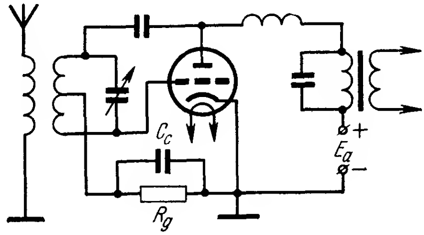


Fig. 8.12. Self-quenching superregenerative detector

generative detector is explained by the fact that the signal level at the receiver output depends, in the final analysis, on the grid voltage in the oscillatory state. The grid voltage due to the self-oscillation of the stage may be several volts, while the signal being received is only several microvolts in many cases. The weak incoming signal actually serves only to control the superregenerative receiver.

Hence, superregenerative reception is not just an improvement on the regenerative method, but is a method in its own right.

In multi-stage receivers, the superregenerative curcuit is ordinarily used in one of the intermediate stages preceding the detector, or in the detector itself.

Superregeneration must not be used in stages coupled to the aerial, since the emission from the superregenerative circuit will interfere with operation of other receivers.

Now let us consider operation of the self-quenching superregenerator of Fig. 8.12. It uses a parallel-feed tapped-coil tuned circuit. R_g and C_c are so selected that the time constant $\tau_g = R_g C_c$ is greater than the time constant τ_{tc} of the tuned circuit. With the time constants thus selected, the oscillation in the tuned circuit will reach a steady-state amplitude before the negative bias at the grid does so. Therefore, the negative grid bias will continue to increase for some time after the oscillation has reached a steady-state amplitude. As this happens, the operating point is shifted to the region of the characteristic with a lower mutual conductance. At some intermediate value of mutual conductance, the feedback provided in the circuit is insufficient to maintain oscillation in the tuned circuit, the condition $R_{eq} < 0$ is no longer met, and the oscillation is stopped. At that instant, C_c begins to discharge through R_g, the negative bias decreases, and the operating point is shifted into the region of a higher mutual conductance. Thus, the discharge of the capacitor resets the circuit, and the circuit again begins to generate a series of r.f. bursts. In the "with-signal" state, the build-up of oscillation is accelerated, and the repetition rate of bursts increases, affecting variations in the average anode current. If the applied signal is amplitude-modulated, the average current will follow the envelope of the modulated wave.

Review Questions

1. What is superregeneration?
2. Why is it that the noise level in a superregenerative receiver is higher than it is in a regenerative one?
3. Does the mode of operation of a superregenerative circuit affect the additional gain?
4. Draw up the diagram of a superregenerative circuit using a transistor.

SUMMARY

1. Application of controllable positive feedback in a receiver can considerably increase its sensitivity and selectivity.
2. The gain of a regenerative stage increases with the amplitude of the signal.
3. Maximum regenerative amplification is obtained when the stage is close to the critical point, i.e. when the circuit is about to oscillate. However, this is an unstable condition.
4. Superregeneration maintains operational stability and also offers higher amplification.

CHAPTER IX
SUPERHETERODYNE RECEPTION

48. General

As distinct from a tuned r.f. (TRF) receiver where all pre-detection amplification is done at the incoming signal frequency, a superheterodyne receiver is one in which the incoming signal frequency is changed by a frequency changer stage to a different, lower frequency before pre-detection amplification is completed.

This new frequency is called the *intermediate frequency (i.f.)*, and it is constant, irrespective of the frequency to which the receiver is tuned.

The intermediate frequency is produced by a *heterodyne process* and is always above the audible limit, that is, super-sonic. Hence, the receiver is called a *super-sonic heterodyne*, or a *superheterodyne*, or simply a *superhet*.

A superhet offers a number of advantages over a TRF receiver, namely higher selectivity and sensitivity which remain constant over the entire range of frequencies, and better stability.

The higher selectivity of a superhet comes from the fact that the frequency-conversion process lowers the radio frequency to be handled. As will be recalled, selectivity is a function of the relative amount off resonance, $\Delta f/f_0$, which, with the absolute amount off resonance held constant, increases as the frequency is decreased. Besides, since the i.f. is a fixed frequency, the i.f. amplifiers are fixed-tuned, and this allows them to be very carefully designed for better frequency response.

The high sensitivity of a superhet is also due to the frequency-conversion process, since the i.f. amplifier can secure a higher and more stable gain. This is also true of the constancy of sensitivity and selectivity over the frequency range, because the i.f. amplifier operates at a fixed frequency and its performance remains basically unchanged at any frequency within the range.

The stability of a superhet is improved owing to the fact that amplification is effected at three frequencies—radio, intermediate, and audio, instead of two in a TRF receiver. The reduced number of stages operating at the same frequency minimizes the danger of self-oscillation due to parasitic feedback.

Frequency conversion, like detection, is a non-linear process, and is performed by non-linear components. In principle, any type of detector may be used for frequency conversion.

Frequency conversion, as has been noted, is a heterodyne process, that is, one of mixing or beating two frequencies together to get the intermediate frequency, f_i. The two frequencies are the incoming signal frequency, f_s, and the local-oscillator frequency, f_o. Sometimes, f_o is supplied from a separate valve (*a local-oscillator valve*), and the heterodyning is done in a *mixer valve*. Sometimes, the mixer and the local oscillator are in a single valve which constitutes a *frequency changer* (also known as a *frequency converter*).

Figure 9.1 shows a simple mixer circuit in which a diode is used as the non-linear component. In order to separate the i.f. signal, the arrangement includes a resonant circuit tuned to the frequency $f_o - f_s$. Diode mixers are widely used on ultra-short waves. Frequency conversion on these waves is dealt with in detail in Chapter XIV. The present chapter discusses frequency conversion on long, medium, short and metre waves. On these waves, multi-grid valves are used as frequency changers. There are two forms of frequency conversion with multi-grid valves: *single-grid injection* and *double-grid injection*.

Single-grid injection uses amplifying valves (usually pentodes) with a single control grid. Double-grid injection uses valves with two control grids.

In the former case, the incoming signal and the local-oscillator output are injected at the same grid of the mixer valve.

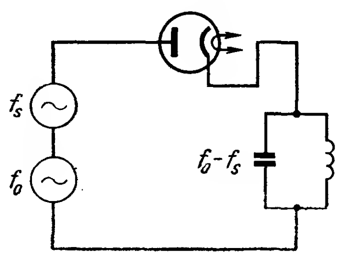


Fig. 9.1. Mixer circuit using a diode

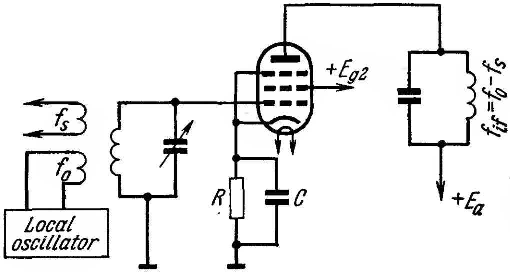


Fig. 9.2. Simple single-grid-injection mixer circuit using a pentode

In the latter case, the incoming signal is injected at one grid, and the local-oscillator output, at the other.

Single-grid Injection. Fig. 9.2 shows a pentode mixer with single-grid injection. The incoming signal frequency, ω_s, and the local-oscillator frequency, ω_o, are injected at the same control grid. The valve is made to operate as an anode-bend detector by applying to its grid the necessary bias voltage from R connected in the cathode circuit. The two frequencies heterodyne, or beat together, in the valve, and the output (anode) current contains beat frequencies which are the sum and difference of the two original ones:

$$n\omega_o \pm m\omega_s$$

where m and n are any integers. Frequency conversion uses the difference frequency, $\omega_o - \omega_s$. This is the intermediate frequency which is extracted by the resonant circuit connected in the anode circuit of the mixer valve and tuned to the i.f.

The valve characteristic may be described, with a certain degree of approximation, by a quadratic polynomial of the form

$$i_a = I_0 + av_g + bv_g^2 \tag{9.1}$$

The grid voltage is

$$v_g = V_{go} \cos \omega_o t + V_{gs} \cos \omega_s t \tag{9.2}$$

Substituting the expression for v_g in Eq. (9.1) and using the trigonometric expansion

$$\cos^2 \alpha = \frac{1}{2} + \frac{1}{2} \cos 2\alpha$$

and

$$\cos \alpha \cos \beta = \frac{1}{2} [\cos (\alpha + \beta) + \cos (\alpha - \beta)]$$

we obtain

$$i_a = I_0 + aV_{go}\cos\omega_o t + aV_{gs}\cos\omega_s t + \frac{1}{2}bV_{go}^2 + \frac{1}{2}bV_{gs}^2$$
$$+ \frac{1}{2}bV_{go}^2\cos 2\omega_o t + \frac{1}{2}bV_{gs}^2\cos 2\omega_s t + bV_{go}V_{gs}\cos(\omega_o - \omega_s)t$$
$$+ bV_{go}V_{gs}\cos(\omega_o + \omega_s)t \qquad (9.3)$$

As is seen from Eq. (9.3), besides the frequencies ω_o and ω_s, the anode current contains direct components (the first, fourth and fifth terms), the second harmonics $2\omega_o$ and $2\omega_s$ (the sixth and seventh terms), and the difference and sum frequencies (the last terms), which dominate all other terms.

The non-linearity of frequency conversion is confirmed by the fact that the frequencies $\omega_o \pm \omega_s$ are due to the quadratic term bv_g^2 in Eq. (9.1).

It should be noted that the difference frequency may be obtained in two different ways:

$$f_i = f_{o1} - f_s$$
$$f_i = f_s - f_{o2}$$

In the former case, the local oscillator is tuned to a frequency above the incoming signal frequency, $(f_o = f_s + f_i)$; it is usually used on long, medium and short waves. In the latter case, the local oscillator is tuned to a frequency below the incoming signal frequency, $(f_o = f_s - f_i)$; this is sometimes used in the VHF-UHF band.

The mixer stage, that is, the mixer valve and the associated tuned circuit, provides both amplification and selectivity.

Gain of the Mixer. The i.f. current is determined by the penultimate term of Eq. (9.3)

$$i_{if} = bV_{go}V_{gs}\cos(\omega_o - \omega_s)t = I_{if}\cos(\omega_o - \omega_s)t$$

where

$$I_{if} = bV_{go}V_{gs}$$

is the i.f. current amplitude.

Multiplying both sides of the equation by R_{0e}, we obtain the amplitude of the i.f. voltage across the tuned circuit

$$V_{if} = I_{if}R_{0e} = bV_{go}V_{gs}R_{0e}$$

The ratio of the i.f. voltage across the tuned circuit to the signal voltage at the mixer grid is known as the *mixer* or *conversion gain*.

Then

$$K_c = \frac{V_{if}}{V_{gs}} = bV_{go}R_{0e} \tag{9.4}$$

Since the mixer gain is a dimensionless quantity, the product bV_{go} may describe only the slope of the curve and have the dimensions of conductance. We shall call the product bV_{go} the *conversion conductance*, g_c, and re-write the expression for the mixer gain thus:

$$K_c = g_c R_{0e} \tag{9.5}$$

Mathematical analysis borne out by experiments shows that the relation between the conversion conductance and the static mutual conductance of the mixer valve may be written thus:

$$g_c = g_{m(max)}/(3 \text{ or } 4) \tag{9.6}$$

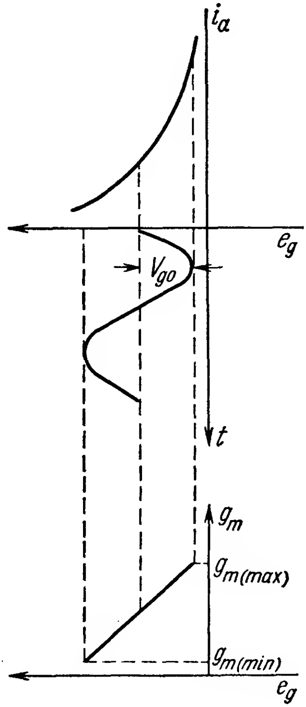


Fig. 9.3. Effect of the local-oscillator voltage on the mutual conductance of the valve in frequency-conversion with single-grid injection

where $g_{m(max)}$ is the static mutual conductance in the linear region of the valve characteristic, that is, within the amplification region.

Operation of the mixer stage is markedly affected by the mode of operation of the local oscillator. Thus, the amplitude of the l.o. output voltage, V_{go}, controls the conversion conductance, and this in turn controls the gain of the mixer stage. For a better insight in the matter, consider the plot of Fig. 9.3, which shows the non-linear region of the anode-grid characteristic used for frequency conversion, and the waveform of the l.o. voltage applied to the mixer grid.

The l. o. voltage, whose amplitude is very large in comparison with the signal amplitude, shifts the operating point along the characteristic curve from the region with a minimum mutual conductance, $g_{m(min)}$, into the region of a maximum value, $g_{m(max)}$. For clarity, the lower plot of Fig. 9.3 shows this change in g_m as a straight line, on the assumption that the anode-grid characteristic obeys the square law.

Thus, a maximum conversion conductance may be obtained only when variations in the voltage during a complete cycle of the local oscillator occupy the whole of the non-linear region, that is, when the mutual conductance varies from $g_{m(min)}$ to $g_{m(max)}$. This condition is satisfied when the peak value of the l.o. voltage is five to ten volts, according to the type of valve.

Going back to Eq. (9.5), it should be noted that the gain of a mixer stage is one-third to one-fourth of that of an amplifier stage using the same valve and the same tuned circuit. Eq. (9.5) holds for a mixer with a single-tuned circuit as load. When the load is a double-tuned circuit, the gain will be given by

$$K_c = \frac{\eta}{\eta^2+1} g_c R_{0e} \tag{9.7}$$

The *selectivity of the mixer* is determined by the type of resonant circuit in the anode lead. Most practical mixers use double-tuned band-pass filters. As already noted, the use of double-tuned circuits in fixed-tuned amplifiers provides for both high selectivity and flat gain over the entire bandwidth. Quantitatively, the selectivity is given by Eq. (6.21) in the case of double-tuned circuits, and by Eq. (6.11) for single-tuned circuits.

Double-grid Injection. This form of frequency conversion uses hexodes, heptodes (pentagrids), octodes, etc. Without going into details of each of these valves, consider the principle of double-grid injection.

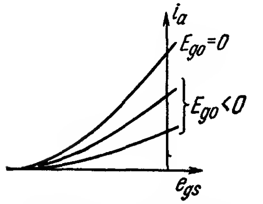


Fig. 9.4. Family of characteristics of a heptode

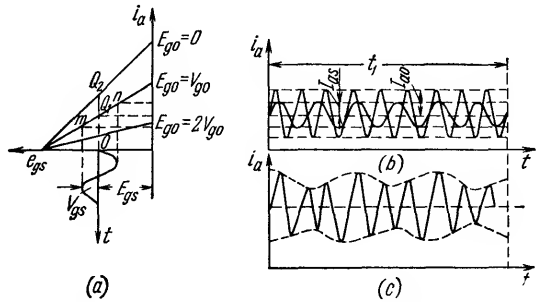


Fig. 9.5. Frequency conversion with double-grid injection

Valves with two control grids have two families of anode-grid characteristics, one $i_a=f(e_{gs})$ with E_{go} held constant for the signal grid, and the other, $i_a=\varphi(e_{go})$ with E_{gs} held constant, for the oscillator grid which accepts the output of the local oscillator.

Figure 9.4 shows typical characteristics of a heptode

$$i_a=f(e_{gs})$$

The characteristics are plotted for different fixed values of the oscillator grid voltage and are nearly linear. In each case, the mutual conductance, while remaining constant with any signal-grid voltage, varies as the oscillator grid voltage is changed. $E_{go}=0$ corresponds to the highest mutual conductance; the gradient of the curve decreases as E_{go} is decreased. When an a. c. voltage acts upon the oscillator grid, the mutual conductance follows changes in the oscillator voltage.

Operation of a valve with two control grids is illustrated in Fig. 9.5. The plot at (*a*) gives the idealized characteristics of the mixer valve. Two a. c. voltages of amplitudes V_{gs} and V_{go} are simultaneously applied to the control grids. The quiescent operating point Q_1 is determined by the grid biases $-E_{gs}$ and $-E_{go}=V_{go}$.

When no a. c. voltage is applied to the oscillator grid, the anode current is controlled only by the a.c. voltage at the signal grid. In this case, variations in the anode current are represented by the region mQ_1n of the average static characteristic, and the amplitude of the alternating component I_{as} is shown in plot (*b*). In the same plot, I_{ao} is the amplitude of the anode current when it is controlled only by the local-oscillator voltage V_{go}.

Now the valve characteristic is shifted bodily from the position corresponding to $E_{go}=0$ to the position corresponding to $-E_{go}=2V_{go}$, and the operating point shifts along the vertical straight line OQ_1Q_2.

The local-oscillator frequency and the incoming signal frequency beat together to produce sum and difference (or beat) frequencies in the anode circuit. The envelope of the difference frequency provides the required intermediate frequency, while the sum frequency is discarded.

In plot (*c*), the beat-frequency waveform has been obtained by algebraic addition of the ordinates of two sinusoids, each of which represents the alternating component of anode current due to the incoming signal and the local-oscillator signal separately.

Referring to plot (*b*), six cycles of i_{as} and nine cycles of i_{ao} occur during the time t_1, while three cycles are obtained in the beat-frequency envelope. This is graphic proof that the frequency-conversion process produces a difference frequency which can be isolated by a suitable filter.

Valves using double-grid injection are characterized by conversion parameters which can be found from static characteristics. These are: conversion conductance

$$g_c=\frac{I_{ac}}{V_{gs}} \qquad V_c=0 \tag{9.8}$$

valve amplification factor

$$\mu_c=\frac{V_c}{V_{gs}} \qquad I_{ac}=0 \tag{9.9}$$

a. c. anode resistance

$$R_{ac}=\frac{V_c}{I_c} \qquad V_{gs}=0 \tag{9.10}$$

or

$$R_{ac}=\frac{\mu_c}{g_c}$$

Review Questions

1. Why does a superheterodyne receiver offer better sensitivity and selectivity?
2. Why does a superheterodyne receiver have better stability?
3. Why is it that, other things being equal, a mixer stage has a lower gain than an amplifier stage?

4. What variable can be used to control the conversion conductance?
5. How does the mixer gain depend on the amplitude of the local-oscillator voltage?
6. What is the disadvantage of single-grid injection?
7. Which part of a mixer stage determines its selectivity?

49. Mixer and Frequency-changer Circuits

The operation of a frequency changer determines to a considerable degree the performance of the superheterodyne receiver. Above all, the changer affects the frequency range. For good coverage, the local oscillator must generate a steady signal over the whole range of frequencies being received. The amplitude of the local-oscillator voltage must remain more or less constant over the range, because it controls the conversion conductance and, consequently, the mixer gain over the range. This calls for careful design of the local-oscillator circuit, proper selection of the mixer valve, and maintenance of the optimum operating voltages.

Let us analyse in greater detail the factors controlling the frequency stability of the local oscillator.

By definition, superheterodyne reception is one in which the incoming signal frequency and the local-oscillator frequency heterodyne to produce the difference (or intermediate) frequency: $f_i = f_o - f_s$.

It is to this fixed intermediate frequency that the resonant circuits in the mixer stage and in the i.f. amplifier are tuned. When the local oscillator generates the assigned frequency, the intermediate frequency is that to which the tuned circuits are tuned, and the signal frequency spectrum is symmetrical about the axis of the resonance curve of the i.f. amplifier (Fig. 9.6*a*). A different situation is observed when the local oscillator is

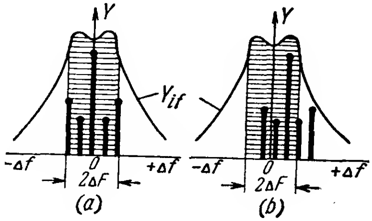


Fig. 9.6. Signal frequency spectrum
(a) at exact tuning of the oscillator; (b) when the oscillator is detuned

unstable. A deviation of the oscillator frequency from its assigned value will bring about a corresponding change in the intermediate frequency, f_i, so that it will move away from the frequency to which the i.f. amplifier is tuned, and the signal frequency spectrum will be unsymmetrical about the axis of the resonance curve (Fig. 9.6*b*). The extreme frequencies of the spectrum will be attenuated, and additional frequency distortion will take place. If the oscillator drifts considerably, the incoming signal frequency spectrum may fall outside the bandwidth, and reception of the signal will be impossible.

The frequency stability of the local oscillator depends on its circuit configuration and some external factors. It also depends on the degree of coupling between the r.f. tuned circuit and the tuned circuit of the local oscillator. If the coupling is tight, any change in the tuning of the r.f. tuned circuit will affect the reactance reflected from the r.f. circuit into the local-oscillator circuit, and there will be a change in the frequency of the local oscillator (known as *pulling*).

Mixer Circuits with Single-grid Injection. Mixer curcuits with single-grid injection employ pentodes operating as anode-bend detectors.

The simple mixer circuit with single-grid injection shown in Fig. 9.2 has a very serious shortcoming in that the oscillator-tuned circuit and the r. f. tuned circuit are coupled to each other. This coupling impairs the frequency stability of the local oscillator.

This disadvantage is minimized in the mixer circuit of Fig. 9.7*a*. In this circuit the oscillator voltage is applied to the mixer valve through a coil L_k connected to the cathode. The coupling between the tuned circuits is through the interelectrode capacitance C_{gk}, which is comparatively small, and the pulling of the local oscillator is reduced.

Fig. 9.7*b* shows a mixer circuit using a pentode with a high mutual conductance. In this case the necessary conversion conductance can be obtained when the local oscillator is loosely coupled to the control grid. The typical values of the coupling capacitor are 1 to 4 picofarads. Mixer circuits with single-grid injection are seldom used. Their application is chiefly limited to radio receivers which, owing to special requirements, have to employ the same type of valve in all stages.

Mixer and Frequency-changer Circuits Using Double-grid Injection. In mixers and frequency changers using double-grid

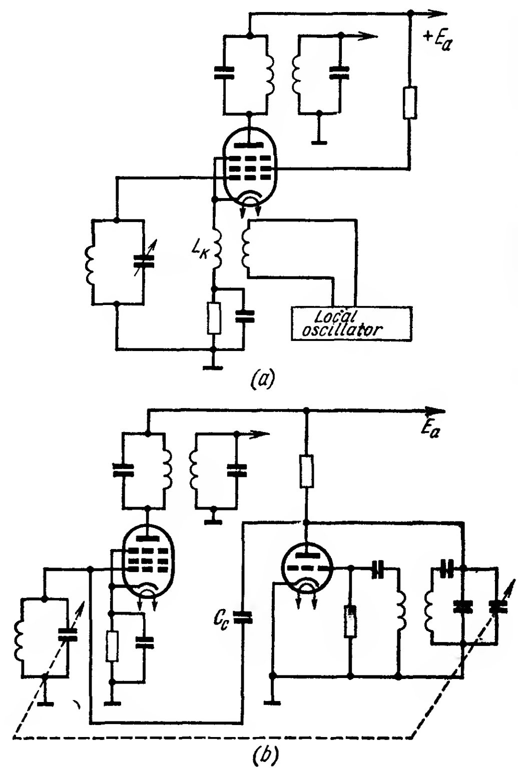


Fig. 9.7. Pentode mixer circuit

injection, the coupling between the r.f. tuned circuit and the local oscillator is considerably reduced by the screen grid or grids between the oscillator and signal grids.

Some valves using double-grid injection (such as heptodes) may be of two kinds. One is designed as a mixer valve; the other combines the functions of a mixer and local-oscillator valve. Valves of this kind are known as *frequency changers* (see Sec. 48).

Other valves with double-grid injection, such as triode-hexodes and triode-heptodes, can be used only as frequency-changers.

Heptode Frequency Changer. Heptodes are valves with five grids, and are frequently referred to as *pentagrids*.

A typical heptode frequency-changer circuit is shown in Fig. 9.8. The incoming signal is taken from the left-hand r.f. tuned

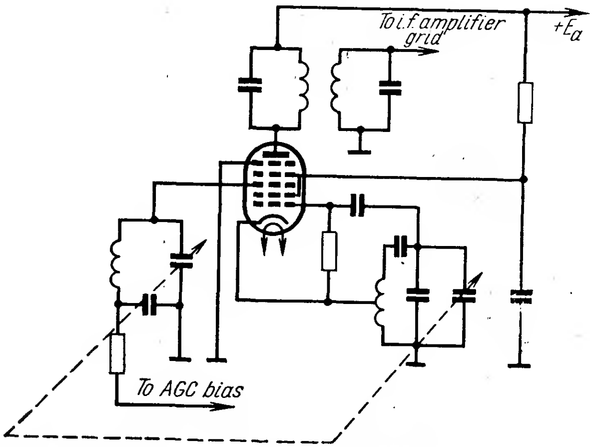


Fig. 9.8. Heptode frequency-changer circuit

circuit and fed to the injector grid $G3$. The cathode, $G1$, and $G2$ act as a triode in the local oscillator which operates as a feedback (or self-) oscillator. $G2$ and $G4$ are connected together and act as the anode of the local-oscillator triode. At the same time, they act as screen grids isolating the injector grid from the oscillator grid, and are therefore held at r.f. earth potential.

Since, however, the anode of the local-oscillator valve is at r.f. earth potential, the local oscillator is of necessity arranged into a cathode-coupled circuit, with the tuned circuit placed between $G1$ and earth.

The tuned circuits contain series capacitors, called *padders*, which are used for alignment of *tracking*, or the proper frequency relationship between the r.f. and local-oscillator tuned circuits.

The anode circuit of the frequency changer contains an i. f. transformer, which is a band-pass filter. It secures the requisite selectivity and passes the intermediate frequency.

Heptode Mixer. The incoming signal is applied to the first grid of the mixer (Fig. 9.9); a separate local oscillator feeds its voltage to the third grid. The second and fourth grids are the screen grids, and the fifth is the suppressor.

The output of the local oscillator, which is arranged into a cathode-coupled circuit, is taken from the lower part of L_o and is applied to the oscillator grid in the mixer via an RC network.

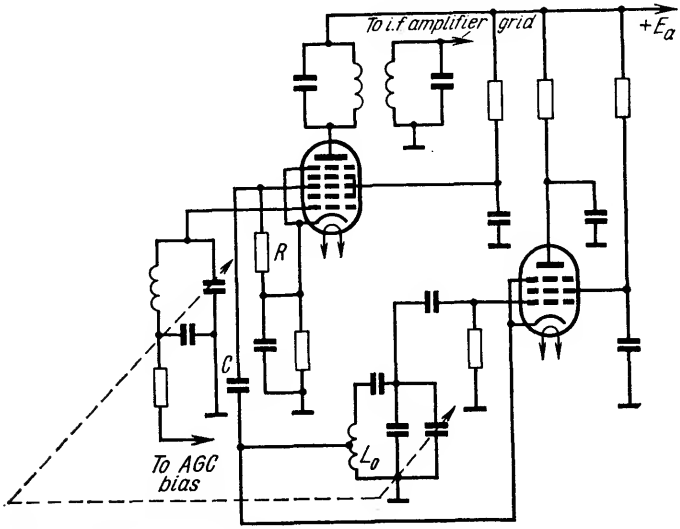


Fig. 9.9. Heptode mixer

Triode-Heptode Frequency Changer. The circuit diagram of a triode-heptode frequency changer appears in Fig. 9.10. The triode is used in a local oscillator, and the heptode acts as a mixer. The local oscillator is an inductively-coupled feedback oscillator.

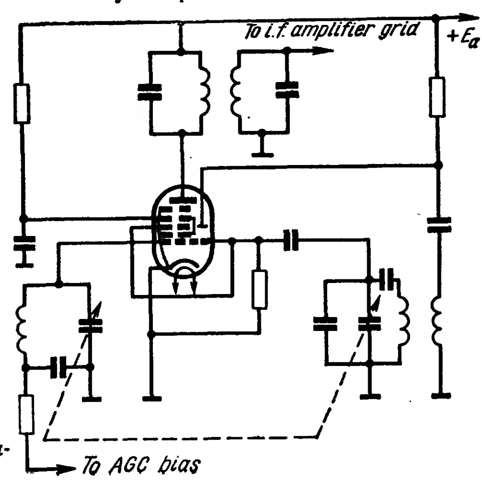


Fig. 9.10. Triode-heptode frequency changer

Review Questions

1. How does the frequency instability of the local oscillator affect operation of the mixer or frequency changer?
2. Can a pentode be used in a frequency changer with double-grid injection?
3. Why is it that the local oscillator in a heptode frequency changer is arranged into a cathode-coupled circuit?
4. How will an increase in coupling between the tuned circuits of the i.f. transformer affect the bandwidth of the frequency changer?

50. Transistor Mixers and Frequency Changers

The non-linearity of semiconductor devices makes them suitable for use in mixer and frequency changers on a par with valves. Both semiconductor diodes and transistors are used for the purpose. The circuit of a semiconductor diode does not differ in any respect from the valve diode circuit shown in Fig. 9.1. Frequency conversion by means of transistors is similar to the single-grid injection method.

Figure 9.11 shows a mixer employing two transistors. This circuit resembles the valve mixer of Fig. 9.7. The incoming signal voltage V_s, taken from the tuned circuit K_1 is applied through a coupling coil L_2 to the base of the transistor connected into a common-emitter circuit. The local oscillation is taken from the bottom part of the coil in the tuned circuit K_2 and is injected

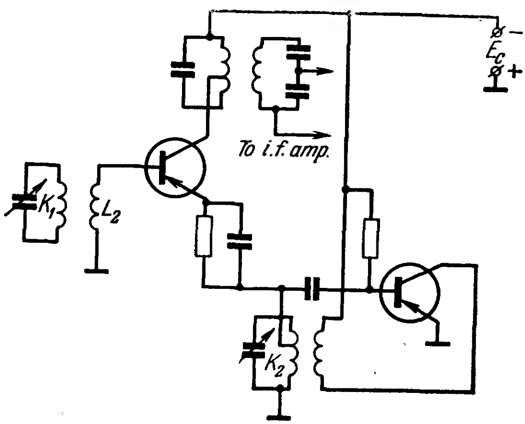


Fig. 9.11. Transistor mixer

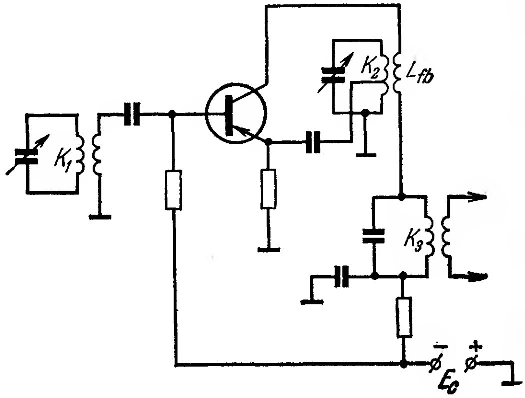


Fig. 9.12. Transistor frequency changer

into the emitter circuit of the mixer transistor. The local oscillator is a self-oscillator with inductive feedback. The collector (output) circuit of the mixer transistor contains an i.f. transformer.

Figure 9.12 shows the circuit of a transistor frequency changer in which the same transistor functions as mixer and local oscillator. The r.f. signal voltage is taken from the tuned circuit K_1 and is applied to the base of the transistor connected into the common-emitter circuit. The local oscillator uses inductive feedback to the tuned circuit K_2 in the emitter circuit. The feedback coil L_{fb} is connected in the collector circuit together with the output tuned circuit K_3 tuned to the intermediate frequency. In this circuit, the incoming signal frequency and the local-oscillator frequency beat together.

Review Questions

1. What form of valve frequency conversion can be accomplished with a transistor circuit?
2. Why is it that in transistor frequency changers the signal is applied to the i. f. amplifier via a coupling coil?
3. Which circuit in a transistor frequency changer governs its selectivity?

51. Particulars of Superheterodyne Reception

Spurious Responses. Of all spurious responses, the more important is *image* or *second-channel interference*. As already noted, the intermediate frequency is produced by beating together the incoming signal frequency and the local oscillation.

Suppose that the local oscillator has been tuned to receive a signal of frequency f_{s1} such that

$$f_o = f_i + f_{s1}$$

However, the receiver can respond also to another signal, f_{s2}, if it is such that

$$f_{s2} - f_o = f_i \tag{9.11}$$

that is, when the same i.f. response is obtained as with the wanted signal

$$f_o - f_{s1} = f_i \tag{9.12}$$

The unwanted signal which causes interference, if not eliminated, is called the *image* for the reason that it is symmetrical about the local-oscillator frequency like a mirror image of the wanted signal (Fig. 9.13). The image is removed from the wanted signal by twice the intermediate frequency:

$$\left.\begin{array}{r} f_{s2} - f_o = f_i \\ f_o - f_{s1} = f_i \\ \hline f_{s2} - f_{s1} = 2f_i \end{array}\right\} \tag{9.13}$$

For example, when $f_{s1} = 3{,}920$ kilohertz and $f_i = 465$ kilohertz, the image will be $f_{s2} = f_{im} = 3{,}920 + 930 = 4{,}850$ kilohertz.

Another form of spurious response is *adjacent channel interference*. As its name implies, it arises when the unwanted signal is very close to the wanted one in frequency, so that about the same i.f. response is produced by the local oscillator.

Then there is *intermediate-frequency interference* which arises when the unwanted signal picked up by the aerial is at, or near, the i.f. of the receiver. If this unwanted signal succeeds in

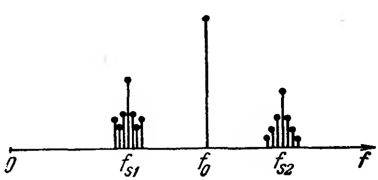


Fig. 9.13. Image channel interference

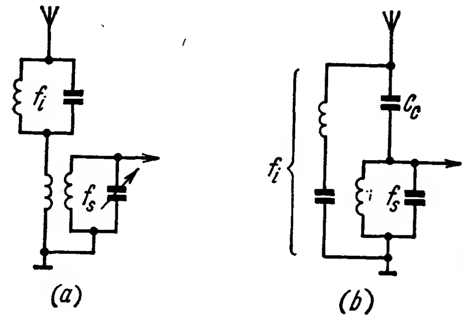


Fig. 9.14. I.f. rejector circuits

breaking through to the signal grid of the mixer, it will receive, there and thereafter, as much amplification as the wanted signal.

Spurious responses may be suppressed only in the circuits preceding the mixer, i.e. in the aerial-input circuit and in the radio-frequency amplifier.

I.f. interference can be eliminated by filters connected at the receiver input. Figure 9.14*a* shows an *i.f. rejector* connected into the aerial-input circuit. This is a parallel resonant circuit tuned to resonate at f_i. At resonance, it offers a high impedance to any i.f. signal, thereby blocking its passage to the mixer. Figure 9.14*b* shows another type of i.f. rejector. This is a series resonant circuit, connected across the receiver input and tuned to the intermediate frequency. At resonance, its impedance drops to a very low value, and the receiver input is short-circuited for the intermediate frequency.

Heterodyne Whistle. Harmonics of the local oscillator may beat with signals far removed from the desired frequency to produce output at, or near, the intermediate frequency, which will go through the receiver in the same way as an ordinary signal.

Suppose there is a beat frequency

$$f_b = 2f_s - f_o$$

If $f_s = 931$ kilohertz and $f_i = 465$ kilohertz, the frequency of the local oscillator will be

$$f_o = f_s + f_i = 931 + 465 = 1{,}396 \text{ kilohertz}$$

whence

$$2f_s - f_o = 1{,}862 - 1{,}396 = 466 \text{ kilohertz}$$

Detection will produce heterodyne beats of frequency $f_b - f_i = 466 - 465 = 1$ kilohertz, which will appear as "birdies" in the speaker.

Selectivity. In contrast to the selectivity of a straight (TRF) receiver, that of a superheterodyne receiver is expressed in terms of two quantities.

1. *Adjacent-channel selectivity* (sometimes called *selectance*), defined as the ability of a superhet to discriminate against adjacent-channel signals.

Adjacent-channel selectivity is provided by all selective circuits of the receiver. Yet, in view of the broad bandwidth of the resonant circuits preceding the frequency changer, particularly on short waves, it may be said that adjacent-channel attenuation is chiefly provided by the intermediate-frequency amplifier.

2. *Image-channel selectivity*, defined as the ability of a superheterodyne receiver to discriminate against image frequency.

The image signal can be suppressed only in the circuits preceding the frequency changer, which is therefore done by the aerial-input circuit and the radio-frequency amplifier.

In Chapter V the selectivity of the aerial-input circuit was given by

$$Y = \frac{I}{I_0} = \frac{1}{\sqrt{1+x^2}}$$

where $x = Q_{ef}\left(\frac{f}{f_0} - \frac{f_0}{f}\right)$ is the generalized frequency separation. Referring to Fig. 5.1, the output voltage is taken from the tuned-circuit capacitor and is fed to the grid of the next stage. Hence, for correct evaluation, the selectivity must be represented by the ratio of voltages

$$Y_V = \frac{V_{out}}{V_{out\,0}} = \frac{I\frac{1}{\omega C}}{I_0\frac{1}{\omega_0 C}}$$

Putting $\frac{I}{I_0} = Y$, we obtain

$$Y_V = Y\frac{\omega_0}{\omega}$$

As a result, the voltage selectivity of the tuned circuit takes the form

$$Y_V = \frac{1}{\sqrt{1+x^2}}\frac{f_0}{f} \tag{9.14}$$

The only difference between the current and voltage selectivity equations is the relative frequency.

When the frequency separation between the wanted and the unwanted signals is small, say $\frac{f-f_0}{f_0} \cong 0.1$, the ratio $\frac{f_0}{f}$ may be considered to be close to unity, and $Y \cong Y_V$.

In other words, the adjacent-channel selectivity of the aerial-input circuit is given by Eq. (5.4), and

$$x = Q_{ef}\frac{2\Delta f}{f_0}$$

The image channel selectivity of the aerial-input circuit should be determined from Eq. (9.14). In the special case of short waves, where the frequency separation $\Delta f = 2f_i$ is small in comparison with f_0, the current selectivity equation should be used.

At large values of absolute frequency separation, $\Delta f = f - f_0$, which is the frequency separation of the image channel, the generalized frequency separation is $x \gg 1$. Therefore, the image channel selectivity equation may be written as:

$$Y_{im\ in} = \frac{1}{x}\frac{f_0}{f}$$

Noting that $f = f_0 + 2f_i$ and writing the image channel selectivity as a reciprocal of Y

$$d_{im\ in} = \frac{1}{Y_{im\ in}} = x\frac{f}{f_0} = Q_{ef}\left(\frac{f_0+2f_i}{f_0} - \frac{f_0}{f_0+2f_i}\right)\frac{f_0+2f_i}{f_0} \quad (9.15)$$

After simplification, the design equation for the image channel selectivity of the aerial-input circuit reduces to

$$d_{im\ in} = Q_{ef}\left[\left(\frac{f_0+2f_i}{f_0}\right)^2 - 1\right] \quad (9.16)$$

It should be noted that for a tuned radio-frequency amplifier the current and voltage selectivity is the same. Hence, the selectivity equation (5.4) applies to radio-frequency amplifiers for both small and large frequency separation between the wanted and unwanted signals.

With $x \gg 1$, the image-channel selectivity of a single-stage tuned amplifier according to Eq. (5.4) will be

$$d_{im\ rf\ amp} \cong x = Q_{ef}\left(\frac{f_0+2f_i}{f_0} - \frac{f_0}{f_0+2f_i}\right) \quad (9.17)$$

The image channel selectivity of the aerial-input circuit and the radio-frequency amplifier with similar tuned circuits is, in

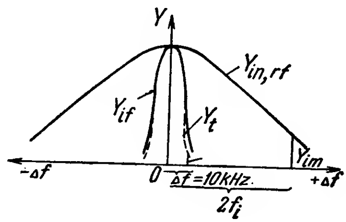


Fig. 9.15. Resonance characteristics of a superheterodyne receiver

accordance with Eqs. (9.15) and (9.17), given by

$$d_{im}=\left[Q_{ef}\left(\frac{f_0+2f_i}{f_0}-\frac{f_0}{f_0+2f_i}\right)\right]^n\frac{f_0+2f_i}{f_0} \tag{9.18}$$

where n is the number of tuned circuits in the aerial coupler and radio-frequency amplifier.

The selectivity of a superheterodyne receiver may be represented by its resonance characteristics. Figure 9.15 shows the resonance curve of the aerial-input circuit and the radio-frequency amplifier ($Y_{in\text{-}rf}$), the resonance curve of the i.f. amplifier Y_{if} and the overall resonance curve Y_t, which is obtained by multiplying the ordinates of the curves. When $\Delta f=2f_i$ the ordinate Y_{im} on the resonance curve $Y_{in\text{-}rf}$ represents the image channel selectivity. The adjacent-channel selectivity is found from the Y_t curve at $\Delta f=10$ kilohertz (which is the frequency separation for communication and broadcast stations by international agreement) and is represented by the segment Y. On short waves, the resonance curve $Y_{in\text{-}rf}$ is gently sloping and has practically no effect on the adjacent-channel selectivity. On medium and, especially, long waves, this curve is very sharp, and the overall bandwidth of the receiver is therefore considerably narrowed. In calculating a superhet, it should be sought that the reduction in bandwidth due to r.f. circuits should be kept at 10 to 15 per cent.

Choice of Intermediate Frequency. The value of intermediate frequency is pivotal to the sensitivity, selectivity, gain, and bandwidth of the superhet.

For example, a low intermediate frequency offers the advantages of high sensitivity due to high i.f. stable gain, and also good adjacent channel rejection. Furthermore, for a given quality factor of the i.f. tuned circuit, Q_{if}, a low intermediate frequency results in a narrow bandwidth, $2\Delta F=f_i/Q_{if}$ and, as a consequence, a sharp resonance curve.

On the other hand, good image rejection is provided by a high intermediate frequency. Referring to Fig. 9.15, an increase in frequency separation, $\Delta f = 2f_i$, improves the rejection of the image frequency (the ordinate Y_{im} decreases).

Obviously, one cannot secure high sensitivity, high adjacent channel selectivity, and high image rejection at one and the same value of intermediate frequency. This is why the choice of the intermediate frequency is a matter of balance between conflicting considerations.

To begin with, it must not be that used by any transmitters within the operating range. Otherwise, it might be picked up directly in the receiver wiring or in the aerial, and result in heterodyne whistle.

In broadcast and some communication receivers, the intermediate frequency is 465 kilohertz which falls in between the medium- and long-wave bands.

In communication receivers operating in the long-wave range, it is good practice to use 110 kilohertz. Then, for a wanted signal frequency of 150 kilohertz, the image-channel frequency will be $f_{im} = 370$ kilohertz which is twice the wanted signal frequency and will be easily rejected by the aerial-input circuit.

Short-wave communication receivers secure good image-channel rejection by using an intermediate frequency of 1,600 kilohertz.

To reduce image response, communication receivers with a narrow bandwidth of 500 hertz, intended for reception of telegraph signals, employ two intermediate frequencies, or so-called *double-frequency conversion* (Fig. 9.16). The first intermediate frequency is high; the second intermediate frequency is sufficiently low to obtain a narrow bandwidth.

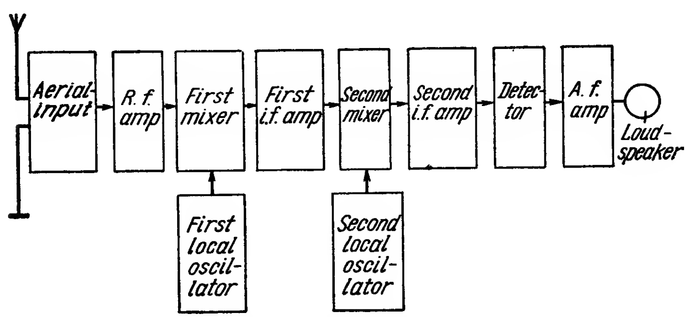


Fig. 9.16. Block diagram of a double-conversion superheterodyne receiver

Review Questions

1. When is the image frequency lower than the wanted signal frequency?
2. What circuits reject signals at or near the intermediate frequency?
3. What circuits provide image rejection in a superhet?
4. When do the aerial-input circuit and the r.f. amplifier affect adjacent-channel selectivity?
5. How does an increase in the intermediate frequency affect adjacent channel selectivity?
6. Which of the receiver resonant circuits control its bandwidth in the long-wave range?

52. Ganged Tuning of Superheterodyne Receivers

In broadband superheterodyne receivers the aerial-input circuit, radio-frequency amplifier and local oscillator are tuned with variable tuning capacitors. Most communication and all broadcast receivers have the variable capacitors of all the tuned circuits mounted on one shaft. This allows the receiver to be tuned with a single control.

This is *ganged tuning*. The number of ganged sections is the same as that of tunable circuits. So, there are two-, three- and even four-gang capacitors (Fig. 9.17).

This operating convenience, however, involves some difficulties. The point is that the tuning of the local oscillator must always be separated exactly the intermediate frequency from the signal frequency. The variable capacitor of the local oscillator is therefore ganged to the r.f. tuning capacitors so that they move

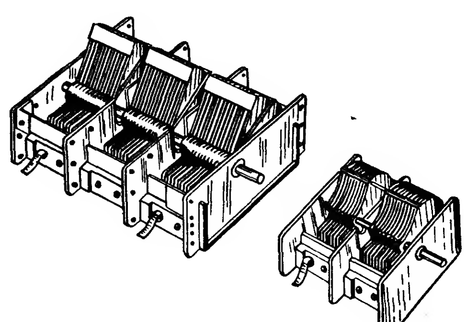


Fig. 9.17. Two-and three-gang capacitors

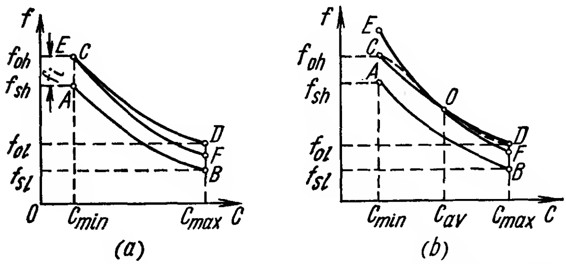


Fig. 9.18. Ganging of radio-frequency and local-oscillator circuits
CD— required oscillator tuning; EF—actual oscillator tuning; AB—aerial-input circuit and r.f. amplifier tuning

together. The r.f. tuned circuits are made as nearly identical as possible. The local oscillator circuit, clearly, cannot be the same. Because of this, the same change in the setting of the variable capacitors will produce different changes in the r.f. and local oscillator tuning.

Consider how changes in the setting of the variable capacitors affect the tuning.

With a fixed inductance L, the coverage from f_h to f_l is determined by the limits of capacitance, C_{min} to C_{max}, of the ganged capacitor.

The tuning of radio-frequency circuits is given by

$$f_s = \frac{1}{2\pi\sqrt{LC}} \tag{9.19}$$

Figure 9.18 relates the resonant frequencies of the receiver tuned circuits to the capacitance of their variable capacitors. Eq. (9.19) is represented by curve AB. In accordance with the superheterodyne principle the requisite tuning of the local oscillator may be expressed as

$$f_o = f_s + f_i = \frac{1}{2\pi\sqrt{LC}} + f_i \tag{9.20}$$

Eq. (9.20) is represented in Fig. 9.18*a* by curve CD, parallel to curve AB.

Let us determine how the actual tuning of the local oscillator varies. Suppose that at a minimum capacitance C_{min} and the corresponding value of the tuned-circuit inductance L_o the local-oscillator frequency is represented by point E coinciding with point C. In this case the tuned circuits are said to be *tracking* at the high-frequency end of the range.

As the capacitance is varied, the local oscillator frequency will change in conformity with the equation

$$f_o = \frac{1}{2\pi\sqrt{L_o C}} \qquad (9.21)$$

Equation (9.21) differs from Eq. (9.20), and, consequently, the actual variations in the tuning of the local oscillator are other than the desired ones.

The third curve EF, representing Eq. (9.21), passes below curve CD because $L_o < L$, and for each value of C in Eq. (9.21) there will be an ordinate f which is smaller than the respective ordinate in Eq. (9.20).

The discrepancy between curves CD and EF shows that at all points, except the upper one, the difference frequency $f_o - f_s$ is other than the assigned value, and, therefore, the receiver cannot provide the desired coverage over the range.

Example 9.1. A radio receiver covers a frequency range from $f_h = 1{,}600$ kilohertz to $f_l = 400$ kilohertz. The intermediate frequency is $f_i = 200$ kilohertz. Find the difference frequency $f_o - f_s$ at the low-frequency end of the range, if the tuned circuits are tracking at the high-frequency end of the range.

Solution. From the statement of the problem, the frequency at the low-frequency end of the range is four times down. Consequently, the capacitance of the variable capacitors must increase 16 times.

Since the tuned circuits are tracking at the high-frequency end of the range (point C), the frequency of the local oscillator is $f_o = 1{,}600 + 200 = 1{,}800$ kilohertz.

As the capacitance of the local oscillator increases 16 times, its frequency goes down to one-fourth of its value

$$f_{o\,l} = \frac{1{,}800}{4} = 450 \text{ kilohertz}$$

Thus, the difference frequency at the low-frequency end of the range (point F)

$$f_{o\,l} - f_{s\,h} = 450 - 400 = 50 \text{ kilohertz}$$

differs from the assigned value of the intermediate frequency $f_i = 200$ kilohertz.

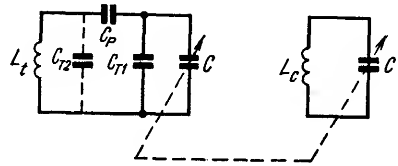


Fig. 9.19. Tracking of the gang-tuned local oscillator and aerial-input circuit

A much better result will be obtained when curves *CD* and *EF* coincide at point *O* (Fig. 9.18*b*) corresponding to the mid-frequency band of the range. Then the discrepancy at the extreme ends of the range will be smaller.

In practical cases, the tuned circuits are made to track at three points of the range: at the beginning, at the end and in the middle. At other points, the frequency of the local oscillator is allowed to differ slightly from its exact value. The requisite alignment is secured by connecting two or more capacitors in the local-oscillator circuit, as shown in Fig. 9.19.

The capacitance of the parallel-connected capacitor C_{T1}, (or C_{T2}) called a *trimmer*, is low, and can be varied within tens of picofarads. The series-connected capacitor C_p called a *padder* has a capacitance of 1,000 picofarads when used along with standard variable capacitors (17 to 500 picofarads) on medium and long waves. On short waves, its capacitance may be several thousand picofarads.

Let us determine the effect of the padder and trimmer on the capacitance of the local-oscillator circuit at two extreme settings, C_{min} and C_{max}, of the tuning capacitor.

(1) $C = C_{min}$.

The capacitance of the parallel connection

$$C_{par} = C_{min} + C_{T1}$$

has increased in comparison with C_{min}.

The total capacitance of the local-oscillator circuit is

$$C_{tot} = \frac{C_{par} C_p}{C_{par} + C_p}$$

Since C_p is usually much greater than C_{par}, we can neglect C_{par} in the denominator, so that

$$C_{tot} \cong C_{par} > C_{min}$$

The increased capacitance decreases the frequency of the tuned circuit. Consequently, point *E* on curve *EF* in Fig. 9.18*b* is shifted down. With proper alignment, that is, proper adjustment of C_p and C_{T1}, point F should coincide with point C.

(2) $C = C_{max}$.

The capacitance of the parallel connection increases only slightly due to C_{T1}. Noting that $C_{max} \gg C_{T1}$

$$C_{par} \cong C_{max}$$

The total capacitance of the local-oscillator circuit

$$C_{tot} = \frac{C_{max} C_p}{C_{max} + C_p}$$

is considerably lower in comparison with C_{max}, because C_{max} and C_p are comparable in magnitude. A decrease in capacitance is accompanied by an increase in frequency. With proper alignment, that is, proper adjustment of C_p, point F on curve EF must coincide with point D (see Fig. 9.18*b*).

The actual tuning curve of the local oscillator (the dotted line) coincides with the desired tuning curve at three points, while at other points it approaches curve CD.

Figure 9.20 shows an approximate curve of the tracking error. Here, f_l and f_h are the limiting frequencies of the range while f_1, f_2 and f_3 are the settings at which the r.f. and local oscillator tunings are exactly right *(perfect tracking)*.

The tracking error, defined as the difference between the actual and the required tuning of the oscillator, is given by

$$\Delta f = (f_s + f_i) - f_o \tag{9.22}$$

The error may be a maximum, Δf_{max}, at the ends of the range and between the points of perfect tracking. It is important to determine the limits of the tracking error already in the design stage for the reasons explained below.

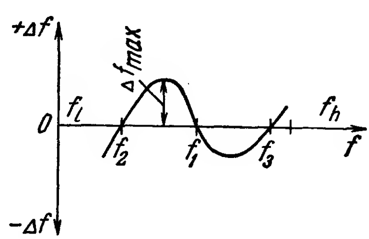


Fig. 9.20. Tracking-error curve

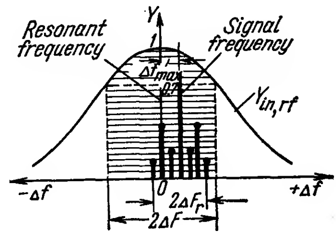


Fig. 9.21. Determining the limits of the tracking error from the resonance curve of the aerial-input circuit and r.f. amplifier

Let us see what happens when the receiver is being tuned at a point of imperfect tracking. The operator tries to secure maximum output, which is obtained when the intermediate frequency amplifier is tuned to resonance. The intermediate frequency will have its assigned value only when imperfect tracking is corrected by a respective change of the local oscillator frequency. Because of ganged tuning, however, this will detune the radio-frequency circuits by an amount equal to the tracking error.

The limits of the tracking error may be determined from the plot of Fig. 9.21. Suppose that the signal frequency along with its sideband spectrum, limited by the bandwidth of the receiver, is shifted from the resonant frequency of the radio-frequency circuits by an amount equal to the maximum tracking error Δf_{max}. Since the aerial-input circuit and r.f. amplifier have a broad bandwidth, the tracking error will only slightly affect operation of the receiver. However, if the side frequencies fall even slightly outside the bandwidth the sensitivity and selectivity of the receiver will be impaired, and frequency distortion will be considerably increased.

Thus, the limit of the tracking error is decided by the shape of the combined resonance curve of the aerial-input circuit and radio-frequency amplifier. On long and medium waves, a sharp resonance curve limits the tracking error to several kilohertz; on short waves, it may be tens of kilohertz.

According to Fig. 9.21, the acceptable limit of the tracking error may be expressed as

$$\Delta f_{max} + \Delta F_r < \Delta F \tag{9.23}$$

where ΔF_r is a half-bandwidth of the receiver.

Review Questions

1. Which circuits of a superhet may be gang-tuned?
2. When will a single padder capacitor suffice in ganged tuning?
3. On what frequencies may the tracking error be a maximum?

53. Calculation of a Frequency Changer

The complete procedure includes calculation of the local-oscillator parameters, conditions for self-oscillation, and the mixer. When a receiver is being designed, it is usually sufficient to calculate the tuned-circuit components of the local oscillator and mixer. If necessary, the conditions for self-oscillation of the local oscillator may be calculated by the methods explained in texts on radio transmitters.

Calculation of the Local-oscillator Parameters. This is done from considerations of ganged tuning. The computation may be performed analytically or graphically. The analytical method is cumbersome and calls for extra care.

What follows is one of approximate graphical methods, applicable to the local-oscillator circuit of Fig. 9.19 where C is the capacitance of each tuned circuit in the aerial coupler and r.f. amplifier, comprising the capacitance C_v of the variable capacitor and the shunt capacitance C_s which is connected in parallel with C_v, such that

$$C = C_v + C_s$$

We assume that the local oscillator uses a variable capacitor C_v identical with those in the aerial-input circuit and the r.f. amplifier. C_P is the capacitance of the padder, C_{T1} is the capacitance of the trimmer.

Given:

1. Frequency range, f_l-f_h.
2. Intermediate frequency, f_i.
3. Tuned-circuit inductance L of the aerial-input circuit and r.f. amplifier.

To Find:

1. Inductance L_o of the local oscillator.
2. C_P.
3. C_{T1}.

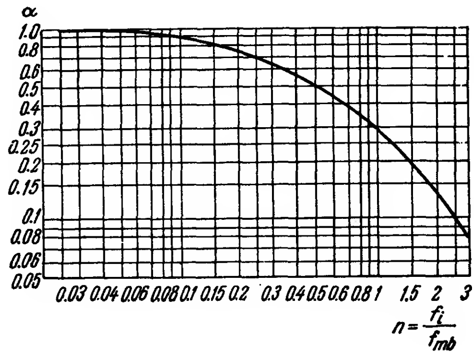


Fig. 9.22. Curve for determining the inductance of the local-oscillator circuit

Design Procedure:

1. Select a variable capacitor and assume that the shunt capacitance is the same, C_s, as in the aerial-input circuit and the r.f. amplifier.

2. Find the ratio $n = \frac{f_i}{f_{mb}}$, where

$$f_{mb} = \frac{f_h + f_l}{2}$$

is the midband frequency.

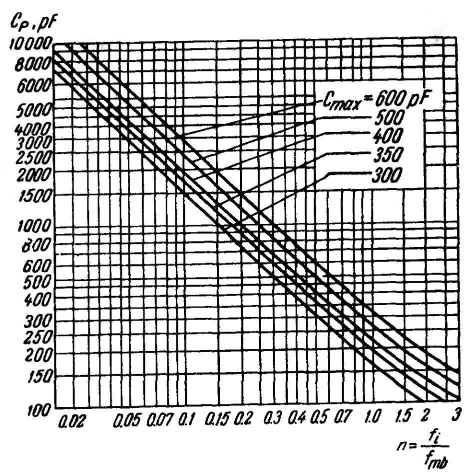


Fig. 9.23. Curve for determining the capacitance of the padder C_P

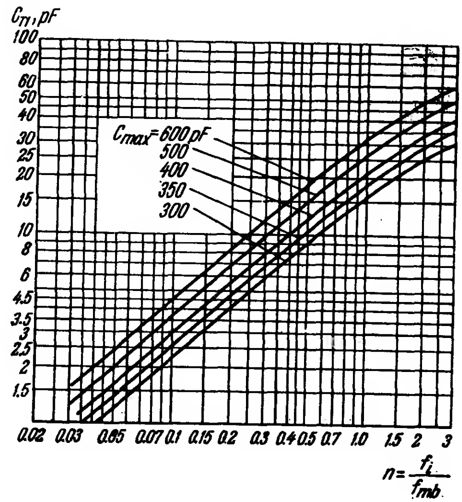


Fig. 9.24. Curve for determining the capacitance of the trimmer C_{T1}

3. Determine the maximum capacitance of the local-oscillator circuit

$$C_{max} = C_{v\,max} + C_s$$

4. Determine the inductance of the local oscillator

$$L_o = \alpha L$$

The value of α can be taken from Fig. 9.22.

5. Find the capacitance of the padder C_p from the plot of Fig. 9.23.

6. Find the capacitance of the trimmer C_{T1} from the plot of Fig. 9.24.

Calculation of the Mixer. The frequency changer uses the same i.f. transformer (band-pass filter) as the i.f. amplifier, which is usually calculated before the frequency changer.

Given:

1. Intermediate frequency, f_i.
2. Resonant resistance of the i.f. transformer (R_{0e}).
3. Conversion gain, K_c.
4. I.f. transformer coupling parameter, η.

To Find:

1. Type of valve.
2. Conversion gain K_c (as a check).

Design Procedure:

1. Select the circuit configuration and type of valve for the mixer.
2. Select the method of connection of the i.f. transformer primary to the anode circuit of the mixer valve so that the shunting effect of the valve would not exceed 25 per cent.

For this purpose, find the coupling parameter p (the tapping-down factor)

$$p \leqslant \frac{1}{2} \sqrt{\frac{R_{a(con)}}{R_{0e}}}$$

where $R_{a(con)} = (1.5 \text{ to } 2)\, R_a$, and R_a is the a.c. anode resistance of the mixer valve.

If $p \geqslant 1$, the i.f. transformer primary may be directly connected to the anode circuit. When $p < 1$, use should be made of transformer or tapped-coil coupling. If inductive coupling is not desirable, another type of valve with a higher value of R_a should be chosen.

3. Determine the inductance of the anode coil (only for inductive coupling):

(a) for transformer coupling

$$L_a = \left(\frac{p}{k}\right)^2 L$$

where the coefficient of coupling is assumed to be

$$k = 0.4 \text{ to } 0.6$$

(b) for tapped-coil coupling

$$L_a = pL$$

4. Find the conversion gain of the mixer

$$K_c = \frac{\eta}{\eta^2 + 1} g_c R_{0e} p$$

When the tuned circuit is directly connected to the anode circuit, $p = 1$.

The computed value of K_c must exceed its specified value.

SUMMARY

1. In contrast to a straight (TRF) receiver, the incoming-signal frequency in a superhet is converted to a lower (intermediate) frequency before detection.

2. Since most of the pre-detection amplification is at the relatively low intermediate frequency, superhets compare favourably with all other types of receiver in selectivity, sensitivity, and stability.

3. Frequency conversion is effected by a non-linear process taking place in a frequency-changer (or mixer) valve or transistor as its mutual conductance follows variations in the local-oscillator voltage.

4. Valves used as frequency changers are characterized by the following conversion parameters: g_c, $R_{a(con)}$, and μ_c. The most important parameter is the conversion conductance g_c which is one-third to one-fourth of the maximum static mutual conductance of the valve.

5. The gain of the frequency-changer stage depends on the type of valve and anode load resistance. Its value is from one-third to one-fourth of the gain of an amplifier stage using similar valves and load resistances.

6. The selectivity of the frequency-changer is mainly determined by the type of resonant circuit connected into its anode lead.

7. Frequency conversion may use single-or double-grid injection.

8. A major disadvantage of the superheterodyne receiver lies in spurious responses, above all image response.

9. In the superheterodyne receiver two types of selectivity are distinguished: image selectivity and adjacent-channel selectivity. The former is secured by the aerial-input circuit and r. f. amplifier, and the latter, mainly by the intermediate-frequency circuits.

10. Ganged tuning of the superheterodyne receiver is made possible by the tracking of the tuned circuits, secured by connecting padder and trimmer capacitors into the local-oscillator circuit. Perfect tracking is obtained at three points of the range. Imperfect tracking at other points of the range does not interfere with normal reception.

Problems

9.1. Find the gain of the frequency changer of Problem 6.6, if the pentode operates as a mixer.

Answer: $K_c \cong 25$.

9.2. The pentode characteristic is described by the equation

$$i_a = I_0 + av_g + bv_g^2$$

Find the conversion conductance if $V_{go} = 5$ volts and $b = 0.1$ mA/V^2.

Answer: $g_c = 0.5$ milliampere per volt.

9.3. Determine the gain of a frequency changer loaded into an i.f. transformer at critical coupling, if $L = 8{,}000$ microhenrys; $Q = 50$; $\eta = 1$; and $f_i = 120$ kilohertz.

Answer: $K_c = 115$.

9.4. Determine the image selectivity of a superheterodyne receiver using a single-tuned coupling circuit for two cases as follows:

(a) $f_s = 1{,}500$ kilohertz; $f_i = 465$ kilohertz; $Q = 50$;
(b) $f_s = 12{,}500$ kilohertz; $f_i = 465$ kilohertz; $Q = 50$.

Use the current and voltage selectivity equations.

Answer:

(a) Voltage selectivity $d_{im} = 81$. Current selectivity $d_{im} = 50$ gives a wrong selectivity value.

(b) Voltage selectivity $d_{im} = 7.5$. Current selectivity $d_{im} = 7$. In the short-wave range, the current selectivity only slightly differs from the voltage selectivity.

9.5. Find the image response of a receiver whose r.f. section consists of an aerial-input circuit and one amplification stage, using the data of Problem 9.4.

Answer:

(a) $d_{im\,rf} = 4{,}050$;
(b) $d_{im\,rf} = 52$.

CHAPTER X

RADIO INTERFERENCE AND ITS SUPPRESSION

54. General

Radio interference is any type of disturbance acting on the receiver and causing unwanted responses in the terminal equipment, such as undesirable sounds or crackling in the loudspeaker, white or black spots on the screen of a cathode-ray tube, or false operation of a relay. More important still, interference may completely drown the wanted signal.

Interference may be divided into periodic and aperiodic.

Periodic interference has a definite carrier frequency. An example is the action of several signals on the same receiver. Periodic interference is chiefly suppressed by the selective devices of the receiver, such as described in the preceding chapters.

Aperiodic interference covers random disturbances, or those having no definite repetitive characteristic.

In this chapter we shall discuss aperiodic interference.

Any aperiodic interference is made up of impulses which bring about bursts of damped oscillation in the resonant circuits of the receiver.

Interference impulses may be of short duration and spaced widely apart in time, or they may come in an extremely rapid succession. The final effect, or response, will be different in each case.

In the former case, when the spacing between impulses is large, each burst of oscillation caused in the first tuned circuit by each pulse builds up a considerable voltage at the final stage. As a result, crackling sound or clicks following at the same rate as the interference will be heard on, say, the earphones. This form of interference is called *impulse noise*.

In the latter case, the burst of oscillation due to one impulse has no time to die out before another produces a burst of its own. Accordingly, it is called *continuous (or white) noise*.

The voltage at the final stage will be a sum of voltages with random phase relationships. Because of phase difference, the resultant voltage may be lower than in the case of impulse noise. In the terminal equipment (speaker or earphones), this form of interference appears as rustling, hissing, etc.

Aperiodic interference is classified in three types according to its source:

(1) atmospherics,
(2) man-made interference,
(3) receiver (internal) noise.

Atmospherics and man-made interference are mainly of the impulse type, while receiver noise is continuous.

Review Questions

1. What type of interference is receiver noise?
2. What noise is classed as aperiodic interference?

55. Atmospherics

Atmospherics, also referred to as *statics* or *strays*, are due to natural weather phenomena and electrical charges existing in the atmosphere.

The main source of statics is lightning. A lightning discharge, having a voltage of several million volts and a current of hundred thousand amperes, would build up an electromagnetic field propagating over hundreds of kilometres and producing radio interference. Statics may also be generated by all types of moving electrical charges in the atmosphere, such as cosmic rays, ultraviolet rays, nuclear radiation, etc.

Lightning discharges are caused by the movement of air masses. This movement are in turn determined by the intensity of sun light. The level of atmospherics also varies with the season of the year, the time of the day and the conditions of wave propagation. In summer, statics are stronger than in winter. They are also stronger at low than at high latitudes.

The intensity of statics varies from band to band. To clarify the matter we shall discuss the frequency spectrum of impulse noise.

Each noise impulse acting upon the receiver may be represented by an infinite series of elementary sinusoidal voltages whose frequencies are spread from zero to infinity. The amplitudes of

Fig. 10.1. Intensity of impulse noise as a function of frequency

these elementary voltages, as shown in Fig. 10.1, decrease as the frequency goes up. Atmospherics are stronger at low radio frequencies (long and medium waves), and are less pronounced at higher radio frequencies (short and ultra-short waves). The intensity of interference depends upon the bandwidth of the receiver; the wider the bandwidth, the stronger the interference.

The main objective of noise suppression is to secure a signal-to-noise ratio necessary for normal reception of the wanted signal.

This may be done in different ways. Wherever possible, short and ultra-short waves (on which the statics level is comparatively low) are used. Highly directional aerials also reduce the interference level to a considerable degree. Narrowing the receiver bandwidth is another method of decreasing the interference. In the latter case, it is good practice to introduce a device for varying the bandwidth of the receiver.

Impulse noise may be partly suppressed by a noise-limiting stage incorporated in the receiver.

Radio reception in aircraft is subject to a peculiar form of interference due to electrostatic charges accumulating on the metal parts of a flying airplane. These charges constantly change in magnitude, giving rise to circulating currents which generate noise in the receiver aerial. To decrease these currents, the metal parts of the plane should have reliable electrical contact with one another, or—as radio engineers say—reliable *bonding* should be provided.

Review Questions

1. How does an increase in the bandwidth affect the response to atmospherics?
2. In what frequency band is the receiver less sensitive to atmospherics?

56. Man-made Noise

Man-made noise is due to the electromagnetic fields generated by various electrical machines and devices. This type of interference is set up chiefly by abrupt current changes accompanied by an electrical spark or arc. The main sources of man-made interference are: electric power generators, telegraph sets, all types of switches, automobile and aircraft ignition systems, tram-cars, trolley-buses, and X-ray installations. It is evident that the level of man-made interference will be much higher in large cities where a great number of electrical machinery operates. In this respect, radio receivers in the countryside are in much better conditions.

Man-made interference finds its way into receivers mainly via the aerial or via the a. c. mains.

The interference picked up by the aerial can be minimized with the aid of a loop aerial. In inverted-L and T-aerials interference is reduced by use of screened and balanced downleads. Interference from a. c. mains is blocked by decoupling filters connected into the primary winding of the power transformer.

Figure 10.2 shows two forms of decoupling filters; one is a choke-capacitor filter, the other uses capacitors. In *(a)*, the filter passes the 50-hertz supply current and filters out the interference components of higher frequencies. In *(b)*, capacitors bypass the primary winding for the r. f.

However, it is difficult to suppress man-made interference at the receiver. Therefore, this interference should be controlled at the source.

For this purpose the ignition systems in internal-combustion engines (automobiles, aircraft, ships, etc.) are thoroughly screened. In electrical installations supplied with power from common mains it is necessary to insert special r. f. filters. Sparking commutators must be fitted with spark-quenching circuits. Electrical-installation codes envisage a number of measures for suppressing interference.

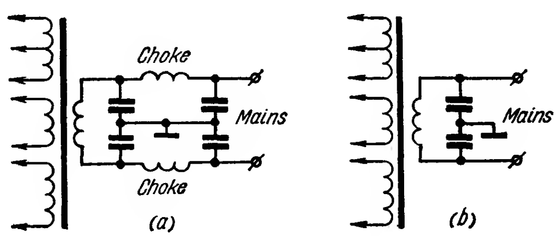


Fig. 10.2. Circuits preventing the penetration of man-made noise into a radio receiver via supply mains

Review Questions

1. How can man-made interference reach the receiver?
2. What devices can be used to prevent the entry of noise voltages from the supply mains?

57. Receiver (Internal) Noise

Even in the "no-signal" state, one can hear hissing or rustling sound in the terminal device of a sensitive receiver. This is *receiver*, or *internal*, *noise*.

Receiver noise is the continuous form of interference and is characterized by a voltage whose average value varies within narrow limits.

Internal noise is a major limiting factor for receiver sensitivity. In valve receivers, it is predominant in the VHF and UHF bands where external noise becomes insignificant because its spectral density falls off rapidly (see Fig. 10.1). In transistor receivers, internal noise reduces sensitivity already at the lower frequencies.

Receiver noise may originate in a number of ways and is divided into *thermal noise*, *valve noise*, or *transistor noise*, which are treated separately in the following sub-sections.

Thermal Noise. Noise of this type, also known as *Johnson* or *circuit noise*, is the result of the random motion of electrons which occurs in any circuit due solely to thermal agitation, that is, even in the absence of an applied emf. Over a period of time, these *random* or *fluctuation currents* (because the motion of electrons is a current) cause no net potential difference or voltage across the circuit. Instantaneously, however, they do, and the result is very small *fluctuation* or *noise voltages*. They are so small that ordinary voltmeters will not register them. Yet, when applied to a sensitive amplifier or receiver, they produce a response comparable with that due to the wanted signal.

Thermal noise voltages have components at all frequencies from zero to infinity. The magnitude of the noise voltage remains fixed over the entire frequency spectrum, but increases with the ambient temperature and the resistance of the circuit. The noise voltage in the output of the receiver is proportional to the bandwidth of the receiver.

The value of the noise voltage across an impedance $R+jX$ is given by:

$$V_n^2 = 4kTR\,\Delta f \tag{10.1}$$

where V_n = r. m. s. value of the noise voltage

k = Boltzmann's constant (1.38×10^{-23} joules/degrees K)
T = temperature of the circuit or component in °K
R = resistance of the circuit or component in ohms
Δf = frequency band of the noise voltage, which is close to the bandwidth $2\Delta F$ of radio receivers with a large number of tuned circuits, in hertz.

In practical calculations it is convenient to use the following equation:

$$V_n = 0.125 \sqrt{R\Delta f} \tag{10.2}$$

where V_n is in microvolts, R in kilohms, and Δf in kilohertz.

To determine the noise voltage across a tuned circuit, R_{0e} should be substituted for R in Eq. (10.2).

Example 10.1. Find the noise voltage across a resistor $R = 0.5$ megohms at $T \cong 300$ °K and $\Delta f = 10$ kilohertz.

Solution.

$$V_n = 0.125 \sqrt{R\Delta f} = 0.125 \sqrt{500 \times 10} = 8.9 \text{ microvolts}$$

It should be noted that the noise voltage at the output of a radio receiver is mainly due to the noise voltage in the input circuits because the noise voltage developed in those circuits is amplified by all the stages of the receiver. The noise voltage generated in the following stages has a much smaller effect.

Example 10.2. Find the noise voltage at the output of a five-stage amplifier. The gain of each stage is $K = 10$. The noise voltage at the input of each stage is $V_n = 5$ microvolts.

Solution. The noise voltage at the amplifier output due to the noise at its input is

$$V'_{n\ out} = K_1K_2K_3K_4K_5V_n = 10^5 \times 5 \times 10^{-6} = 0.5 \text{ volt}$$

The noise voltage at the amplifier output due to the noise at the second-stage input is

$$V''_{n\ out} = K_1K_2K_3K_4V_n = 10^4 \times 5 \times 10^{-6} = 0.05 \text{ volt}$$

Since the noise voltage in the subsequent stages is very small, it may be disregarded. Then the overall noise voltage at the amplifier output will be

$$V_{n\ out} \cong V'_{n\ out} + V''_{n\ out} = 0.5 + 0.05 = 0.55 \text{ volt}$$

Valve Noise. Electrons are emitted by a hot cathode at random, so that there is a random fluctuation of current between the cathode and anode (anode current). This random fluctuation contains components at all frequencies. The resultant noise, known as *shot noise*, is thus similar in its effects to thermal noise. Fluctuations in the anode current produce a fluctuation voltage, which is the noise voltage appearing across the anode load.

The magnitude of noise voltage introduced by a valve is usually specified by quoting a resistance which, when placed at the input to the same valve, now assumed to be noiseless, will yield the same noise voltage at the output. This resistance is known as the *equivalent noise resistance*, R_n.

For valves with control grids it is found convenient to place this fictitious R_n in the grid circuit (as the input to the valve). The noise voltage in the grid circuit may be found from Eq. (10.1)

$$V_n^2 = 4kTR_n\Delta f \tag{10.3}$$

Theoretical research, confirmed by experiments, shows that the equivalent noise resistance of a triode is given by

$$R_n = \frac{2.5 \text{ to } 3}{g_m} \tag{10.4}$$

If g_m is expressed in milliamperes per volt, the noise resistance R_n will be in kilohms.

Example 10.3. Find the equivalent noise resistance of the 6C1П triode.

Solution. From a valve manual we find that g_m is 2.2 milliamperes per volt. The equivalent noise resistance is given by Eq. (10.4)

$$R_n = \frac{2.5}{g_m} = \frac{2.5}{2.2} = 1.14 \text{ kilohms}$$

There is a further source of noise in multi-grid valves in which the anode current is shared or *partitioned* by the grids positive to the cathode. This sharing or *partition* is random, so that a fluctuation akin to the shot effect is caused. This is *partition noise*. Thus, multi-grid valves are noisier than tetrodes. The noise level in a pentode is from 3 to 5 times higher than that in a triode.

The noise resistance is extremely high in multi-grid frequency-changer valves, because of which they are not used in the VHF and UHF bands.

The noise resistance of some Soviet valves is given in Table 10.1.

TABLE 18.1

Valve type	Particulars	Noise resistance, R_n, ohms
6Ж1Б	Miniature pentode	2,180
Same	Same, triode-connected	180
6Н14П	Bantam dual triode (US equivalent, 6J6)	700
6Н15П		470
6Н16П		140-170
6Ж8С	Screen-grid, sharp cut-off glass pentode (US equivalent, 6F8)	960
6С2С	Glass triode (US equivalent, 6J5-GT)	960
6С1П	Bantam triode (US equivalent, 6C4)	1,140
6С2П	Bantam triode	680
6С3П	Bantam triode	150
6Н9С	Glass dual triode (US equivalent, 6SL7)	1,560
6К4	Screen-grid, remote cut-off r. f. pentode	4,000
6Ж9П	Screen-grid, sharp cut-off bantam pentode	530
6К9С	Screen-grid, remote cut-off glass pentode (US equivalent, 6U7G)	11,000
6К1П	Screen-grid, remote cut-off, UHF bantam pentode	13,000
6А7	Frequency-changer valve for double-grid injection (US equivalent, 6SA7)	200×10^3
6Ж1П	Screen-grid, sharp cut-off bantam r. f. pentode (US equivalent, 6AK5)	2,130
6Ж2П	Screen-grid, sharp cut-off bantam r. f. pentode	4,340
6Ж3П	Screen-grid, sharp cut-off r. f. bantam pentode (US equivalent, 6AG5)	1,370
6Ж5П	Screen-grid, sharp cut-off r. f. bantam pentode	640
6Ж9П	Screen-grid, sharp cut-off r. f. bantam pentode	390
6К4П	Screen-grid, remote cut-off bantam pentode (US equivalent, BA6)	3,500
6А2П	Frequency-changer valve for double-grid injection (US equivalent, 6BE6)	240,000

Transistor Noise. Transistor noise consists of $1/f$ noise, shot noise, thermal noise, and partition noise.

The $1/f$ noise components arise from the properties of the semiconductor lattice. They are chiefly confined to the collector region and account for a noise voltage of 50 to 100 microvolts. Their designation "$1/f$" reflects the fact that their power is inversely proportional to frequency. In practical junction transistors, $1/f$ noise becomes negligible at very low frequencies. The $1/f$ noise components can be kept to a minimum by making the direct collector current as low as practicable.

Shot noise in transistors is due to the fact that the transport of carriers across the collector and emitter junctions is a random process. As measured at room temperature, the shot-noise voltage across the collector junction is 50 microvolts and across the emitter junction, about one microvolt. Shot noise contains components of a uniform frequency spectrum.

Thermal noise is due to thermal agitation in the base bulk resistance on heating. As measured at room temperature, the thermal-noise voltage is about one microvolt. It contains components of a uniform frequency spectrum, too.

Partition noise is due to the random sharing or partition of current between the base and collector. Since the partition-noise voltage of a transistor is proportional to its alpha current gain, there is a slight increase in the partition-noise voltage with frequency.

Thus, transistor noise consists of components which are non-uniformly distributed over the frequency spectrum. It is more pronounced at low frequencies up to 30 kilohertz and at high frequencies over 100-500 kilohertz (for alloy transistors).

The level of noise in transistors varies with the source resistance. The lowest noise in a transistor stage will be obtained with a source resistance, R_g, equal to anywhere from 500 to 1,000 ohms, which corresponds to perfect impedance matching between the source and the input resistance of the transistor operated common-emitter.

It may be noted that the circuit configuration does not affect the noise properties of a transistor stage.

Noise Factor. The total noise voltage $V_{\Sigma n}$ of a circuit is given by

$$V_{\Sigma n} = \sqrt{V_{n1}^2 + V_{n2}^2 + V_{n3}^2 + \ldots} \tag{10.5}$$

where V_{n1}, V_{n2}, etc., are the noise voltages due to each circuit component.

However, the total noise voltage is not the only criterion of the noise properties of a radio receiver.

Another criterion is the *noise factor* (sometimes called the *noise figure*), N.

The noise factor (Fig. 10.3) can be stated as the ratio of the signal-to-noise ratio at the input to that at the output of the

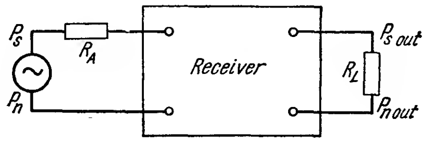


Fig. 10.3. Concerning the noise factor of a receiver

receiver

$$N = \frac{P_s/P_n}{P_{s\,out}/P_{n\,out}} \tag{10.6}$$

where P_s and $P_{s\,out}$ are, respectively, the signal levels at the input and the output, while P_n and $P_{n\,out}$ are the noise levels at the input and output.

An ideal noiseless receiver introduces no extra noise and has the same signal-to-noise ratio at the output as at the input. Its noise factor is therefore unity. A practical receiver adds some noise to the signal, the signal-to-noise ratio at the output deteriorates, and the noise factor becomes less than unity.

Thus, the noise factor specifies the extent to which a practical receiver causes deterioration of the signal-to-noise ratio in comparison with an ideal receiver.

The noise factor increases with frequency. In the frequency range from 200 to 10,000 megahertz, it varies from unity to a few tens. At frequencies exceeding 10,000 megahertz, $N > 50$. Usually, the factor N is expressed in decibels as

$$N_{db} = 10 \log_{10} N$$

For instance, the noise factor $N = 50$ in decibels is $N_{db} = 10 \log_{10} 50 = 17$ db.

The noise factor can also specify the noise quality of individual stages and amplifiers.

From the foregoing we may conclude that the noise level in highly sensitive receivers can mainly be reduced by the proper choice of the bandwidth and the use of valves with a minimum noise resistance in the early stages.

The methods of noise suppression discussed in the present chapter do not solve all the problems concerned with interference-free reception. Such reception can be best obtained by replacing amplitude modulation with frequency or some other type of modulation.

Review Questions

1. What type of interference limits the sensitivity of the receiver in the VHF-UHF band?
2. In what receivers does internal noise appear in all frequency ranges?
3. What circuit parameter controls thermal noise in a tuned circuit?
4. Which of the valve parameters indirectly specifies the noise quantity of the valve?
5. Which stages in a receiver control its noise quality?
6. Why is a practical receiver noisier than an ideal one?

SUMMARY

1. Interference may be periodic and aperiodic. Aperiodic interference is of two kinds: impulse noise and white (continuous) noise. Atmospherics and man-made noise are mainly of the former type, while internal (receiver) noise is white noise.
2. Atmospherics and man-made noise are most pronounced on long, medium and, partly, short waves. At very- and ultra-high frequencies internal (receiver) noise predominates.
3. Radio reception is possible only when the signal voltage is some specified level above the noise voltage.
4. Interference may be minimized by use of decoupling filters at the receiver input and by suppressing interference at the source.
5. The noise quality of a radio receiver may be specified by quoting the resultant noise voltage due to "noisy" components of the circuit, and by the noise factor.
6. The level of internal (receiver) noise may be reduced by the proper choice of the bandwidth and valves with a minimum noise resistance for the early stages.

CHAPTER XI

RECEPTION OF FREQUENCY-MODULATED SIGNALS

58. General

With advances in the art of radio reception, the requirements for receiver sensitivity, bandwidth and selectivity grow more and more stringent. However, as we already know, an increase in sensitivity and bandwidth is inevitably accompanied by the deterioration of the signal-to-noise ratio, and the actual sensitivity of the receiver decreases.

A way out is offered by frequency modulation (FM). The outstanding feature of FM is that it permits reception which is free from interference and noise. With FM, the signal-to-noise ratio at the receiver output is more than a hundred times higher than at the output of an amplitude-modulation (AM) receiver. In brief, frequency modulation consists in that an r. f. carrier is

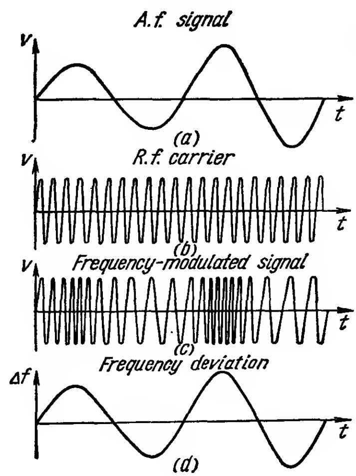


Fig. 11.1. Frequency modulation

shifted, or *deviated*, to a higher or lower frequency at a rate decided by the frequency of the audio (modulating) signal, while the amplitude of the carrier remains constant.

Figure 11.1 is a representation of frequency modulation.

When a modulating signal (waveform *a*) is applied, the carrier frequency (waveform *b*) is increased during one-half cycle of the modulating signal and decreased during the half-cycle of opposite polarity. This is indicated in the plot (waveform *c*) by the fact that the r. f. cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative.

The change in the carrier frequency (waveform *d*) is called the *frequency deviation*, Δf. As is seen, the frequency deviation is controlled by the amplitude of the modulating signal. As shown in the plots of Fig. 11.1, the amplitude of the modulated wave does not change during modulation.

A frequency-modulated wave, modulated by a single tone of frequency F, may be described as

$$v = V_m \sin(\omega_0 t + M_f \sin \Omega t) \tag{11.1}$$

where: ω_0 = angular frequency of the r. f. carrier

$\Omega = 2\pi F$ = angular frequency of the audio (modulating) signal

$M_f = \frac{\Delta f_{max}}{F}$ = modulation index defined as the ratio of the maximum frequency deviation to the modulating frequency.

The width of an FM channel depends on the value of the modulation index M_f. When $M_f \ll 1$, the FM channel contains frequency ω_0 and two side frequencies $\omega_0 \pm \Omega$. As M_f increases, more side-bands are added to the FM channel, and it widens. For practical cases, it may be considered that when $M_f > 1$ the FM channel is mainly determined by the maximum frequency deviation Δf_{max} and is equal to $2\Delta f_{max}$.

In broadcasting, the maximum frequency deviation Δf_{max}, which corresponds to the peak amplitude of the modulating signal, is 75 kilohertz. This means that a radio station occupies a 150-kHz channel. Since each station is allocated a channel of 250 kilohertz, frequency modulation may be used on ultra-short waves (the VHF band) only.

FM receivers are immune to interference mainly because the signal amplitude remains unchanged by modulation. Let us compare the signal-to-noise ratio at the input of FM and AM recei-

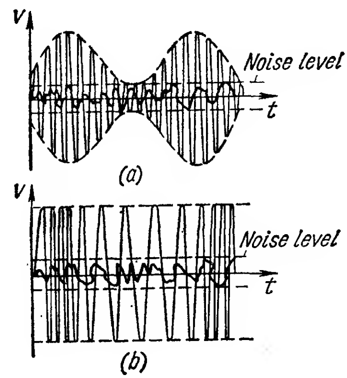


Fig. 11.2. Relation between signal and noise in AM and FM

vers. Assume that the amplitude of the FM signal equals the peak amplitude of the AM signal (Fig. 11.2), and that the interference level is the same in both cases.

As is seen from Fig. 11.2*a*, in AM the signal-to-noise ratio is constantly changing. At large amplitudes, the signal is considerably greater than the noise so that the latter affects reception but slightly. Conversely, at low amplitudes, the signal and noise may be equal in level, and the noise can make reception impossible. Consequently, for AM reception, the minimum amplitude of the signal must be several times the interference level.

The situation is quite different in FM reception. As is seen in Fig. 11.2*b*, the signal-to-noise ratio remains constant and equal to the value obtained at the peak of amplitude modulation.

A good deal of their freedom from noise in FM receivers comes from refinements in their circuitry.

The point is that external and internal noise causes amplitude variations in the signal in sympathy with the interference signal, which is, in effect, unwanted amplitude modulation.

In receivers, interfering noise is kept to a minimum by a process called *limiting* and by circuits called *limiters*. Figure 11.3 shows how this is done in AM and FM receivers.

In AM receivers, however, partial elimination of variations in the amplitude of the signal caused by interference upsets the modulation, and, ultimately, leads to non-linear distortion in the audio signal.

In contrast, the amplitude limiters used in FM receivers eliminate all amplitude variations in the signal without distorting

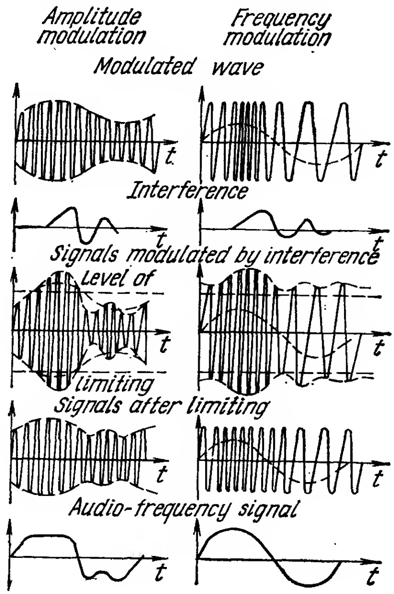


Fig. 11.3. Limiting in AM and FM

the intelligence contained in the modulation. Thus, limiting is an effective method of removing amplitude variations from FM signals and gives FM receivers their well-known freedom from noise.

FM is widely used in radio broadcasting, television and microwave radio relay systems.

The difference between AM and FM makes an FM receiver somewhat different from its AM counterpart. For one thing, FM receivers normally operate in the VHF band. For another, the r. f. section of the receiver (before the detector) should have a much broader bandwidth.

Modern FM receivers are chiefly superheterodyne receivers owing to the high sensitivity and good selectivity they offer in the VHF band.

Functional Units of an FM Superheterodyne Receiver. As compared with AM receivers, an FM receiver has an additional limiter stage and employs a *frequency discriminator* instead of an amplitude detector (Fig. 11.4). The purpose of the limiter is to eliminate all amplitude variations in the signal. The limiter may be a separate stage, or its function may be performed by the last i. f. stage.

The frequency discriminator converts the constant-amplitude, frequency-modulated signal into an audio signal whose frequency is the frequency at which the deviation of the FM occurs and whose amplitude is proportional to the magnitude of this frequency deviation.

The i. f. amplifier of an FM receiver, in contrast to that of an AM receiver, must amplify signals over a comparatively broad bandwidth (150-200 kilohertz). Therefore, an FM i. f. amplifier must have more stages than the usual narrow-band i. f. amplifier.

Let us establish the relation between the gain and bandwidth of a single-stage FM amplifier.

As will be recalled,

$$K_0 = g_m R_{0e}$$

Noting that the resonant resistance of a tuned circuit is

$$R_{0e} = \rho Q = \frac{Q}{2\pi f_0 C}$$

and the bandwidth is

$$2\Delta F = \Delta f_{3\mathrm{db}} = \frac{f_0}{Q}$$

we obtain

$$K_0 = \frac{g_m Q}{2\pi C f_0} = \frac{g_m}{2\pi C \Delta f_{3\mathrm{db}}} \tag{11.2}$$

Thus, the broader the bandwidth, the lower the gain of a single stage is.

As a rule, the i. f. amplifier of an FM receiver has at least three amplifier stages.

The intermediate frequency in such receivers is from several megahertz to tens of megahertz. In order to secure high fidelity of sound reproduction, the bandwidth of the a. f. section is usually broadened to 15 kilohertz.

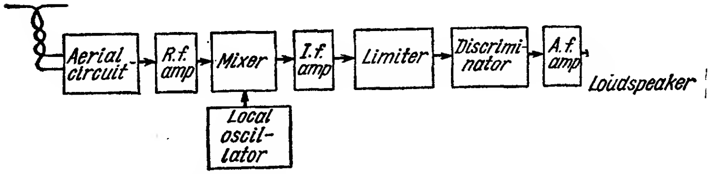


Fig. 11.4. Block diagram of an FM superheterodyne receiver

Review Questions

1. What factor controls the magnitude of the frequency deviation?
2. Why does limiting distort the modulation in AM signals?
3. How does limiting affect the signal-to-noise ratio?
4. Why is FM used mainly in the VHF band?
5. How does the bandwidth affect the stage gain of an FM receiver?

59. The Limiter

There are two main causes for amplitude variations in the FM signal. The first one is that the r. f. and i. f. stages do not have a frequency response which is perfectly flat across the top and has sharp cut-offs. Frequencies close to the carrier are amplified more than those removed from it.

The other cause is external and internal interference.

The limiter may be a separate stage specifically designed for that purpose. In that case, it bears a close resemblance to a diode detector and may have any one of many configurations.

More frequently, the last i. f. stage serves as the limiter. The circuit of such a stage is shown in Fig. 11.5. This circuit differs from that of the usual i. f. stage only in that R_g and C_g are incorporated in the grid circuit to provide for the limiting action. Figure 11.6 illustrates the operating principle of a *clipper-limiter*, or *slicer*, also known as the *grid-leak limiter*.

When the signal voltage is applied to the grid of the valve, a current appears in the grid circuit. This current builds up a negative bias voltage across R_g thereby reducing the stage gain. The magnitude of the grid bias is determined by the magnitude

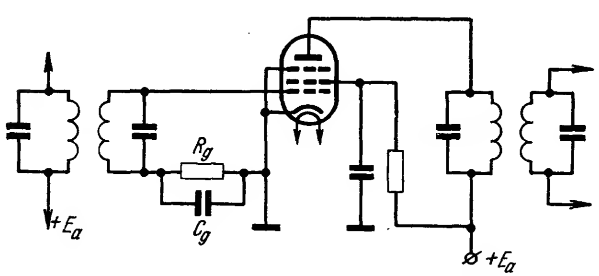


Fig. 11.5. I.f. amplifier stage operating as a limiter

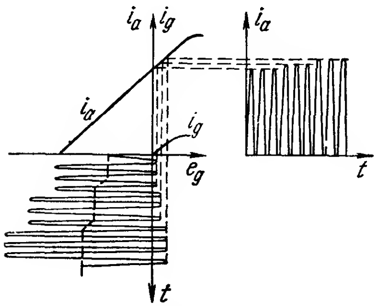


Fig. 11.6. Operating principle of a clipper-limiter

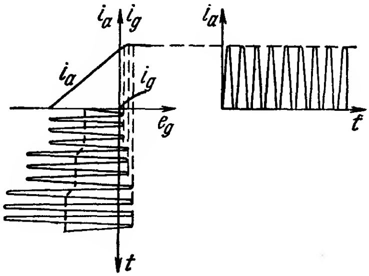


Fig. 11.7. Clipping-limiting with reduced voltages at anode and screen grid of the limiter valve

of the positive signal peak from the i. f. amplifier. The greater the amplitude of the input signal, the greater is the negative grid bias.

Owing to this, the positive peaks of the pulsating anode current are said to be "clipped at the grid", or "flattened by grid current". The negative peaks of the signal voltage are clipped or flattened by anode current cut-off (or simply, by "cut-off"). The result is that a horizontal slice of the signal is taken.

This method of limiting is effective only when high-amplitude signals are applied to the grid. To limit low-amplitude signals, the voltage at the anode and screen grid of the limiter stage is decreased to 30 to 50 volts. In this case, the valve characteristic is shifted to the right, and its upper bend in the saturation region is shifted more to the left (Fig. 11.7) than in the case shown in Fig. 11.6.

The operation of the limiter under such conditions is more efficient. The positive peaks are flattened by both grid current and the anode bend.

Two limiter stages are used where a stronger limiting action is essential.

Review Questions

1. What current controls the grid bias in a grid-leak limiter?
2. Why are the anode and screen-grid supply voltages brought down in a grid-leak limiter?

60. The Frequency Detector (Discriminator)

Frequency detection is the process of removing the intelligence (the a.f. signal) from an FM wave. This is done by a circuit called a *frequency discriminator*. The FM signal reaching the discriminator is converted into an amplitude-modulated signal, which is then detected by an ordinary amplitude detector.

The simplest method of converting an FM signal into an AM signal is based on *staggered tuned circuits* (Fig. 11.8).

Assume that the resonant frequency of the final i.f. amplifier differs from the carrier frequency of the incoming signal by a value exceeding the maximum frequency deviation Δf_{max} (waveform *a*).

Then, as the signal frequency varies with respect to the carrier frequency by $\pm\Delta f$ (waveform *b*), the voltage across the tuned circuit will change by $\pm\Delta V_m$ (waveform *c*).

Waveform *d* shows the amplitude-modulated r.f. signal obtained by the conversion process. Applying this wave to the usual amplitude detector, we obtain an a.f. signal.

Practical discriminators mostly use *centre-tuned circuits* (the *Foster-Seeley discriminator*) or *ratio detectors*.

The Foster-Seeley Discriminator. The circuit of this discriminator is shown in Fig. 11.9.

As is seen, the primary of the double-tuned transformer connected as the anode load of the limiter is coupled to a centre-tapped secondary, and both are tuned to resonate at the centre frequency of the i.f. amplifier. The signals developed by the two halves of the secondary are rectified by a push-pull detector ordinarily using a double diode. The audio-frequency output is

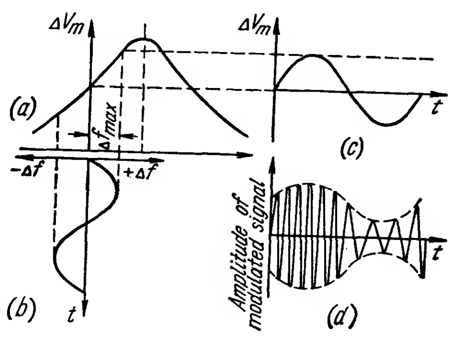


Fig. 11.8. FM detection by stagger-tuned circuits

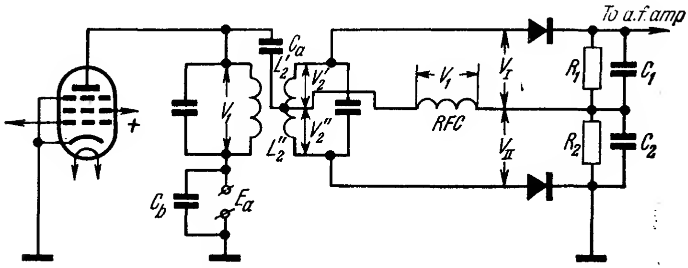


Fig. 11.9. Forster-Seeley discriminator

equal to the difference between the voltages, V_{I} and V_{II}, built up across R_1 and R_2, the diode loads.

As is seen, the voltage across each arm is made up of the voltage V_1 across *RFC* and half of the voltage across the secondary (V_2' or V_2'')

$$\left.\begin{aligned}\dot{V}_{\text{I}} &= \dot{V}_1 + \dot{V}_2' \\ \dot{V}_{\text{II}} &= \dot{V}_1 + \dot{V}_2''\end{aligned}\right\} \tag{11.3}$$

The voltage V_1 across *RFC* is equal to the primary voltage because the left-hand end of the choke is connected for the r. f. through a coupling capacitor C_a to the top terminal of the transformer primary, and the right-hand end of the choke is connected to the lower terminal of the transformer primary through C_2, common return, and a bypass capacitor C_b.

The input voltages V_{I} and V_{II} are vectors and, in order to determine them, it is necessary to establish phase relationships between V_1 and V_2', and V_1 and V_2''.

Consider the equivalent circuit of the transformer secondary and the vector diagram of voltages and currents at resonance (Fig. 11.10). E_M is the emf of mutual induction determined as the no-load voltage across L_2. As will be recalled, an emf of mutual induction is opposite in phase to V_1 across the primary.

The secondary current I_2 is in phase with E_M and builds up across L_2' and L_2'' the voltages V_2' and V_2'', which are 90° out of phase with I_2. V_2' and V_2'' are equal and opposite with respect to the centre tap. As a consequence, V_1 is in quadrature with V_2' and V_2''. When the signal frequency differs from the centre frequency the transformer will not be at resonance, and the phase shift between the vector voltages will be other than 90°.

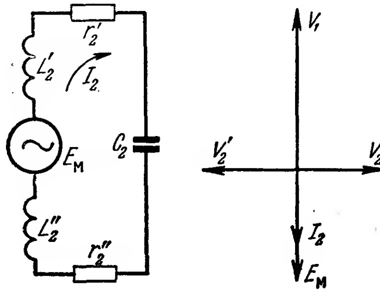


Fig. 11.10. Equivalent circuit of the transformer secondary and vector diagram of voltages and currents at resonance

Vector diagrams of voltages acting in each arm of the push-pull detector are shown in Fig. 11.11. Vector diagram *a* holds for an unmodulated wave, when the frequency deviation $\Delta f = 0$, and, consequently, when the tuned circuits are at resonance. In this case, the total voltages V_I and V_{II} are equal and build up equal voltages across R_1 and R_2. The output voltage is equal to the difference between these voltages and is zero.

Vector diagram *b* holds when the frequency deviation is positive, $+\Delta f$, and diagram *c* when the frequency deviation is negative, $-\Delta f$. In both cases the transformer is off resonance and the phase shift between V_1 and V_2 is greater or smaller than 90°. This upsets the equality between V_I and V_{II} and between the output voltages across the diode loads, so that an audio-frequency voltage appears at the output. The amplitude of this voltage is proportional to the maximum frequency deviation of the input signal.

The Ratio Detector. Present-day FM receivers widely use a discriminator circuit known as the *ratio detector*. The main advantage of the ratio detector is that it does not respond to any amplitude variations in the i.f. signal. As a result, it permits the elimination of a limiter stage in the FM receiver.

As compared with the Foster-Seeley discriminator, the ratio detector has its diodes connected in opposition, and the detector

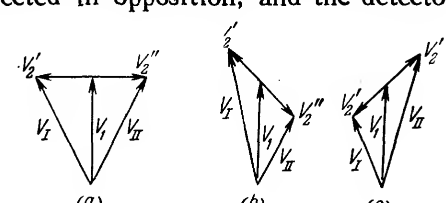


Fig. 11.11. Vector diagrams illustrating detection of FM signals

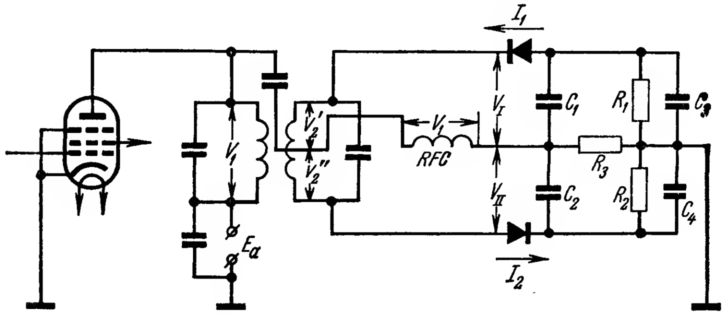


Fig. 11.12. Ratio detector

output voltage is developed across a load resistor common to both diodes (Fig. 11.12).

The arm voltages V_I and V_{II} in the ratio detector are similar to those in the Foster-Seeley circuit, and the phase-shift diagrams for that discriminator also hold for the ratio detector.

When the i. f. signal is at its centre frequency (no frequency modulation), the transformer secondary is at resonance, the arm voltages V_I and V_{II} are equal (Fig. 11.11*a*), and the currents through the diodes are $I_1 = I_2 = I$. Since, however, the diodes are connected in opposition, these currents cancel out in the load resistor R_3, the voltage across R_3 is zero, and no signal is applied to the audio amplifier.

When the i. f. signal is above or below the centre frequency (frequency modulation), the arm voltages are unequal (Fig. 11.11*b* and *c*), and different currents are flowing through the diodes. In one diode, the current goes up by ΔI_1, while in the other it goes down by $\Delta I_2 = -\Delta I_1$. Thus, the diode currents are

$$I_1 = I + \Delta I_1$$
$$I_2 = I - \Delta I_2$$

where I is the diode current when the i. f. signal is at its centre frequency.

The current through R_3 is the difference between the diode currents

$$I_{af} = I_1 - I_2 = I + \Delta I_1 - I + \Delta I_2 = \Delta I_1 + \Delta I_2 = 2\Delta I$$

As is seen, it is decided by the sum of the increments ΔI_1 and ΔI_2 proportional to the amplitude of the a. f. signal.

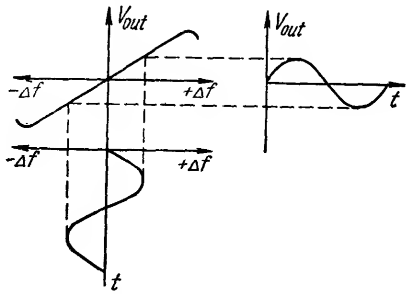


Fig. 11.13. Frequency response of frequency discriminator

The a. f. voltage appearing across R_3 is applied to an audio amplifier. C_1 and C_2 provide bypasses for the r. f. around R_3.

Consider the limiting action of the ratio detector. R_1 and R_2 are diode loads in the respective arms and are bypassed for the a. f. by C_3 and C_4.

If C_3 and C_4 are chosen sufficiently large, the voltages across R_1 and R_2 will remain practically constant with time. Under the circumstances, the operating angle θ of the linear detector and, as a consequence, the voltage transfer factor will be functions of the incoming signal amplitude. For low-amplitude signals K_d will increase; for high-amplitude signals the operating angle θ will increase and K_d will decrease. The result will be a marked reduction in the effect of parasitic amplitude modulation on the output signal.

The operation of the discriminator is additionally explained by the characteristic shown in Fig. 11.13. The output voltage equal to the difference between the voltages across R_1 and R_2 is laid off as ordinate, and the frequency deviation as abscissa To minimize non-linear distortion in FM detection the linear part of the characteristic should exceed the double amplitude of the frequency deviation $2\Delta f_{max}$.

Review Questions

1. What circuit element may be used instead of the r. f. choke in a frequency discriminator?
2. What is the phase angle between I_2 and V_2' and V_2'' in the vector diagram of Fig. 11.11?
3. What type of distortion will occur in a frequency discriminator if the operating region exceeds the linear portion of the frequency response?

4. Which characteristic of a frequency discriminator will be affected by an increase in the gradient of its frequency response?

SUMMARY

1. The FM wave is an r. f. wave whose amplitude is constant but whose frequency changes in accordance with the audio-frequency modulating signal.
2. The main advantage of FM receivers is their freedom from noise and interference.
3. FM receivers are intended for operation in the VHF band and their r. f. circuits have a broad bandwidth.
4. FM receivers are mainly superheterodyne sets with a limiter and discriminator added to their circuitry.
5. The limiter eliminates amplitude variations in the incoming signal caused by interference and other factors. The limiter is usually the final stage of the i. f. amplifier.
6. The discriminator converts frequency-modulated signals into amplitude-modulated signals which are then detected.

CHAPTER XII

RECEIVER CONTROLS

61. General

One of the conditions for good reception is the constancy of the voltage or power fed to the terminal component (loud-speaker, relay, etc.).

Variations in the receiver output voltage may be caused by a number of factors. Some of them are external to the receiver. Other factors originate within the receiver.

External factors affect the input voltage. The voltage reaching the receiver input, other conditions being equal, depends on the power of the transmitting station, the distance from the transmitter, and the conditions of radio wave propagation. The higher the power of the transmitter and the closer it is located to the receiver, the higher the field intensity will be at the point of reception and the greater the voltage reaching the input of the receiver.

The effect of the distance from the transmitter is particularly noticeable in mobile radio stations. As an aircraft flies away from its base, the field intensity decreases. The effect of the distance is very noticeable in the VHF band because the field intensity is inversely proportional to the square of the distance.

In the HF and VHF bands the field intensity depends greatly on the conditions of radio wave propagation. Fading, frequently observed in the HF band, can change the field intensity at the reception point within broad limits. In a receiver designed for reception of weak signals, an increase of hundreds or even thousands of times in the input voltage might overload the radio-frequency stages, detector, and particularly, audio-frequency stages, giving rise to considerable non-linear distortion.

The internal factors causing variations in the output voltage of a radio receiver are as follows:

(1) variations in the voltage gain of the aerial-input circuit;

(2) variations in the gain of the r.f. amplifier with frequency and between bands;

(3) imperfect tracking (misalignment) of the local oscillator circuit and the circuit tuned to the signal frequency;

(4) local-oscillator frequency drift in operation;

(5) unstable supply voltages.

Variations in the local-oscillator frequency change the intermediate frequency. If the intermediate frequency were shifted outside the bandwidth, the gain of the i.f. stages would sharply decrease and the distortion would greatly increase.

In modern receivers, the local oscillator is gang-tuned along with the aerial-input circuit and r.f. stages. Therefore, alignment of the local oscillator to secure the requisite value of the intermediate frequency might detune the circuits tuned to the signal frequency. This would inevitably decrease the voltage gain and the stage gain of the r.f. amplifier.

To eliminate or minimize these factors, the receiver must incorporate circuits to control the gain by varying receiver sensitivity within certain limits.

In some cases, in addition to controlling receiver sensitivity, it is necessary to vary the bandwidth of the receiver.

For reception of telephone signals, the bandwidth must be at least 6 to 8 kilohertz. In telegraphy, it is sufficient to have a bandwidth of two or three kilohertz. If it were made broader, noise due to external sources would increase and the stability of the channel would be lowered, particularly when receiving weak signals. Therefore, when changing over from telephony to telegraphy, the bandwidth of the receiver should be made narrower.

In broadcast receivers, it is frequently necessary to vary the bandwidth at the higher or lower frequencies of the audio signal. This control makes it possible to change the tone of sound reproduced by the loudspeaker.

Thus, for normal operation of the receiver, it must incorporate the following controls:

(1) receiver sensitivity control;

(2) bandwidth control;

(3) local-oscillator frequency control.

These types of control may be either manual or automatic, or both.

Automatic control is usually used in the early stages of the receiver, while manual control is resorted to in the succeeding stages. Automatic control should not follow manual control, because the effect of manual control in the preceding stages will be minimized by automatic control in the succeeding stages of the receiver.

Review Questions

1. What is the purpose of controllable gain in receivers?
2. Why should stages with automatic gain control precede those with manual control?
3. When may manual control prove insufficient?

62. Manual Gain Control

Manual gain control is employed in the radio-frequency circuits (before the detector), and in the audio-frequency circuits (after the detector). Gain control before the detector helps to avoid overloading the detector and r.f. stages by strong signals.

Manual gain control in the r.f. stage is performed by varying the grid bias voltage (Fig. 12.1). To effect gain control within wide limits, the stage should employ a variable-mu valve. The current characteristic of a variable-mu valve has a long tail (the portion within the negative values of grid voltage), so that cut-off occurs later, or is said to be *remote*. By varying the bias voltage it is possible to vary the mutual conductance at the operating point of the valve within wide limits and, consequently, the stage gain

$$K = g_m Z_a$$

Figure 12.2 shows the anode-grid characteristic of the 6K1П screen-grid remote cut-off pentode (US equivalent, 6K7). As the bias voltage is varied from —2 to —20 volts the mutual conductance is decreased from 1.55 to 0.05 milliampere per volt, i.e. to 1/31th of its original value.

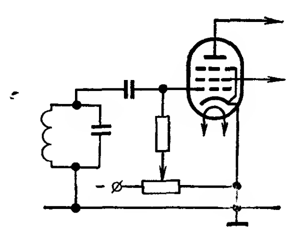


Fig. 12.1. Receiver gain control by varying grid bias

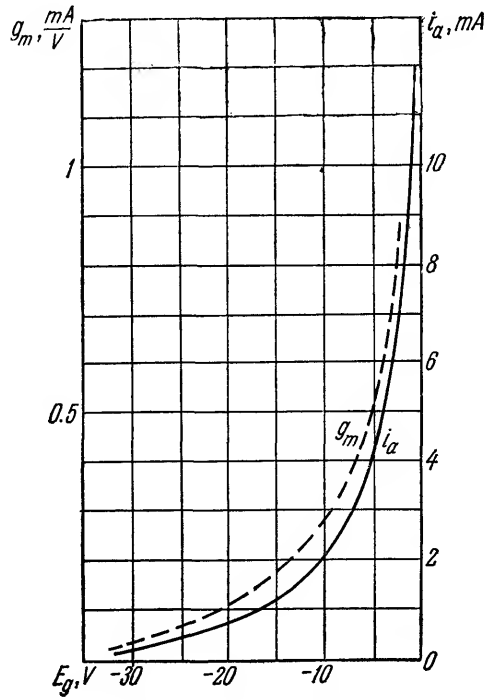


Fig. 12.2. Dependence of anode current and mutual conductance of 6К1П valve on grid voltage

Manual gain control in the audio-frequency section is usually performed at the detector output, i.e. at the input to the first stage of the audio-frequency amplifier (Fig. 12.3). This protects the first i.f. stage and all the following ones from overloading by strong signals.

It should be noted that ordinary volume control attenuates all audio frequency equally, and the fidelity of sound reproduction is considerably impaired. This is explained by the fact that the sensitivity of the human ear to sound waves at different frequencies rises with the loudness of sound. When the loudness is decreased, the human ear becomes insensitive to the higher and, particularly, the lower audio frequencies. Sound reproduction will be natural only when the loudness is the same at the microphone and at the speaker.

To improve the fidelity of reproduction at all levels of loudness, radio receivers employ so-called *compensated volume control*

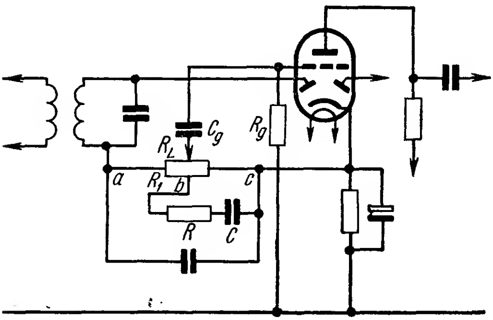


Fig. 12.3. Compensated volume control with a single compensation network

(also called *loudness control*) which simultaneously adjusts the loudness and the frequency response of the audio-frequency amplifier. A simple circuit for compensated volume control with one compensation network is shown in Fig. 12.3. When the potentiometer is set at point *a*, the compensating network has practically no effect on the frequency response, because R_1 has a considerably higher value than the resistance between points *b* and *c*. Therefore, at this setting of the potentiometer, i.e. at the higher sound level, the frequency response is flat. As the potentiometer moves from point *a* to point *b*, decreasing the sound level, the resistance between points *b* and *c* has a growing effect on the frequency response. At point *b*, the voltage taken from the potentiometer will be directly proportional to the resistance between *b* and *c*. Owing to the compensating *RC*-network the resistance between *bc* increases as the frequency decreases, and the frequency response goes up in the low-frequency region. Further adjustment

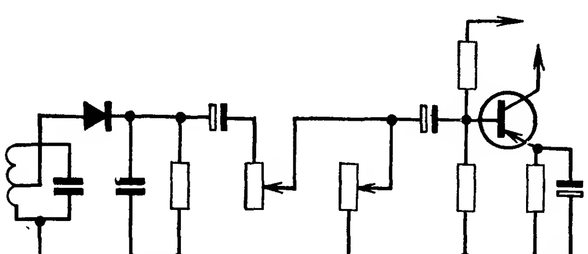


Fig. 12.4. Manual gain control at the detector output in a transistor receiver

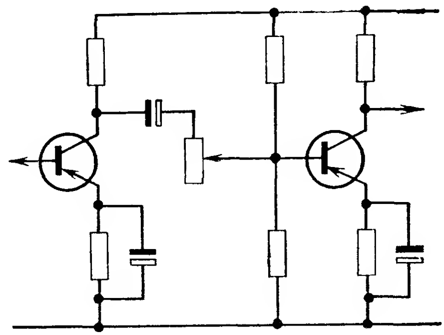


Fig. 12.5. Manual gain control between the stages of the a.f. amplifier

of the potentiometer from *b* to *c* does not change the frequency response.

In transistor receivers, manual gain control is provided either at the detector output or between the stages of the a.f. amplifier. While in valve receivers manual gain control consists in bringing down the a.f. voltage applied to the grid of the a.f. amplifier valve, in transistor receivers this is done by varying the current in the input circuit of the a.f. amplifier. The gain control circuit should be arranged so as not to affect the values of the d.c. supply voltages.

Figure 12.4 shows a circuit for manual gain control at the detector output.

Figure 12.5 shows a simple circuit for manual gain control between the stages of the a.f. amplifier.

The methods of gain control discussed above are effective only when the operator has time to follow changes in receiver operation. If the incoming signal sharply changes in amplitude due to fading or when tuning from one station to another, the operator is frequently unable to provide adequate compensation. This is where automatic gain control comes in.

Review Questions

1. List methods of manual gain control in the r.f. amplifier.
2. Can manual gain control use the same methods in the a.f. amplifier as in the r.f. amplifier?
3. List methods of manual gain control in transistor receivers.
4. What is the purpose of compensating networks in loudness control?

63. Automatic Gain Control

Automatic gain control (abbreviated AGC) is a method (and also a circuit) of automatically obtaining a substantially constant output despite a variation in the signal amplitude at the input.

To give automatic control of gain, a voltage proportional to the amplitude of the carrier of the incoming signal is fed back to the r.f. amplifier, frequency changer, and i.f. amplifier in such a way that the gain is reduced when the amplitude of the received signal increases, and vice versa.

This voltage is developed in the detector stage or in a separate stage immediately following the i.f. amplifier.

The following types of AGC are mainly used:

(1) normal (or simple) AGC;

(2) delayed AGC;

(3) amplified AGC.

A circuit for *normal* (or *simple*) *automatic gain control* is shown in Fig. 12.6. The d.c. component of the voltage appearing across the detector load resistor is applied through R_fC_f to the grid of the valves in the stages covered by AGC. Thus the negative grid bias voltage of these valves is made up of AGC bias voltage and the d.c. voltage taken from the detector load.

As the amplitude of the incoming signal increases, the d.c. voltage across the detector load and the negative grid bias of the controlled valves increase. The greater the amplitude of the signal, the higher the negative bias and the greater the reduction of the gain. As a result, the amplitude characteristic of a receiver with AGC ceases to be linear (Fig. 12.7).

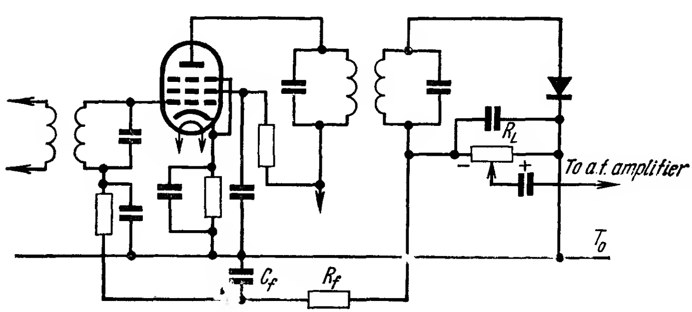


Fig. 12.6. Simple AGC

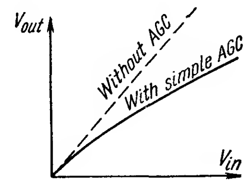


Fig. 12.7. Amplitude characteristic of a radio receiver with simple AGC

The filter R_fC_f and other filters connected in the grid circuit of the controlled stages decrease parasitic inter-stage coupling and prevent demodulation which might occur in AGC circuits employing no filters.

If there were no filter, the grid circuit of each stage would accept a d.c. negative voltage along with an audio-frequency voltage opposite in phase to the envelope of the modulated signal.

Then an increase in the amplitude of the modulated signal would cause an increase in the control voltage; this would unavoidably reduce the maximum amplitude of the modulated signal. With a decrease in the amplitude of the modulated signal (the amplitude of the carrier frequency remains unchanged), the control voltage would decrease and the gain of the receiver would go up. Thus, the effective depth of modulation would be lower.

The demodulation effect may be eliminated by increasing the time constant of the filter. If the time constant is sufficiently large, the filter capacitor will have no time to discharge as the voltage across the detector load varies (which happens at the rate of the audio-frequency), and the control voltage will depend only on the amplitude of the carrier.

The time constant of the filter R_fC_f in broadcast receivers is 0.05 to 0.2. If the time constant were made greater the circuit would lag behind the changes in the amplitude of the carrier caused by fading.

A major disadvantage of simple AGC is the decrease in stage gain in the presence of both strong and weak signals, which fact impairs the receiver sensitivity. This is easily eliminated by delayed AGC.

Delayed AGC operates on practically the same principle as simple AGC. The only difference is that the development of the AGC voltage is delayed until the signal exceeds a certain value. For this purpose, the normal AGC circuit is interrupted by a

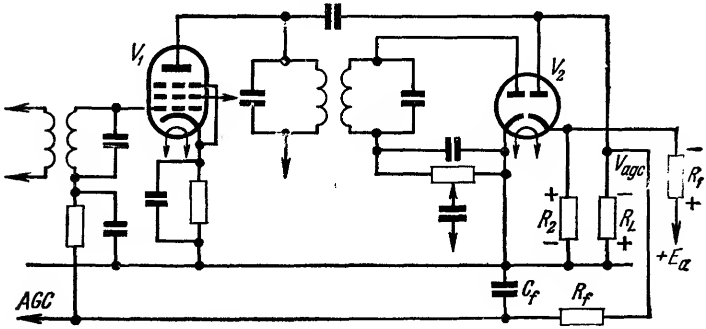


Fig. 12.8. Delayed AGC circuit

second diode which accepts both the r.f. signal and a negative d.c. delay voltage.

Figure 12.8. shows one of possible delayed AGC circuits. The left-hand diode of V_2 is used to detect the incoming signals, while the right-hand one provides automatic gain control.

The r.f. voltage applied to the anode of the second diode is taken from the primary of the i. f. output transformer, i.e. directly from the anode of the i.f. amplifier valve.

The anode circuit of the AGC detector contains a load resistor, R_L, and a filter network, R_fC_f. The cathode circuit contains R_2 of a voltage divider connected to the common supply source of anode voltage. The polarity of the voltage drop produced across R_2 is shown in Fig. 12.8.

In the absence of a signal, the anode of the second diode is negative to the cathode. The cathode potential is equal to the voltage drop across R_2 and is called the *delay voltage* E_d. The value of this voltage depends on the operating conditions of the main detector. In absolute value, the delay voltage must be equal to, or greater than, the amplitude of the r.f. wave necessary for normal operation of the main detector. If, for instance, the receiver detector and all the following stages are so designed that, during reception, the detector must be supplied with a voltage of at least 2 volts, the delay voltage at the anode of the second diode must also be at least 2 volts. If the incoming signal has a lower amplitude, the resultant potential at the anode of the second diode will be negative and there will be no anode current flowing. There will be no current flowing through the load resistor R_L and the grid circuits of the controlled valves will receive no additional

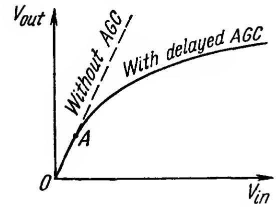


Fig. 12.9. Amplitude characteristic of a radio receiver employing delayed AGC

negative bias. When the amplitude of the incoming signal exceeds the delay voltage, the anode of the AGC detector will become positive, a current will flow through its load resistor, and a negative AGC voltage, V_{agc}, will be built up across this resistor.

The value of the AGC voltage is given with sufficient accuracy by

$$V_{agc} = V - E_d$$

where V is the signal voltage amplitude at the detector; and E_d is the delay voltage.

Thus, the gain is automatically reduced only on signals exceeding the normal operating voltage of the detector.

The amplitude characteristic of a radio receiver employing delayed AGC is shown in Fig. 12.9. When weak signals are being received, AGC does not function and the amplitude characteristic is linear (portion OA). Past point A the amplitude characteristic becomes curved. The departure of the characteristic from a straight line depends on the number of controlled stages and the value of the negative voltage applied to the grids of the controlled valves.

As already noted, the r.f. voltage in the circuit of Fig. 12.8 is fed to the AGC detector anode from the primary and not the secondary of the i. f. output transformer. If the AGC detector were connected to the secondary, its Q-factor would be considerably smaller than the Q of the primary because it would be shunted by the input resistances of the two detectors, and the filter would be unbalanced. When the detectors are connected to the two sides of the i. f. transformer separately, the symmetry is upset but slightly.

To plot the amplitude characteristic of a radio receiver with AGC, it is necessary to know the total gain of all the stages up to the detector for each value of the input voltage. The gain of each controlled stage depends on its anode load resistance and mutual conductance.

It is not difficult to prove that the total gain of a receiver up to the detector is proportional to the product of the mutual conductance of each valve

$$K_{total} = K_1 K_2 \ldots K_n = g_{m_1} R_{0e_1} \times g_{m_2} R_{0e_2} \ldots g_{m_n} R_{0e_n} = A g_{m_1} g_{m_2} \ldots g_{m_n}$$

Thus, the shape of the amplitude characteristic depends on changes in the product of mutual conductances.

The amplitude characteristic is computed as follows. An arbitrary value of AGC voltage V_{agc} is assumed, and the effective mutual conductance is found at the operating point of each valve. After this the product of mutual conductances and the change in gain, as compared with the original gain, are determined.

A few words should be said about power supply for, and the tuning of, receivers with AGC.

AGC will be improved if the circuit does not use self-bias. Bias for the grids of controlled valves should preferably be taken from a common voltage divider.

If self-bias were used, the bias voltage would depend on the valve current. As the negative voltage at the grid of the controlled valve were increased, the anode current and the voltage across the self-bias resistor would be reduced. As a result, the effective value of the AGC voltage would be decreased.

Nor is it advisable to feed the screen grid voltage through separate dropping resistors. A decrease in control grid potential would bring down the screen grid current. The voltage drop across the dropping resistor would go down, and the screen grid voltage would rise. As a result, the mutual conductance of the valve would increase. Thus, in this case, too, the effect of the AGC voltage would be minimized.

It is best to feed the screen-grid voltage either from individual voltage dividers or from a divider common to all screen grids.

As to the tuning of receivers with AGC the main difficulty lies in the fact that a small amount off tune will only distort and not reduce the signal in strength, and sharp tuning is thus hard to obtain. This is because the reduction in gain off tune is made up for by the effect of the AGC voltage.

Sharp tuning in receivers with AGC is facilitated by tuning indicators. The most popular among them is an electron-ray tube, called the "magic eye". This is a combination of a triode and simplified cathode-ray tube. The target of the CRT section is given a coating of a phosphor, a fluorescent material which gives off fluorescence when bombarded by electrons.

The indicator circuit is shown in Fig. 12.10. The target is connected to the positive side of the power supply directly and

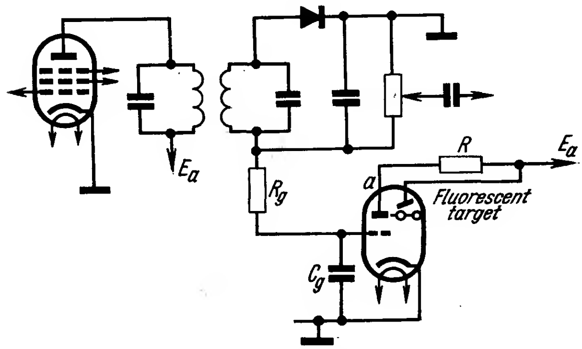


Fig. 12.10. Circuit for a "magic eye"

the triode anode, through a high-value resistor R. Therefore, the anode is always at a lower potential than the target. The triode grid is held at a variable negative direct potential applied through R_gC_g from the anode load of the detector. When no signal is applied, the triode grid is at zero potential, and a considerable current flows in the anode circuit, bringing down the anode potential. As this happens, the potential of the ray-control electrode connected to the triode anode, placed in front of the target, will also decrease. When it is at low potential the ray-control electrode forms a broad shadow pattern on the target, that is, an area not illuminated by electrons as they are deflected by the ray-control electrode.

When a signal appears across the detector load, a negative voltage will be present at the triode grid. The anode current in the triode will decrease and the anode potential increase. This will in turn cause the shadow pattern on the target to contract. When the receiver is tuned exactly to resonance, the maximum negative voltage is developed across the detector load, and the width of the shadow is a minimum.

Review Questions

1. Why is AGC used in the r.f. and i.f. stages?
2. Can AGC be used in the a.f. stages?
3. What determines the number of stages that can be covered by AGC?
4. What is the effect of AGC on receiver sensitivity?
5. Does AGC affect the reliability of radio communication?

64. Bandwidth Control

With bandwidth control it is possible to increase the selectivity of the receiver, to lower the noise level and to vary the tone reproduced by the loudspeaker. The bandwidth can be controlled in the i.f. and a.f. stages. Usually it is controlled manually. In some cases automatic control is used; it automatically narrows the bandwidth when the interference level becomes high.

Bandwidth control in the i.f. stages involves variations in the resonance properties of the i.f. transformers either by varying coupling between the tuned circuits or by shunting one of the tuned circuits with a resistance.

With capacitive coupling between the tuned circuits, the bandwidth is controlled by varying the coupling capacitance. With inductive coupling, the bandwidth is controlled by varying the distance between the coils. In some cases, band switching is used instead of continuous bandwidth control. In band switching, an additional coupling coil is introduced into the tuned circuit. A band switching circuit is shown in Fig. 12.11, where a wider band is obtained by connecting in the secondary tuned circuits additional coils L_c coupled to the primary tuned circuits, so that interstage coupling is increased.

Bandwidth control in the audio-frequency stages is based on changing the frequency response of these stages. In this case, the bandwidth is controlled either by connecting reactive components affecting the frequency response into the circuit, or by using feedback in anti-phase with the signal, or by employing audio-frequency filters with a limited pass-band.

Figure 12.12 shows one of the most commonly used circuits for reducing the bandwidth at the higher frequencies. The band-

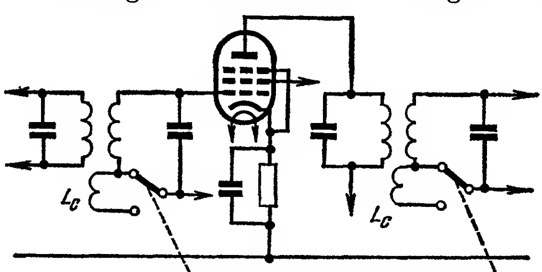


Fig. 12.11. Bandwidth control by changing coupling between the tuned circuits

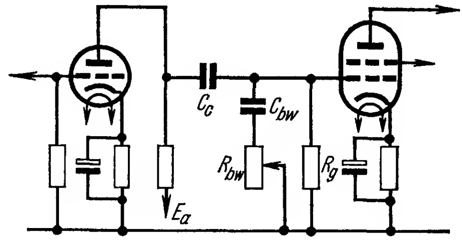


Fig. 12.12. Bandwidth control in the high-frequency range

width at the higher frequencies is controlled by a network, R_{bw} and C_{bw}, connected in parallel with the anode load.

At a low value of R_{bw}, the capacitance C_{bw} produces a considerable shunting effect in the high frequency range, and the frequency response in that range drops sharply. With an increase in R_{bw}, the shunting action of C_{bw} decreases, the frequency response in the high-frequency range flattens, and the bandwidth broadens.

Figure 12.13 shows the frequency response of a stage with different values of R_{bw}.

In the circuit of Fig. 12.14 bandwidth control is effected by variations in the voltage gain β and the phase φ_β of the feedback circuit. The capacitance of C_{bw} is so selected that the impedance of $C_{bw}R_{bw}$ at the audio frequencies is considerably higher than R_g, and the negative feedback signal is comparatively low.

As the frequency increases, the reactance of C_{bw} and the impedance Z_{bw} are decreased. This results in an increase in the feedback factor and the phase shift approaches 180°. Thus, the

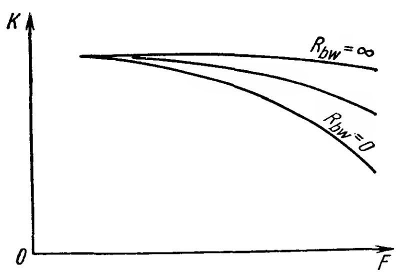


Fig 12.13. Frequency response of an amplifier with various values of R_{bw}

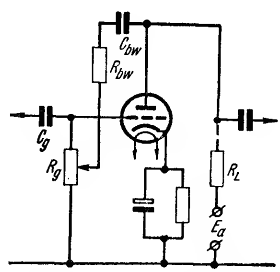


Fig. 12.14. Bandwidth control by means of negative feedback

stage gain at higher frequencies is smaller than at lower frequencies.

When the setting of the potentiometer R_g is reduced, the factor β is decreased and the frequency response is flattened in the high-frequency range.

In the extreme lower position of the potentiometer the amplifier has the broadest bandwidth.

Review Questions

1. How can the bandwidth be controlled in the r.f. stages?
2. Can the bandwidth in the i.f. stages be controlled continuously?
3. How can the bandwidth be controlled in the a.f. stages?
4. Can the bandwidth be controlled by varying the capacitance of the bypass capacitor in the self-bias circuit?

65. Automatic Frequency Control

As we know, the intermediate frequency f_i in a superheterodyne receiver is the difference in frequency between the signal and the local oscillator

$$f_i = f_o - f_s$$

Variations in any of the two frequencies change the intermediate frequency and detune the i.f. stages. This would inevitably reduce the power output of the receiver and decrease the reliability of radio communication.

As a rule, the signal frequency of broadcast and other stationary radio transmitters is sufficiently stable. But the frequency of the local oscillator in the receiver can vary within considerable limits with time (frequency drift).

These variations in the frequency of the local oscillator can be eliminated by an automatic control. With automatic frequency control (abbreviated AFC), any variations in the local-oscillator frequency from a value necessary for obtaining the intermediate frequency are compensated automatically. The block diagram of a receiver with AFC is shown in Fig. 12.15.

The AFC circuit comprises a frequency discriminator and a reactance-valve stage connected to the tuned circuit of the local oscillator. The frequency discriminator develops a direct voltage proportional to the deviations of the intermediate frequency from its assigned value, and in opposite polarity.

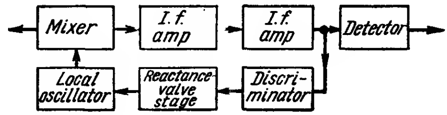


Fig. 12.15. Block diagram of a radio receiver with automatic frequency control (AFC)

The reactance-valve stage, which accepts the 'frequency-correcting voltage' from the discriminator, brings the local oscillator back to the desired frequency.

In a reactance-valve circuit, the anode is coupled to the grid of the same valve in such a way that the voltage fed back into the grid circuit is nearly 90° out of phase with the alternating anode voltage. With such feedback, the valve thus acts as a variable reactance.

Whether the reactance valve acts as a variable inductance or as a variable capacitance, depends on how feedback is coupled in.

Consider operation of the reactance-valve circuit. Fig. 12.16*a* shows one of the most commonly used reactance-valve circuits. The anode of the reactance valve accepts a direct voltage and an r.f. signal. Part of the r.f. voltage is fed through the feedback circuit $C_{fb}R_{fb}$ to the grid.

The values of C_{fb} and R_{fb} are so selected that the following condition is satisfied at all frequencies

$$\frac{1}{\omega C_{fb}} \gg R_{fb} \tag{12.1}$$

As a result, the current in the feedback circuit will lead the r.f. voltage applied to the anode by approximately 90°. Since the feedback voltage applied to the grid is taken from R_{fb} it will be in phase with the current in the feedback circuit, and

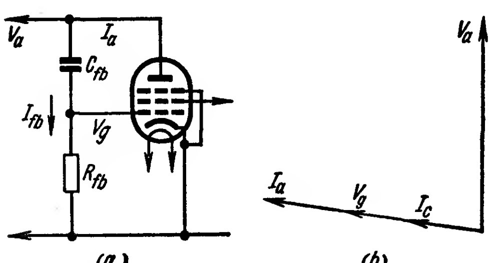


Fig. 12.16. Reactance valve acting as a variable capacitance. Vector diagram of currents and voltages acting in the circuit

the alternating anode current will be in phase with the voltage applied to the grid.

The vector diagram of the voltages and currents in the circuit is shown in Fig. 12.16*b*. As is seen, the anode current vector I_a leads the anode voltage vector by approximately 90°. Thus the output impedance of the valve

$$Z_{out}=\frac{\dot{V}_a}{\dot{I}_a} \tag{12.2}$$

is capacitive in its effect. In other words, the valve acts as a variable capacitance.

Determine the relation between the equivalent output capacitance of the reactance valve and the parameters of the circuit and valve.

The alternating anode current mainly depends on the grid voltage and the mutual conductance of the valve

$$\dot{I}_a=g_m\dot{V}_g \tag{12.3}$$

The grid voltage is

$$\dot{V}_g=R_{fb}\dot{I}_{fb} \tag{12.4}$$

The current through the feedback circuit is

$$\dot{I}_{fb}=\frac{\dot{V}_a}{R_{fb}+\frac{1}{j\omega C_{fb}}}\cong\frac{\dot{V}_a}{\frac{1}{j\omega C_{fb}}}=j\omega C_{fb}\dot{V}_a \tag{12.5}$$

Substituting (12.4) and (12.5) in (12.3), we obtain

$$\dot{I}_a=jg_m\dot{V}_aR_{fb}\omega C_{fb}$$

Therefore

$$Z_{out}=\frac{\dot{V}_a}{\dot{I}_a}=\frac{\dot{V}_a}{jg_mR_{fb}\omega C_{fb}\dot{V}_a}=-j\frac{1}{\omega C_{fb}R_{fb}g_m}$$

The product $C_{fb}R_{fb}g_m$ has dimensions of capacitance; let it be called the equivalent capacitance C_{eq} such that

$$C_{eq}=C_{fb}R_{fb}g_m=\tau_{fb}g_m \tag{12.6}$$

Thus, the output impedance Z_{out} of the valve is capacitive in its effect while the equivalent capacitance C_{eq} depends on the time constant of the feedback circuit and the mutual conductance of the valve.

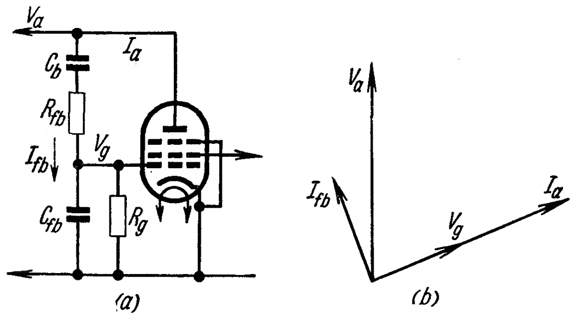


Fig. 12.17. Reactance valve acting as a variable inductance and the associated vector diagram

Consider another reactance-valve circuit (Fig. 12.17*a*). In this arrangement, the feedback circuit consists of R_{fb} and C_{fb}. The blocking capacitor C_b and the grid-leak resistor R_g perform auxiliary functions in the circuit and have little effect on the character and value of feedback.

The values of C_{fb} and R_{fb} are so selected that the following inequality is satisfied at all operating frequencies:

$$R_{fb} \gg \frac{1}{\omega C_{fb}}$$

A vector diagram of the voltages and currents acting in the circuit is shown in Fig. 12.17*b*.

The feedback current vector I_{fb} slightly leads the anode voltage vector. The vector of the grid voltage taken from C_{fb} lags behind the current by 90°. The anode current vector I_a is in phase with the grid voltage vector V_g. It is not difficult to see that the anode current vector lags behind the anode voltage vector by nearly 90°. Thus, the output impedance of the valve is inductive in its effect, that is, the valve acts as a variable inductance. It may be shown that the equivalent inductance L_{eq} is given by

$$L_{eq} = \frac{C_{fb} R_{fb}}{g_m} = \frac{\tau_{fb}}{g_m}$$

Consequently, the equivalent inductance L_{eq} is directly proportional to the time constant of the feedback circuit and inversely proportional to the mutual conductance.

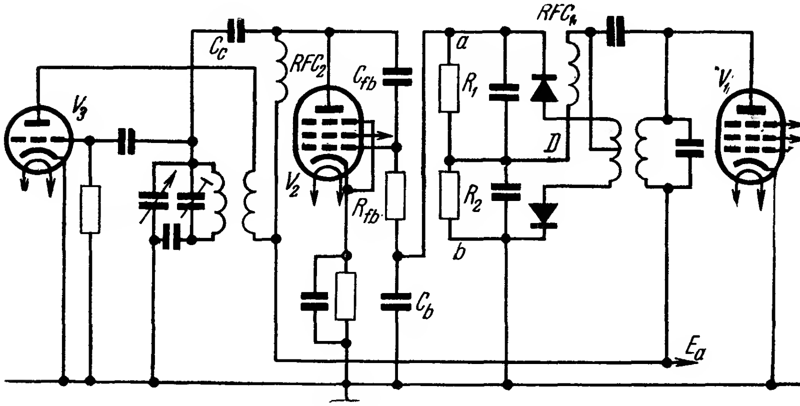


Fig. 12.18. Receiver circuit with AFC

The equations derived for the equivalent capacitance and inductance of the reactance valve suggest the most convenient method for controlling the frequency of the local oscillator. The anode circuit of the reactance valve should be connected to the tuned circuit of the local oscillator. The latter will produce a composite signal whose frequency will vary in accordance with the bias on the reactance valve. The bias can be taken from the output of the discriminator.

Fig. 12.18 shows a suggested AFC circuit. The i.f. voltage from the anode tuned circuit of the valve V_1 is applied to the input of the discriminator D employing semiconductor diodes.

When the intermediate frequency deviates from its assigned value, a d.c. voltage, proportional to the deviation appears across R_1 and R_2. This voltage is injected into the grid circuit of the reactance valve V_2, connected into the circuit of Fig. 12.16. As already shown, the valve V_2 in such a circuit acts as a variable capacitance.

The anode circuit of the reactance valve V_2 is connected to the local oscillator via a coupling capacitor C_c. That is the equivalent capacitance of the reactance valve is, in effect, connected in parallel with the tuned-circuit capacitor of the local oscillator.

The discriminator circuit must be arranged so that when the i.f. frequency decreases, the potential at point a goes negative.

Suppose that heating causes an increase in the total capacitance of the local oscillator. This entails a decrease in the frequency of the local-oscillator signal. As the local-oscillator frequency goes down, the intermediate frequency also goes down

$$f_i = f_o - f_s$$

As a result, the discriminator develops a negative 'frequency-correcting' voltage across *ab*.

The negative grid bias on the valve increases, reducing the mutual conductance at the operating point and the equivalent capacitance C_{eq} of the reactance valve. Accordingly, this decreases the total capacitance of the tuned circuit, and the local-oscillator frequency and the i.f. are restored to their assigned values.

AFC not only restors the intermediate frequency when the local-oscillator frequency drifts, but also eliminates all inaccuracies in local-oscillator alignment. The latter feature is important in push-button-tuned receivers and when the tuned circuits of a multi-band superheterodyne receiver are not tracking perfectly.

Review Questions

1. How can the reactance valve be replaced in an AFC circuit?
2. Does AFC affect the reliability of radio communication?
3. Is AFC necessary in receivers with a crystal-controlled local oscillator?
4. Does the bandwidth of the i.f. amplifier affect operation of the AFC circuit?

SUMMARY

1. For normal operation, receivers incorporate circuits for control of sensitivity or gain, and bandwidth.
2. Receiver sensitivity may be controlled manually or automatically.
3. Automatic gain control is performed in the stages preceding the detector by varying the mutual conductance of the valves.
4. Manual gain control is usually performed in the receiver detector, i.e. at the input of the audio-frequency amplifier.
5. Bandwidth control may be effected in the i.f. and a.f. stages.

6. Bandwidth control makes it possible to decrease interference and to vary the tone reproduced by the receiver.
7. The intermediate frequency of a radio receiver can be maintained constant by automatic frequency control (AFC).
8. For its operation, the reactance-valve stage in the AFC system depends on variable reactive feedback.

CHAPTER XIII

CALCULATION OF SUPERHETERODYNE RECEIVERS OPERATING ON LONG, MEDIUM AND SHORT WAVES

66. General

The present chapter deals with the procedures for calculation of communication and broadcast AM superheterodyne receivers operating on long, medium and short waves.

A receiver design is based on specifications. However, the specifications are not sufficient to formulate all the requirements that the receiver should meet. Therefore, the design procedure involves preliminary and final calculation.

Preliminary calculations include approximate selection of the receiver circuitry and determination of variables necessary for the calculation of each stage.

Final calculations include the computation of the individual stages of the receivers.

The design procedure ends with the calculation of the overall characteristic of the receiver and drawing up its circuit diagram complete with components lists.

The design specifications of a receiver include the following:

1. Application of the radio receiver and type of installation.
2. Type of operation (telephony, telegraphy).
3. Frequency range, f_l-f_h.
4. Number of bands and maximum-to-minimum frequency ratio of the band.
5. Lowest sensitivity, in microvolts.
6. Adjacent-channel selectivity, that is, the attenuation of adjacent-channel signals d (in db) at Δf kilohertz off resonance.
7. Image response d_{im} in db.
8. I.f. interference response d_{if} in db.
9. Type of terminal load (earphones, loudspeaker, printer, etc.), and its impedance.
10. Power output P_{out} in watts or output voltage V_{out} in volts and the limit of non-linear distortion γ in per cent.

11. Modulating-frequency range F_l-F_h in hertz. The highest modulating frequency F_h simultaneously characterises the sideband of the radio frequency section of the receiver ($\Delta F = F_h$).

12. Frequency distortion M (in db) of the receiver at the highest modulating frequency F_h. The value of M characterises total frequency distortion of the entire receiver.

13. Requirements for manual and automatic gain control. Different types of manual and automatic control may be incorporated in the receiver according to its expected application:

(a) Manual bandwidth control is employed in telephone-telegraph communication receivers and broadcast sets.

(b) AGC is specified in terms of range of control as follows: with the input voltage varying α times, the output voltage should change not more than β times. This type of control is used in most modern receivers.

(c) AFC is specified in terms of the AFC factor.

14. Aerial parameters (resistances and reactances).

For long, medium and, partly, short waves, a typical aerial used with stationary receivers is assumed to have an effective height $H_{ef} = 4$ m; $r_A = 25$ ohms; $L_A = 20$ microhenrys; and $C_A = 150$ to 300 picofarads.

15. Phono-jack sensitivity. It is specified only for broadcast receivers in which provisions are made for record playing.

16. Types of valves.

17. Reliability requirements, especially for vehicular installations.

18. Type of power supply and limits of power consumption, especially for battery-powered receivers.

19. Tuning indication. In broadcast receivers this is done by a "magic eye". Some communication receivers employ pointer-type tuning indicators.

20. Mechanical requirements:

(a) overall dimensions;
(b) protection from atmospheric exposure and pressurization;
(c) shock-mounting;
(d) controls;
(e) adaptability to mass production.

21. Safety requirements.

67. Preliminary Calculations

As part of preliminary calculations, the number of bands is determined (if it has not been specified in advance) and also the limiting frequencies of the bands, the number of stages, types of valves and the number of tuned circuits are decided upon. The suggested procedure of preliminary calculations is as follows.

Choice of the Number of Bands. The specified frequency range has usually to be broken down into bands when the variable capacitor cannot cover it with a single fixed tuned-circuit inductance.

Normally, the ratio of the highest to lowest frequency within a range (or band) for which the same coil and tuning capacitor can be used, K_r or K_b, is 1.2 to 3.

If K_r is greater than three, the frequency range should be divided into bands of an equal f_h to f_l ratio.

Circuit Configuration for the R.F. Section. This is done in agreement with the specified values of sensitivity, bandwidth, adjacent-channel selectivity, and image response, which are strongly affected by the value of the intermediate frequency. The choice of the intermediate frequency is explained in Chapter IX.

Choice of the Number of R.F. Tuned Circuits. This is usually controlled by the specified image response. The effect of these tuned circuits on the bandwidth of the receiver should also be considered. The tentative number of tuned circuits n may be found from the image response d for a single tuned circuit.

Thus, in the 150-1,600 kilohertz range, d will be:
25-40 db at $f_i = 465$ kilohertz;
20-30 db at $f_i = 110$-115 kilohertz;
In the 1,600-15,000 kilohertz range, d will be:
10-25 db at $f_i = 465$ kilohertz.

To determine the final number of tuned circuits it is necessary to find their Q and check the found number of tuned circuits against its value.

The Q is determined from the following considerations:

1. From the consideration of image response d_{im}. Using Eq. (9.18)

$$d_{im} = \left[Q_{ef}\left(\frac{f_0+2f_i}{f_0} - \frac{f_0}{f_0+2f_i}\right)\right]^n \frac{f_0+2f_i}{f_0}$$

we get

$$Q'_{ef}=\frac{\sqrt[n]{\dfrac{d_{im}f_0}{f_0+2f_i}}}{\dfrac{f_0+2f_i}{f_0}-\dfrac{f_0}{f_0+2f_i}}$$

In the short-wave range, where $2f_1 \leqslant (0.1-0.15)\,f_0$, the factor $\frac{f_0+2f_i}{f_0}\cong 1$, and it may be neglected. Then the equation for Q_{ef} takes a simpler form

$$Q'_{ef}=\frac{\sqrt[n]{d_{im}}}{\dfrac{f_0+2f_i}{f_0}-\dfrac{f_0}{f_0+2f_i}}$$

The value of Q_{ef} should be determined for the highest frequency of the band, where the resonance curve of the tuned circuit is flat-topped most.

2. From the considerations of bandwidth and frequency distortion M'_K. This quantity is the ordinate of the resonance curve at the limit of the bandwidth.

For n tuned circuits this ordinate is

$$Y_{\Delta F}=M'_K=\left(\frac{1}{\sqrt{1+\left(\dfrac{2\Delta F}{f_0}Q''_{ef}\right)^2}}\right)^n$$

After some manipulations the Q_{ef} is found to be

$$Q''_{ef}=\frac{f_{0\,min}}{2\Delta F}\,\frac{\sqrt{1-\sqrt[n]{M'^2_K}}}{\sqrt[n]{M'_K}}$$

The frequency distortion should be 0.6 to 0.8 in the long-wave band from 2,000 to 750 metres (150 to 400 kilohertz); if $n=2$, $M'_K=0.8$.

In the band extending from 500 to 2,000 (or 3,000) kilohertz, $M'_K=0.7$ to 0.9. In the short-wave band (above 3,000 kilohertz), $M'_K=0.9$ to 0.95.

The Q''_{ef} should be calculated for the bandwidth $2\Delta F$ such that

$$2\Delta F \geqslant 2\,(\Delta F_{spec}+\Delta f_{track}+\Delta f_o)$$

where $2\Delta F_{spec}$ is the specified bandwidth; Δf_{track} is the tracking-error limit which should be 10-20 kilohertz for short waves and 1-5 kilohertz for medium and long waves; Δf_o is the expected drift of the local-oscillator frequency.

Assuming that variations in the local-oscillator frequency are five to ten parts in ten thousand, we have

$$f_o = (0.5 \text{ to } 1)\times 10^{-3} f_o$$

From the values of Q'_{ef} and Q''_{ef}, found from the considerations of image response and bandwidth, the final value of Q_{ef} is found from the following inequality:

$$Q''_{ef} > Q_{ef} > Q'_{ef} \qquad (*)$$

With this Q_{ef} the resonance curve will be sharper than is required for the specified image response, but not so broad as is necessary for the bandwidth $2\Delta F$.

If the Q''_{ef} happens to be less than the Q'_{ef}, it is necessary to begin with a value of M'_K which is smaller than the former value and to find a new value of Q''_{ef}. If the condition (*) is not satisfied in this case either, the number of tuned circuits n should be increased.

The obtained Q must be realizable. As a guide, for tuned circuits using air-core coils wound with single-conductor wire, the Q-factor will be from 25 to 50. For tuned circuits using ferrite-core coils or coils wound with Litz wire, it is possible to obtain a Q of 50 to 100. In tuned circuits with gap-free cores, the Q can be 250.

After this, the adjacent-channel selectivity is determined from the Q_{ef}:

$$d' = \left[\sqrt{1+\left(\frac{2\Delta f}{f_{0\,max}} Q_{ef}\right)^2}\right]^n$$

In the short-wave region, d' is clos to unity.

Then the actual frequency distortion is determined for one tuned circuit:

$$M_K = \frac{1}{\sqrt{1+\left(Q_{ef}\frac{2\Delta F_{spec}}{f_{0\,min}}\right)^2}}.$$

Frequency Distortion. The limits of frequency distortion specified in terms of M must be distributed between the r.f., i.f., and a.f. sections of the receiver.

Since the frequency separation is comparatively narrow (not more than 10 kilohertz for long, medium and short waves), it should be assumed that the frequency distortion of the audio-frequency section may be 1 to 2 db.

The frequency distortion of the r.f. and i.f. sections M_{hf} is

$$M_{hf\,\mathrm{db}} = M_{\mathrm{db}} - M_{af\,\mathrm{db}}$$

The obtained value of M_{hf} should be distributed between the r.f. and i.f. sections.

For the incoming-signal (radio) frequency section, i.e. the aerial input circuit and r.f. amplifier

$$M_{rf} = (M_K)^n$$

where M_K is the frequency distortion of a single-tuned circuit.

The frequency distortion of the i.f. section,

$$M_{if} = \frac{M_{hf}}{M_{rf}}$$

or, in decibels:

$$M_{if\,\mathrm{db}} = M_{hf\,\mathrm{db}} - M_{rf\,\mathrm{db}}$$

The frequency distortion should be distributed in each band.

Number of I.F. Transformers. In most receivers for long, medium and short waves, the i.f. amplifier and the frequency changer employ double-tuned bandpass filters as i.f. transformers. The number of such filters is determined on the basis of the desired adjacent-channel response and specified bandwidth of the receiver.

The calculation should provide a margin of 15 to 20 per cent in response, in case the selectivity should be impaired by imperfect tracking of the tuned circuits. Besides the adjacent-channel selectivity of the r.f. tuned circuits should be taken into consideration

$$d = \frac{(1.15 \text{ to } 1.2)\, d_{spec}}{d'}$$

where d_{spec} is the specified adjacent-channel selectivity.

Receivers most frequently use two or three i.f. transformers, one in the frequency changer and two in the i.f. amplifier. It is rare practice to use a single i.f. transformer in the receiver.

Some help in fixing the number of i.f. transformers can be had from Table 13.1.

TABLE 13.1

Bandwidth $2\Delta F$, kHz	Adjacent-channel selectivity d, db	Number of transformers m
10	20-24	2
	25-28	3
8	26-30	2
	30-38	3
7	30-35	2
	36-45	3
6	35-40	2
	40-60	3

Table 13.1 holds for tuned circuits with a Q of less than 120, an intermediate frequency of 465 kilohertz, and a frequency distortion of 0.6 to 0.7 at the i.f.

The value of m is selected from Table 13.1 for the specified bandwidth $2\Delta F$ and adjacent-channel selectivity.

Number of Stages in the R.F. Section. This is found from the sensitivity $V_{s\,min}$ such that the i.f. voltage at the detector input will provide for linear detection. In this case, the best choice is diode detection. The normal operation of a diode detector is obtained when the i.f. voltage applied to its input is $V_d = 2$ to 5 volts.

It should also be remembered that in receivers with a sensitivity of 300 to 400 microvolts, the value of V_d is from 1.5 to 2 volts. In highly sensitive receivers, using a "magic eye" or another type of tuning indicator, $V_d = 3$ to 5 volts.

The overall gain of the r.f. and i.f. sections taken together should be

$$K''_{hf} = \frac{V_d}{V_{s\,min}}$$

Allowing a gain margin of 25 to 40 per cent, we have

$$K'_{hf} = (1.25 \text{ to } 1.4) K''_{hf}$$

Once the necessary gain of the r.f. section, the number of r.f. tuned circuits n and the number of i.f. transformers m are known, the number of stages that will secure such amplification can be decided upon.

TABLE 13.2

Circuit or stage	Gain, K			
	long and medium waves	short waves	$f_i=465$ kHz	$f_i=110$ to 115 kHz
Aerial-input circuit	2-4	3-8	—	—
One r.f. stage	20-40	5-25	—	—
Frequency changer	—	—	15-40	20-30
One i.f. stage	—	—	50-150	100-140

Table 13.2 gives approximate values of gain in the various circuits from which we can find the gain K_{hf}

$$K_{hf}=K_{in}K_{rf\,amp}K_c\,(K_{if\,amp})^{m-1}$$

As already noted, the number of r.f. tuned circuits n rarely exceeds 2. If $n=1$, there is no r.f. amplifier, and a single-tuned circuit is connected to the receiver input. If $n=2$, then, as a rule, one stage of r.f. amplification is used along with a single-tuned aerial-input circuit.

It should be noted that in some cases two r.f. tuned circuits may be arranged as an r.f. transformer.

The obtained value must satisfy the following condition:

$$K_{hf}\geqslant K'_{hf}$$

In cases where K_{hf} is 30 to 50 times K'_{hf}, it is necessary to use valves of low mutual conductance and tuned circuits with a small value of R_{0e} in the final design of the i.f. amplifier.

If K_{hf} is smaller than K'_{hf} at least in one of the bands, it is necessary to use valves of high mutual conductance. In cases where insufficient amplification calls for an additional stage, it is recommended to increase the number of stages in the i.f. amplifier. When only a single-tuned circuit is used at the input, it may be worth while to incorporate an untuned r.f. amplifier into the receiver.

Circuit Configuration for the A.F. Section of the Receiver. 1. Select circuit configuration and type of valve for the final stage, so as to obtain the specified power output, P_{out}.

2. Determine the number of stages and types of valves for the voltage amplifier so as to secure the necessary voltage gain

$$K_v = \frac{V_{gf}}{V_{in\,min}}$$

where V_{gf} = drive voltage to be applied to the grid of the final stage to obtain the specified power output

$V_{in\,min}$ = minimum input voltage to be applied to the grid of the first a.f. amplifier valve.

The drive voltage V_{gf} is given by

$$V_{gf} = \frac{1+\alpha}{\mu} \sqrt{\frac{2PR_a}{\alpha}}$$

where α is the load factor which is 2 to 4 for triodes, and from 0.1 to 0.15 for beam tetrodes and pentodes; $P = \frac{P_{out}}{\eta_t}$ is the power output of the final valve; and η_t is the output transformer efficiency, which is 0.8 to 0.9.

For a Class A push-pull power amplifier the computation is performed for one arm of the circuit and one-half of the power.

The value of $V_{in\,min}$ depends on the expected application of the receiver. For communication receivers

$$V_{in\,min} = K_d V_d$$

where K_d is the voltage gain of the diode detector, equal to 0.6-0.8, and V_d is the voltage at the detector input, whose value is given in (2) of the present section.

For radio-gramophones the value of $V_{in\,min}$ is decided by the sensitivity of the gramophone pickup. A crystal pickup develops a voltage of 0.5 volt, while an electromagnetic unit, 0.15-0.25 volt. The minimum voltage developed by the pickup gives the value of $V_{in\,min}$.

Having determined K_v, we can now determine the number of stages in the voltage amplifier and the types of valves to be used in the receiver.

With a push-pull power amplifier there must be a transformer or phase-inverter stage.

The phase-inverter may use two or one valve with a split load. In the former case, each arm has the same gain as an *RC*-coupled stage using the same valve. In the split load circuit, the gain of one arm is 0.8-0.9.

3. The frequency distortion of the audio-frequency section (M_{af}) is divided between the voltage amplifier (M_{va}), power amplifier

(M_{pa}) and detector (M_d)

$$M_{af\,\mathrm{db}} = M_{va\,\mathrm{db}} + M_{pa\,\mathrm{db}} + M_{d\,\mathrm{db}}$$

It should be noted that the power amplifier introduces a greater distortion than a voltage amplifier emloying an *RC*-coupled circuit.

From the preliminary design, a complete block diagram of the receiver may be drawn up.

68. Final Calculations

In this phase of design procedure, the stages of the selected circuit configuration and the selectivity characteristics of the receiver are computed. The point of departure is the specifications and the values obtained in preliminary calculations.

The following procedure may be suggested:

1. Determine the parameters of the r.f. tuned circuits.
2. Calculate the aerial coupling to the tuned circuit.
3. Calculate the radio-frequency amplifier.
4. Calculate the intermediate-frequency amplifier.
5. Calculate the frequency changer.
6. Calculate and plot the selectivity characteristic.
7. Calculate the diode detector circuit.
8. Calculate the AGC circuit.
9. Calculate the audio-frequency power amplifier.
10. Calculate the audio-frequency voltage amplifier.
11. Determine the power consumed by the radio receiver.

Since the procedures for calculation of the various stages have been discussed in the preceding chapters, the present section will deal only with the parameters of the r.f. tuned circuits and the selectivity characteristics.

Calculation of the R.F. Tuned Circuits (Aerial-input Circuit and R.F. Amplifier)

Given: (from preliminary calculations)

1. Frequency band, f_h-f_l.
2. Q_{ef} of the loaded tuned circuit.
3. Band factors, K_b (maximum-to-minimum frequency ratio within each band).

To Find:

1. Capacitance range, C_{min}-C_{max}, of the variable capacitor.
2. Capacitance of the trimmer capacitor, C_T.
3. Tuned-circuit inductance.

Design Procedure:

1. Select a variable capacitor to secure the specified minimum and maximum capacitances. For communication receivers operating in a single-frequency range refer to Table 13.3.

TABLE 13.3

f, kHz	C_{max}, pF	C_{min}, pF
up to 300	450-700	12-25
300- 1,500	250-500	10-15
1,500- 6,000	150-250	8-12
6,000-30,000	50-150	6-10

For multi-band broadcast receivers it is best to use standard gang tuning capacitors.

2. Find the equivalent capacitance C_{eq} necessary for obtaining the required value of the band factor K_b

$$C_{eq} = \frac{C_{max} - K_b^2 C_{min}}{K_b^2 - 1}$$

The obtained value of C_{eq} must be positive and lie between 20 and 70 picofarads. If this value is exceeded, another gang capacitor should be used.

3. Find the capacitance of the trimmer

$$C_T = C_{eq} - C'_{eq}$$

where C'_{eq} is arbitrarily chosen to be 25 to 40 picofarads for long and medium waves, and 15 to 20 picofarads for short waves. The obtained value is rounded off to the closest standard value.

4. Find the tuned-circuit inductance for each band

$$L=\frac{2.53\times10^{10}}{C_{max}-C_{min}}\,\frac{f_h^2-f_l^2}{f_h^2 f_l^2}$$

where L is in microhenrys, C in picofarads and f in kilohertz.

5. Find the Q of the unloaded tuned circuit

$$Q\cong\frac{Q_{ef}}{0.8}$$

The Q is one of the starting points for the construction of the coil. In the long-wave range, the Q may be so low that it may be difficult to build a coil. In this case, the Q of the coil should be brought down deliberately by connecting a resistance either in series or in parallel with the tuned circuits.

Calculation and Plotting of the Selectivity Curve. The selectivity of a receiver is calculated as the product of selectivity responses (ordinates) of the respective response curves of the aerial-input circuit, r.f. amplifier, frequency changer and i.f. amplifier.

In view of the similarity between the aerial-input circuit and the r.f. amplifier, and also between the frequency changer and the i.f. amplifier, two resonance curves are needed at the incoming-signal frequency, and at the intermediate frequency.

The resonance curves at the incoming-signal frequency for a receiver using an aerial-input circuit with a single-tuned circuit can be obtained by use of the equation:

$$d=\sqrt{1+\left(\frac{2\Delta f}{f_0}Q_{ef}\right)^2}\,\frac{f_0+\Delta f}{f_0}$$

If the receiver also has an r.f. amplifier, the following equation is used:

$$d=\left[\sqrt{1+\left(\frac{2\Delta f}{f_0}Q_{ef}\right)^2}\right]^n\frac{f_0+\Delta f}{f_0}$$

The selectivity curves of the entire receiver are obtained by adding together the values of d_1 and d_2 in decibels with the same amount, Δf, off resonance at f_h and f_l, and tabulating the results.

On short waves the value of d_1 in decibels, as measured at 0 to 10 kilohertz off resonance, is nearly zero. Because of this, the overall selectivity curve merges with the resonance curve of the frequency changer and i.f. amplifier.

Resonance and selectivity curves should be plotted for the highest and lowest frequencies of each band; only one graph, for the highest frequency of the band, will suffice for short waves. The curves should preferably be plotted on a logarithmic scale, laying off selectivity as ordinate both in relative units and in decibels.

When the logarithmic scale is used, divisions are made on each axis; these divisions correspond to the decimal logarithm of the attenuation. The attenuation is written against the divisions. A slide-rule scale may be employed for the logarithmic scale.

Calculation of Power Consumption. The power consumed by the receiver must be calculated in order to determine the requirements for a rectifier or cells and batteries.

1. Filament power:

(a) with filaments in parallel

$$P_f = E_f(I_{f1} + I_{f2} + I_{f3} + \ldots)$$

(b) with filaments in series

$$P_f = I_f(E_{f1} + E_{f2} + E_{f3} + \ldots + E_{f\,bar})$$

where E_f = filament voltage in each valve
$E_{f\,bar}$ = voltage across the baretter or dropping resistor
I_f = filament current.

2. Power consumed by the anode and screen grid circuits:

$$P_a = E_{a\,max}(I_{a1} + I_{a2} + \ldots + I_{s\cdot g1} + I_{s\cdot g2} + I_{s\cdot g3} + \ldots)$$

3. Total power consumption of the receiver is

$$P_\Sigma = 1.2(P_f + P_a)$$

where the factor 1.2 takes care of the necessary power reserve.

CHAPTER XIV

RADAR RECEIVERS

69. General

By usage, the word "radar" applies to radio equipment and techniques used to detect the presence and to determine the distance *(range)* of objects *(targets)* by means of reflected radio waves.

Radar systems operate in the microwave region, mostly in the decimetre and centimetre (UHF and SHF) bands.

At these frequencies, qualitatively new mechanisms are at work, which limit the use of ordinary valves and tuned circuits.

Among other things, vacuum valves are subject to "transit-time" effects, when the transit time of an electron is comparable with the period of the signal on the grid of a triode.

As a result, the alternating component of the anode current tends to lag behind the phase of the grid voltage. The grid loses its control properties, and the valve no longer amplifies the signal.

The inductance and capacitance of conventional tuned circuits in the microwave region are so insignificant that the physical dimensions of the circuits become comparable with the wavelength. Due to this, radiation of energy takes place, IR loss in tuned circuits increases, and the Q is sharply reduced.

Obviously, the usual valves and tuned circuits cannot be used in the microwave region; special valves intended for this range only, and tuned circuits with distributed parameters are used instead. For example, in SHF radars, the local oscillators use *klystrons* which combine a valve and a tuned circuit in one unit.

The maximum stable gain of receivers in the microwave region is also limited.

So far we have specified receiver sensitivity in terms of the minimum input voltage required to produce a specified output voltage having a definite signal-to-noise ratio.

This concept is still applicable to VHF receivers, but in the UHF band and higher it can be seldom used. Instead, the concept of *power sensitivity* is introduced.

The point is that a radar aerial, which is the signal source at the receiver input, is described in terms of effective area defined as the ratio of the maximum power dissipated by the load to the energy flux (power) density of the incident electromagnetic wave. Hence, a meaningful definition of a signal at the receiver input is in terms of power and not of voltage. It is the power of the echo signal that decides the operating range of radar.

There are *effective power sensitivity* and *available power sensitivity*. For their understanding, let us introduce a new quantity, the *signal-to-noise power ratio*, α_p defined as

$$\alpha_p = P_{s\ out}/P_{n\ out} \tag{14.1}$$

that is, the ratio of the maximum power in the load of the aerial to the energy flux density of the incident wave. Obviously, a discernible response will be obtained only when $P_{s\ out}$ is sufficiently above $P_{n\ out}$. The signal power just sufficient to produce such a response is the effective sensitivity of a radar receiver, P_{ef}.

However, no matter what we might do to improve the response, the inherent or irreducible noise will set a limit to the minimum value of α_p. For radar receivers with visual display, for example, the limit is $\alpha_p = 1$, when the input signal power is equal to the irreducible noise power. This is the available sensitivity of the receiver.

Let us establish the relationships involved mathematically. When

$$\alpha_p = P_{s\ out}/P_{n\ out} = 1$$

Eq. (10.6) for the noise factor reduces to

$$N = \frac{P_s/P_n}{P_{s\ out}/P_{n\ out}} = P_s/P_n,$$

whence

$$P_s = NP_n = P_{r\ min} \tag{14.2}$$

where $P_{r\ min}$ symbolizes the minimum received (discernible or detectable) power.

The noise source at the receiver input may be represented as a generator of voltage V_n and of internal resistance R_A (that of

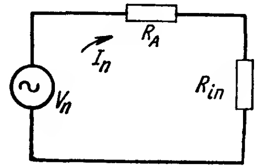


Fig. 14.1. Equivalent circuit of an aerial as a noise source

the dummy aerial used) connected to the load input resistance R_{in} as shown in Fig. 14.1. As will be recalled (see Sec. 12), power transfer from source to load is a maximum when the load resistance (in our case, R_{in}) is equal to the generator resistance (in our case, R_A), or

$$R_{in} = R_A$$

Then the noise source, like any generator, gives a maximum power of $V_n^2/4R_A$, which is the available noise power

$$P_{n\,av} = V_n^2/4R_A \tag{14.3}$$

Substituting the expression for V_n^2 from Eq. (10.1) in Eq. (14.3) gives

$$P_{n\,av} = \frac{4kTR_A\,\Delta f}{4R_A} = kT\,\Delta f \tag{14.4}$$

Substituting $P_{n\,av}$ for P_n in Eq. (14.2), we obtain

$$P_{r\,min} = NkT\,\Delta f \tag{14.5}$$

where Δf is the effective bandwidth of the receiver, $2\Delta F$.

Numerically, the effective power sensitivity P_{ef} is α_p times the available power sensitivity $P_{r\,min}$

$$P_{ef} = \alpha_p NkT\,\Delta f \tag{14.6}$$

The sensitivity of radar receivers may be expressed either directly in milliwatts or in decibels relative to the one-milliwatt level.

Depending on application and frequency range, the sensitivity of existing radar receivers is 10^{-9} to 10^{-11} milliwatts. By way of example, this power would be developed by a scale weight of 0.75 gram lowered at a rate of one millimetre per year.

Decibels relative to one milliwatt shows what fraction of one milliwatt the input signal is. For example, an input signal power $P_s = 10^{-8}$ milliwatts expressed that way will be

$$P_{s\,\mathrm{db}} = 10\log_{10}(1/P_s) = 10\log_{10}10^8 = 80\ \mathrm{db}$$

In the microwave region, the effect of external noise may be completely disregarded, and the available power sensitivity of radar receivers is decided by internal noise. This, coupled with the noisiness of conventional valve amplifiers, has led to the development of basically different types of amplifier, such as the travelling-wave valve type, parametric amplifiers, and quantum-mechanical amplifiers.

There are three categories of radar systems: *pulse*, *CW (continuous-wave)*, and *coherent-pulse* radars. The layout and arrangement of the circuits and components vary from type to type.

Review Questions

1. What electric quantity is used to specify the sensitivity of radar receivers in the SHF (centimetric) band?
2. Which of the sensitivities is defined by the minimum received (detectable or discernible) signal power?
3. Why can external interference be neglected in the microwave region?
4. How will an increase in the bandwidth of a radar receiver affect its sensitivity?

70. Pulsed-radar Receivers

Pulse (or pulsed) radar, which has been the basic one for years, is used to detect the presence of various objects, such as aircraft or ships, for plan-position presentation, and some other purposes.

In pulse radars, sharp bursts of radio energy are sent out. When these bursts, or *transmitted pulses*, encounter a reflecting object, they are reflected as discrete echoes which are detected by the radar receiver during the intervals between the transmitted pulses.

The fact that the echoes are picked up during the intervals between the transmitted pulses makes it possible to use the same aerial for both transmission and reception. Separation of the pulses in the aerial circuit is effected by a *duplexer*.

Most often, the receivers for pulsed-radar systems are superhets. The first unit within such a receiver is a TR (transmit-receive) tube which is part of the duplexer serving to decouple the receiver automatically from the aerial during the transmitting period.

The r.f. section of the receiver is arranged and uses the various stages according to the frequency of the received signals. In metric

(VHF) and decimetric (UHF) radars, the r.f. section contains an aerial-input circuit and an r.f. amplifier. In the UHF band, the aerial-input circuit and the r.f. amplifier use lighthouse tubes and resonant lines as the tuned circuits. At wavelengths shorter than 25 centimetres the first stage of the receiver usually is a crystal frequency changer, while the local oscillator is a reflex klystron. Incidentally, this band may as well use "straight" (TRF) receivers with an r.f. amplifier built around a travelling-wave tube. The most commonly used arrangement of radar receivers operating in the centimetric (SHF) band is shown in the block-diagram of Fig. 14.2. In this band, r.f. energy is conveyed by means of coaxial and waveguide transmission lines which are mechanically coupled to the TR tube and the frequency changer.

The i.f. amplifier usually has two separate sub-units, an i.f. preamplifier and a main i.f. amplifier. The reason for this separation is that the r.f. section of many radars, namely the magnetron r.f. pulse generator, the duplexer, and the mixer, is removed from the rest of the equipment to a distance of several metres, and the i.f. signal is fed to the i.f. amplifier over an r.f. cable. Both the crystal mixer and the r.f. cable attenuate the signal and introduce additional noise. If the signal were transmitted without preamplification, its signal-to-noise ratio would seriously deteriorate. Provision of a low-noise i.f. preamplifier between the crystal mixer and the r.f. cable improves the signal-to-noise ratio at the input of the main i.f. amplifier and the effective power sensitivity of the receiver.

The receiver also contains a separate AFC circuit. It has a mixer operating from the common local oscillator. The transmitter

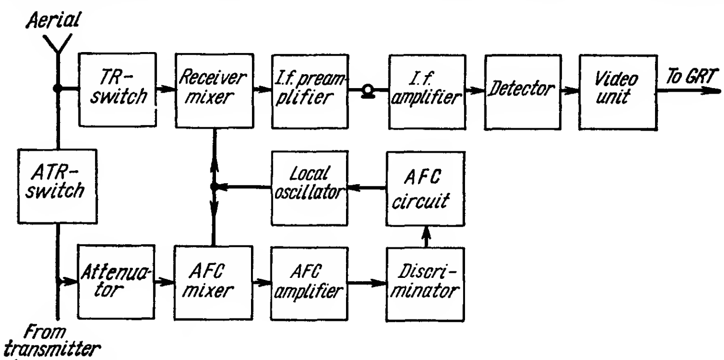


Fig. 14.2. Block-diagram of a pulse radar receiver using a separate AFC channel

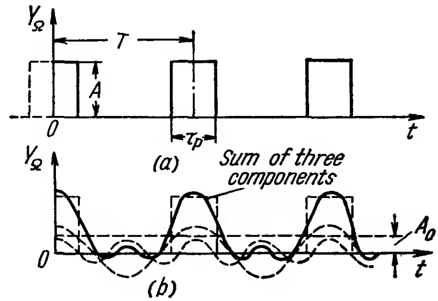


Fig. 14.3. Expansion of periodic d. c. pulses into a Fourier series

signal is coupled into the AFC circuit through a power divider or an attenuator. From the mixer output, the signal goes to the AFC amplifier and then to a frequency discriminator.

A very important aspect of radar-receiver design is the bandwidth.

The bandwidth of a radar receiver is determined by the frequency spectrum radiated by the transmitter and by the level of internal noise.

The transmitted radio-frequency pulses are produced by amplitude modulation of the carrier with rectangular d.c. pulses.

Figure 14.3*a* shows a train of such d.c. pulses. These "video pulses", as they are called, may be represented as a sum of sinusoidal and cosinusoidal components like any waveform. This is the statement of the *Fourier theorem*. If the waveform is repetitive, the components are discrete and may be represented as terms in a sum called the *Fourier series*.

With the rectangular pulses shown in Fig. 14.3*a* we can simplify the computation by placing the waveform symmetrically about the zero time axis. Then the function becomes "even", and all the sinusoidal terms reduce to zero, so that the Fourier series only contains the cosine terms:

$$Y_\Omega = A_0 + A_1 \cos \Omega t + A_2 \cos 2\Omega t + A_3 \cos 3\Omega t + \ldots + A_n \cos n\Omega t \ldots \quad (14.7)$$

where A_0 = d.c. component

$A_1, A_2, A_3, \ldots, A_n$ = amplitudes of the first, second, third and nth harmonics, respectively

$\Omega = 2\pi F$ = angular pulse repetition frequency, equal to the first-harmonic frequency.

This series may be shown graphically. If we represent all the components of the series as cosinusoids and add them together, the resulting curve will reproduce the shape of the pulse being studied. The curve representing the sum of the first three components of the series (Fig. 14.3*b*) approaches the rectangular pulses to a certain de‸ree.

In order to identify the part of the series (14.7) that contributes to acceptable reproduction of the pulse waveform, it is necessary to analyse the relation between the amplitude A_n of each harmonic and its frequency nF. This relation, $A_n = \varphi(nF)$, is called the *amplitude spectrum* of a rectangular video pulse and is described by the equation:

$$A_n = \frac{2A}{\pi n} \sin \pi\tau_p nF \tag{14.8}$$

where A is the height of the pulse.

Equation (14.8) is shown graphically in the plot of Fig. 14.4. As is seen, the amplitudes of the harmonics decrease sinusoidally with increasing harmonic order n and, consequently, harmonic frequency nF. That is, the amplitudes of the harmonics rise and fall with alternation of the sign so that the zeroes of the spectrum occur at frequency intervals

$$\pi\tau_p nF = \pi,\ 2\pi,\ 3\pi,\ \text{etc.}$$

corresponding to the following frequencies:

$$nF = \frac{1}{\tau_p},\quad \frac{2}{\tau_p},\quad \frac{3}{\tau_p},\ \text{etc.}$$

Thus, the spectrum of a repetitive video pulse is not continuous but is a sum of discrete harmonics whose frequencies are multiples of the fundamental frequency F.

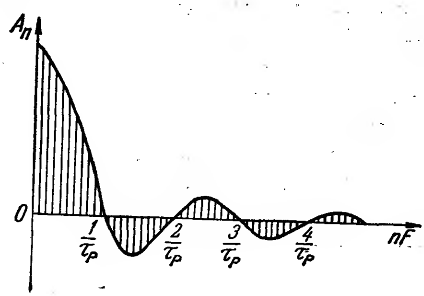


Fig. 14.4. Amplitude spectrum of a repetitive rectangular video pulse, composed of discrete harmonics

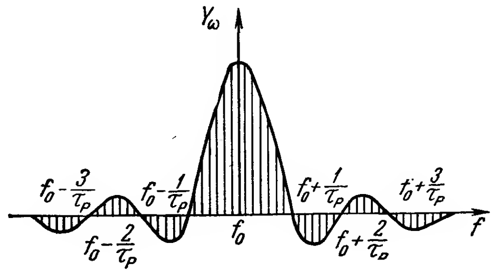


Fig. 14.5. The spectrum of rectangular radio pulses

It has been established that acceptable reproduction of a rectangular pulse will result if the bandwidth of the receiver extends out to the second zero of the spectrum $\left(\frac{1.5}{\tau_p}\right)$.

For an insight into the spectrum of r.f. (radio) pulse, suppose that in addition to the carrier frequency f_0, amplitude modulation causes each component of the video pulse of frequencies F, $2F$, $3F$, ..., nF to produce two side frequencies

$$f_0 \pm F;\ f_0 \pm 2F;\ f_0 \pm 3F;\ \ldots;\ f_0 \pm nF$$

Graphically, the spectrum of a radio pulse consists of two symmetrical halves, each representing the spectrum of a video pulse (Fig. 14.5). Thus, a radio pulse has a spectrum twice as wide as that of a video pulse. In other words, acceptable reproduction of an r.f. pulse will be obtained if the receiver bandwidth is $2\times\frac{1.5}{\tau_p}$. That is,

$$2\Delta F = \frac{3}{\tau_p} \tag{14.9}$$

Or, graphically, the bandwidth should extend out to the third zero of the spectrum.

Reproduction of the pulse waveform is not the only factor controlling the choice of the receiver bandwidth. An equally important consideration is that the noise level at the output increases and the effective sensitivity of the receiver decreases as the bandwidth is made broader. It has been proved that the highest signal-to-noise ratio at the receiver output is obtained with an optimum bandwidth given by

$$2\Delta F_{opt} = \frac{1.37}{\tau_p} \tag{14.10}$$

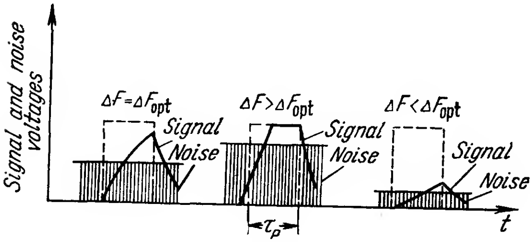


Fig. 14.6. Effect of bandwidth on the pulse shape and signal-to-noise ratio

Research work has shown that the optimum bandwidth is

$$\underset{(\text{MHz})}{2\Delta F_{opt}} = \frac{1.0 \text{ to } 1.37}{\underset{(\mu\text{sec})}{\tau_p}} \tag{14.11}$$

depending on the shape of the resonance curve of the receiver.

Figure 14.6 shows in approximate terms the effect the bandwidth has on the pulse waveform and the signal-to-noise ratio at the output of the receiver. The best signal-to-noise ratio is secured with the optimum bandwidth. In this case, the signal amplitude approaches that of a rectangular pulse. If the bandwidth exceeds its optimum value, the pulse waveform is improved, but the signal is more difficult to detect on the screen because of the higher noise level. With a bandwidth less than the optimum one, neither the signal waveform nor the signal-to-noise ratio meet the specified requirements.

The bandwidth of radar receivers usually extends out to the fifth zero on the spectrum plot, or

$$2\Delta F = \frac{1 \text{ to } 5}{\tau_p} \tag{14.12}$$

depending on the intended application of a given radar system. In precision height- and distance-finding radars, the bandwidth is broader than, for instance, in early-warning radars.

Review Questions

1. Is it necessary to use an i.f. preamplifier in receivers with an r.f. amplifier?
2. Which frequency is controlled by AFC?
3. Why is the frequency spectrum of a radio pulse twice as wide as that of a video pulse?

4. How does a decrease in the pulse duration affect the spectrum of the radio pulse?
5. What should the bandwidth of a pulsed-radar receiver be to secure a maximum signal-to-noise ratio?

71. CW Radar Receivers

In CW (continuous-wave) radar systems, a continuous flow of radio energy is sent out. The reflected wave is distinguished from the outgoing signal by a slight change in radio frequency. This change in frequency can arise from the relative motion between the radar and the reflecting object *(Doppler radar)* or it may be imposed on the transmitted wave by frequency modulation *(FM radar)*.

Doppler radar uses what is known as the Doppler effect which consists in that motion of the target relative to the radar causes an apparent change in the frequency of the reflected wave by

$$F_d = f_t 2u_r/c \tag{14.13}$$

called the *Doppler-shift frequency*. In Eq. (14.13), f_t is the frequency of the outgoing signal, u_r is the radial component of the velocity of the target, and c is the velocity of light. As is seen, the Doppler-shift frequency is a function of the speed of the target and may vary within broad limits.

In a Doppler-radar receiver, this frequency is detected as a difference frequency (beats) at the output of a non-linear circuit (a detector) by mixing the reflected signal and the leakage signal (from the transmitter).

The appearance of a Doppler-shift frequency (usually as an audio output) points to the presence of a moving target in the area illuminated by the radar. The value of the Doppler-shift frequency gives a measure of the speed of the target.

Doppler radar has many and varied uses. In contrast to it, FM radar is employed mainly in distance- and altitude-measuring systems. In FM radar, the radiated wave is frequency-modulated. The range (or altitude) is measured by beating the returning wave with the one being radiated.

Frequency modulation is effected by causing the transmitted signal to sweep in frequency over a small range about the carrier value. The reflected wave retains the frequency with which it left the transmitter and returns to the receiver only after the frequency of the transmitter has changed by an appreciable amount.

This amount can be detected as a beat frequency and indicated aurally or otherwise, being proportional to the range (or altitude) of the target

$$F_r = 4r\Delta f F_M/c \tag{14.14}$$

where Δf is the swing of the modulating frequency, F_M is the modulating frequency, c is the velocity of light, and r is the distance of the target.

FM radar can be used to detect and determine the range of fixed and moving targets. In the former case, the FM radar has found its widest use as an absolute altimeter for aircraft. In the latter case, the equipment develops both a range-shift and a Doppler-shift frequency, F_r and F_d, so that both the distance and speed of targets can be conveniently measured.

Since the transmitter and receiver of a CW radar are operating simultaneously and continuously, it is impractical to have a common aerial system. Usually two similar structures are used. As a result, the receiver does not incorporate a TR-tube, indispensable in the duplexer when a single aerial is utilized for both transmission and reception.

Figure 14.7 shows a block-diagram for a CW radar. The first mixer accepts the outgoing (leakage) signal of frequency f_t, while the second mixer accepts the returning signal of frequency f_s, differing from f_t by the Doppler-shift frequency F_d. At the same

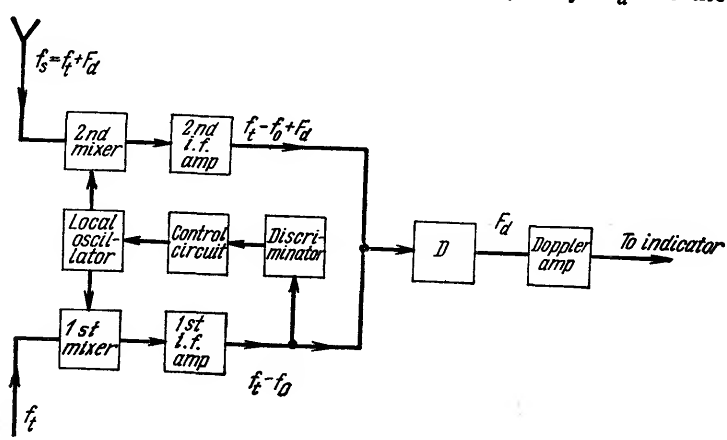


Fig. 14.7. Block-diagram of a CW radar receiver

time, the two mixers are fed with the signal of frequency f_o from klystron local oscillator.

The difference frequencies, f_t-f_o and $f_t-f_o+F_d$, appearing in the output of the two mixers are amplified in two i.f. amplifiers and are then applied to a detector, D. The output of the detector contains a difference frequency equal to F_d. The Doppler-shift frequency is then amplified by a Doppler-shift frequency amplifier, and is applied to a frequency-meter. There is also an AFC circuit to control the frequency of the local oscillator so that the difference frequency f_t-f_o is held unvarying. The AFC circuit comprises a discriminator and a control circuit.

The arrangement shown in the block-diagram of Fig. 14.7 may also be used in FM radar equipments.

The bandwidth of the output amplifier in Doppler radar receivers is decided by the range of Doppler-shift frequencies and may be from several to tens of kilohertz. The bandwidth of the r.f. section in such a receiver is made broader by the frequency drift of the transmitter and local oscillator.

Review Questions

1. How does an increase in the speed of the target affect the Doppler-shift frequency?
2. Why is the modulating frequency varied linearly in FM radar?
3. What will happen to the frequency-meter if the AFC circuit were omitted from the receiver shown in the block-diagram of Fig. 14.7?

72. Coherent-pulse Radar Receivers

The coherent-pulse system is the basis of moving-target indicators (MTI) in which the display of radar information is limited primarily to moving objects.

In fact, moving targets might be detected and ranged by the continuous-wave Doppler radar discussed in the previous section. However, the equipment involved would be prohibitively complicated. A way out is offered by the coherent-pulse method which is a combination of pulsed and Doppler techniques.

When a moving target is illuminated with r.f. pulses, the Doppler-shift frequency cannot be directly detected as the difference in frequency between the outgoing and returning waves

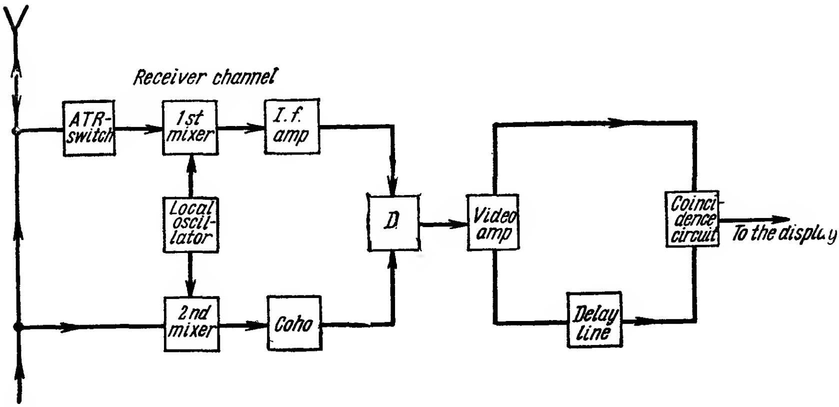


Fig. 14.8. Block-diagram of a coherent-pulse radar receiver

because the transmitted pulse is very short. Instead, one measures the phase shift between the returning pulse and the continuous wave generated in the receiver by a local oscillator. Thus, the signal of the local oscillator serves as a reference which is synchronized with the r.f. pulse generator. An oscillator whose phase is locked to the phase of another oscillator is called *coherent* ("Coho"), and the method is called coherent-pulse operation.

A simplified block-diagram of a coherent-pulse radar receiver is shown in Fig. 14.8. The r.f. section comprises a receiver channel and the coherent-oscillator (Coho) channel. The former controls the phase of the returning signal, and the latter the phase of the reference wave, locked to the phase of the transmitter oscillator.

The shift in phase between the outgoing and incoming signals is detected at an intermediate frequency which is obtained separately in the receiver and Coho channels. In the former, the i.f. is obtained by heterodyning the returning signal, and in the latter by heterodyning the leakage (transmitted) signal in the respective mixers, with the local wave furnished by a local oscillator.

The coherent local oscillator, Coho, generates the intermediate frequency whose phase, as already noted, is locked to that of the r.f. pulse oscillator in the transmitter.

Owing to the difference in phase between the returning echoes and the steady output of the Coho, beats result in the output of the detector, with an amplitude dependent on the phase difference.

It is important to note that the echoes from stationary and moving targets differ markedly. For stationary targets, the phase difference is fixed, and the signal at the detector output will be steady. For moving targets, the phase difference is changing, and a rapidly fluctuating (amplitude-modulated) signal appears.

In the arrangement shown in the block-diagram of Fig. 14.8, the echoes from moving targets are displayed as follows. The output from the detector is fed to a video amplifier, from which the output is divided between a delay line and a coincidence circuit. In the delay line, the output is stored for the duration of a pulse period and is reversed in polarity. In the coincidence circuit, the echoes from two sequential pulses are mixed. Those due to stationary targets cancel each other completely, while those due to a moving target will not be similarly cancelled, and are transmitted to the display.

Review Questions

1. Define a coherent signal.
2. How is the phase shift detected in the coherent-pulse method?
3. Why does the phase difference for moving-target echoes remain fixed?

73. Microwave Valves

For proper understanding of radar, it is essential to elucidate the factors limiting the use of conventional valves in the microwave region. These factors, which may safely be neglected at the lower frequencies are parasitic energy-storage circuit elements, such as interelectrode capacitance and lead inductance, dielectric losses in the envelope and base, and finite transit time of electrons between electrodes.

Let us dwell on lead inductance. As will be recalled, inductance is a property of single conductors and not only of coils. Treating an electrode lead as a straight conductor, its inductance is given by

$$L = 2l\left(\ln\frac{4l}{d} - 1\right)\times 10^{-3} \text{ microhenrys}$$

For example, a lead 1.5 centimetres long and 0.1 centimetre in diameter has an inductance of about 0.01 microhenry.

At lower frequencies, the inductive reactance X_L of valve leads is so small as to be neglected. In the UHF band, X_L becomes considerably larger, as seen from the following numerical comparison:

at $f = 3{,}000$ kilohertz:

$$X_L = 2\pi fL = 6.28 \times 3{,}000 \times 10^3 \times 0.01 \times 10^{-6} = 0.19 \text{ ohm}$$

at $f = 3{,}000$ megahertz:

$$X_L = 2\pi fL = 6.28 \times 3{,}000 \times 10^6 \times 0.01 \times 10^{-6} = 190 \text{ ohms}$$

The role played by interelectrode capacitance and lead inductance can be understood from reference to the circuit diagram of an ordinary single-stage amplifier. Figure 14.9 shows only the inductance of the cathode lead, L_k because the inductances of the other leads do not materially affect amplifier operation. The input to the amplifier is the circuit across *og*. This is, in effect, a voltage divider comprised of series-connected capacitive reactance X_{gk} and inductive reactance X_L. The actual voltage V_{gk} applied between the cathode and grid will be determined by the ratio of these reactances. As the frequency increases, the capacitive reactance X_{gk} will decrease and the inductive reactance X_L will increase, and the voltage applied to the grid will be considerably less than the input voltage V_g.

Putting $C_{gk} = 2$ picofarads, then at $f = 3{,}000$ kilohertz the capacitative reactance will be

$$X_{gk} = \frac{1}{\omega C_{gk}} = \frac{1}{6.28 \times 3{,}000 \times 10^6 \times 2 \times 10^{-12}} \cong 27 \text{ ohms}$$

Taking $X_L = 190$ ohms, the voltage applied directly to the grid will be $V_{gk} \approx 0.15\, V_g$. At a higher frequency, the valve input (*gk*) will be, in effect, short-circuited, and the valve will no longer act as an amplifier.

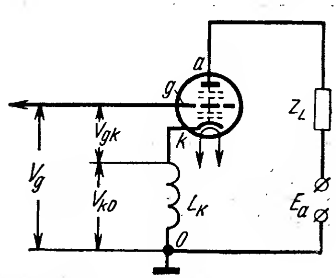


Fig. 14.9. Circuit explaining the influence of cathode lead inductance upon operation of a valve

Now consider the effect of the transit time of electrons.

At the lower frequencies, including short waves, the period of oscillation applied to the grid is many times the transit time of electrons from cathode to anode. Therefore, a fast change in the grid voltage causes a similar fast change in the anode current.

As the frequency is increased, the period of oscillation becomes comparable with the transit time, and the anode current no longer follows the grid voltage instantaneously.

The transit time of electrons in a valve is determined by the spacing between the electrodes and the voltages at which they operate. Electrons travel through a valve with a varying speed. In the cathode-grid space, where the grid voltage is insignificant, the speed of electrons is comparatively low. Past the grid, the electrons travel with a much higher speed due to the relatively high screen-grid and anode potentials. Thus, the electrons spend most of their transit time on covering the cathode-grid space, and this time may be considered, with sufficient approximation, to be the transit time of electrons in the valve.

For valves with cylindrical electrodes the transit time τ can be determined from the equation

$$\tau = 0.255 \times 10^{-7} \frac{R}{\sqrt{V_g}}$$

With a grid radius $R = 0.2$ centimetre, and with a grid voltage $V_g = 2.5$ volts, the transit time is

$$\tau = 0.255 \times 10^{-7} \frac{0.2}{\sqrt{2.5}} \cong 3.3 \times 10^{-9} \text{ second}$$

It is easy to calculate that a wavelength of 10 metres corresponds to a period of $T = 33 \times 10^{-9}$ second which is ten times the transit time of electrons, and the anode current begins to lag in phase behind the grid voltage.

With a wavelength of 1 metre, the transit time and the period of oscillation are equal. In this case the transit time is such that the electrons reach the grid out of phase (in reverse polarity), and are thrown back to the cathode. It is obvious that on this and even on slightly longer wavelength the valve loses its amplifying properties.

The most important aspect of this transit-time effect is that it causes a decrease in the input resistance of the valve.

Review Questions

1. How is lead inductance affected by an increase in lead diameter?
2. Which space in a valve accounts for the greater part of the transit time?
3. Why does the transit-time effect become more pronounced with decrease of wavelength?

74. Input Resistance of Valves at Microwave Frequencies

In Section 33 it has been shown that the input resistance of valves operating on long, medium and, partly, short waves and loaded into a resistance is close to infinity, so that the shunting effect of the valve on the tuned circuit of the preceding stage can be neglected. As will be shown shortly, in the microwave region the input resistance is low and, therefore, affects markedly the operation of the preceding stages.

Let us examine the effect of the cathode lead inductance on the input resistance of the valve in the microwave region. To begin with, we plot a vector diagram, using the circuit of Fig. 14.9. We also assume that the anode load impedance is resistive: $Z_L = R$.

Noting the diagram of Fig. 6.16 (see Sec. 33), plotted for resistive load, we conclude that the grid current I_g actually leads the grid voltage V_{gk} by 90° (Fig. 14.10).

If the total resistance of the anode circuit $(R_a + R)$ is many times the inductive reactance of the cathode lead X_L, the anode current I_a of the valve is resistive and is in phase with V_{gk}. The voltage across the cathode-lead inductance V_{k0} leads the anode current by 90°. The input voltage is

$$\dot{V}_g = \dot{V}_{gk} + \dot{V}_{k0}$$

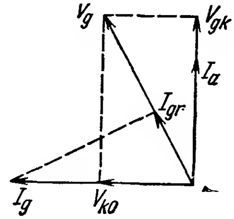


Fig. 14.10. Vector diagram for determining the input resistance of a valve, taking into account the effect of cathode lead inductance

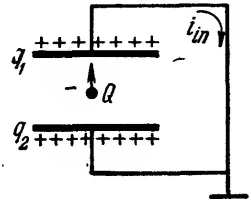


Fig. 14.11. Induction of current by moving charges

On the diagram of Fig. 14.10, the phase difference between V_g and I_g is less than 90°. Accordingly the resistive component of the grid current, I_{gr}, as a projection of the vector I_g on the vector V_g, has a finite value, and the input resistance R_{in} ceases to be infinity.

Consider the effect of the finite transit time on the input resistances of a valve.

At the lower frequencies, it is convenient to think of the currents in the valve circuits as flowing only when electrons actually arrive at the electrodes. In the microwave region, one is forced to take into account so-called *induced currents*.

According to the law of electrostatic induction, a point charge Q, located mid-way between two short-circuited plates (Fig. 14.11), induces in them similar image charges q_1, q_2, which are opposite in polarity to Q, and whose sum is

$$|q_1+q_2|=|Q|$$

When the charge Q is moved towards one of the plates, the value of induced charges changes: one increases while the other decreases. As a result, an induced current i_{in} begins to circulate around the external circuit.

Let us see how such a current is induced in the grid circuit of a valve (Fig. 14.12).

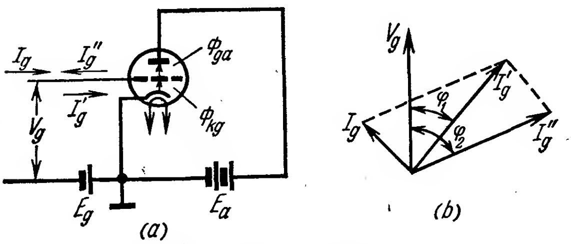


Fig. 14.12. Circuit and vector diagram for determining the input resistance of a valve, taking into account the transit time of electrons

When the grid is negative, the electrons flying between the grid wires induce electric charges on the grid. The electrons moving from cathode to grid build up the induced charge on the grid, and an induced current is flowing in the grid circuit in one direction. As the electrons move from grid to anode, the grid charge decreases, and there appears an induced grid current flowing in the opposite direction. The resultant grid current is then the difference between these induced currents.

With constant voltages across the valve electrodes, when the electron stream is the same at any cross-section of the valve, the resultant grid current is zero because the induced current, caused by the electron stream flowing to the grid, will be equal to the induced current due to the electron stream leaving the grid.

Consider the case where the grid accepts an alternating voltage of a frequency such that the transit time of electrons may be disregarded in comparison with the period of oscillation. The space charge at the cathode releases electrons as discrete sheets moving to the grid in step with variations in the grid voltage. Each electron sheet covers the cathode-anode space so rapidly that the instantaneous grid voltage has no time to change. As a result, the grid currents induced by each sheet of electrons approaching the grid and departing from it are equal, and there is no resultant grid current.

At such a frequency it may be considered that the electron stream at any cross-section of the valve, from cathode to grid, remains in phase with the grid voltage.

At microwave frequencies each electron sheet covers the cathode-anode space during a considerable part of the r.f. period so that the instantaneous grid voltage has time to change. Therefore, at each cross-section of the valve, from cathode to grid, the electron stream lags behind the grid voltage in phase. As the stream moves farther away from the cathode, the phase difference increases, reaching its maximum near the grid.

To plot a vector diagram, consider the two electron streams shown by the arrows in Fig. 14.12*a*. Assume that the stream Φ_{kg}, approaching the grid, lags behind V_g by an angle φ_1, equal to the average phase difference in the cathode-grid space. The other stream, Φ_{ga}, which is moving away from the grid, lags behind V_g by an angle φ_2 which is the same across the grid-anode region (the transit time of electrons from grid to anode can be neglected because of the high voltage applied to the anode). The angle φ_2

represents the maximum lag of the electron stream right at the grid: $\varphi_2 > \varphi_1$.

In the vector diagram of Fig. 14.12*b*, V_g is the datum vector. The current I'_g induced in the grid circuit (by the approaching electron stream Φ_{kg}) lags behind V_g by φ_1. The stream Φ_{ga} departing from the grid gives rise to the grid current I''_g, which lags behind V_g by φ_2.

The resultant grid current is

$$\dot{I}_g = \dot{I}'_g - \dot{I}''_g$$

As is seen from the diagram, I_g leads V_g by an angle smaller than 90°, which is another way of saying that there is a decrease in the resistance R_{in}.

To sum up, the input resistance of a conventional valve operating at microwave frequencies is reduced by both parasitic energy-storing circuit elements and the finite transit time.

The input resistance of the valve is given by

$$R_{in} = \frac{K}{f^2} \tag{14.15}$$

Values of K for some of the Soviet-made valves are given in Table 14.1. In using the data of Table 14.1 and Eq. (14.15), R_{in} should be expressed in megohms and the frequency in megahertz.

TABLE 14.1.

Valve type	Coefficient K, megohms×MHz^2
6С1П, bantam triode	160
6С5Д, triode	200
6Н3П, bantam dual triode	200
6Н14П, same	100
6Ж1П, bantam screen-grid sharp-cutoff pentode	70
6Ж2П, same	36
6Ж3П, same	37
6Ж1Б, same, miniature	62

By way of example, R_{in} of the conventional 6K3 pentode is 5 megohms at $f = 2$ megahertz ($\lambda = 150$ metres). In this case,

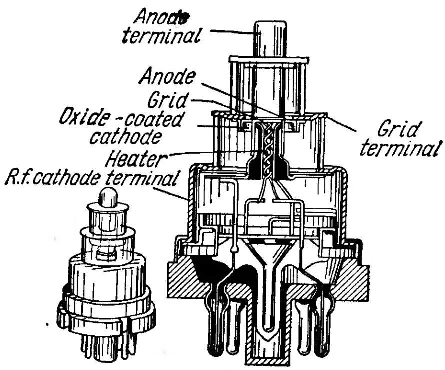


Fig. 14.13. External view and cross-section of a lighthouse valve

the effect of R_{in} on the operation of the amplifier may be neglected. If we increase the frequency to 200 megahertz ($\lambda = 1.5$ metres), the input resistance will be $R_{in} = 500$ ohms. With such values of the input resistance, valves of the conventional design cannot be used at microwave frequencies because the valve input would be almost short-circuited. Instead, special types of valves are used.

In acorn and bantam valves the interelectrode capacitance is reduced because the electrodes are very small. The inductance of the electrode leads is decreased owing to the absence of valve bases, and also owing to the short and thick leads used in these valves. Their electron transit time is also reduced due to the closer electrode spacing.

At frequencies higher than 500 megahertz special triodes with disc terminals are employed. These valves are known as *disc-seal* or *lighthouse* valves. An overall view and cross-section of such a valve are shown in Fig. 14.13. In this valve, the anode, grid, and cathode are brought out to discs sealed into, and separated by, glass cylinders. The discs fit onto lengths of coaxial lines used as tuned circuits. In this way, the grid-cathode and grid-anode capacitances are incorporated into the resonant circuit, and their stray effect is minimized. The same is true of the lead inductance which is insignificant with disc terminals. The electrode spacing is negligible. Disc-seal or lighthouse valves may be used at frequencies up to 3,000 megahertz.

Review Questions

1. Why does a decrease in the input resistance of a valve show as a decrease in the phase angle between the vectors V_g and I_g?
2. What features in valve construction minimize the transit-time effects?
3. How is the lead inductance minimized in disc-seal valves?

75. Tuned Circuits of Radar Receivers

The use of conventional tuned circuits consisting of inductors and capacitors in radar receivers is limited to frequencies under 300-350 megahertz. In the decimetre and, even more so, the centimetre bands tuned circuits of a different kind have to be employed. Tuned circuits in which the capacitance C is lumped in a capacitor, and the inductance L in a coil, the so-called *lumped-constant* tuned circuits, are unsuitable.

Let us see why such lumped-constant circuits cannot be used at microwave frequencies.

The resonant frequency of a resonant circuit is given by

$$\omega_0 = \frac{1}{\sqrt{LC}}$$

and its Q factor, by

$$Q = \frac{\sqrt{\frac{L}{C}}}{R}$$

As follows from these equations, an increase in frequency may be obtained by reducing the inductance or capacitance of the tuned circuit. However, it is more advantageous to reduce the capacitance, because this will improve the Q of the tuned circuit.

The capacitance of a tuned circuit is made up of the capacitance of the capacitor and the shunt capacitance C_s, defined as

$$C_s = C_{out} + C_{in} + C_w$$

where C_{out} = valve output capacitance
C_{in} = input capacitance of the next valve
C_w = capacitance of the wiring.

The minimum capacitance to which the tuned-circuit capacitance can be reduced is C_s. Further increase in frequency cannot be obtained by a reduction in capacitance; it is necessary to decrease the inductance. The reduction in the coil inductance involves a reduction in the number of coil turns. With a wavelength of one metre the coil consists only of one turn.

A circuit in which a physical capacitor is absent and the coil is represented by one turn is no longer a lumped-constant tuned circuit. The electromagnetic field in it is distributed over the entire tuned circuit, including the connecting wires along which small capacitances and inductances are distributed. The fact that electromagnetic energy is not limited to the coil and capacitor, but is distributed along the entire circuit whose dimensions are comparable with the wavelength, results in electromagnetic radiation and in increased radiation-loss resistance.

Thus, a decrease in the tuned-circuit inductance lowers the Q of the circuit, firstly, because the Q is directly proportional to inductance and secondly, due to an increase in radiation-loss resistance. This is why distributed-constant tuned circuits are used at microwave frequencies. Such tuned circuits are segments of parallel-wire transmission lines or coaxial lines. This type of tuned circuit has found wide application in the decimetre band and is used in the radio-frequency amplifiers and aerial-input circuits of radar receivers.

Still better performance, particularly in the centimetre range, is offered by a special kind of tuned circuits known as *cavity resonators*. They are used on frequencies higher than 3,000 megahertz.

A cavity resonator is a metallic chamber in which oscillating electromagnetic energy is generated and stored. By making the chamber of proper dimensions, it is possible to give the circuit a very high Q at microwave frequencies—thousands or even tens of thousands, which is much higher than that of tuned circuits using segments of coaxial lines.

Cavity resonators owe their high Q to the absence of an external field and, as a consequence, reduced radiation loss.

Cavity resonators are simple and robust in design, a factor enhancing their stability, although they come in a variety of shapes, such as toroids, cylinders, parallelepipeds and cubes. Radar receivers use cavity resonators in the aerial-input circuits, local oscillators and frequency changers.

The resonant frequency of cavity resonators is determined by their shape, dimensions and method of excitation. At frequencies above 1,000 megahertz cavity resonators measure only several centimetres in length and width.

The resonant frequency of a cavity resonator can be changed with adjusting screws which protrude into the cavity, or by changing the shape of the cavity.

Electromagnetic energy is coupled in and out of cavity resonators by means of either coupling loops or probes.

Review Questions

1. What types of lines can be used as tuned circuits?
2. Why do cavity resonators have a high Q?
3. How can the resonant frequency be changed in cavity resonators?

76. Aerial-input Circuits in the VHF and UHF Bands

The choice of the aerial-input circuit for any frequency range is mainly decided by the amount of coupling of the aerial to the input tuned circuit. Let us see what factors determine the aerial coupling at microwave frequencies.

The aerial-input circuit, being the first element of a radio receiver, controls the noise quality of the receiver more than any other stage. Therefore, the aerial-input circuit must, above all, secure a maximum signal-to-noise ratio. With loose aerial coupling, the signal-to-noise ratio is insignificant. With tight coupling approaching the optimum one the signal becomes a maximum, while the noise voltage remains almost unchanged.

The noise voltage developed in the aerial-input circuit depends on its resonant resistance. For optimum coupling, the aerial resistance reflected into the tuned circuit must be equal to the resistance of the tuned circuit. Thus, the choice of coupling is governed by matching the aerial to the input tuned circuit. While on long, medium and short waves, simple untuned aerials are used, those for microwave frequencies are of the special tuned type, with energy transferred from them over transmission lines.

In the VHF band, the transmission lines are of either the parallel-wire or the coaxial type. The high-frequency end of the VHF band and the UHF band use coaxial feeders. In the SHF band energy is conveyed from the aerial to the receiver by waveguides.

The primary function of a transmission line is to convey r.f. energy with a minimum loss. One of the basic types of losses in transmission lines is that due to radiation. This loss occurs

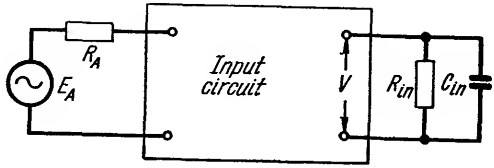


Fig. 14.14. Equivalent circuit of the aerial-input circuit

only when standing waves are present. Hence, a very important aspect of transmission-line design is the reduction of standing waves, or provision of the travelling-wave mode. This is done by proper construction, termination, and adjustment of the line.

To begin with, the line must be short. Next, it must be terminated into an impedance equal to its wave impedance. On the other hand, its input impedance should be matched to the output impedance of the aerial, so that

$$R_A = \rho = R_{in}$$

In selecting the configuration for the aerial-input circuit it is important to take into account the considerable shunting effect of the input resistance of the next stage, because it impairs the properties of the tuned circuit. Figure 14.14 shows an equivalent circuit for the circuits preceding the first receiver stage. The aerial is represented by a generator of emf E_A and of internal resistance R_A. The input resistance R_{in} and the input capacitance C_{in} of the next stage (valve amplifier or mixer) serve as the input-circuit load.

The aerial-input circuit must be properly matched at both ends. The equivalent circuit shows only the resistance of the aerial, although in the general case it is an impedance. This simplification is acceptable when the resonant frequency of the aerial is equal to the frequency of the incoming signal (a tuned aerial). Under such conditions, the aerial reactance is zero. Even if these conditions are not fulfilled, the reactive component of the aerial impedance can be compensated for by adjusting the input tuned circuit.

Let us determine the gain of a perfectly matched aerial-input circuit.

According to Eq. (14.3), the maximum power delivered by a generator to a matched load is

$$P_{max} = \frac{E_A^2}{4R_A} \qquad \text{(a)}$$

The power delivered by the aerial will be dissipated by the effective resistance of the input tuned circuit R_{0e} and the input resistance R_{in} of the next stage. Let the total load resistance, that is, the resistance of the tuned circuit shunted by R_{in}, be R'_{0e}. Then

$$P_{max}=\frac{V^2}{R'_{0e}} \tag{b}$$

where V is the voltage at the circuit output (see Fig. 14.14).

Equating (a) and (b) gives

$$\frac{E_A^2}{4R_A}=\frac{V^2}{R'_{0e}} \tag{c}$$

By definition, the voltage gain of the input circuit is

$$K=\frac{V}{E_A}$$

Noting Equality (c), the maximum gain of the input circuit is

$$K_{max}=\frac{1}{2}\sqrt{\frac{R'_{0e}}{R_A}} \tag{14.16}$$

The transmission line can be matched to the input curcuit by use of either transformer or tapped-coil coupling. The former is used with balanced transmission lines, the latter with unbalanced coaxial lines.

Transformer-coupled Aerial-input Circuits. Transformer-coupled aerial-input circuits may have

(a) an untapped secondary,
(b) a tapped-down secondary.

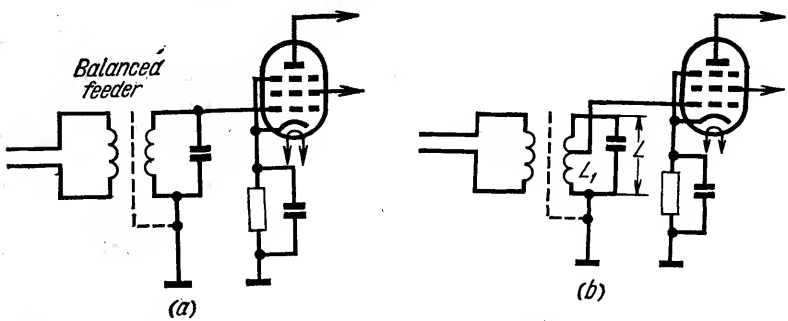


Fig. 14.15. Input circuits using transformer coupling
(a) untapped secondary; (b) tapped-down secondary

A transformer-coupled circuit with an untapped secondary is shown in Fig. 14.15*a*. An electrostatic screen (shown by the dotted line) decreases the capacitive coupling between the coils. The inductance, L_c, of the primary and the coefficient of coupling k necessary for impedance matching are given by

$$\left.\begin{aligned} L_c &= \frac{R_A}{\omega} \\ k &= \sqrt{\frac{2}{Q}} \end{aligned}\right\} \tag{14.17}$$

The Q of the input tuned circuit is the one determined by the loss resistance coupled in by the input resistance R_{in} of the next stage. At microwave frequencies, the coupled-in, or reflected, resistance Δr is many times the tuned-circuit loss resistance r. Therefore,

$$Q = \frac{\rho}{\Delta r}$$

where

$$\Delta r = \frac{\rho^2}{R_{in}}$$

Hence,

$$Q = \frac{R_{in}}{\rho} \tag{14.18}$$

This expression gives only the Q of the input tuned circuit without allowance for the resistance coupled in from the aerial circuit. At optimum coupling, the aerial circuit reflects the same value of resistance into the tuned circuit as the input of the next stage. Therefore, the loss resistance in the tuned circuit doubles, and the loaded Q is halved

$$Q_{ef} = \frac{Q}{2} = \frac{R_{in}}{2\rho} \tag{14.19}$$

Eq. (14.19) may be used for determining the selectivity and bandwidth of the aerial-input circuit for optimum coupling.

In accordance with Eq. (14.16), the maximum gain of the aerial-input circuit using untapped transformer coupling to the transmission line is given by

$$K_{max} = \frac{1}{2}\sqrt{\frac{R_{in}}{R_A}} \tag{14.20}$$

The resistance

$$R'_{0e} = \frac{R_{0e}R_{in}}{R_{0e} + R_{in}} \cong R_{in}$$

because

$$R_{in} \ll R_{0e}$$

A transformer-coupled circuit with a tapped-down secondary is shown in Fig. 14.15*b*. When the tuned circuit is coupled to the valve grid in this way the effect of R_{in} and C_{in} on the input circuit is reduced.

The new values of R'_{in} and C'_{in} are related to their values in the previous case thus:

$$\left.\begin{aligned} R'_{in} &= \frac{R_{in}}{p_g^2} \\ C'_{in} &= p_g^2 C_{in} \end{aligned}\right\} \tag{14.21}$$

where $p_g \cong \frac{L_1}{L}$ is the tapping-down ratio on the grid side.

For such a circuit, the gain K at optimum coupling is given by Eq. (14.20). The loaded or effective Q, with allowance for the coupled-in resistance, is found from Eq. (14.19)

$$Q_{ef} = \frac{R'_{in}}{2\rho}$$

Noting Eq. (14.21), we obtain

$$Q_{ef} = \frac{R_{in}}{2p_g^2\rho} \tag{14.22}$$

From Eq. (14.22) it follows that the effective or loaded Q of a tapped-down circuit may be considerably increased as compared with the untapped-down circuit by reducing the value of p_g.

The maximum gain is secured by matching k and p_g.

Example 14.1. Find the maximum gain and the effective Q of an aerial-input circuit for: $R_A = 75$ ohms, $R_{in} = 2{,}000$ ohms, $L = 0.3$ microhenry, and $f = 50$ megahertz.

Solution. The maximum gain is

$$K_{max} = \frac{1}{2}\sqrt{\frac{R_{in}}{R_A}} = \frac{1}{2}\sqrt{\frac{2{,}000}{75}} = 2.6$$

To find the effective Q, determine ρ:

$$\rho = 2\pi f L = 6.28 \times 50 \times 10^6 \times 0.3 \times 10^{-6} = 94 \text{ ohms.}$$

The effective Q for an untapped tuned circuit ($p_g = 1$) is

$$Q_{ef1} = \frac{R_{in}}{2\rho} = \frac{2{,}000}{2 \times 94} \cong 10$$

For $p_g = 0.5$

$$Q_{ef2} = \frac{R_{in}}{2p_g^2\rho} = \frac{2{,}000}{2\times 0.5^2\times 94} = 40$$

As is seen from Example 14.1, the effective Q can be varied within broad limits by adjusting p_g.

Tapped-coil Coupled Input Circuit. The inner conductor of the coaxial transmission line (Fig. 14.16) is connected to a tap on the coil L, while the outer conductor is earthed.

In this circuit optimum coupling is obtained by selecting a suitable tap on the tuned-circuit coil. As in the circuit of Fig. 14.15, the new value of R_A' is related to its value in an untapped circuit as

$$R_A' = \frac{R_A}{p_A^2} \tag{14.23}$$

where $p_A \simeq \frac{L_A}{L}$ is the tapping-down ratio of the coil on the aerial side.

The maximum gain K_{max} of the tapped-coil coupled circuit is found from Eq. (14.20). In this case the optimum tapping-down ratio is given by

$$P_{A\,opt} = \sqrt{\frac{R_A}{R_{in}}} \tag{14.24}$$

The effective Q_{ef} of the loaded tuned circuit at optimum coupling is given by Eq. (14.19).

Aerial-input Circuit with a Resonant Line. The tuned circuit in the arrangement of Fig. 14.17 is a segment of a coaxial line a quarter of a wavelength long, short-circuited at the far end.

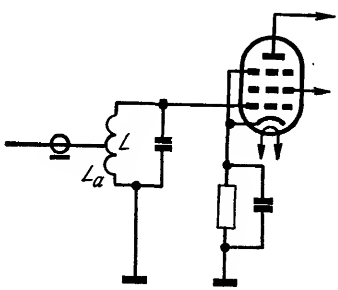


Fig. 14.16. Aerial-input circuit with tapped-coil coupling

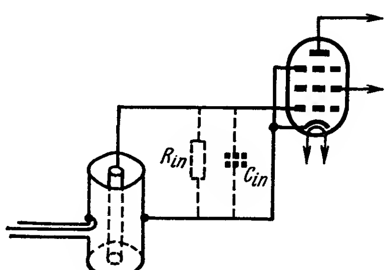


Fig. 14.17. Aerial-input circuit with resonant line

The line input is connected to the grid circuit of the valve and, consequently, is loaded into R_{in} and C_{in}. Coupling to the transmission line is by a coupling loop.

Aerial-input circuits using resonant lines are employed in UHF radar receivers. Physically, the aerial-input circuit in UHF radars is a self-contained unit consisting of a lighthouse valve, segments of coaxial lines of the input and anode tuned circuits.

In the SHF band, the input circuit, as a rule, is incorporated into the r.f. unit.

Review Questions

1. Why should the transmission line be terminated into an impedance equal to its characteristic impedance?
2. Why is the effective Q of a transformer-coupled input circuit halved at optimum coupling?
3. How does the effective Q of a transformer-coupled circuit vary with a decrease in the tapping-down ratio?
4. What value of the tapping-down ratio gives a maximum voltage-transfer ratio in a tapped-coil coupled input circuit?

77. Microwave Valve Amplifiers

R.f. valve amplifiers in radar receivers are advantageous as long as they are capable of amplification and help to reduce receiver noise.

It has been shown that at microwave frequencies the input resistance goes down as the frequency goes up. Indeed, R_{in} falls to a small fraction of the resonant resistance R_{0e} of the tuned circuit, so that the resultant anode load resistance R'_{0e} in a circuit with untapped transformer coupling may fairly accurately be put equal to R_{0e}. Then the stage gain is

$$K_0 = g_m R_{in} \tag{14.25}$$

With a low value of R_{in} the gain is less than unity, and such an 'amplifier' is useless.

There is a certain limiting frequency for each type of valve at which the valve ceases to amplify. Let us find this frequency for $K_0 = 1$. In this case, Eq. (14.25) can be rewritten in the following way:

$$1 = g_m R_{in}$$

so that

$$R_{in} = \frac{1}{g_m} \qquad \text{(a)}$$

According to Eq. (14.15), the input resistance

$$R_{in} = \frac{K}{f^2} \qquad \text{(b)}$$

the values of K being shown in table 14.1.

Equating (a) and (b) gives

$$\frac{1}{g_m} = \frac{K}{f^2}$$

whence, the limiting frequency is

$$f_{lim} = \sqrt{g_m K 10^3} \qquad (14.26)$$

In Eq. (14.26) the frequency is in megahertz if K is in megohms·MHz^2 and the mutual conductance is in milliamperes per volt.

Table 14.2 gives the values of limiting frequencies for several types of valves. The values are found by Eq. (14.26).

In a circuit with tapped-coil (or tapped-transformer) coupling the shunting effect of the valve input is decreased and the limiting frequency is higher in comparison with the data of Table 14.2. Practical valves of conventional construction cease to amplify at the lower end of the VHF band.

TABLE 14.2

Valve type	f_{lim}, MHz	λ_{lim}, metres	Valve type	f_{lim}, MHz	λ_{lim}, metres
6С1П	600	0.5	6Ж1П	600	0.5
6С5Д	3,000	0.1	6Ж3П	400	0.75
6К1П	300	1.0	6Ж4	250	1.2
6К3	200	1.5			

Bantam, super-miniature and acorn valves may be used in the high-frequency range of the UHF band. Lighthouse triodes operate up to 3,000 megahertz.

Single-stage Amplifier with a Tapped-down Secondary. As a rule, parallel-feed circuits are used at microwave frequencies.

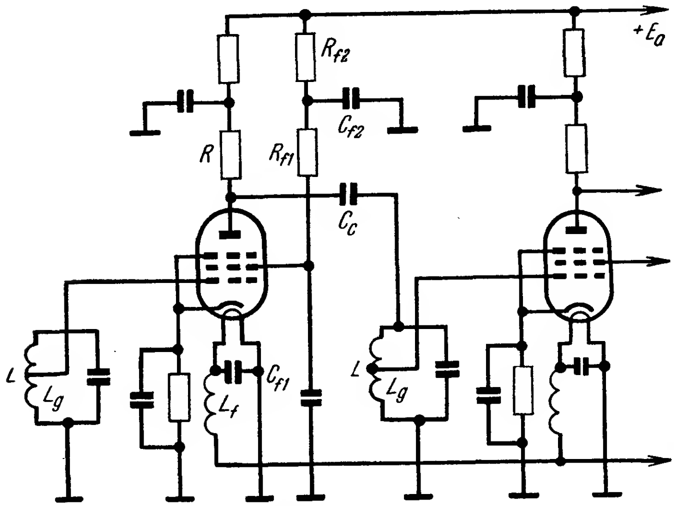


Fig. 14.18. Single-stage microwave amplifier with tapped-down transform coupling

In the circuit of Fig. 14.18, R in the anode lead replaces a choke. C_c is a coupling capacitor. To reduce the shunting effect of R_{in} and C_{in} on the tuned circuit, the grid of the next stage is connected to a tapping on the tuned-circuit coil.

When a common supply source is employed, parasitic coupling in the amplifier is eliminated by use of decoupling filters in the supply circuit. Thus, the anode circuit uses an L-type filter, and the screen grid circuit utilizes a two-section L-type filter, the first section of which $R_{f1}C_{f1}$ filters out radio frequencies, while the second section $R_{f2}C_{f2}$ the lower frequencies. Similar filters are also used in the filament circuits.

Depending on the resonant frequency, the tuned-circuit capacitance may consist of the capacitance of the capacitor and the shunt capacitance C_s, or only of C_s. In the latter case, the tuned-circuit capacitor may be omitted in the diagram.

According to Eq. (6.13), the stage gain is

$$K_0 = g_m R'_{0e} p_g \tag{14.27}$$

where $p_g = \frac{L_g}{L}$ is the tapping-down ratio on the grid side, and

R'_{0e} is the load resistance of the anode circuit consisting of R and $\frac{R_{in}}{p_g}$ in parallel.

The resonant resistance of the tuned circuit $R_{0e} \gg R_{in}$ is neglected and

$$R'_{0e} = \frac{R\frac{R_{in}}{p_g^2}}{R + \frac{R_{in}}{p_g^2}} \tag{14.28}$$

The maximum stage gain at

$$p_{g\,opt} = \sqrt{\frac{R_{in}}{R}}$$

is given by

$$K_{max} = \frac{g_m}{2}\sqrt{RR_{in}} \tag{14.29}$$

Connection of the grid to a tap on the tuned-circuit coil also affects the selectivity and bandwidth of the amplifier. Using Eq. (6.12) for the bandwidth and noting that

$$Q_{ef} = \frac{R'_{0e}}{\rho} = R'_{0e}\omega_0 C \tag{14.30}$$

we obtain

$$2\Delta F = \frac{f_0}{Q_{ef}}\sqrt{\frac{1}{\sqrt[n]{M^2}} - 1} = \frac{1}{2\pi C R'_{0e}}\sqrt{\frac{1}{\sqrt[n]{M^2}} - 1} \tag{14.31}$$

The value of the tuned-circuit capacitance C is given by

$$C = C_c + C_{out} + \frac{C_w}{2} + p_g^2\left(C_{in} + \frac{C_w}{2}\right) \tag{14.32}$$

where C_c is the capacitance of the capacitor; and C_w is the total capacitance of wiring.

Connection of the grid to a tap on the tuned-circuit coil is mainly used in order to obtain a comparatively narrow bandwidth. If a broad bandwidth is desired, untapped connection should be preferred.

Single-stage Amplifier with a Series Inductance. The circuit of Fig. 14.19 makes use of parallel feed. The tuned circuit consists of a coil L connected between the anode of the first valve and the grid of the next valve and of a capacitance comprising two series-connected capacitances C_1 and C_2. The capacitances C_1

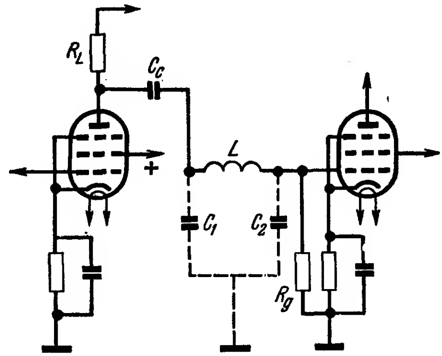


Fig. 14.19. Single-stage UHF amplifier with a series inductance

and C_2 represent, in effect, the input and output capacitances of the valves, with the wiring capacitance taken into consideration.

The advantage of the circuit of Fig. 14.19 over the usual circuits in which C_1 and C_2 are connected in parallel is that the series connection of capacitances brings down the total tuned-circuit capacitance considerably. This raises the limiting resonant frequency of the tuned circuit. This type of circuit can operate at frequencies up to 375 megahertz.

In the circuit of Fig. 14.19 both the anode and the grid are connected to a tap on the total capacitance.

For the anode, the tapping-down ratio is

$$p_a = \frac{C_2}{C_1 + C_2}$$

For the grid, this ratio is

$$p_g = \frac{C_1}{C_1 + C_2}$$

Therefore, the gain of the circuit is

$$K_0 = g_m R'_{0e} p_a p_g \tag{14.33}$$

Earthed-grid Amplifier. The usefulness of bantam, ultra-miniature and acorn pentodes as amplifying valves is limited to the high-frequency end of the UHF band because their input resistance drops to a low value and they introduce excessive noise. Besides, the inductance of the screen-grid lead in such valves begins to tell even in the low frequency range of the VHF band;

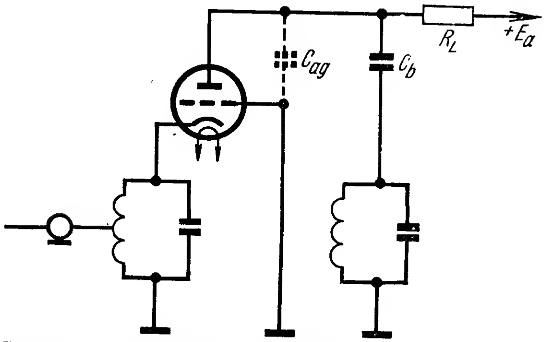


Fig. 14.20. Earthed-grid amplifier

the screen grid is no longer earthed for the r.f. and ceases to serve as a screen.

Those are the reasons why pentodes have been fully replaced by triode valves in the low-frequency range of the UHF band. Yet, although they have a lower noise level than pentodes, triodes tend to self-oscillate at higher frequencies if connected in the conventional manner owing to parasitic feedback between the input and output circuits of the valve through the anode-grid capacitance. A way out for UHF amplification is offered by an amplifier circuit in which the grid is earthed.

The unusual feature of the earthed-grid amplifier circuit (Fig.14.20) is that the tuned input circuit is connected between the cathode and the earthed grid, and the output circuit between the anode and earth. Just as in many other microwave circuits, the anode is parallel-fed and the electron stream is controlled in the usual way. The alternating input voltage is applied to the cathode which varies in potential relative to the earthed grid. This variation affects the space current in the normal way, and the anode potential varies accordingly. The earthed grid placed between the cathode and anode acts as an electrostatic screen. As a result, the anode-cathode capacitance is brought down to a few hundredths of its usual value. This isolation of the input and output circuits sharply diminishes parasitic feedback through C_{ak}, and the stability of the amplifier is considerably improved. C_{ag} is connected in parallel with the output circuit. The gain of the earthed-grid amplifier is

$$K_0 = g_m R'_{0e} p_g \qquad (14.34)$$

where R'_{0e} is the resistance of the anode load adjusted for the shunting effect of R_L and R_{in}.

In a multi-stage earthed-grid amplifier, the shunting effect of one stage upon another is greater than in the earthed-cathode circuit. To clarify this point, let us determine the input resistance of an earthed-grid stage. From the circuit of Fig. 14.20, it is seen that alternating anode current flows through the tuned input circuit. In other words, the anode current flows in the cathode-grid circuit even at negative bias.

By definition, the input resistance of the valve

$$R_{in} = \frac{V_{in}}{I_{in}}$$

where

$$V_{in} = V_g; \quad I_{in} = I_a$$

Noting that the alternating anode current is

$$I_a \cong g_m V_g$$

we have

$$R_{in} \cong \frac{V_g}{V_g g_m} \cong \frac{1}{g_m} \tag{14.35}$$

Thus the input resistance of the earthed-grid triode is independent of frequency and is small in value.

Owing to the low values of R_{in}, the earthed-grid circuit is particularly adapted for super-high frequencies, where R_{in} is greater than the input resistance of the pentode.

In multi-stage earthed-grid amplifiers, the grid of a next valve should be connected to a tap on the output tuned-circuit coil of the preceding one. If it were connected to the untapped coil, the stage gain would be unity. To prove this, we write

$$R'_{0e} \cong R_{in} \cong \frac{1}{g_m}$$

With $p=1$, the gain is

$$K_0 = g_m R'_{0e} = g_m R_{in} = g_m \frac{1}{g_m} = 1$$

A two-stage earthed-grid amplifier is shown in Fig. 14.21.

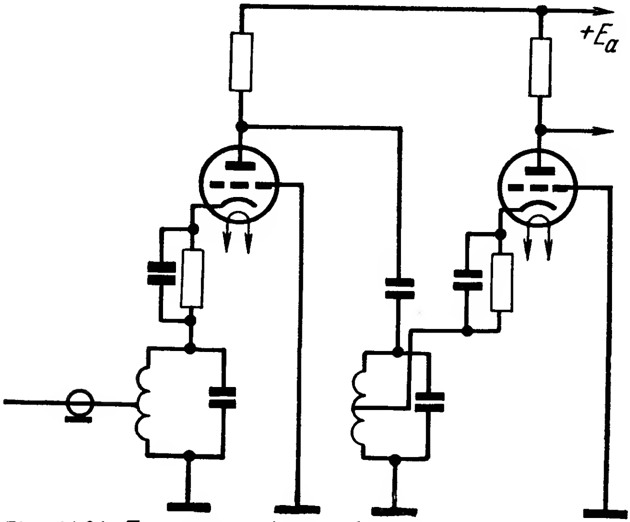


Fig. 14.21. Two-stage eartnea-grid amplifier

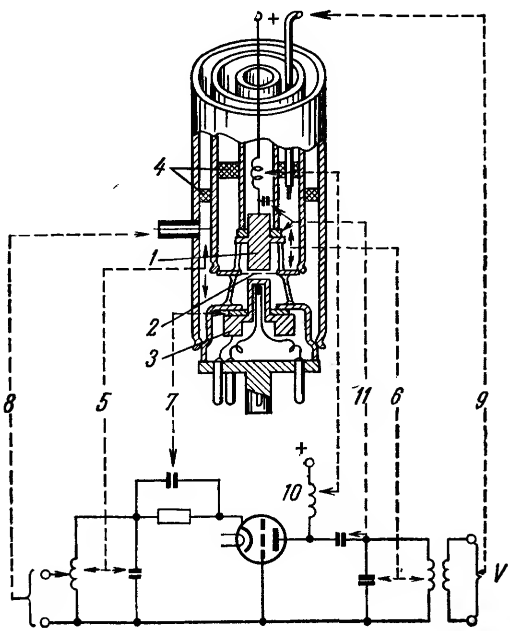


Fig. 14.22. Construction of an earthed-grid triode stage and its circuit diagram

In the low-frequency range of the VHF band and, partly, in the UHF band, earthed-grid amplifiers can employ bantam, ultra-miniature and acorn triodes, along with the usual tuned circuits or those made from segments of transmission lines. In the UHF band, particularly in its low-frequency range, lighthouse valves and coaxial-line segments may be used.

Figure 14.22 shows a sketch and a circuit diagram of an earthed-grid amplifier using a lighthouse valve. Three concentric lines are put on the disc terminals (anode *1*, grid *2* and cathode *3*) of the lighthouse valve. Each pair of lines forms a tuned circuit. Between the concentric lines are placed short-circuiting plungers *4*. One of the circuits formed by the outer and middle cylinders and connected between the cathode and earthed grid is the output tuned circuit *5*. This tuned circuit is conductively (autotransformer) coupled to the output transmission line *8*. Coupling is effected by connecting the inner conductor of the feeder to the middle cylinder.

The second tuned circuit *6* connected between the anode and grid is formed by the middle and outer cylinders. The output tuned circuit is coupled to the succeeding circuits by mutual inductance. This coupling is effected via the transmission line *9* which passes inside the plunger. The middle conductor of the transmission line terminates in a coupling loop. The inner cylinder is coupled to the anode by a capacitor *11*, since there is an insulating spacer located between the cylinder and the anode terminal. The d.c. voltage is fed to the anode through a choke *10*.

Grid bias is built up across the resistor in the cathode circuit. The capacitance *7*, by-passing this resistor, is created between the radio-frequency cathode terminal and the cathode lead connected to the valve pin (for the direct anode current).

Figure 14.23 shows the construction of a tunable single-stage amplifier employing a lighthouse valve and operating at 1,000 megahertz. The segments of concentric lines are tuned by means of plungers, the ends of which carry short-circuiting springs. The input tuned circuit is adjusted by a bridge which is a block rigidly connected to the tuning plunger. The output cable is conductively coupled to the inner cylinder of the input line through a hole in the bridge block.

When the bridge is shifted along the line, the coupling contact also shifts. The output tuned circuit is adjusted by rotating a knob which causes the inner plunger to shift. The feeder is inductively coupled to the output tuned circuit by means of a

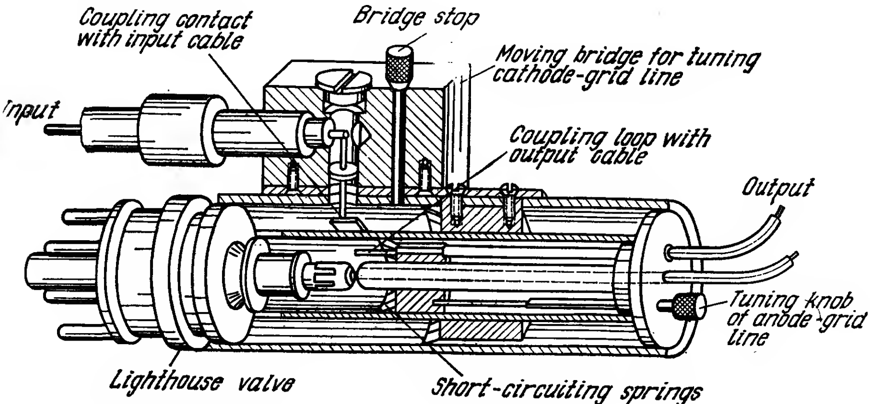


Fig. 14.23. Construction of lighthouse valve amplifier stage for 1,000 MHz

coupling loop. The latter moves together with the tuning plunger; as a result, the coupling is maintained constant over the range.

Cascode Amplifier. A triode amplifier, specifically designed for work in the VHF band and the low-frequency end of the UHF band, is shown in Fig. 14—24. Commonly called the *cascode amplifier*, it uses an earthed-cathode triode amplifier followed by an earthed-grid triode output stage.

At the input to the earthed-cathode stage, there is a double-tuned coupling circuit formed by L_1 and L_2 and the respective circuit capacitances. The anode load of the first stage is a single tuned circuit with L_3.

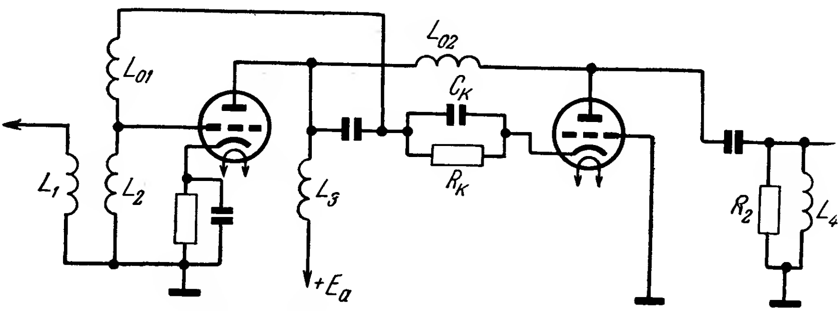


Fig. 14.24. Cascode amplifier

The earthed-grid stage is a parallel-feed circuit. The R_kC_k network connected into the cathode of the second valve provides automatic bias to the earthed grid through the series circuit $L_{01}L_2$.

L_{01} connected between the anode and grid of the first valve neutralizes the anode-grid capacitance C_{ag} of the valve. Its value is such that it resonates at the intermediate frequency with C_{ag}, thereby reducing the noise factor. D.c. voltage is applied to the anode of the second valve from a common supply source via L_{02}. Besides, L_{02} neutralizes the anode-cathode capacitance of the second valve.

In terms of gain, the cascode amplifier is equivalent to a single triode stage. To prove this, we write the gain of the first stage

$$K_1 \cong g_{m1}R_{in2}$$

According to Eq. (14.35), the input resistance of the earthed-grid stage is

$$R_{in2} = \frac{1}{g_{m2}}$$

whence

$$K_1 = \frac{g_{m1}}{g_{m2}}$$

The gain of the second stage is

$$K_2 = g_{m2}R_{0e2}$$

where R_{0e2} is the load of the second stage.

The gain of both stages is

$$K_{tot} = K_1K_2 = \frac{g_{m1}}{g_{m2}}K_2$$

Substituting the expression for K_2, we obtain

$$K_{tot} = g_{m1}R_{0e2} \tag{14.36}$$

i.e. the gain of the circuit is determined by the mutual conductance of the first triode and by the load of the second.

In such a circuit, the first stage does not provide voltage amplification. However, it does provide power amplification because its input resistance is higher than that of the earthed-grid stage. This facilitates matching to the preceding stage.

The noise factor of the cascode amplifier is practically the noise factor of the first stage.

The point is that the noise of the first stage is amplified in power to a proportion such that the noise in the second stage is negligible in comparison.

The cascode amplifier is widely used in the early stages of i.f. amplifiers in SHF radar receivers.

Review Questions

1. Why is it that at microwave frequencies the anode load is mainly the input resistance of the next stage?
2. Why does the gain of a microwave valve amplifier decrease as the frequency increases?
3. How does an increase in the tapping-down on the grid side affect the bandwidth?
4. What gain is provided by the optimum tapping-down?
5. What limits the use of the earthed-grid amplifier at lower radio frequencies?
6. How are the valves connected for d.c. in the cascode circuit?

78. Travelling-wave Valve Amplifier

One of the basic parameters of SHF radar receivers, the effective power sensitivity, is limited by the large value of the noise factor. In turn, the high noise factor of a receiver in this range is determined by the high noise level and low power gain of the mixer, which is the first stage of the receiver.

The receiver noise factor may be decreased by employing an amplifier with a low noise level and high power gain in the r.f. unit. Such amplifiers with conventional valves are limited to the UHF band because of the finite transit time of electrons.

More useful in this frequency range are valves which utilize the transit-time effects for their operation. One such valve, which is, in effect, a self-contained amplifier, is the travelling-wave valve.

The construction of the travelling-wave valve is shown in Fig. 14.25. An evacuated glass envelope has a cathode and an electrostatic electron gun assembly *1* at one end, and a helix *2* of copper wire fixed along the axis of the envelope. A collector electrode, or anode, *3* at the opposite end of the envelope collects the electrons in the beam. A longitudinal magnetic field is produced by a solenoid carrying direct current and mounted

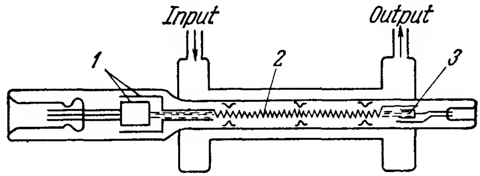


Fig. 14.25. Schematic of a travelling-wave valve

outside the envelope. The field causes the electrons emitted by the cathode to travel parallel to the axis of the helix, and this is magnetic focusing.

The action of the valve may be explained briefly as follows. The UHF signal to be amplified is impressed on the helix at the input end and gives rise to a current wave which travels with a velocity close to that of light. The electromagnetic field so created may be said to have two components, a transverse one moving along the wire of the helix, and a longitudinal one moving along the axis of the helix.

Obviously, the velocity of the longitudinal wave will be determined by the pitch of the helix. If the length of the helix wire is k times the axial length of the helix, then the velocity of the longitudinal wave will be

$$u = cl/k$$

where c is the speed of light, l is the circumference of a turn, and k is the pitch of the helix.

The velocity of the electron beam is arranged to be very slightly greater than that of the longitudinal component of the electromagnetic field produced by the UHF signal. Under these conditions, a continuous interaction takes place along the helix between the electron beam and the electromagnetic wave. A map of the field for each instant of time is shown in Fig. 14.26*a*. As is seen, there is an electric field around each turn of the helix. Variations in the magnitude and phase of the field intensity E are shown in the plot of Fig. 14.26*b*. Consider how the electromagnetic field due to the UHF signal interacts with the electron beam.

Since the velocity of the beam is approximately the same as that of the field, the field may be said to be stationary relative to the beam. Owing to this, the field acts on each electron continuously over nearly the entire length of the helix. Let the positive half-cycles of the field shown in the plot of Fig. 14.26*b* accelerate the beam, and the negative half-cycles

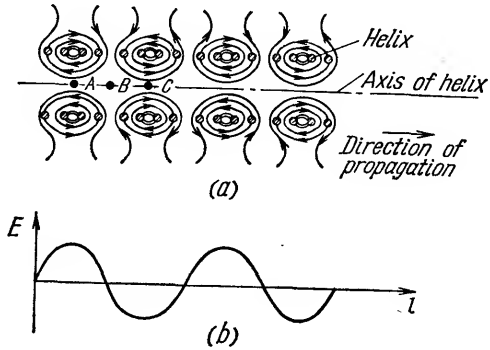


Fig. 14.26. Map of electrical field due to electromagnetic wave travelling along the helix

retard it. Then, electron A (see Fig. 14.26*a*) will be accelerated all the time. Electron C will be retarded, while electron B will be unaffected by the field. Thus, some electrons in the beam are speeded up, while others are slowed down. This is known as *velocity modulation*. Velocity modulation results in that the electrons are separated into groups, or "bunched". This bunching action builds up in amplitude until after a certain distance the packets of electrons begin to pass energy to the electromagnetic wave of the UHF signal. This transfer of energy increases down the remaining length of the helix, thereby providing an amplified signal at the output end.

Travelling-wave valves have a low noise figure and a high power gain, so that the overall noise figure of the receiver can be markedly reduced.

Travelling-wave valve amplifiers find use in both the UHF and SHF bands.

Review Questions

1. What type of electromagnetic field focuses the electron beam in the travelling-wave valve?
2. What field effects the velocity modulation of the electron beam in the travelling-wave valve?

79. Parametric Amplifiers

This type of amplifier owes its name to the fact that it uses variable reactances, C or L, usually defined as circuit parameters.

While normal amplifiers convert power from a d.c. source into power at some signal frequency, a parametric amplifier converts power at one frequency into power at the signal frequency. Parametric amplifiers are of interest in microwave radar because they have a low noise figure.

The physical principles involved in parametric amplification may be seen in the following example. Consider an LC circuit as shown in Fig. 14.27. In this circuit we shall assume that the capacitor plates can be physically pulled apart and pushed together. We shall further assume that the tuned circuit accepts a sinusoidal signal voltage V_s of frequency f_s, such that the capacitor voltage V_C and the capacitor charge q_C will vary also sinusoidally. Pulling the capacitor plates suddenly apart when the signal voltage is a maximum will increase the capacitor voltage $V_C = q_C/C$, because the capacitance is reduced while the charge remains the same. Moving the plates together when the voltage across the capacitor is zero will cause no change in the energy of the system. A graphic interpretation of parametric amplification is presented in Fig. 14.28.

The upper plot shows variations in the signal voltage; the middle plot shows the moving together and pulling apart of the plates; the lower plot represents variations in the capacitor voltage. The plates are pulled apart at points marked by x's on the curve, when V_C is a maximum. As is seen, the capacitor voltage is steadily built up in jumps. That is, the signal voltage is amplified.

Pulling the plates apart involves the expenditure of mechanical energy because the presence of the electric field causes an attractive force to exist between them. No energy is expended

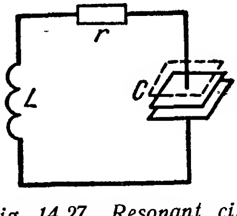


Fig. 14.27. Resonant circuit with a variable capacitance

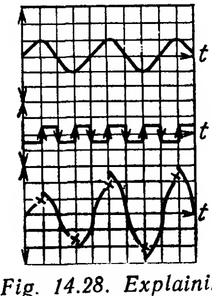


Fig. 14.28. Explaining parametric amplification

when the plates are moved together, because this is done when the capacitor voltage is zero and there is no electric field to give rise to forces of repulsion.

Thus, the work done on pulling the plates apart builds up the energy of the system, and this is accompanied by an increase in the capacitor voltage. In other words, mechanical energy is converted into the energy of the electric field.

It appears that every half-cycle of oscillation some additional energy is pumped into the tuned circuit, and the source of this additional energy is called, quite appropriately, a *pump*. This occurs at what is called the *pump frequency* which, as follows from the principle of parametric amplification, must be twice the incoming-signal frequency.

In practical parametric amplifiers, the non-linear reactance in which the pump voltage is mixed with the signal voltage is usually a *varactor diode*.

In a varactor, the barrier (or depletion) layer acts as the dielectric of a capacitor, because it is almost completely devoid of charge carriers, while the P- and N-region on each side act as capacitor plates (this is the reason why varactors are often called *capacitor diodes*). The application of an alternating voltage from a pump circuit will periodically displace the P- and N-regions, varying the capacitance of the varactor.

Going back to parametric amplification, it should be noted that the pumping of energy through a variable capacitance and the associated increase in voltage may be regarded as occurring due to the addition of a negative resistance to the tuned circuit, which results in a reduction of losses and improvement of the Q-factor of the tuned circuit.

This interpretation of parametric amplification gives grounds for treating it as regeneration. In contrast to conventional regeneration, however, as caused by positive feedback, this is *parametric regeneration*, because it stems from variations in the capacitance of the tuned circuit.

Regeneration in a parametric amplifier is a function of the pump energy. If the negative resistance coupled into the tuned circuit exceeds its loss resistance, the amplifier will lapse into self-oscillation, and the result will be a *parametric oscillator*.

In valve and transistor amplifiers, the main source of noise is the controlled stream of charge carriers. Parametric amplification does not involve any control of the charge stream, and so the noise level in parametric amplifiers is negligible.

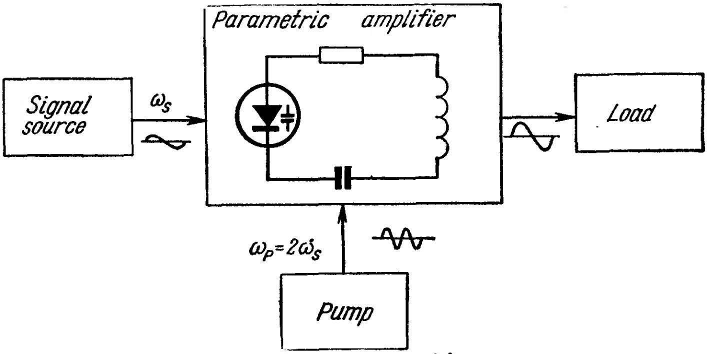


Fig. 14.29. Single-tuned parametric amplifier

Single-tuned Parametric Amplifiers. As is seen from the block-diagram of Fig. 14.29, the key units are a parametric single-tuned circuit, a signal source, a pump, and a load.

A single-tuned parametric amplifier can operate as both a regenerative and a superregenerative circuit. In the latter case, the pump should be modulated by the quench frequency.

In the single-tuned parametric amplifier, the load is connected to the same tuned circuit that accepts the signal voltage. This results in an increased noise factor because the thermal noise originating in the load can affect the tuned circuit and go back to the load amplified. To prevent this, the load is usually connected to the parametric amplifier through a *circulator*, a unidirectional device. Single-tuned parametric amplifiers are used in the VHF, UHF, and SHF bands.

In the VHF band, the tuned circuits are of the conventional lumped-constant type; in the UHF band, use is made of coaxial resonant lines; in the SHF band these are cavity resonators and waveguides. In all bands, the variable capacitance is a varactor, although use may be made of the collector junction in a transistor.

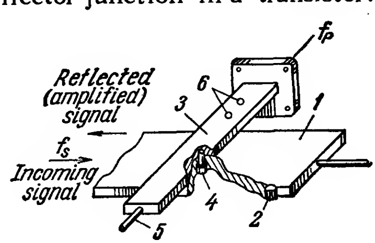


Fig. 14.30. Sketch of a single-tuned parametric amplifier for the SHF band

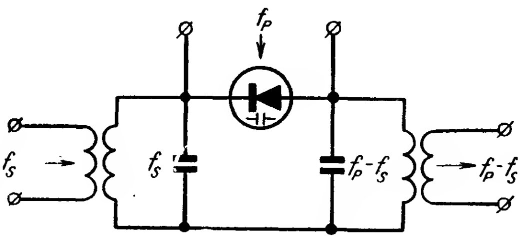


Fig. 14.31. Explaining operation of a negative-resistance parametric converter

Figure 14.30 shows a sketch of a parametric amplifier for the SHF band, which consists of two waveguide segments arranged in a criss-cross manner. The signal is injected into a waveguide chamber *1* which is tuned with a plunger *2* placed at the end of the waveguide. The pump voltage is applied to another waveguide chamber *3* from a pump oscillator and is mixed with the signal voltage in a varactor *4*. The pump waveguide chamber has a tuning piston *5* and tuning screws *6*. The amplified signal is extracted from the chamber *1* and is fed through a circulator to the succeeding stages of the receiver.

Double-tuned Parametric Amplifiers. A double-tuned parametric amplifier contains two tuned circuits coupled by a common non-linear capacitance varying at the pump frequency f_p.

The varactor mixes two frequencies, the pump frequency f_p and the signal frequency f_s. Its output contains a sum or a difference frequency, $f_p \pm f_s$, called the *output frequency*. The first tuned circuit is tuned to resonate at f_s, and the second at the output frequency f_2. When $f_2 > f_s$, the circuit is an *up-converter*; when $f_2 < f_s$, it is a *down-converter*.

Consider operation of a parametric amplifier (Fig. 14.31) in which the output tuned circuit is tuned to resonate at the difference frequency $f_2 = f_p - f_s$. The output frequency is then called the *idler frequency*, and the output tuned circuit is referred to as the *idler circuit*. The action of the pump gives parametric regeneration, and negative resistance is coupled into each tuned circuit, because of which the entire circuit is called a negative-resistance parametric converter.

Negative-resistance parametric converters have a lower noise factor than single-tuned parametric amplifiers. Besides, they need no circulator. For efficient operation, it is essential that the idler frequency be five to ten times the signal frequency. Since the idler frequency is determined by the pump frequency, the limit

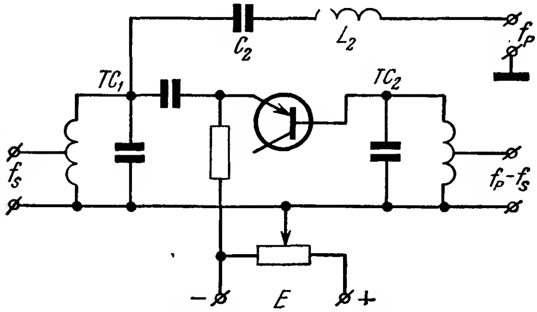


Fig. 14.32. Circuit diagram of a parametric converter for the VHF band

of operation for a parametric converter is decided by the frequency of the pump generator. Existing low-power microwave generators are limited to a pump frequency of ten to fifteen gigahertz (one gigahertz, symbolized GHz, is equal to one thousand megahertz). In other words, parametric converters can operate at frequencies not over 1 to 1.5 gigahertz.

Figure 14.32 shows the circuit diagram of a parametric converter for about 50 megahertz. The signal to be amplified is impressed on part of the coil in the first tuned circuit TC_1. The amplified output at 226 megahertz is coupled out from a tap on the coil in the second tuned circuit and is fed to a next stage.

The parametric capacitance is the capacitance of the collector junction in a power triode to which a negative bias E is applied. The pump voltage at $f_p = 276$ megahertz comes from a pump generator through a filter C_2L_2. The pump generator may be any r. f. oscillator developing a power of several tens of watts.

Figure 14.33 suggests a circuit for a parametric converter operating in the UHF band. The signal cavity is a coaxial resonator *1* which can be tuned with a piston *2*. The idler cavity is a waveguide chamber *5* with a tuning piston *6*. The

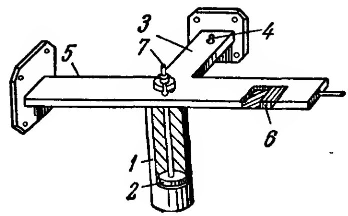


Fig. 14.33. Sketch of a parametric converter for the UHF band

idler frequency lies in the SHF band. The varactor *7* is placed at the junction of the two cavities. The pump generator is connected to the transverse waveguide *3*, tunable with a screw *4*.

A comparison of parametric amplifiers and converters is presented in Table 14.3.

TABLE 14.3

Circuit	Input frequency, MHz	Output frequency, MHz	Gain, db	Bandwidth, MHz	Noise figure, db
Parametric amplifier	500	—	15	6	1.5
	4,500	—	22	11	2.6
	11,500	—	10	25	3.6
Parametric converter	460	8,495	20	2	1.5
	500	8,500	18	5	1.2
	750	8,750	25	8	1.5
	925	8,800	20	30	1.5

Review Questions

1. What is the function of the pump generator in a parametric amplifier?

2. What is the output frequency in a negative-resistance parametric converter?

3. Why should the pump frequency be twice the signal frequency in a single-tuned parametric amplifier?

80. Induced Emission Amplifiers

In these circuits, amplification is effected by the stimulated emission of radiation. This is a quantum-mechanical process performed by *masers*.

For an insight into operation of masers, it is essential to learn some of the properties of quantum-mechanical particles and the key postulates of quantum theory.

Quantum theory deals with molecules, ions, atoms and electrons. A system where these particles interact is called a quantum-mechanical system. The internal energy of a quantum-mechanical system or of its particles can change. One of the causes of such changes is an exchange of energy with the electromagnetic field.

When a system absorbs energy, it rises to a higher energy level. When it gives up energy, it descends to a lower energy

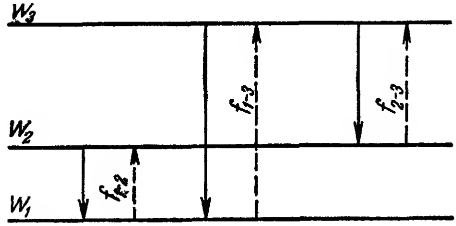


Fig. 14.34. A quantum system with three energy levels

level. As an example, Fig. 14.34 shows a quantum system with three energy levels, W_1, W_2 and W_3.

One of the key postulates of quantum theory is that an atom or molecule does not emit or absorb energy continuously. Rather, it does so in a series of steps, each step being the emission or absorption of a unit quantity (*quantum*) of electromagnetic energy, called a *photon*.

The energy of a photon is

$$W = hf \tag{14.37}$$

where $h = 6.625 \times 10^{-34}$ joules-sec is Planck's constant, and f is the frequency of the incident photon (the electromagnetic field). Thus, the energy of the photon is completely decided by its frequency.

Two adjacent energy levels are separated by a few gigahertz to a few tens of gigahertz. This implies that if an atom "sees" an electromagnetic wave of the proper frequency, it can be stimulated to change to a higher or a lower level. The process is called *transition*, and the frequency of the field causing it is called the *transition frequency*. Referring to Fig. 14.34, the energy-level separation, respectively, is $W_2 - W_1$, $W_3 - W_1$, and $W_3 - W_2$. The corresponding transition frequencies are

$$f_{1-2} = (W_2 - W_1)/h$$
$$f_{3-1} = (W_3 - W_1)/h$$
$$f_{2-3} = (W_3 - W_2)/h$$

For a quantum system, the lowest energy level is most stable, and the system will always tend towards it. This is why the number (or the *population*) of lowest-energy particles is the largest. However, the particles cannot remain in that minimum-energy state because of continuous thermal agitation which drives them to a higher state.

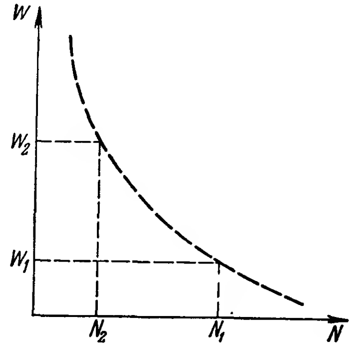


Fig. 14.35. Plot of energy level vs. population

The populations of the energy levels at thermal equilibrium may be specified by Maxwell-Boltzmann statistics. A plot of energy level versus population N is shown in Fig. 14.35. As is seen, the population at a higher energy level, say W_2, is smaller than that of the lowest energy level W_1 ($N_2 < N_1$). As the temperature is raised, the population of the higher levels approaches that of the lowest level (Fig. 14.36).

Quantum systems specified by Maxwell-Boltzmann statistics always absorb energy from the field because the population of the lower energy levels is greater than that of the higher energy levels. In order that such a system can emit energy, it is necessary to invert its populations, that is, to cause more particles to change from a lower to a higher state.

There exist several methods for population inversion. The most commonly used ones are by separation of upper-state particles with an electrostatic or electromagnetic field and by pumping in an external electromagnetic energy.

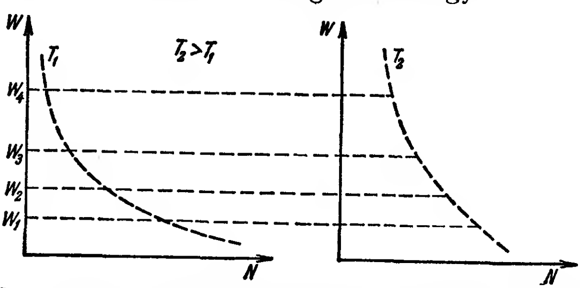


Fig. 14.36. Variations in population distribution with temperature

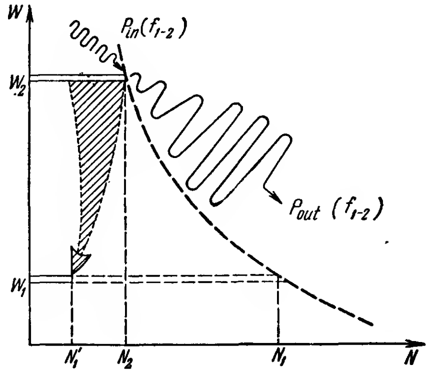


Fig. 14.37. Approximate plot of induced emission

In a quantum system with the inverted population, a weak electromagnetic wave of the proper transition frequency will excite the transition of particles from the upper to the lower energy level. This is *induced transition*, and the emission of energy associated with it is *induced emission*. It is the basis of maser operation.

An approximate plot of induced emission is shown in Fig. 14.37. The dotted line represents the energy state of the system at a temperature T_1, W_2 and W_1 are the allowed energy levels with a transition frequency f_{1-2}. The respective populations are N_2 and N_1, such that $N_1 > N_2$. Population inversion can be effected by separating the lower-state particles so that their population is reduced to N_1', such that $N_1' > N_2$.

Then a weak signal P_{in} at frequency f_{1-2} induces the population of the energy level W_2 to change to the lower level W_1 in an avalanche fashion. As this happens, an induced radiation P_{out} is emitted at the same transition frequency f_{1-2}.

Induced-emission amplifiers have a gain of tens to thousands. The exceptionally high gain of masers is due to the fact that they generate *coherent* radiation. That is, the emitted photons are all in the same phase with one another and with the external field.

In contrast to conventional amplifiers in which noise is caused by minor variations in the electron stream, induced-emission amplifiers are free from such streams and, as a consequence, from the associated noise. This is why induced-emission amplifiers may be regarded as having the lowest noise figure ever attained.

There are two main types of masers, *gas masers*, and *solid-state masers*. In the former, the material is a gas; in the latter, a paramagnetic crystal. Microwave systems ordinarily use solid-state masers, therefore the discussion that follows will be about them.

As will be recalled, all materials may be classed into ferromagnetic, paramagnetic, and diamagnetic.

The ultimate source of the magnetic properties of matter is the orbital motion of the electrons around the nucleus and the spin of the electron about its own axis, both of which produce a magnetic dipole moment.

Ferromagnetism is mainly due to the electron-spin magnetic moment. In those atoms where the electrons form complete shells around the nucleus, the magnetic contributions of the individual spins or individual orbital motions neutralize one another and result in diamagnetic atoms. Copper, silver, and carbon are examples of diamagnetic materials. In the atoms of some materials, the magnetic moments (which may be orbital or spin moments) of the individual electrons do not compensate one another, and a net magnetic moment results. Examples of paramagnetic materials are rare-earth salts, iron salts, cobalt, nickel, etc.

When placed in an external magnetic field, the paramagnetic atoms orient themselves as elementary magnets. When a paramagnetic atom is oriented with the field, there is no interaction

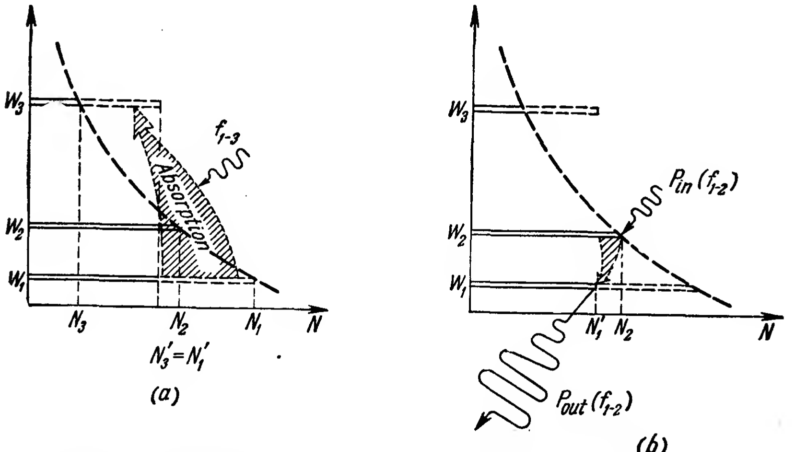


Fig. 14.38. Amplification in a three-level paramagnetic maser
(a) excitation; (b) induced emission

between the two, and the energy of orientation is zero. When a paramagnetic atom is oriented against the field, the energy of orientation is a maximum. The atom tends to orient itself in the direction of the field, and the interaction is strongest.

Like any quantum system, a paramagnetic system is characterized by a set of discrete energy levels. The population distribution of the energy levels is likewise specified by Maxwell-Boltzmann statistics. To obtain maser action from a paramagnetic system, the population distribution must be inverted. The most common mode of operation is three-level population inversion.

Referring to Fig. 14.38, the dotted curve shows the Maxwell-Boltzmann distribution prior to excitation. The three energy levels are chosen such that W_3 is above W_2, and W_2 is above W_1. The population distribution is such that $N_1 > N_2 > N_3$. The difference between the populations may be rather considerable if the active element is maintained at a sufficiently low temperature so as to isolate the paramagnetic atoms from thermal agitation.

Population inversion is accomplished by use of a pump energy, Fig. 14-38*a*. The pump frequency $f_{1=3}$ corresponds to the transition from W_1 to W_3. As the energy supplied by the microwave pump is absorbed by the system, some of the particles change to W_3, and its population rises to N'_3, that of W_1 drops to $N'_1 = N'_3$. As a result, there will be an excess of high-energy particles at the second level in comparison with the first level, that is, $N_2 > N'_1$, and the system will be ready for induced radiation at frequency $f_{1=2}$ (Fig. 14.38*b*). If now a weak signal of transition frequency $f_{1=2}$ is injected into the input, it will cause induced radiation at a much higher power level.

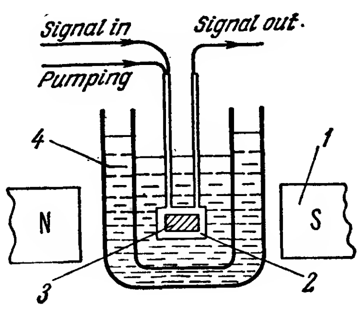


Fig. 14.39. Sketch of a three-level paramagnetic maser with an external magnet

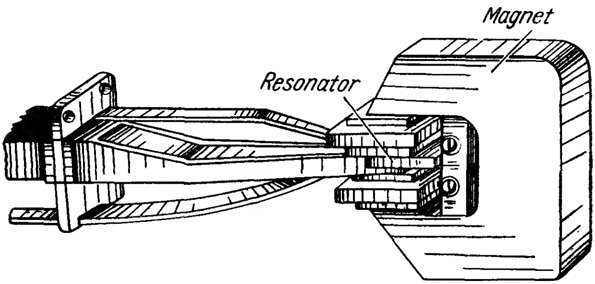


Fig. 14.40. Resonator head of a 3-cm paramagnetic maser

Figure 14.39 shows a sketch of a three-level paramagnetic maser with an external magnet. The paramagnetic crystal *3* is placed in a cavity resonator *2*, and both are immersed in a cryostat *4*, made up of two Dewar flasks, one filled with liquid helium and the other with liquid nitrogen. The cryostat is placed between the pole-pieces of a permanent magnet *1*.

Paramagnetic masers for use in the 3-cm band incorporate an internal magnetic system, in which case the cryostat encloses the resonator holding the crystal, and the permanent magnet.

The resonator head of such a maser is shown in Fig. 14.40. The active (paramagnetic) element is a ruby crystal doped with chromium. The pumping source operates at a wavelength of 1.25 cm.

Paramagnetic masers have a very low noise figure and a bandwidth of tens of megahertz. They are employed in radar and space communication.

Review Questions

1. What characterizes the energy state of a quantum system?
2. What accompanies an increase in the energy of a quantum system?
3. How does the energy of a photon vary with the frequency of the electromagnetic field?
4. How are population inversion and induced radiation related?
5. What is the function of pumping energy in a paramagnetic maser?

81. Mixers and Frequency Changers for the VHF and UHF Bands

In radar receivers, the frequency changer (mixer and local oscillator) is a far more important component unit than it is in receivers operating at lower frequencies.

In the VHF-UHF bands, the circuits preceding the frequency changer provide insignificant amplification, while there are no such circuits in the SHF band at all. This is why the internal noise of the frequency changer and its gain have a marked effect on the noise quality and, consequently, sensitivity of the receiver. Indeed, the noise quality of a valve intended for use in a microwave frequency changer is the governing factor in its selection.

Mixer circuits with double-grid injection cannot be used in microwave receivers because multi-grid valves introduce excessive noise and tend to self-oscillate. Therefore, the mixer valves for the VHF and UHF bands are mainly pentodes, triodes, and diodes. In the SHF band, use is made of crystal mixers. Since SHF mixers and frequency changers possess some specific features, they will be discussed separately. This section will be devoted to mixer and frequency changer circuits for the VHF and UHF bands.

Pentode and Triode Mixers. *Pentode mixers* are mainly used in the VHF band and differ only slightly from the circuits using single-grid injection discussed in Chapter IX. The most commonly used configuration is the cathode-coupled circuit of Fig. 9.7*a*.

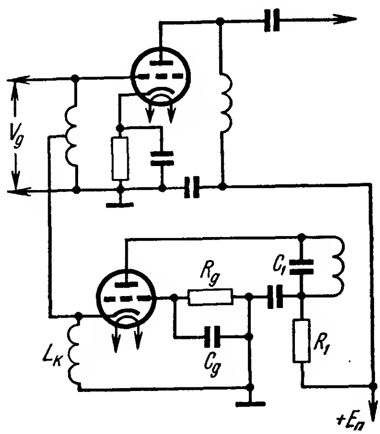


Fig. 14.41. VHF vacuum-triode frequency-changer

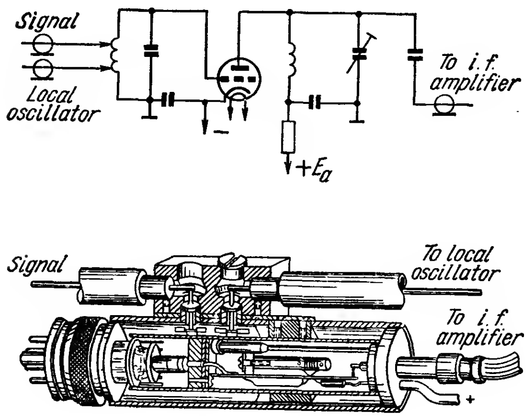


Fig. 14.42. Circuit diagram and construction of a lighthouse triode mixer

Triode mixers are used within a broader frequency range up to 1,000 megahertz. They are less noisy than pentode circuits.

Figure 14.41 shows the circuit schematic of a VHF frequency changer. It consists of a triode mixer and a separate local oscillator. The local oscillator is loosely coupled to the mixer through a tap on the mixer coil. The local oscillator is a circuit widely used at frequencies up to 350 megahertz. It has its tuned circuit placed between anode and cathode and utilizes capacitative feedback through the interelectrode capacitances of the valve. Grid bias for the local-oscillator valve is furnished automatically from the network R_gC_g. The local-oscillator voltage V_o develops across L_k connected in the cathode lead. The network R_1C_1 is a decoupling filter. The valves are bantam, super-miniature and acorn triodes.

In the UHF band, mixers are disc-seal (lighthouse) valves in conjunction with tuned circuits made from coaxial resonant lines. The circuit diagram and a sketch of a mixer using a disc-seal valve appears in Fig. 14.42. It looks very much like an amplifier using the same type of valve (see Fig. 14.23).

The input tuned circuit tuned to resonate at the incoming-signal frequency is made up of two concentric tubes connected to the cathode and grid. The output i.f. tuned circuit consists of a conventional inductor and an adjustable capacitor.

The two are placed inside the middle cylinder. The i.f. signal is fed to the i.f. amplifier over a coaxial feeder. The incoming-signal frequency and the local-oscillator output are coupled to the input tuned circuit through separate taps on the tuned-circuit coil. The respective feeders are connected to the inner cylinder through holes in the terminal block by means of spring terminals.

The input tuned circuit is tuned by moving a short-circuiting piston permanently attached to the block. This type of mixer is effective at frequencies up to 1,000 megahertz. The local oscillator operating close to 1,000 megahertz also uses a disc-seal valve.

Diode Mixers. Diode mixers are employed in the UHF band at frequencies up to 2.85-3 gigahertz.

The circuit of a diode mixer is shown in Fig. 14.43. It contains the receiver input circuit, a short-circuited segment of a coaxial line. The r.f. signals from the aerial and local oscillator are applied to the mixer over two coaxial cables. The i.f. signal develops across the *LC* tuned circuit connected to the anode of the diode. The R_1C_1 circuit provides automatic bias to the anode, necessary for diode operation with current cutoff.

Diode mixers use special diodes with low noise level, small interelectrode capacitance and short electron transit time.

The efficiency of a diode mixer is specified in terms of voltage transfer factor in contrast to grid-valve mixers whose efficiency is specified in terms of gain. The voltage transfer factor of a diode mixer, like that of a diode detector, is less than unity.

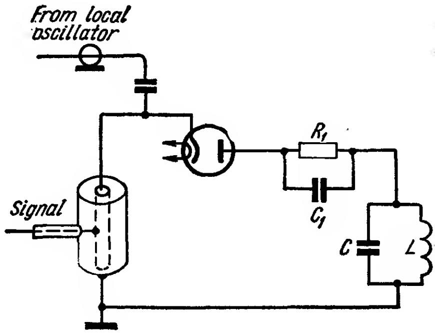


Fig. 14.43. Diode mixer

Review Questions

1. Why is it that at microwave frequencies preference should be given to frequency conversion with single grid injection?
2. What types of tuned circuits are used at the input and output of a disc-seal valve mixer?
3. What frequency is the LC-circuit tuned to in a diode mixer?

82. Single-ended Frequency Changers for the SHF Band

SHF radar receivers use crystal mixers and crystal frequency changers.

The crystal diode has a number of advantages over the vacuum diode, such as reduced time necessary for charges to pass between the electrodes, lower interelectrode capacitance and noise level. As a result, the circuit has better efficiency and reduces receiver noise.

A crystal mixer is specified in terms of operating wavelength, power transfer factor, noise level and output resistance.

The power transfer factor K_{pm} is defined as the ratio of the i.f. power at the mixer output to the r.f. signal power at the input under conditions of a matched load

$$K_{pm}=\frac{P_{out}}{P_{in}}$$

The value of K_{pm} usually does not exceed 0.25.

The noise properties of the crystal mixer are specified in terms of equivalent noise temperature t_{eq}. The relative noise temperature is a quantity showing how many times it is necessary to increase the absolute temperature of an equivalent noisy resistance so that the noise level in it would be equal to the noise level of the crystal mixer. The value of t_e is usually 2 to 3.

The *output resistance* of the mixer is defined as the ratio of the i.f. voltage at its output to the i.f. current.

These parameters of the most commonly used diodes of Soviet manufacture are given in Table 14.4.

It should be noted that the operation of a semiconductor diode depends to a large degree on the power of the signal from the local oscillator. As the power increases, the noise level increases, too. Besides, when the crystal mixer accepts too much power it is overloaded. As a result, the crystal burns at the point of contact with metal. The maximum power applied to

TABLE 14.4

Parameters \ Diode type	ДГ-С1	ДК-С1	ДГ-С2	ДК-С2	ДГ-С3	ДК-С3	ДГ-С4	ДК-С4	ДК-С5	ДК-С7
Operating wavelength, cm	9.8	9.8	9 8	9.8	3.2	3.2	3.2	3.2	2	3.2
Power transfer factor K_{pm}, db	8.5	8.5	6.5	6.5	8.5	8.5	6.5	6.5	8	7
Equivalent noise temperature, t_{eq}	3	2.7	3	2	3	2.7	3	2.2	2.5	2
Output resistance R_{if}, ohms	400	400	400	400	400	400	400	400	400	400

a semiconductor diode from the local oscillator should not exceed one milliwatt, otherwise the diode will be destroyed. In crystal mixers, a check of the applied power is provided by milliammeters which measure the direct component of the mixer current.

Since SHF radar receivers have no r.f. amplifier, the crystal mixer is the input stage of the receiver. Because of this the noise level of the crystal mixer and of the first i.f. stage determines the noise quality of the receiver as a whole. The noise factor of such a receiver is given by

$$N=\frac{t_{eq}+N_{if}-1}{K_{pf}K_{pp}K_{pm}} \tag{14.38}$$

where N_{if} = i.f. amplifier noise level
K_{pf} = power transfer factor of the feeder
K_{pp} = power transfer factor of the TR switch.

In calculations, K_{pf} is taken equal to 0.9; foı some types of TR switches the values of K_{pp} are given in Table 14.5.

In design, SHF mixers radically differ from VHF and UHF valve mixers. In the SHF mixer, the crystal diode is located inside a coaxial line, waveguide or cavity resonator, and special types of valves, klystrons and lighthouse valves, with coaxial tuned circuits, are used as the local oscillators. With lighthouse valves, the intermediate frequency is obtained by heterodyning with one of the local oscillator harmonics, usually the third,

$$f_i=3f_o-f_s$$

TABLE 14.5

TR switch type	Particulars	Frequency range, MHz	K_{pp}	Q_p
721B	Narrow-band with external resonator, untuned	2,750-3,300	0.7	300
1B27	Narrow-band with external resonator, tuned	2,400-3,750	0.7	350
724B	Narrow-band with external resonator, untuned	8,600-9,700	0.6	200
1B24	Narrow-band with internal resonator, tuned	8,500-9,600	0.8	300
1B26	Narrow-band with internal resonator	23,420-24,580	0.7	200

Coaxial Crystal Mixer. In 10-cm radar systems ($f = 3{,}000$ megahertz) the crystal diode is placed into a coaxial cavity. The construction of the mixer is shown in Figs. 14.44 and 14.45.

The mixer is made up of three segments of coaxial line. The input side of the line begins at a flange *1*, which is connected to the flange of a TR tube. The r.f. signal is applied to the mixer input via a coupling loop *2* located in the TR resonator. The cartridge containing the diode *4* is located at the opposite end of the line and is connected to its inner conductor *3*. The main coaxial line, through which the signal is fed, is

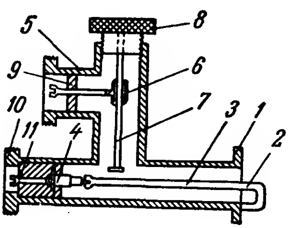


Fig. 14.44. Construction of a pencil-type coaxial crystal mixer for 3,000 megahertz

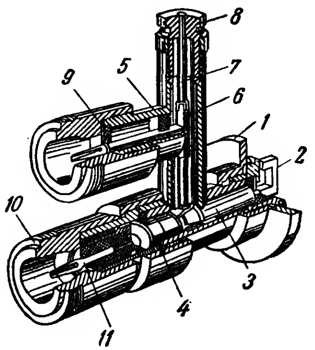


Fig. 14.45. Sketch of a pencil-type coaxial crystal-mixer for 3,000 megahertz (Notation is the same as in Fig 14.44)

coupled to a T-junction. The local oscillator signal is applied through the horizontal part of the T-junction *5*, into which a pin connected to a flexible conductor is screwed. The inner conductor in the horizontal part of the T-junction terminates in a bushing *6*. The probe *7* passing inside the bushing is used to adjust coupling between the local oscillator and mixer. The coupling is effected through the capacitance between the inner conductor of the line and the probe plate and is adjusted by moving the probe with a knob *8*.

The local-oscillator load is a 50-ohm disc resistor *9*. The crystal diode, connected to the inner conductor of the line is in an electromagnetic field due to the incoming signal and local-oscillator voltages, so that an i.f. current flows in the diode circuit as in any non-linear element. The i.f. signal is coupled out of the mixer over a flexible concentric cable joined to the mixer with the aid of a connector. The connector is screwed onto the threaded part of the line *10* so that its inner conductor touches the face of the crystal cartridge through a jack *11*.

Single-cavity Crystal Mixer. At frequencies above 3,000 megahertz, microwave energy is conveyed from one part of a system (say, the aerial) to another (the receiver) through hollow pipes called *waveguides*. Commonly used waveguides are either rectangular or circular in cross-section. The transfer of microwave energy in a waveguide occurs as the propagation of electromagnetic waves through the interior space bounded by the walls of the waveguide.

A uniform plane electromagnetic wave has the electric field E at right angles to the magnetic field H, and the direction of propagation is at right angles to the plane containing both E and H. The direction of propagation forms a right-handed system of axes with the vectors E and H, as shown in Fig. 14.46.

This mode of propagation is encountered in ordinary transmission lines and is called the *transverse electromagnetic mode* (*TEM*). In a waveguide, the resultant wave travelling longitudinally down the guide has a component of either the electric or magnetic field in the direction of propagation. Therefore, transmission may be either *transverse electric* (*TE*) or *transverse magnetic* (*TM*). In the former case, the electric field lies wholly in the transverse plane at right angles to the direction of propagation (the XY plane in Fig. 14.46), while there is a magnetic component along the direction of propagation (the Z axis). This type of wave is sometimes called an *H-wave*. In the latter,

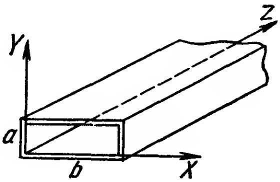


Fig. 14.46. Sketch of rectangular waveguide

it is the magnetic field that is wholly in a plane transverse to the direction of propagation, and there is a component of electric field in the direction of propagation. This form of wave is sometimes called an *E-wave*.

Two subscripts are used to designate a particular mode in either kind of modes designated as TE or TM. The first subscript designates the number of half-wave variations, or half-sinusoids in the electric field distribution along the dimension a in Fig. 14.46, and the second, the number of half-wave variations along the dimension b. If one of the subscripts is zero, the field varies only along one of the sides of the waveguide. Thus, a TE_{01} mode designates a field pattern in which the electric field is always transverse to the direction of propagation and in which the electric field has zero half-wave variations across the narrow dimension and one across the wide.

Waveguide mixers mainly use the TE mode. The TE_{01} mode is excited by a probe which should be arranged parallel with the electric field vector E. In practical circuits, the probe is an extension of the inner conductor of the local-oscillator output coaxial line.

The construction of a waveguide mixer for 10 gigahertz (3 cm) is shown in Fig. 14.47. At the input to the mixer there is a TR switch *1*, coupled to the mixer waveguide *3* through a diaphragm or iris *2*. The crystal diode *4*, mounted in a cartridge *9*, is transverse to the axis of the waveguide and longitudinal with the electric field vector E. Impedance matching between the crystal and the waveguide is effected by a piston *5* placed at a quarter of a wavelength from the axis of the crystal. As a result, the wave reflected from the short-circuited end of the waveguide (the piston) arrives at the crystal in phase with the direct wave, so that the signal energy is fully absorbed by the crystal.

The local oscillator is a klystron *6*, the output coaxial line of which is inserted into a socket provided in the broad wall

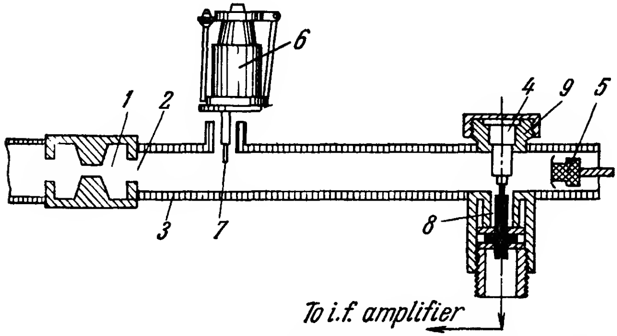


Fig. 14.47. Construction of 10,000-MHz single-cavity mixer

of the waveguide. The waveguide is excited by a probe *7* which is an extension to the inner conductor of the klystron output lead. The amount of energy injected into the waveguide (the degree of coupling) is determined by the length of the probe inserted into the waveguide.

It is important that the local-oscillator signal should not find its way into the cavity of the TR tube. Because of this, the input impedance of the waveguide in the direction of the TR tube should be very high. Such an impedance is obtained by placing the probe at a distance equal to an odd number of quarter-wavelengths, $\frac{(2n-1)\lambda}{4}$, from the klystron.

In the crystal diode, the electromagnetic waves due to the incoming signal and the local oscillator are mixed together to produce an intermediate frequency which is coupled out through a quarter-wave choke *8* and a coaxial cable to an i.f. amplifier. The coaxial cable is connected to the mixer by a pin screwed onto the threaded end of the crystal holder.

With this type of mixer, the klystron local oscillator operates under adverse conditions owing to the presence of a reactive load. The point is that the energy absorbed by the crystal is only a small fraction of the energy generated by the klystron. The remainder is wasted as standing waves in the klystron line made up of the probe and the coupling element of the cavity resonator within the klystron. Standing waves are equivalent in their effect to a reactive load. This upsets operation of the klystron and may even kill its oscillation. This disadvantage is non-existent in double-cavity crystal mixers.

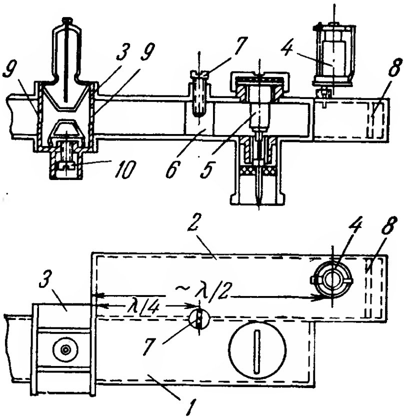


Fig. 14.48. Construction of double-cavity mixer

Double-cavity Crystal Mixer. The construction of a double-cavity crystal mixer is shown in Fig. 14.48. The mixer consists of a signal cavity *1* and a local-oscillator cavity *2*. The narrow sides of the cavities are adjacent and form a common wall. The cavities are excited, respectively, by the incoming signal through a TR tube *3* and by a klystron *4*. The crystal diode *5* is placed in the signal cavity. The local-oscillator frequency is applied to the diode through a coupling aperture *6* in the common wall. The amount of power fed through the coupling aperture is adjusted by a screw *7*. The unused power from the local oscillator is absorbed by a strip attenuator *8*, serving as a matched load for the local oscillator, preventing the generation of standing waves in the klystron line and, consequently, eliminating the shortcoming of the single-cavity mixer.

The TR tube employed in the mixer has an integral resonator.

The signal is injected through the glass windows *9* in the TR tube and transferred into the cavity. The resonator is tuned with a screw *10*.

Review Questions

1. What type of klystron is used in the mixers of 3-cm radar equipments?
2. How is the crystal diode arranged relative to the electric-field vector in a cavity mixer?

3. How does an increase in the power of hte local-oscillator signal affect operation of the crystal diode?
4. Why does the local-oscillator signal produce standing waves in the cavity mixer?

83. Balanced Mixers for the SHF Band

Radar receivers operating at 10 gigahertz mainly use balanced mixers which reduce receiver noise to one-half or even one-third of that in single-ended, or unbalanced, mixers.

In the SHF band, mixer noise comes from the crystal diode and the local oscillator. In the balanced circuit, it is reduced owing to the suppression of local oscillator noise.

Figure 14.49 shows the circuit of a balanced mixer. It is called so because it has a balanced input, a balanced output, and diodes connected in push-pull. The local-oscillator signal is applied to the diodes in antiphase, and the incoming signal in phase.

In the diagram of Fig. 14.49, the polarities of voltages and the directions of currents are shown for a single half-cycle of the signal and local-oscillator inputs. Since the signal is applied to the diodes in phase, the i.f. currents appear to be in opposite directions in the transformer primaries. The secondary coils L_2' and L_2'' are connected in opposition, and the electromotive forces induced in them will add together. Since the local-oscillator voltage is fed to the diodes in antiphase, for the polarities shown the oscillator currents and the accompanying noise currents will pass through only the upper primary coil, so that the noise emf induced in the secondary circuit will cancel each other.

The key component in the structure of the balanced microwave mixer is a *hybrid junction*, also called the *magic-tee* (Fig. 14.50). The hybrid junction consists of the main waveguide and two branch guides connected at right angles to the main one. One of the branch guides is joined to a broad side and the other to

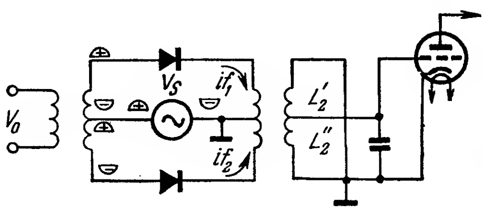


Fig. 14.49. Schematic of balanced microwave mixer

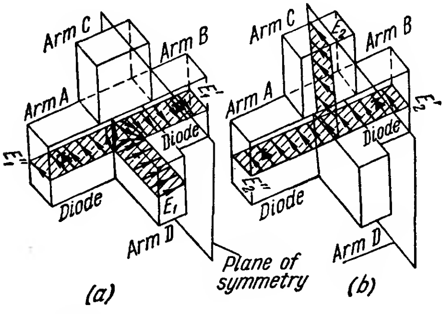


Fig. 14.50. Propagation of electrical field in the arms of a magic-tee

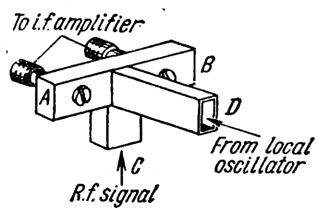


Fig. 14.51. Balanced microwave mixer

a narrow side. The hybrid junction has four arms. A and B are output arms, while C and D are input arms. The input arms accept signals from oscillators and the output arms are coupled to loads. For normal operation of the hybrid junction the output arms should be properly terminated. That is, the arms A and B must be of equal length and be loaded into similar matched loads.

Let us see how the hybrid junction operates when excited in a TE mode by an oscillator connected only to arm D. In Fig. 14.50a the propagation of the electric field is illustrated by an imaginary plane in which the lines of force are indicated by arrows. The electric field E_1, established by the oscillator in arm D, is divided at the junction into two fields, E_1' and E_1''. At equal distances from the plane of symmetry these fields in the arms A and B are 180° out of phase.

Thus, in terms of phase inversion the hybrid junction is equivalent to the input transformer in the circuit of Fig. 14.49.

Now let us take a case where the oscillator is connected to the arm C (Fig. 14.50b). The propagation of the electric field is indicated by lines of force in two mutually perpendicular planes. At the junction of the arms the field E_2 is divided into two fields, E_2' and E_2''. However, the fields in the arms remain in phase at equal distances from the plane of symmetry.

When the arms C and D are excited simultaneously, the electromagnetic waves in the output arms A and B will be in phase when set up by one oscillator, but they will be in phase opposition when they are due to the other oscillator.

These properties of the hybrid junction make it suitable for use as a balanced microwave mixer.

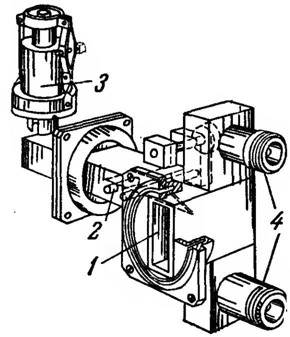


Fig. 14.52. External view of a balanced microwave frequency changer

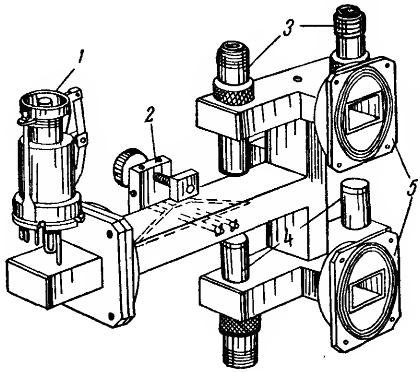


Fig. 14.53. Double balanced microwave frequency changer

The construction of a mixer with a balanced output is shown in Fig. 14.51. As is seen, both arms are excited in a *TE* mode. In this case, the oscillator signal is fed into the arm *D*, and the incoming signal into the arm *C*.

Figure 14.52 shows a complete frequency changer (including a balanced mixer and a local oscillator). The input arms are terminated in flanges which connect them to a klystron local oscillator and the incoming-signal guide. The aerial signal is injected into port *1*. The opposite port joined to the klystron *3*, holds a dissipative attenuator *2* which serves to control the power fed from the local oscillator to the mixer crystal. The crystal cartridges *4* are located in the output arms of the mixer.

SHF radar receivers with double-channel AFC use frequency changers employing double-balanced mixers.

In fact, this is a combination of two balanced mixers coupled to a common local oscillator. An example of such a frequency changer is shown in Fig. 14.53 where *1* is a reflex klystron; *2* is an attenuator; *3* and *4*, crystal catridges. The signal and local-oscillator ports are at the flanges *5*.

Review Questions

1. What noise component is suppressed in a balanced microwave mixer?
2. What is the hybrid junction of Fig. 14.49 equivalent to?
3. What radar receivers use double balanced mixers?

84. I.F. Amplifiers for VHF and UHF Receivers

The i.f. amplifier of a radar receiver should amplify r.f. signals of both short and long duration. Consequently, in contrast to the i.f. stages of ordinary receivers, the amplifier should be able to pass a band of frequencies from one to a few tens of megahertz.

Owing to the high sensitivity of radar receivers, the i.f. gain is 10^4 to 10^6. The i.f. amplifiers of radar receivers use high-mu pentodes and transistors.

The wide-band i.f. amplifiers used in radar receivers may be *synchronously tuned* amplifiers (amplifiers with the *single tuned* circuit in each stage tuned to the same frequency), *double stagger-tuned* amplifiers, *triple stagger-tuned* amplifiers, and amplifiers using mutually coupled (double-tuned) circuits between stages. Consider each of these amplifier circuit configurations.

Synchronously Tuned Amplifiers. These are the usual tuned-circuit amplifiers. Their advantage is the fact that the tuning is standard; their disadvantage consists in the low stage gain and the difficulty of obtaining a wide bandwidth.

From tuned-amplifier theory we know that the overall frequency response of n amplifier stages is

$$Y_{overall} = Y^n$$

where Y is the ordinate of the resonance curve of one stage.

With a large number of stages, the overall resonance curve of the amplifier becomes sharper and the bandwidth narrower. This means that in order to obtain the required bandwidth in the amplifier, each stage should have a broad bandwidth. With a broad bandwidth the Q of the tuned circuit and the stage gain are low.

Double Stagger-tuned Amplifiers. In these amplifiers, the odd stages are tuned to one frequency and the even ones to another. As a result, each pair is stagger-tuned. A block diagram of such an i.f. amplifier and the resonance curves of one pair of stages are shown in Fig. 14.54. The overall frequency response for one pair is obtained by multiplying the ordinates of the resonance curves of the individual stages; the overall resonance curve can have one or two humps. Stagger-tuned i.f. amplifiers have a flat frequency response and a high stage gain.

Triple Stagger-tuned Amplifiers. The number of stages in amplifiers of this type is a multiple of 3. In each triple group,

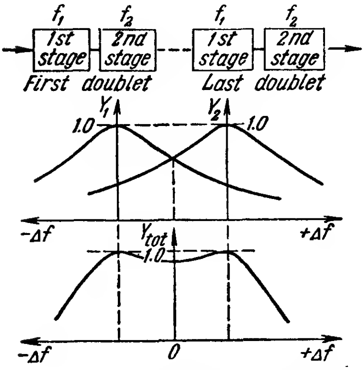


Fig. 14.54. Block diagram of wide-band double stagger-tuned i.f. amplifier and resonance curves of one doublet of stages

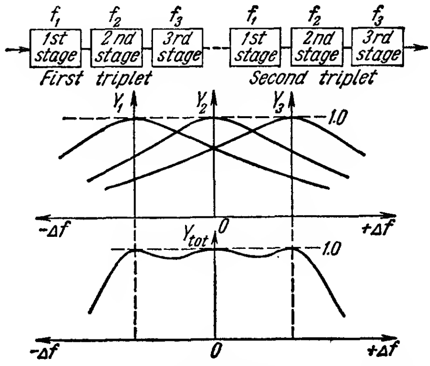


Fig. 14.55. Block diagram of a wide-band triple stagger-tuned i.f. amplifier and resonance curves of one triplet

one is tuned to the centre frequency, while the other two are tuned to the frequencies removed an equal amount from the centre frequency. The block diagram of an i.f. amplifier and its resonance curves are shown in Fig. 14.55.

Such a circuit provides a still flatter frequency response and a greater gain in comparison with the double stagger-tuned circuit.

The triple stagger-tuned arrangement makes the i.f. amplifier more difficult to tune as compared with other circuit configurations.

Figure 14.56 shows a stage with a single-tuned circuit. This configuration is typical of i.f. amplifiers employing single-tuned circuits. In contrast to the usual tuned amplifier, it uses parallel feed. The coil L in the grid circuit speeds up the recovery of amplification after strong interference. The tuned-circuit capacitance is made up of the input and output capacitances of the valves and wiring. R_{sh} connected across the decoupling choke L_{ch} broadens the bandwidth. $L_H C_H$ is the decoupling filter in the heater circuit of each valve.

Double-tuned Amplifiers. These amplifiers do not differ in any respect from the band-pass amplifiers discussed in Chapter VI. Such circuits offer a flat frequency response and considerable gain per stage.

The circuit of a double-tuned amplifier is shown in Fig. 14.57. To broaden the bandwidth shunting resistors R_{sh1}

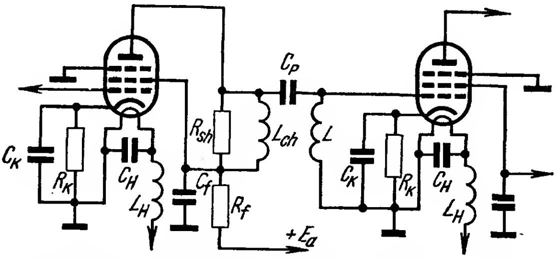


Fig. 14.56. Basic circuit of i.f. amplifier stage employing a single tuned circuit

and R_{sh2} are connected across each side of the double-tuned circuit.

In VHF and UHF receivers the triode or pentode mixer is usually loaded into a tuned circuit similar to one in the succeeding i.f. amplifier stages. Therefore, the mixer stage may be considered to be an amplifier stage which is a part of the entire i.f. circuit, differing from the other stages only in the mutual conductance and output capacitance of the valve.

In wide-band i.f. amplifiers it is important to secure the stability of frequency response, bandwidth and gain.

On the other hand, in wide-band amplifiers operating at a high intermediate frequency (tens of megahertz) the capacitance of the tuned circuits becomes small and is chiefly determined by the interelectrode capacitances of the valves. Because of this, valve replacement in such amplifiers may considerably change amplifier parameters.

Of the circuit configurations discussed the best stability of gain and frequency response is offered by the double-tuned circuit arrangement.

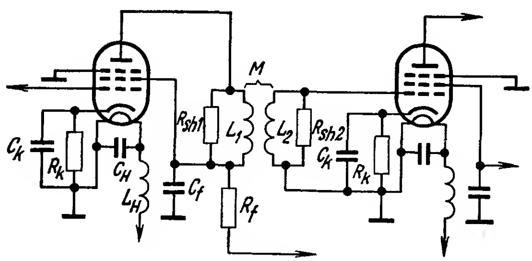


Fig. 14.57. Basic circuit of double-tuned i.f. amplifier

Review Questions

1. What factor affects the shape of the resonance curve of a double stagger-tuned amplifier?
2. How will a decrease in R_{sh} affect the bandwidth of the circuits shown in Figs. 14.56 and 14.57?
3. How will the shape of the resonance be affected by a reduced spacing between the coils in the double-tuned circuit?

85. I. F. Amplifiers for SHF Receivers

From analysis of the functional units of radar receivers we have established that an i.f. amplifier for the SHF band consists of a preliminary amplifier and a main amplifier. In its circuitry, the main i.f. amplifier does not differ in any respect from the i.f. amplifiers employed in the VHF and UHF receivers discussed earlier. On the other hand, the i.f. preamplifier, being the input of the i.f. amplifier, mainly determines the noise quality of the receiver; this is the decisive factor in the selection of the circuit configuration for it.

The i.f. preamplifier comprises a coupling network which may be single- or double-tuned, and a low-noise amplifier. The first two stages of the i.f. preamplifier (see Fig. 14.24) typically are earthed-grid circuits. The third and, usually, the last stage of the i.f. preamplifier have a cable transmission line to convey the i.f. signal to the main i.f. amplifier. The coupling network which applies the i.f. signal from the crystal mixer to the grid of the first amplifier valve is in fact a matching transformer.

Correct selection of the coupling network and of its components affects not only the energy transfer to the input of the first valve, but also the noise factor of the i.f. amplifier. These requirements are generally met by a double-tuned circuit which is widely used in radar receivers. In the most commonly used circuit configuration the coupling between the primary and secondary side is by mutual inductance (Fig. 14.58).

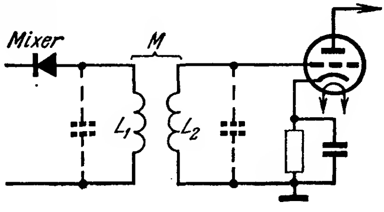


Fig. 14.58. Double-tuned coupling input stage network of i.f. amplifier, employed in a SHF receiver

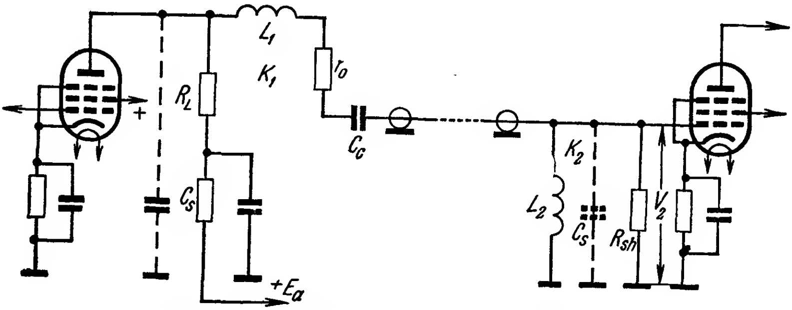


Fig. 14.59. An i.f. preamplifier stage with a coaxial transmission line

With a balanced mixer, the circuit of Fig. 14.49 is employed.

The double-tuned coupling network, as compared with the single-tuned one, secures a broader bandwidth at the same noise level, which is very important for a wide-band receiver. Such a coupling network usually has an unsymmetrical resonance curve, because each tuned circuit is shunted by a different impedance, one by that of the mixer output, and the other by that of the next valve input. The difference in parameters between valves is also responsible for the lack of symmetry.

Consider operation of the i.f. preamplifier stage employing a coaxial transmission line shown in Fig. 14.59. The transmission line is a length of r.f. cable inserted in the anode circuit of the valve in series with the tuned-circuit coil through a coupling capacitor. This cable conveys the r.f. energy from the i.f. preamplifier to the main i.f. amplifier. The input of the first valve in the i.f. amplifier is connected to the output of the cable. For maximum power transfer the cable must be loaded into an impedance equal to the wave impedance of the feeder, i.e.

$$\rho_f = R'_{0e2}$$

where R'_{0e2} is the equivalent resonant resistance of the tuned circuit K_2, with allowance for the fact that it is shunted by the shunt resistance R_{sh} and the input resistance R_{in} of the valve. R'_{0e2} may be found from Eq. (6.5).

Then the condition for a perfect match is

$$\rho_f = \frac{R_{0e2}}{\frac{R_{0e2}}{R_{in}} + \frac{R_{0e2}}{R_{sh}} + 1} \tag{14.39}$$

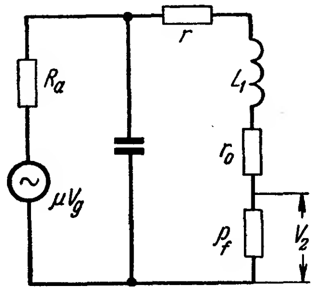


Fig. 14.60. Equivalent circuit of a stage with a coaxial transmission line

The matching condition is satisfied by the selection of the proper value of R_{sh}

$$R_{sh} = \frac{R_{0e2}}{\frac{R_{0e2}}{\rho_f} - \frac{R_{0e2}}{R_{in}} - 1} \tag{14.40}$$

where the resonant resistances of both tuned circuits may be assumed to be equal, $R_{0e1} = R_{0e2}$.

In analysis of the resonance properties of such a stage it should be remembered that the tuned circuit K_2 is shunted by the low wave impedance of the cable. Due to this, its bandwidth, as compared with that of the whole i.f. amplifier, is widened 3 to 4 times. Therefore, in the equivalent circuit, the stage along with the tuned circuit K_2 may be replaced by the wave impedance of the cable ρ_f. The equivalent circuit thus obtained (Fig. 14.60) is a single-tuned circuit, which, as far as the resonance properties are concerned, is the same as the stage with a cable transmission line.

The effective or loaded Q of such a tuned circuit is:

$$Q_{ef} = \frac{1}{2\pi f_0 C_1 (r + r_0 + \rho_f)} \tag{14.41}$$

where:

$C_1 = C_c + C_{out} + C_w$

r = loss resistance of the coil (these losses are frequently neglected)

r_0 = resistance required to secure the necessary value of Q_{ef}.

To determine the stage gain K_0, the output voltage at resonance is found

$$V_2 = I_c \rho_f$$

The current in a parallel resonant circuit at resonance is $I_c = Q_{ef} I_a$. In turn, the anode current will be $I_a = \frac{\mu V_g}{R_a}$.
After substitutions, the gain will be

$$K_0 = \frac{V_2}{V_g} = \frac{\mu}{R_a} Q_{ef}\rho_f = g_m Q_{ef}\rho_f \tag{14.42}$$

Now we are in a position to discuss the design of the i.f. amplifier in greater detail. In the SHF band, this involves the calculation of the noise factor of the i.f. amplifier, design of the double-tuned input stage, and design of the amplifier proper.

Review Questions

1. What is the main function of the i.f. preamplifier in SHF receivers?
2. How is the output impedance of the mixer matched to the input impedance of the 1st stage in the i.f. amplifier?
3. What type of resonant circuit is equivalent to a stage with a coaxial transmission line?

86. The Noise Factor of I.F. Amplifiers in SHF Receivers

As mentioned in Sec. 77, the noise factor of an i.f. amplifier with the cascode arrangement at its input depends on the noise of the first stage. In turn, the noise of the first stage is determined to a considerable degree by the components of the input stage and by its operating conditions. In analysis of the noise properties of an i.f. amplifier it is usual to consider coupling network noise, first-valve noise, and also fluctuation noise in the output resistance of the crystal mixer.

Thus, all the "noisy" components form a noise circuit shown in Fig. 14.61. For convenience, it is drawn up in terms of conductances. Instead of a constant-voltage generator in series

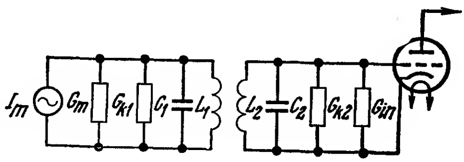


Fig. 14.61. Equivalent noise circuit of i.f. amplifier coupling network

with its internal resistance, it uses a constant-current generator in parallel with its internal conductance, and all resistances are replaced by conductances. The coupling network is a double-tuned circuit.

In the equivalent circuit of Fig. 14.61, and in the subsequent calculations, the following notation is used:

C_1, C_2, L_1, L_2 are the capacitances and inductances of the tuned circuits;

I_m is the current generator representing the mixer;

$G_m = \frac{1}{R_m}$ is the conductance of the mixer;

$G_{K1} = \frac{1}{R_{0e1}}$ is the resonant conductance of the primary;

$G_{K2} = \frac{1}{R_{0e2}}$ is the resonant conductance of the secondary;

$G_{in} = \frac{1}{R_{in}}$ is the input conductance of the first-stage valve;

$m = \frac{V_1}{V_2}$ is the transformation ratio of the coupling network.

The output resistance of the mixer is usually small, about 400 ohms. Therefore the resonant conductance G_{K1} of the primary is negligible in comparison with that, G_m, of the mixer.

According to Vilensky, the minimum noise factor of the circuit in question may be obtained when the output resistance of the mixer is matched to the input resistance of the valve and the resonant conductance of the secondary has a definite optimum value G_{K2opt}. These considerations determine the procedure for the calculation of the noise factor.

Given:

1. Type of valve in the first stage.
2. Type of crystal mixer.

To Find:

1. The optimum resonant conductance of the secondary of the coupling network, G_{K2opt}.
2. Transformation ratio of the coupling network in the case of perfect match, m_m.
3. Minimum noise factor, N_{if}, of the i.f. amplifier.

Design Procedure:

1. Find the optimum resonant conductance of the secondary from considerations of the minimum noise factor

$$C_{K2opt} = G_{in}\left(\frac{1}{\sqrt{R_n G_{in}}} - 1\right) \qquad (14.43)$$

where R_n is the noise resistance of the first valve.

2. Determine the transformation ratio of the coupling network from considerations of matching:

$$m_m = \sqrt{\frac{G_{in} + G_{K2opt}}{G_m}} \qquad (14.44)$$

3. Find the minimum noise factor of the i.f. amplifier for perfect match

$$N_{if} = 2 + 8\sqrt{R_n G_{in}} \qquad (14.45)$$

87. Design of a Double-tuned Coupling Network for I. F. Amplifiers of SHF Receivers

In the design of a double-tuned coupling network, it is necessary to select its parameters and the coupling between them in such a way that the noise factor of the i.f. amplifier has a minimum value (determined in Sec. 86). Besides, the coupling network should have the specified bandwidth.

Given:

1. The optimum resonant conductance, G_{K2opt}, of the secondary.
2. Transformation ratio of the coupling network for perfect match, m_m.
3. Bandwidth, $2\Delta F_{in\ cir}$.

To Find:

1. Capacitances of the tuned circuits.
2. Inductances of the tuned circuits.
3. Q of the tuned circuits.
4. Optimum coupling.
5. The resonance curve and bandwidth.

Design Procedure:

1. Find the total capacitances of the tuned circuits from considerations of stable bandwidth after valve replacement, assuming that the spread in capacitances is $0.1\,(C_{out}+C_{in2})$

$$C \geqslant \frac{C_{out}+C_{in2}}{10\delta}\,\frac{f_i}{2\Delta F} \tag{14.46}$$

where δ is the coefficient allowing for the coupling circuit configuration. For a double-tuned circuit, $\delta = 0.4$ to 0.5.

Assume that $C=C_2$. Determine the capacitance of the tuned-circuit capacitor C_K defined as

$$C_K = C_2 - (C_{in2} + C_{w2})$$

where C_{in2} = input capacitance of the first valve
C_{w2} = capacitance of the wiring taken equal to 3-5 picofarads.

Assuming from structural considerations that the tuned-circuit capacitor C_K has the same capacitance on both sides, the total capacitance of the primary will be

$$C_1 = C_K + C_{xtal} + C_{w1} \tag{14.47}$$

where $C_{xtal} \cong 10$ picofarads is the output capacitance of the crystal diode
$C_{w1} \cong 2\text{-}3$ picofarads is the distributed capacitance of the wiring.

2. Find the optimum unloaded Q of the secondary Q_{2opt}, securing the minimum noise factor in the i.f. amplifier

$$Q_{2opt} = R_{oe2opt}\,\omega_0 C_2 \tag{14.48}$$

where

$$R_{oe2opt} = \frac{1}{G_{K2opt}}$$

3. Find the loaded Q of the secondary, considering the shunting effect of the valve

$$Q_{2ef} = \frac{Q_{2opt}}{1+\dfrac{R_{oe2opt}}{R_{in}}} \tag{14.49}$$

4. Find the loaded Q of the primary, considering the shunting effect of the crystal diode

$$Q_{1ef} = \omega_0 C_1 R_{oe1}$$

Since

$$R'_{0e} = \frac{R_{0e1} R_{mixer}}{R_{0e} + R_{mixer}}$$

and noting that

$$R_{0e1} \gg R_{mixer}$$

we obtain

$$R'_{0e} \gg R_{mixer}$$

Therefore

$$Q_{1ef} = \omega_0 C_1 R_{mixer} \quad (14.50)$$

In the calculation of the coils the unloaded Q of the tuned circuit should be taken equal to Q_{2opt}.

5. Check the bandwidth of the coupling network

$$2\Delta F_{cn} = f_0 \frac{\frac{1}{Q_{1ef}} + \frac{1}{Q_{2ef}}}{\sqrt{2}} \quad (14.51)$$

It should not markedly differ from the value stipulated in the specifications.

6. Determine the inductance of the tuned circuits:

$$L_1 = \frac{1}{\omega^2 C_1}$$

$$L_2 = \frac{1}{\omega^2 C_2}$$

7. Find the coefficient of coupling between the tuned circuits securing the minimum noise factor

$$k_{opt} = \frac{m_m G_{mixer}}{\omega \sqrt{C_1 C_2}} \quad (14.52)$$

The next step in design is the i.f. amplifier usually comprising a preamplifier and a main amplifier. In the Soviet Union, the most commonly used method is one due to Volin discussed in detail in his *I. F. Amplifiers*. His method assumes that an amplifier consists of a number of stages each of which has the same gain that would be had by an amplifier with a bandwidth equal to that of the whole amplifier. The gain of such a stage is referred to as *unit gain*.

Volin's method is equally applicable to amplifiers using single-tuned circuits, double-tuned circuits, and their combinations.

Calculations are reduced to the use of tables and charts. It may be added in passing that in cases involving the cascode configuration and also a stage with a coaxial transmission line to the i.f. preamplifier, both should be treated as those using single-tuned circuits. The unit gain of a stage with a coaxial transmission line is given by

$$K_{unit} = g_m f_0 \rho_f / 2\Delta F \tag{14.53}$$

derived from Eq. (14.42) after substitution of $f_0/2\Delta F$ for Q_{ef}.

For the cascode circuit, the unit gain is

$$K_{unit} = g_{m1} / 2\pi C_2 (2\Delta F) \tag{14.54}$$

In both cases,

$2\Delta F$ = bandwidth of the entire amplifier
g_{m1} = mutual conductance of the primary
C_2 = capacitance of the secondary.

88. Video Detectors

In a radar receiver, a video detector is a stage where the intermediate-frequency signals are converted into signals at the video frequency. In other words, the r.f. pulses are converted into d.c. voltage pulses, known as *video pulses* (Fig. 14.62). The non-linear elements used in video detectors may be valves and crystal diodes.

A video detector may be of the anode-bend, infinite-impedance, or diode type.

In the detection of rectangular pulses it is important to preserve the shape of the envelope of the detected signal. In this

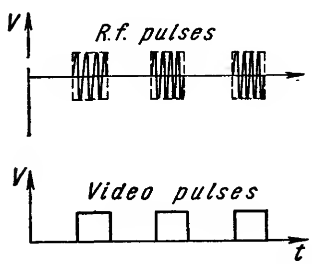


Fig. 14.62. R.f. pulses and video pulses

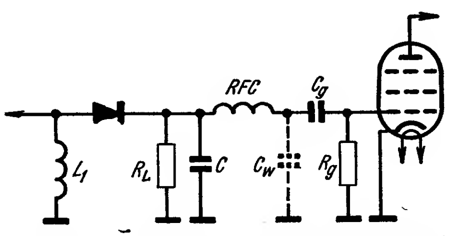


Fig. 14.63. Circuit of a single-ended diode detector

respect, the diode detector is best, because its detection characteristic remains linear within comparatively broad limits, from 1 or 2 volts to tens of volts. However, the diode detector suffers from certain disadvantages. For one thing, in contrast to anode-bend and grid-leak detectors, it does not amplify (its voltage gain is less than unity). Its input resistance is very low, because of which the tuned circuit of the last i.f. stage has to sustain a considerable load.

Yet, the diode detector is widely used for video detection in both single-ended and push-pull configurations.

Single-ended Diode Video Detector. The r.f. signal pulses are taken from the tuned circuit of the last i.f. stage whose coil L_1 is shown in the diagram of Fig. 14.63. The diode load is R_L, bypassed for the intermediate frequency by C. The video pulse is applied to the grid of the following stage through a filter RFC/C_w which blocks the intermediate frequency. Before the signal reaches the grid of the limiter valve, it has to pass through the C_gR_g network which transmits short pulses but stops interfering pulses of long duration.

Let us examine in greater detail the pulse waveform at the detector output, i.e. across the detector load. The detection of a rectangular radio pulse is shown in Fig. 14.64. The diode accepts an r.f. wave whose envelope has a rectangular shape. For simplicity, we assume that the static characteristic of the diode is idealized and passes through the origin of coordinates. Hence, when no signal is applied, the diode current is zero. When an r.f. signal appears, the usual detection process takes place by current cutoff, and r.f. current pulses appear in the anode circuit.

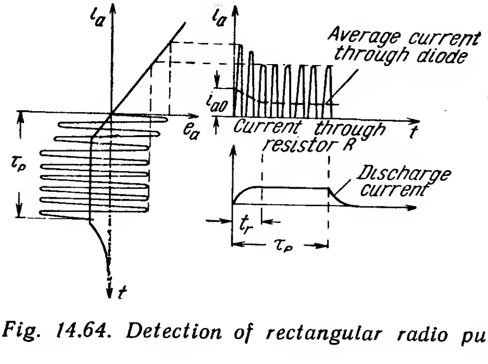


Fig. 14.64. Detection of rectangular radio pulses

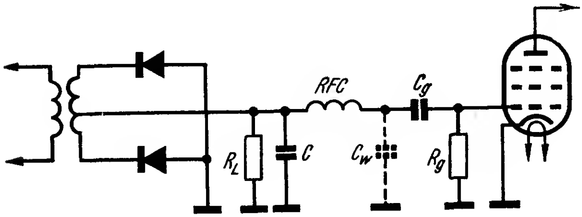


Fig. 14.65. Push-pull diode detector

The highest average current i_{a0} flowing through the diode corresponds to the first anode current pulse. This current only charges the capacitance connected across the load R_L, while the current flowing through the load is equal to zero. On the following pulses, only a part of the detected current goes to charge the capacitance; the remaining part of the current flows through the resistance, building up across it a bias voltage which shifts the operating point to the left. After a certain time interval t_r, which is the *pulse rise time*, the system attains a state of equilibrium, the capacitor is fully charged, and the whole of the diode current is flowing through R_L. The pulse rise time is chiefly determined by the time constant $\tau_1 = R_{a(d)}C$, where $R_{a(d)}$ is the dynamic resistance of the diode. When the pulse ceases, the capacitor C begins to discharge through R_L. As a result, the current will cease flowing through R_L after the time it takes the capacitor to discharge. The pulse fall time is determined by the time constant $\tau_2 = R_LC$.

Thus, the video pulse differs from the i.f. pulse by a steeper leading edge and a more gradual trailing edge. Diodes for video detection should have a low dynamic resistance $R_{a(d)}$ and a low anode-cathode capacitance.

Push-pull Diode Detector. This circuit (Fig. 14.65) differs from the single-ended one in that it uses a balanced input transformer and two diodes. The push-pull circuit is advantageous in that the i.f. voltage at the detector output is considerably lowered, since the first harmonics of the i.f. current from each arm of the circuit flow through the load in opposite directions, cancelling each other. The push-pull circuit has a higher input resistance.

The shortcomings of the circuit are its complexity and lower voltage gain, as compared with the single-ended configuration.

Review Questions

1. What is the major advantage of the diode detector over other types of detectors?
2. What is the advantage of a balanced detector over an unbalanced one?
3. How will an increase in the C of Fig. 14.63 affect the leading edge of the detected pulse?

89. Calculation of a Single-ended Video Detector

Given:

1. Pulse duration, τ_p.
2. Detector input voltage, V_{in}.
3. Pulse rise time, t_{rd}, in the detector.
4. Intermediate frequency, f_i.
5. Equivalent resonant resistance of the i.f. amplifier, R'_{0e}, with allowance for the shunting effect of the detector.

To Find:

1. Type of valve.
2. Circuit parameters C, R_L, RFC, C_g and R_g (see Fig. 14.63).
3. Detector gain, K_d.
4. Detector input resistance, $R_{d\,in}$.
5. Detector output voltage, $V_{d\,out}$.

Design Procedure:

1. Select the valves or crystal diodes from a manual.
2. Determine the capacitance C bypassing the load, assuming that the r.f. voltage is applied directly to the diode. For this purpose, C must be such that

$$C \geqslant 10C_d$$

3. Determine the load resistance such that the pulse due to the discharge of C through R_L has a sufficiently steep trailing edge, that is:

$$R_L \leqslant \frac{t_f}{2.2C} \tag{14.55}$$

where the fall time of the pulse is $t_f \leqslant 0.2\tau_p$.

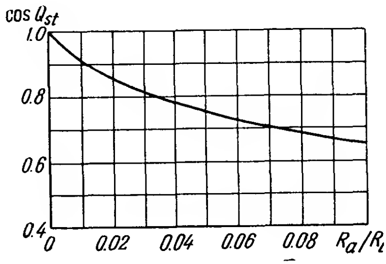


Fig. 14.66. Plot of cos Θ_{st} as a function of R_a/R_L

4. Find R_a/R_L and determine $\cos\theta_{st}$, the steady-state value of $\cos\theta$, from the plot of Fig. 14.66.
5. Find the input resistance of the detector under steady-state conditions from

$$R_{d\,in} \cong \frac{R_L}{2\cos\theta_{st}} \tag{14.56}$$

6. Determine the necessary resonant resistance of the tuned circuit of the preceding stage so that

$$R_{0e} = \frac{R'_{0e}}{1-\dfrac{R'_{0e}}{R_{d\,in}}} \tag{14.57}$$

From the value of R_{0e}, find the damping factor of the unloaded tuned circuit

$$d = \frac{1}{\omega_0 C_K R_{0e}}$$

which must be realisable. If d is too small, it should be increased by changing the parameters of the tuned circuit.
7. Find the pulse rise time

$$t_{rd} \cong 5C\left[2R_a + R'_{0e}\left(1-\frac{2R_a}{R_{d\,in}}\right)\right]\cos\theta_{st} \tag{14.58}$$

The value thus obtained (t_{rd}) should not markedly differ from the specified value. If it happens to be much smaller, the capacitance C should be increased along with R_L, in conformity with Eq. (14.55). If t_{rd} exceeds the specified one, it will be a good plan either to decrease R_l or to use tapped-down connection between the preceding stage and the detector. To decide which is preferable, it is necessary to find the critical value of the coupling parameter

$$m_{crit}=\sqrt{\frac{2R_a}{R'_{0e}\left(1-\frac{2R_a}{R_{d\,in}}\right)}} \tag{14.59}$$

If $m_{crit}<1$, a tapped-down connection should be used.

Using Eq. (14.58), determine the new values of rise time at different values of m, up to $m=m_{crit}$, with R'_{0e} substituted for $R''_{0e}=m^2R'_{0e}$. Regardless of how t_{rd} has been reduced, adjust the previously obtained values of $\cos\theta_{st}$, $R_{d\,in}$ and R_{0e}.

8. Determine the gain of the detector

$$K_d=m\cos\theta_{st} \tag{14.60}$$

9. Find the voltage at the detector output

$$V_{d\,out}=K_dV_{in} \tag{14.61}$$

10. Find the inductance of the r.f. choke

$$L_{RFC}=\frac{1}{4\pi^2f^2C_{RFC}} \tag{14.62}$$

where f is the resonant frequency of the choke

$$f=(0.5\text{ to }0.7)f_i$$

and C_{RFC} is the capacitance of the choke, equal to about two to four picofarads.

11. Determine the parameters of the coupling circuit.

On putting $R_g=0.3$ to 1 megohm, the capacitance will be given by

$$C_g=\frac{\tau_p}{GR_g}$$

where G is the limiting droop of the pulse top, equal to 0.05-0.2.

90. Automatic Control in Radar Receivers

Automatic Gain Control. In Chapter XII we have discussed automatic gain control (AGC) used in CW (continuous-wave) work. AGC in pulse receivers has a number of features of its own. For one thing, in pulse radar receivers the wanted signal exists during a relatively short time, determined by the duty cycle of the pulse train. Between the signal pulses, the receiver may pick up unwanted signals which may be of the same frequency as the wanted one, while their amplitude may be equal to, or exceed,

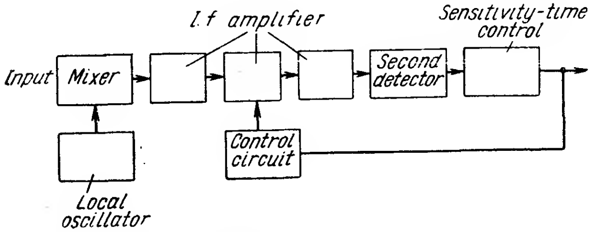


Fig. 14.67. Block diagram of a receiver with sensitivity-time control

that of the signal. To discriminate against such unwanted signals, a more sophisticated AGC system has to be used.

One form of sophistication is what is variously called as *time-gain, gain-time control*, or *sensitivity-time control*. It operates continuously and on a time basis, increasing the gain of the receiver only when an echo is to arrive at the receiver. The block-diagram of a radar receiver using a sensitivity-time-control (STC) circuit is shown in Fig. 14.67.

Now suppose that the receiver picks up both the wanted and a sustained unwanted signal. If the unwanted signal is much weaker than the wanted one, it can be readily discriminated against with manual gain control. If not, there must be some means in the AGC system that would increase the gain quickly so as not to miss the signal and reduce it quickly in order to suppress unwanted signals (also referred to as *clutter*).

When the signal and clutter are allowed to beat together, the beat frequency will have an envelope retaining the waveforms of

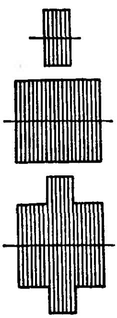


Fig. 14.68. Waveforms of the signal and clutter

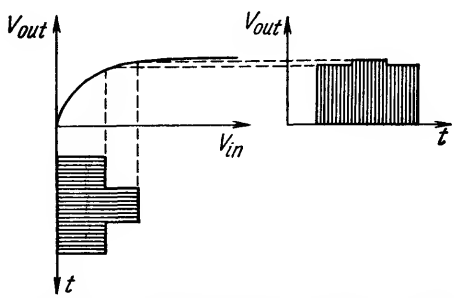


Fig. 14.69. Reception of the signal and clutter with normal AGC

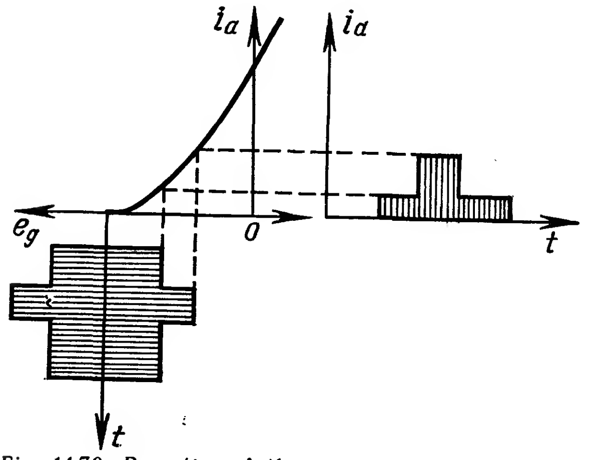


Fig. 14.70. Reception of the signal and clutter with IAGC

both, as shown in Fig. 14.68. If the AGC system were driven by clutter, the signal would be completely drowned at the receiver output, as shown in Fig. 14.69. This is where another form of sophistication is introduced into AGC, known as *instantaneous automatic gain control* (abbreviated to IAGC).

The IAGC circuit responds to variations of the mean clutter level by applying an additional negative bias to the controlled stage so that clutter is reduced and the waveform of the signal envelope is retained (Fig. 14.70).

Obviously, the effect of clutter will be felt only if the difference in intermediate frequency between clutter and signal lies within the passband of the i.f. amplifier. This suggests the form of the IAGC circuit, such as shown in Fig. 14.71 where it covers two stages of the i.f. amplifier.

Referring to Fig. 14.71, the output voltage from the second stage is applied to the input of the detector which generates a negative output voltage. The detector output is fed to the grid of the cathode follower V_3. The voltage developing across the load of the cathode-follower is applied to the grid of V_1 as negative bias. As a matter of record, negative bias may also be additionally applied to the grid of the stage coupled to the detector.

Automatic Frequency Control. Radar receivers may receive either the echoes of the signals transmitted by the same radar station, or the signals radiated by radio beacons.

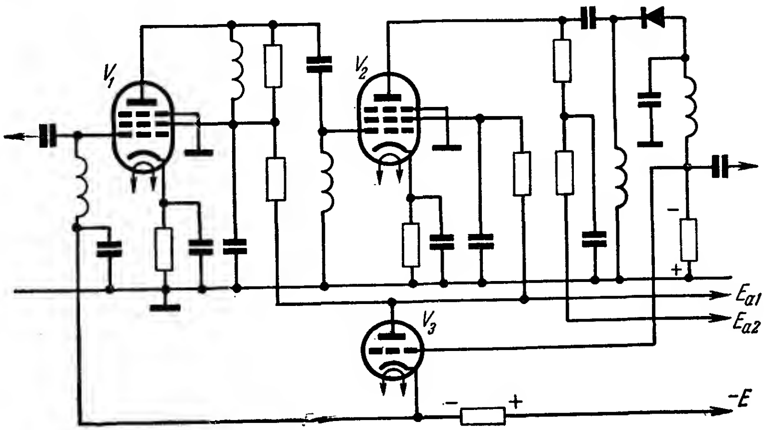


Fig. 14.71. IAGC circuit

In the first group of radar receivers, the function of AFC is to maintain a constant difference in frequency between the transmitter and the local oscillator.

The function of AFC in the second group is that of maintaining the frequency of the local oscillator constant.

If a receiver is designed for both services, its circuitry should have provisions for both functions of AFC.

In pulse SHF radar receivers shown in block-diagram form in Fig. 14.2 there is a separate AFC unit. For its operation, it depends on the fact that the frequency of the klystron local oscillator is a function of the voltage, V_r, at its repeller electrode. However the repeller voltage also affects the power output of the klystron (Fig. 14.72). When the frequency of the local oscillator deviates markedly from its assigned value, the power output of the klystron is considerably reduced. Therefore the operating voltage $V_{r\,(op)}$ at the repeller electrode is selected in such a way that as the klystron frequency varies within the limits necessary for AFC, the power output changes by not more than 50 per cent.

Figure 14.73 shows a simplified AFC circuit for the klystron. The negative voltage at the repeller electrode is made up of the voltage taken from R_2 and the voltage drop across the anode load resistance R_L of valve V_1 which operates as a d.c. amplifier. The control grid of the valve is directly connected to the discriminator output. Hence, any change of voltage at the discriminator

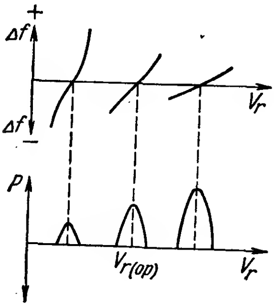


Fig. 14.72. Frequency and power of klystron as functions of repeller voltage

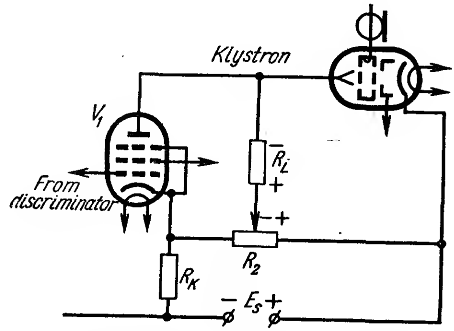


Fig. 14.73. AFC circuit of the klystron with d.c. amplifier

output will change the potential at the grid of V_1, thereby causing a change in the voltage drop across the anode load resistance R_L.

The stage parameters and operating voltages of V_1 are so selected that changes in the voltage drop across R_L will be sufficient to keep the oscillator frequency within the specified limits.

The reliability of AFC depends on the limits within which the oscillator frequency should be controlled and the rate of control. If the transmitter frequency varies more slowly than the AFC circuit can operate, even a considerable shift in frequency will be corrected by the AFC circuit. If the opposite is true, the AFC circuit will be able to perform its function only when the deviation from the assigned frequency is small. Therefore, SHF radar receivers use a circuit which gets around this difficulty. Known as "swept" AFC, it uses a sawtooth generator which swings the local-oscillator frequency over a range determined by its amplitude, so that the intermediate frequency is swept through its assigned value. The sawtooth generator is controlled by pulses from the discriminator. The negative pulses have no effect on it, but the positive pulses terminate the sawtooth wave shortly after it has swept the intermediate frequency through the assigned value. In this way the receiver is held on tune.

The amplitude of the sawtooth wave is chosen so that the total swing does not exceed the acceptance band of the receiver, and the rate of change of the klystron frequency is lower than that of the transmitted pulses.

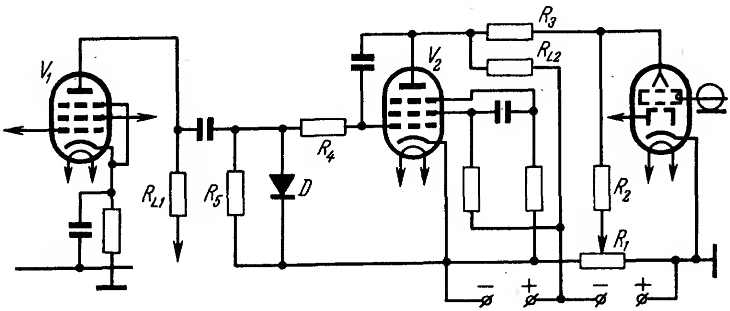


Fig. 14.74. "Swept" AFC circuit using a phantastron sawtooth generator

In some cases, this sawtooth sweeping system is arranged in two steps. When the receiver is well off tune, a large sawtooth voltage swings the klystron frequency slowly and by a wide margin to find the correct frequency. As soon as the correct frequency is found, a second low-amplitude and faster sweep takes over and disables the high-amplitude sweep, while the receiver is constrained to be on tune. The high-amplitude sweep takes over again each time a large frequency shift takes place.

The circuit of a "swept" AFC system is shown in Fig. 14.74. The operating voltage at the repeller electrode of the klystron local oscillator is set mainly with a potentiometer, R_1. The repeller electrode accepts both the negative voltage V_0 and the sawtooth sweeping voltage coming via R_3 from the anode load of the valve V_2 arranged in a phantastron circuit. The phantastron oscillator is controlled by the positive pulses taken from the anode load R_{L1} of the valve V_1. The negative pulses have no effect on V_2, but the positive pulses appearing across R_{L1} cause the diode D to conduct so that a negative voltage develops across R_5. This voltage is fed via R_4 to the control grid of V_2, and the sawtooth wave is terminated.

Review Questions

1. Why do pulse radar receivers use sensitivity-time control?
2. What is the purpose of IAGC?
3. How can the klystron frequency be controlled?
4. What is the purpose of "swept" AFC?

SUMMARY

1. Radar receivers serve to receive weak signals reflected from targets. There are VHF-UHF and SHF radar receivers, differing in circuitry and construction.

2. Radar receivers have a broad bandwidth which is necessary for the reception of short rectangular pulses.

3. Sensitivity of radar receivers in the microwave region is limited by internal (receiver) noise which masks the signal on the indicator screen.

4. The basis of all radar receivers is the superheterodyne principle.

5. Radar receivers use automatic frequency control (AFC). The use of this circuit is necessitated by the frequency instability of the transmitter and of the local oscillator in the receiver.

6. The aerial-input circuits of VHF and UHF receivers not only amplify the signal but also secure a maximum signal-to-noise ratio. This is ensured by matching the aerial to the transmission line and the line to the input circuit of the receiver.

7. Like the aerial-input circuits, the microwave amplifiers must be as noise-free as practicable. Of all amplifier configurations, the noise level is lowest in the earthed-grid amplifier.

8. Frequency conversion in the VHF and UHF bands uses pentode, triode and diode mixers.

9. In the SHF band semiconductor diodes serve as mixers.

10. I.f. amplifiers of radar receivers have a broad bandwidth and high gain.

11. Radio pulses are converted into video pulses by video detectors. Among existing detector circuits use is mostly made of the diode detector.

■ Problems

14.1. Find the available sensitivity $P_{s\,av}$ of a receiver if the noise factor $N = 13$ db, the bandwidth $2\Delta F = 1$ megahertz and $T = 300°$ K.

Answer: $P_{s\,av} = 8.3 \times 10^{-11}$ milliwatt.

14.2. Find the maximum transfer factor of the aerial-input circuit if the input resistance of the valve $R_{in} = 1{,}500$ ohms, $R_A = 50$ ohms.

Answer: $K_{max} = 2.74$.

14.3. Find the loaded Q of the aerial-input circuit if $R_{in} = 1{,}500$ ohms and $\rho = 150$ ohms.

Answer: $Q_{ef} = 5$.

14.4. Determine the maximum gain of a stage employing a 6Ж1П valve, (U.S. equivalent, 6AK5), if the anode load resistance $R_L = 3{,}000$ ohms and $f = 200$ megahertz.

Answer: $K_{max} \cong 6$.

14.5. Find the bandwidth of a stage with $M = 0.7$ and for the data of Problem 14.4, with optimum coupling of the tuned circuit on the grid side. $C_{in} = 4$ picofarads; $C_{out} = 2$ picofarads; $C_w = 8$ picofarads.

Answer: $2\Delta F = 10$ megahertz.

CHAPTER XV

VIDEO AMPLIFIERS

91. General

Video amplifiers are amplifiers which operate in the video-frequency range from d.c. to several megahertz. In some cases, the frequency range of a video amplifier may extend to tens or even hundreds of megahertz.

A video amplifier may be used to amplify pulses in radar and periodic visual signals in television. In a radar receiver the video amplifier is connected between the detector and the CR tube. In TV receivers, it is connected between the detector and the picture tube.

Besides a broad bandwidth, video amplifiers must introduce negligible phase distortion because the eye is very sensitive to changes of phase between the components of a complex light signal.

The electric properties of a video amplifier are evaluated in terms of the frequency and phase characteristics, and the shape of the pulses appearing on the screen. The point is that when a rectangular pulse is applied to the input of a video amplifier (Fig. 15.1), because of the distortion introduced by the amplifier, the screen grid will display pulses with a waveform other than rectangular.

Performance of a pulse amplifier can be analysed in terms of the following pulse characteristics (see Fig. 15.1):

(1) pulse rise time, t_r;
(2) pulse fall or decay time, t_d;
(3) overshoot, ΔV_o;
(4) undershoot, ΔV_u;
(5) pulse sag or droop, ΔV_{st} (Fig. 15.2).

Pulse rise time t_r is the interval between the instants at which the leading edge of the voltage pulse at the output of the amplifier first reaches 10 and 90 per cent of its steady-state

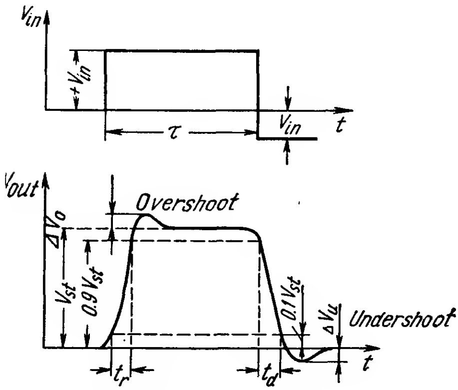


Fig. 15.1. Rectangular pulse and its characteristics

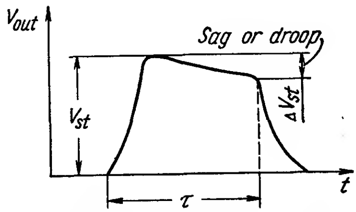


Fig. 15.2. Long rectangular pulse

value V_{st} in response to a step input $+V_{in}$. The total rise time in a multi-stage video amplifier depends upon the pulse rise time of one stage and also the number of stages. The greater the number of stages, the greater the total rise time.

Pulse fall or *decay time* t_d is the interval of time required for the trailing edge of the voltage pulse at the amplifier output to decay from 90 to 10 per cent of its steady state value V_{st} in response to a step input $-V_{in}$.

The rise time t_r and fall time t_d are usually almost equal to each other

$$t_r \cong t_d$$

Overshoot ΔV_o is the initial transient response to a unidirectional change in input, which exceeds the steady-state response.

Voltage overshoot is usually expressed in terms of *percentage overshoot*

$$\delta_v = \frac{\Delta V_o}{V_{st}} 100\%$$

Undershoot ΔV_u is the initial transient response to a unidirectional change in input which precedes the main transition and is opposite in polarity.

It is commonly expressed as *percentage undershoot*

$$\delta_u = \frac{\Delta V_u}{V_{st}} 100\%$$

Pulse droop or *sag* ΔV_{st} is a distortion of an otherwise essentially flat-topped rectangular pulse characterized by a decline of the pulse top. It is expressed in terms of the percentage droop or sag

$$\delta = \frac{\Delta V_{st}}{V_{st}} 100\%$$

Requirements for Video Amplifiers. The requirements that video amplifiers should meet depend on their expected application and function.

The basic requirements are:

(1) The output voltage of the amplifier must be within the limits stipulated in the spesifications.

(2) Rise and fall time of a rectangular pulse at the amplifier output must not exceed the specified values.

(3) Voltage overshoot and undershoot at the amplifier output must be kept to a minimum.

(4) The droop of the longest pulse should not be greater than that specified.

(5) The operating voltages of the amplifier valves and indicator tube between pulses should not noticeably differ from the quiescent (no-signal) state (d. c. reference).

(6) Changes in the amplifier performance caused by valve replacement must not exceed the limits allowed by the specifications.

Examine some of these requirements in greater detail.

1. The range of output voltage between V_{min} and V_{max} depends on the type of indicator used. In indicators using intensity modulation of the beam to indicate the target coordinates the control voltage usually changes from $V_{min} = 20$ volts to $V_{max} = 40$ volts.

In indicators where the beam is deflected in position to indicate the target coordinates, the control voltage usually ranges from 20-50 volts to 100 volts. Thus, the output voltage of the signal must not change more than two to five times.

The voltage that the detector feeds to the video amplifier usually varies within much broader limits (tens or even hundreds of times). To avoid an overload on the amplifier and malfunction of the indicator, the input stage of the video amplifier usually operates as a limiter. The amplitude of the output voltage is limited by causing the input stage to operate in such a way that the amplitude characteristic of the stage, $V_{out} = F(V_{in})$ falls off sharply beginning with a certain value of input voltage.

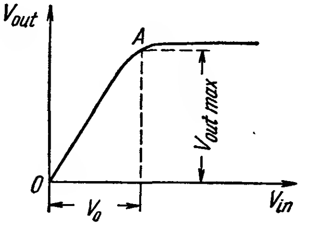


Fig. 15.3. Amplitude characteristic of the limiter

The amplitude characteristic of such a stage is shown in Fig. 15.3. As long as the input voltage rises to V_0, the output voltage varies in proportion to the input voltage.

When the input voltage is V_0, the output voltage is a maximum, $V_{out\ max}$. Then, as the input voltage keeps rising, it increases insignificantly. Knowledge of the amplitude characteristic makes it possible to specify the minimum voltage that may be applied to the input of the stage from the detector. As already noted in tubes with intensity-beam modulation the maximum and minimum control voltages are in the ratio 2-to-1. Therefore, for maximum output voltage the input voltage should be V_0, and for output voltage half as great, the input voltage should be at least

$$V_{in\ min} = 0.5V_0$$

In tubes with a deflected beam in which the maximum and minimum control voltages are in the ratio 2-to-1 to 5-to-1 the minimum input voltage should be

$$V_{in\ min} = \frac{1}{2\text{ to }5} V_0$$

When a negative voltage is applied from the detector, limiting is accomplished simply by cut-off. Figure 15.4 gives the circuit of the limiter stage. Figure 15.5 shows its waveforms.

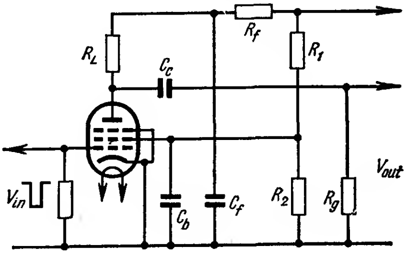


Fig. 15.4. Limiter stage

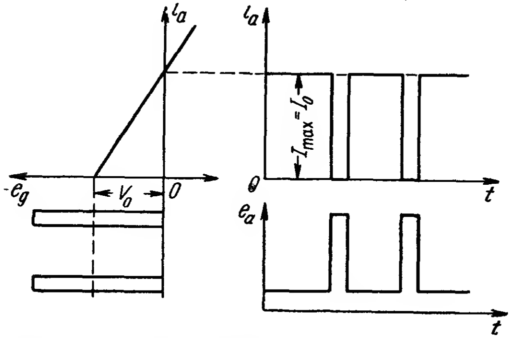


Fig. 15.5. Limiter waveforms

In the no-signal state, the grid voltage is $E_g = 0$, the anode current is a maximum and equal to the quiescent current I_0. Negative pulses are applied to the input of the stage from the detector. When the negative input voltage becomes V_0, the valve is driven to cut-off, the anode current drops to zero, and so does the voltage across the anode load resistance R_L, while the anode voltage becomes a maximum equal to the output voltage of the filter R_fC_f. Even if the negative input voltage continues to increase, the anode voltage of the cut-off valve will remain practically unchanged for the duration of the pulse. Therefore, no matter how much the negative input voltage increases, the output voltage pulse will remain constant in magnitude.

2. The total time of rise t_r and fall t_d of the pulse at the output of the amplifier is usually not more than one-tenth of the pulse duration

$$t_{r\,tot} \cong t_{dtot} = 0.1\tau$$

The time of pulse rise and fall can be even smaller. As will be shown later, it depends upon the anode load resistance R_L and the distributed capacitance C_s of the stage.

3. Overshoots and undershoots are caused by the reactive elements in the circuit and by the droop of the pulse top (in the case of undershoot). The capacitive and inductive reactances present in the various stages may form resonant circuits. The transients arising in them produce voltages which may combine with the pulse voltage. The voltage overshoots should not exceed 10 per cent of the steady-state value of the signal voltage.

4. The droop of the pulse top (see Fig. 15.2) is mainly observed during the passage of long pulses through the stage. A change in the output voltage within the pulse duration can vary the brightness of the image on the tube screen. The droop of the pulse from the output of one of the preceding stages can cause an undershoot in the following stages, thus upsetting operation of the indicator tube. The pulse droop is caused, as a rule, by the low time constants of the coupling circuit, screen-grid circuit, and the self-bias network R_kC_k defined as follows:

— time constant of the coupling circuit

$$\tau_c = R_g C_c$$

— time constant of the screen-grid circuit

$$\tau_s = R_s C_b$$

— time constant the self-bias network

$$\tau_k = R_k C_k$$

As any of these time constants is increased, the droop of the pulse decreases.

If the time constant τ_s is small, a long pulse will cause the blocking capacitor C_b to discharge through R_s, and the screen-grid voltage, as well as the anode current, will decrease. This will cause a decrease in the output voltage. With a small time constant τ_k, the capacitor in the self-bias network charges rapidly and the self-bias network develops a negative feedback voltage. This leads to a decrease in voltage between the grid and cathode and, consequently, to a decrease in anode current and output voltage.

To obtain a minimum droop, the capacitances C_b and C_k should be such that

$$C_b \geqslant \frac{\tau_{max}}{\delta_s} \frac{1}{R_s} \tag{15.1}$$

$$C_k \geqslant \frac{\tau_{max}}{\delta_k} g_m \tag{15.2}$$

where τ_{max} = maximum pulse duration
R_s = dynamic screen resistance of the valve
g_m = mutual conductance of the valve
δ_s, δ_k = limits of pulse droop caused by the screen-grid circuit and the self-bias network.

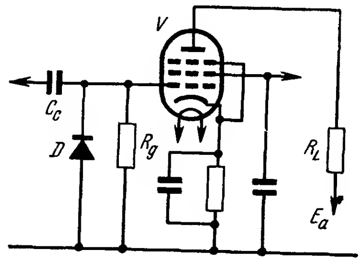


Fig. 15.6. Diode clamping circuit

It can be shown that the total droop caused by all these factors is equal to the sum of the individual droops

$$\delta = \delta_1 + \delta_2 + \delta_3 + \ldots$$

5. The need for maintanence of d.c. (quiescent) reference arises when pulses arrive at the input so closely together in time that the amplifier output has not yet recovered from the first before the second pulse arrives. If no measure were taken, a shift in the d.c. (reference) level would occur, and the indicator would be unable to display low- and medium-amplitude pulses since they would be masked by undershoot.

For this reason, special *clamping circuits* are used to restore a zero reference level voltage common to each pulse at the input of the indicator tube or the video amplifier final stage.

Figure 15.6 shows a diode clamping circuit at the input to the final stage. There is a diode D added to the usual grid circuit components R_g and C_c, and connected in parallel with R_g. When the signal is in positive polarity, the diode is cut off, its input resistance is very high, and the diode has no effect on the circuit. When a negative voltage due to an undershoot is applied to the grid of V_1, the diode is thrown into conduction, the input resistance of the diode sharply decreases, the input of the final stage is nearly short-circuited, and the negative voltage due to undershoot is drastically reduced.

Review Questions

1. What effect is produced by frequency distortion at the lower end of the range on the pulse waveform?
2. Same, at the higher end of the range?
3. What is the cause of overshoot and undershoot?
4. Does non-linear distortion affect the pulse waveform?
5. Why is the total droop equal to the sum of partial droops?

92. Video Amplifier Circuit Configurations

***RC*-coupled Amplifier as Video Amplifier.** Analysis of circuit configurations for voltage amplifiers has shown that the flattest frequency response and the lowest phase distortion are offered by the *RC*-coupled circuit. This is why it is the basis of most broad-band amplifiers, including video amplifiers.

The mid-band gain is given by

$$K_0 = g_m R_L$$

Frequency and phase distortion appearing in *RC*-coupled amplifiers in the low- and high-frequency ranges is

$$M_l = \sqrt{1 + \left(\frac{1}{\Omega_l \tau_g}\right)^2}$$

$$\tan \varphi_l = \frac{1}{\Omega_l \tau_g}$$

$$M_h = \sqrt{1 + (\Omega_h \tau_a)^2}$$

$$\tan \varphi_h = -\Omega_h \tau_a$$

where

$$\tau_g = C_c R_g$$

$$\tau_a = C_s R_e$$

It is not difficult to see that frequency and phase distortion in the low-frequency range may be minimized by increasing the time constant τ_g. In the high-frequency range this can be done by decreasing the time constant τ_a, that is, by decreasing either the capacitance C_s or the anode load resistance. However, C_s is made up of the output capacitance of the valve of the stage, the input capacitance of the following stage, and the distributed capacitance of the wiring, and it cannot be decreased below a certain limit. Therefore, the distortion can be decreased only by reducing the anode load resistance R_L, which unavoidably decreases the stage gain. The gain can be slightly increased by using high-mu pentodes.

Let us see how the parameters of an *RC*-coupled amplifier affect a rectangular pulse passing through it. For this purpose we shall refer to the equivalent circuits of the stage. Figure 15.7*a* shows a complete equivalent circuit of the stage; Figure 15.7*b* and *c* gives the equivalent circuits derived from the circuit of Fig. 15.7*a* for long and short pulses.

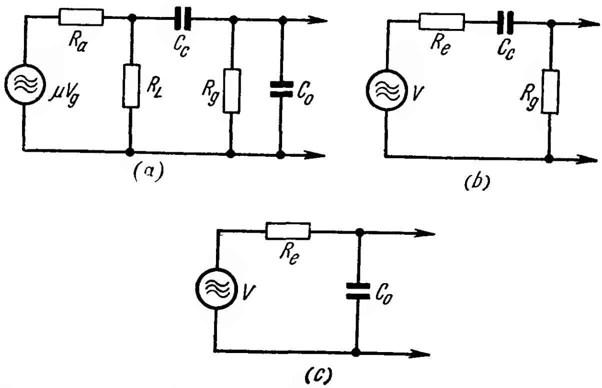


Fig. 15.7. Equivalent circuits of an RC-coupled amplifier
(a) complete circuit; (b) circuit for a long pulse; (c) circuit for a short pulse

In these circuits V is the equivalent generator voltage given by

$$V = \mu V_g \frac{R_L}{R_a + R_L} \cong g_m R_L V_g$$

since

$$R_L \ll R_a, \text{ and } \frac{\mu}{R_a} = g_m$$

R_e is the equivalent resistance of the generator, given by

$$R_e = \frac{R_L R_a}{R_a + R_L} = \frac{R_L}{1 + \frac{R_L}{R_a}} \cong R_L$$

When the d.c. voltage V appears across the generator terminals (see Fig. 15.7*c*), there is a flow of current charging the capacitor. The output voltage (the voltage across the capacitor) is described by the well-known equation for a capacitor charging through a resistor

$$V_{out} = V\left(1 - e^{-\frac{t}{\tau_a}}\right) \qquad (15.3)$$

where

$$\tau_a = C_s R_e \cong C_s R_L$$

A plot of this equation is shown in Fig. 15.8. The lower the time constant τ_a, the faster the capacitor charges and the output voltage reaches a steady-state value. As the pulse ceases at time t_2, C_s discharges through R_e, and the voltage across it

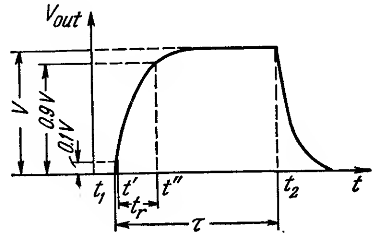


Fig. 15.8. Waveform of output voltage

and, consequently the amplifier output voltage, decreases exponentially

$$V_{out}=Ve^{-\frac{t}{\tau_a}} \tag{15.4}$$

Let us determine the dependence of pulse rise time on the time constant of the anode circuit τ_a.

By definition, pulse rise time t_r, is the interval between the instants, t' and t'', at which the output voltage reaches 10 and 90 per cent of its steady-state value V [see Eq. (15.8)]

$$t_r=t''-t'$$

Substituting the output voltage at t'' and t' into Eq. (15.3) gives

$$V_{out2}=0.9V=V\left(1-e^{-\frac{t''}{\tau_a}}\right) \tag{15.6}$$

$$V_{out1}=0.1V=V\left(1-e^{-\frac{t'}{\tau_a}}\right) \tag{15.7}$$

In Equation (15.6)

$$e^{-\frac{t''}{\tau_a}}=0.1$$

or

$$e^{\frac{t''}{\tau_a}}=10$$

Solving for t'' gives

$$t''=\tau_a \ln 10 \tag{15.8}$$

In Equation (15.7)

$$e^{-\frac{t'}{\tau_a}}=0.9$$

or

$$e^{\frac{t'}{\tau_a}}=\frac{10}{9}$$

Solving for t' yields

$$t' = \tau_a \ln\frac{10}{9} \tag{15.9}$$

Substituting the values of t'' and t' into Eq. (15.5), we obtain

$$t_r = t'' - t' = \tau_a \ln 10 - \tau_a \ln\frac{10}{9}$$

or

$$t_r = 2.2\,\tau_a \tag{15.10}$$

If we assume that the pulse rise time must not exceed 10 per cent of the pulse duration

$$t_r = 0.1\,\tau = 2.2\,\tau_a$$

then the time constant of the anode circuit must be such that

$$\tau_a = 0.0455\,\tau \tag{15.11}$$

It can be shown that the pulse decay time is equal to the rise time

$$t_d = t_r$$

When the amplifier accepts a short pulse, the capacitor C_c, which is usually smaller than C_s, has no time to charge. The voltage between the capacitor plates is insignificant, and almost the whole of the generator voltage appears across C_s, i. e. is applied to the amplifier input.

When the amplifier accepts a long pulse, C_c has time to charge to a certain level. The opposing voltage developing across it reduces the output voltage of the stage, and the pulse will sag or droop.

The relation of pulse droop to the time constant of the grid circuit τ_g can be derived as follows.

When a voltage V appears across the generator terminals (see Fig. 15.7*b*), a current flows in its circuit, varying exponentially

$$i = \frac{V}{R_g} e^{-\frac{t}{\tau_g}} \tag{15.12}$$

The output voltage developing across R_g is given by

$$V_{out} = iR_g = Ve^{-\frac{t}{\tau_g}} \tag{15.13}$$

The higher the time constant τ_g, the smaller will be the change in the output voltage.

If we assume that at the end of the pulse the output voltage must be 90 per cent of its peak value, the necessary value of τ_g can be found thus. From the expression for the output voltage

$$V_{out} = 0.9\,V = Ve^{-\frac{\tau}{\tau_g}}$$

it follows that

$$e^{-\frac{\tau}{\tau_g}} = 0.9$$

or

$$e^{\frac{\tau}{\tau_g}} = 1.11$$

Solving for τ_g gives

$$\frac{\tau}{\tau_g}\log_{10} e = \log_{10} 1.11 = 0.046$$

Therefore

$$\tau_g = \frac{\tau}{0.046}\log_{10} e = \tau\frac{0.434}{0.046} \cong 10\,\tau$$

or

$$\tau_g = 10\,\tau \qquad (15.14)$$

That is, the time constant of the C_cR_g network must be ten times the pulse duration τ.

From Eqs. (15.11) and (15.14) it is possible to determine the basic parameters of the stage—the anode load resistance R_L, the coupling capacitance C_c and the grid-leak resistance R_g.

As a pulse passes through a multistage amplifier, each stage affects the pulse shape. Therefore, if the pulse is to retain its original waveform at the output of a multistage amplifier, the individual stages must meet more stringent requirements than when each operates separately. Accordingly, Eqs. (15.11) and (15.14) for a multistage amplifier must be as follows:

$$\tau_a = \frac{\tau}{20\,N} \qquad (15.15)$$

$$\tau_g = 10\,N\tau \qquad (15.16)$$

where N is the number of stages in the amplifier.

When selecting C_c and R_g, the following must be taken into consideration so as to obtain the requisite value of τ_g.

To decrease the shunting effect of the grid-leak resistor on the anode load, R_g must be such that

$$R_g \gg 25\, R_L$$

To reduce the pulse droop, C_g must be 30 to 50 times C_{in}, the input capacitance of the following stage.

Example 15.1. Determine R_L, C_c and K_0 of the stage if: $\tau = 2\,\mu$ sec, valve 6Ж9П, $C_s = 30$ picofarads, $g_m = 17.5$ milliamperes/volt, $C_{in} = 8.5$ picofarads.

Solution. 1. Find the time constant of the anode circuit

$$\tau_a = \frac{\tau}{20} = \frac{2}{20} = 0.1\ \mu\text{ sec}$$

2. Find the anode load resistance

$$R_L = \frac{\tau_a}{C_s} = \frac{0.1\times 10^{-6}}{30\times 10^{-12}} \cong 3{,}300 \text{ ohms} = 3.3 \text{ kilohms}$$

3. Find the time constant of the coupling network

$$\tau_g = 10\,\tau = 10\times 2 = 20\ \mu\text{sec}$$

4. Select the grid-leak resistance

$$R_g = 25 R_L = 25\times 3.3 = 82.5 \text{ kilohms}$$

or, taking the nearest standard value, $R_g = 82$ kilohms.

5. Find the capacitance of the coupling capacitor

$$C_c = \frac{\tau_g}{R_g} = \frac{20\times 10^{-6}}{82\times 10^{3}} = 0.244\times 10^{-9} = 244 \text{ picofarads}$$

or, taking the nearest standard value, $C_c = 270$ picofarads.

If the following stage employs the same type of valve, the input capacitance of the stage will be 8.5 picofarads

$$\frac{C_c}{C_{in}} = \frac{270}{8.5} = 32$$

which satisfies the condition: $C_c > (30 \text{ to } 50)\, C_{in}$.

6. Determine the stage gain

$$K_0 = g_m R_L = 17.5\times 10^{-3}\times 3{,}300 \cong 58$$

It can be shown that undistorted reproduction of a rectangular pulse by an *RC*-coupled amplifier depends on its frequency range and that the rise and fall time is determined by the high-frequency end of the range, and the pulse droop by the low-frequency end.

The equations for frequency distortion in the high- and low-frequency ranges are given at the beginning of this section.

Writing the lower limiting frequency Ω_l of the range in terms of the time constant of the coupling network C_cR_g and the frequency distortion M_l, we get

$$\Omega_l = \frac{1}{\tau_g \sqrt{M_l^2 - 1}}$$

Substituting Eq. (15.14) for τ_g, we obtain

$$\Omega_l = \frac{1}{10\tau \sqrt{M_l^2 - 1}}$$

That is, the lower limiting frequency must be decreased with the duration of the pulse. In other words, an amplifier with considerable frequency distortion in the low-frequency range is not able to pass long pulses, because the droop will be prohibitively large.

The upper limiting frequency ω_0 of the amplifier is the frequency at which the gain of the amplifier falls below 0.707 times the midband gain (3 db down). Then, at the upper limiting frequency

$$M_h = \sqrt{1 + (\omega_0 \tau_a)^2} = \sqrt{2}$$

Whence

$$\omega_0 \tau_a = 1$$

and

$$\omega_0 = \frac{1}{\tau_a}$$

or

$$f_0 = \frac{1}{2\pi\tau_a}$$

Substituting the value of τ_a from Eq. (15.10), we obtain

$$f_0 = \frac{2.2}{2\pi t_r} = \frac{0.35}{t_r}$$

The dependence of the upper limiting frequency on the pulse rise time t_r implies that the upper limiting frequency must be raised as the pulse rise time increases. Thus, an amplifier with a narrow bandwidth is not able to pass short pulses without distortion because the pulse rise and fall time is increased and the leading and trailing edges of the pulse fall off gradually.

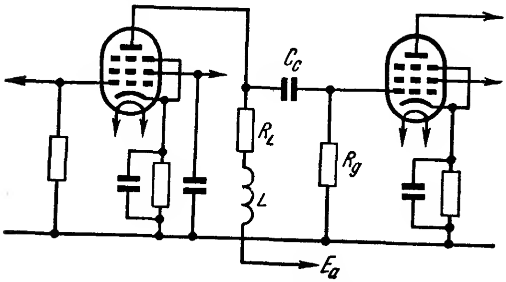


Fig. 15.9. Simple shunt inductive high-frequency compensation

To broaden the bandwidth, video amplifiers include frequency-compensation circuits which operate on short pulses.

Frequency Compensation of Video Amplifiers. Frequency compensation is widely used in television and in all other receivers designed for the reception of pulse signals.

High-frequency Compensation. One of the most commonly used forms of frequency compensation in the high-frequency range is a simple shunt-compensated coupling connection (Fig. 15.9). In this circuit, a compensating coil L is connected in series with the anode load resistance. At low frequencies, the inductive reactance of the coil is very small, and the coil does not affect operation of the stage. As the signal frequency increases, the inductive reactance of the coil is also increased, causing an increase in the anode load impedance Z_L, and in the stage gain. The total increase in the high frequency gain will compensate for the decrease in gain caused by the shunting action of the shunt capacitance C_s.

Phase distortion, too, is decreased in the shunt-compensated video amplifier. As we know, C_s causes a negative phase shift in the output voltage. The compensating coil causes a positive phase shift, and the resultant phase shift at the output of the stage is considerably decreased.

To evaluate the effect of L on the electric properties of the stage, consider its high-frequency equivalent circuit (Fig. 15.10).

The mid-band gain is

$$K_0 = g_m R_L$$

At the high frequencies, the equivalent anode load impedance is Z_L. Therefore, the high frequency gain is

$$K = g_m Z_L \qquad (15.17)$$

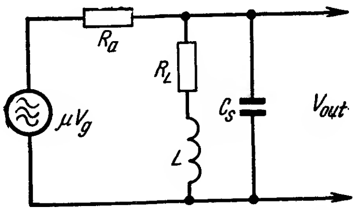


Fig. 15.10. Equivalent circuit of a video amplifier using simple shunt inductive compensation

Frequency distortion at the high frequencies is

$$M_h = \frac{K_0}{K_h} = \frac{g_m R_L}{g_m Z_L} = \frac{R_L}{Z_L} \tag{15.18}$$

Writing out the expression for Z_L, we get

$$Z_L = \frac{(R_L + j\Omega_h L)\dfrac{1}{j\Omega_h C_s}}{R_L + j\Omega_h L + \dfrac{1}{j\Omega_h C_s}} = \frac{R_L + j\Omega_h L}{1 - \Omega_h^2 LC_s + j\Omega_h C_s R_L} =$$

$$= R_L \frac{1 + j\Omega_h \dfrac{L}{R_L}}{1 - \Omega_h^2 LC_s + j\Omega_h C_s R_L}$$

On setting

$$\frac{L}{R_L} = \tau_L$$

$$C_s R_L = \tau_a$$

$$LC_s = LC_s \frac{R_L}{R_L} = \tau_L \tau_a$$

the expression for Z_L may be written as

$$Z_L = R_L \frac{1 + j\Omega_h \tau_L}{1 - \Omega_h^2 \tau_L \tau_a + j\Omega_h \tau_a} \tag{15.19}$$

and Eq. (15.18)

$$M_h = \frac{R_L}{Z_L} = \frac{R_L}{R_L \dfrac{1 + j\Omega_h \tau_L}{1 - \Omega_h^2 \tau_L \tau_a + j\Omega_h \tau_a}} = \frac{1 - \Omega_h^2 \tau_L \tau_a + j\Omega_h \tau_a}{1 + j\Omega_h \tau_L} \tag{15.20}$$

The modulus of frequency distortion is

$$M_h = \sqrt{\frac{(1 - \Omega_h^2 \tau_L \tau_a)^2 + (\Omega_h \tau_a)^2}{1 + (\Omega_h \tau_L)^2}}$$

Removing the brackets in the numerator, and collecting similar terms, we obtain

$$M_h = \sqrt{\frac{1+(\tau_a^2-2\tau_L\tau_a)\,\Omega_h^2+\tau_L^2\,\tau_a^2\Omega_h^4}{1+(\Omega_h\tau_L)^2}} \tag{15.21}$$

Frequency distortion in the high-frequency range will not occur only when the right-hand side of Eq. (15.21) is unity. This condition, in turn, can be fulfilled only when the numerator and denominator are identically equal. This is possible only when the same powers of Ω_h have the same factors, i.e. when the following equalities are satisfied:

$$\tau_a^2 - 2\tau_L\,\tau_a = \tau_L^2 \tag{a}$$

and

$$\tau_L^2\,\tau_a^2 = 0 \tag{b}$$

Condition (b) cannot be met because neither the anode load resistance nor C_s can be equal to zero. Therefore, in order to minimize frequency distortion, it should be sought to satisfy condition (a) at least.

With condition (a) satisfied, the time constant τ_a may be written in terms of the time constant τ_L as

$$\tau_a = (1+\sqrt{2})\,\tau_L \tag{15.22}$$

Substituting the expression for τ_a into Eq. (15.21) we obtain

$$M_h = \sqrt{\frac{1+\Omega_h^2\,\tau_L^2+\Omega_h^4\,\tau_L^4\,(1+\sqrt{2})^2}{1+\Omega_h^2\,\tau_L^2}} \tag{15.23}$$

Using Eq. (15.23), it is possible to determine the necessary value of τ_L and, consequently, of τ_a, from the specified Ω_h and M_h.

Solving Equation (15.23) for $\Omega_h^2\tau_L^2$ gives

$$\Omega_h^2\,\tau_L^2 = \frac{1}{2}\,\frac{M_h^2-1}{(1+\sqrt{2})^2}\left[1+\sqrt{1+\frac{4\,(1+\sqrt{2})^2}{M_h^2-1}}\right]$$

Figure 15.11 shows a plot of $\Omega\tau_L$ as a function of frequency distortion M_h. Entering this graph with M_h, it is possible to find the value of $\Omega_h\tau_L$ and to calculate τ_L

$$\tau_L = \frac{(\Omega_h\tau_L)}{\Omega_h}$$

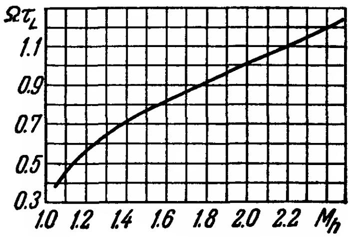


Fig. 15.11. Plot of $\Omega\tau_L$ versus M_h

Once τ_a is found from Eq. (15.22), it is easy to find the anode load resistance

$$R_L = \frac{\tau_a}{C_s}$$

The inductance of the compensating coil is given by

$$L = R_L \tau_L$$

The phase response of a shunt-compensated video amplifier is given by

$$\tan\varphi = -\Omega\tau_a + \Omega\tau_L(1 - \Omega^2\tau_L\tau_a) \qquad (15.24)$$

The phase response of the stage, as a function of $\Omega\tau_a$ and $\Omega\tau_L$, is shown in Fig. 15.12. As is seen, phase overcompensation takes place at certain values of $\Omega\tau_L$ and $\Omega\tau_a$. This is evidenced by the positive phase shift in the stage.

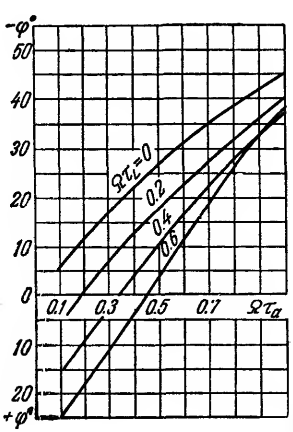


Fig. 15.12. Phase response of the stage, as a function of $\Omega\tau_a$ and $\Omega\tau_L$

Shunt-inductive h.f. compensation also improves the gain of the amplifier.

Let us prove this by the following example.

Example 15.2. A wideband stage is to amplify in a band extending up to 5 megahertz. The equivalent shunt capacitance of the amplifier stage, employing a type 6Ж9П valve, is $C_s \cong 30$ picofarads. The limit of frequency distortion is $M_h = 1.3$. It is required to determine the anode load resistance and stage gain K_0.

Solution. In an uncompensated amplifier, the anode load resistance is given by Eq. (2.28)

$$R_L = \frac{\sqrt{M_h^2 - 1}}{\Omega_h C_s} = \frac{\sqrt{1.3^2 - 1}}{6.28 \times 5 \times 10^6 \times 30 \times 10^{-12}} = 880 \text{ ohms}$$

The gain is

$$K_0 = g_m R_L = 17.5 \times 10^{-3} \times 880 = 15.4$$

At Ω_h the gain is

$$K_h = \frac{K_0}{M_h} = \frac{15.4}{1.3} = 11.8$$

Determine the anode load resistance and gain of the compensated stage under the same conditions.

Determine $\Omega_h \tau_L$ from the plot of Fig. 15.11:

$$(\Omega_h \tau_L) = 0.64$$

Hence

$$\tau_L = \frac{(\Omega_h \tau_L)}{\Omega_h} = \frac{0.64}{6.28 \times 5 \times 10^{-6}} = 2.04 \times 10^{-8} \text{ second}$$

The time constant τ_a is

$$\tau_a = (1 + \sqrt{2})\, \tau_L = 2.41 \times 2.04 \times 10^{-8} = 4.9 \times 10^{-8} \text{ second}$$

As a result, the anode load resistance will be

$$R_L = \frac{\tau_a}{C_s} = \frac{4.9 \times 10^{-8}}{30 \times 10^{-12}} = 1{,}630 \text{ ohms}$$

Find the inductance of the compensating coil

$$L = \tau_L R_L = 2.04 \times 10^{-8} \times 1{,}630 = 33.3 \text{ microhenrys}$$

The stage gain is

$$K_0 = g_m R_L = 17.5 \times 10^{-3} \times 1{,}630 = 28.5$$

As is seen from Example 15.2, frequency compensation has nearly doubled the gain with the same frequency response.

In deriving the design equations for shunt-inductive h.f. compensation, we have sought to secure the desired bandwidth. For a broad-band pulse amplifier, however, the decisive requirement is to maintain the pulse shape. Accordingly, the design equations are written in terms of pulse characteristics as given below.

1. Anode load resistance:

(a) an uncompensated amplifier

$$R_L = \frac{t_r}{2.2C_s} \tag{15.25}$$

(b) a shunt-compensated amplifier with δ_h of not over 1 per cent

$$R_L = \frac{t_r}{1.31C_s} \tag{15.26}$$

(c) a shunt-compensated amplifier with δ_h of not over 6 per cent

$$R_L = \frac{t_r}{1.22C_s} \tag{15.27}$$

In all cases t_r is the puls rise time of one stage:

$$t_r = 0.1\tau$$

defined in terms of the upper limiting frequency of the amplifier as

$$t_r = \frac{0.35}{F_h} \tag{15.28}$$

C_s is the equivalent shunt capacitance defined as

$$C_s = C_{out} + C_{in} + C_w$$

where C_{out} = output capacitance of the valve
C_{in} = input capacitance of the valve in the succeeding stage
C_w = disturbed capacitance of wiring; with good wiring, it does not exceed 10 picofarads.

2. Inductance of the compensating coil:

(a) an amplifier with an overshoot of 1 per cent

$$L = 0.36R_L^2C_s \tag{15.29}$$

(b) an amplifier with an overshoot of 6 per cent

$$L = 0.5R_L^2C_s \tag{15.30}$$

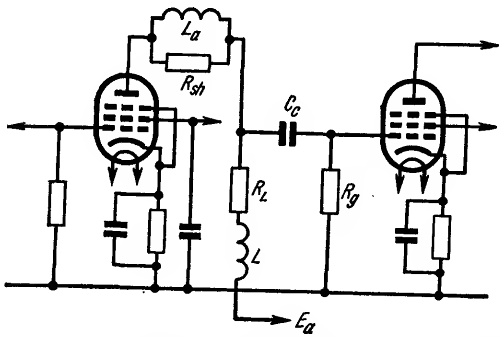


Fig. 15.13. Modified shunt high-frequency compensation

A better form of high-frequency compensation is offered by the modified shunt-coupling connection of Fig. 15.13. In this arrangement, a second compensating coil, L_a, shunted by R_{sh}, is connected between the anode and the anode load resistance R_L. As a result, the high-frequency gain is improved by 20 to 40 per cent, as compared with simple shunt inductive compensation.

This improvement in stage gain stems in part from the fact that the shunt capacitance C_s is divided in two parts (Fig. 15.14)

$$C_1 = C_{out} + 0.5C_w$$

and

$$C_2 = C_{in} + 0.5C_w$$

C_1 connected across the anode and cathode is numerically smaller than the total capacitance C_s. When a rectangular pulse reaches the valve grid, C_1 charges faster than C_s in a shunt-compensated circuit. Therefore, the pulse rise time is lower, which is the same as raising the upper limiting frequency of the amplifier. Besides, by properly selecting the inductance of L_a,

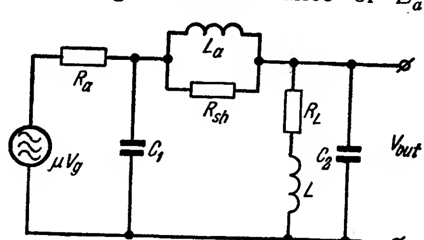


Fig. 15.14. Equivalent circuit of modified shunt compensated amplifier

voltage resonance can be obtained in the series tuned circuit formed by L_a and C_2 near the upper limiting frequency. Hence, the voltage across C_2, i.e. the stage output voltage, will also be increased. Thus modified shunt compensation also broadens the bandwidth and increases the gain of the stage.

The parameters of a modified shunt-compensated amplifier are found from the following equations:

1. Anode load resistance

$$R_L = \frac{t_{r1}}{qC_s} \tag{15.31}$$

The rise time as found from Eq. (15.28) is

$$t_{r1} \leqslant 0.1\tau$$

2. The inductance of the shunt coil is

$$L = aR_L^2 C_s \tag{15.32}$$

3. The inductance of the series coil is

$$L_a = a' R_L^2 C_s \tag{15.33}$$

4. The shunting resistance is

$$R_{sh} = bR_L$$

The values of q, a, a' and b are given in Table 15.1 as functions of the ratio $\frac{C_1}{C_s}$, where $C_1 = C_{out} + 0.5C_w$.

TABLE 15.1

$1-\frac{C_1}{C_s}$	q	a	a'	b	δ_h, %	$\frac{C_1}{C_s}$
0.344	0.934	0.122	0.511	—	4.3	0.656
0.35	0.95	0.122	0.514	50.0	4.1	0.65
0.4	1.04	0.126	0.536	6.6	3.8	0.6
0.437	1.07	0.13	0.554	4.35	3.4	0.563
0.45	1.08	0.132	0.56	3.7	3.3	0.55
0.5	1.09	0.14	0.582	2.75	2.8	0.5
0.55	1.1	0.146	0.61	2.33	2.3	0.45
0.6	1.12	0.148	0.652	2.12	1.9	0.4
0.65	1.15	0.146	0.72	2.0	1.6	0.35

The values of C_1, $\frac{C_1}{C_s}$ and q for some Soviet valves are given in Table 15.2.

TAELE 15.2

Valves	6Ж3П (US 6AG5)	6Ж9П	6Ж11П	6Ж20П-1	6Ж20П-2	6П9 (US 6AG7)	6П14П	6П18П	6П3С (US 6L6)	6П6С (US 6V6 GT)
C_1, picofarads	7	9	10	7	9	12.5	13	12	16.5	12.5
$\frac{C_1}{C_{s\,min}}$	0.35	0.36	0.4	0.35	0.36	0.45	0.45	0.45	0.53	0.45
q	1.15	1 14	1.12	1.15	1.14	1.1	1.1	1.1	1.084	1.1

Figure 15.15 shows high-frequency compensation by means of an RC network. The network capacitance is so selected that its reactance in the mid-band range is much higher than the resistance R. Some of the voltage taken from the anode load resistance is dropped across R, and the stage output voltage is lower than that of the usual amplifier. At the high frequencies the reactance of C decreases, and it bypasses R to a greater extent so that the impedance of the RC network is decreased considerably. This decreases the voltage across the RC network and, consequently, increases the output voltage.

For efficient high-frequency compensation, R must be two or three times R_g. The network capacitance can be selected on the basis of the following relation:

$$\frac{1}{\Omega_h C} \cong R$$

which secures an increase of 30 to 40 per cent in the output voltage at the high frequencies.

Figure 15.16 shows high-frequency compensation employing an RC network in the cathode lead. For its effect, this form of frequency compensation depends on negative current feedback. The values of R and C are so selected that in the mid-band range the capacitive reactance is many times R. At these frequencies a negative feedback voltage is developed across Z_{RC}, decreasing the signal voltage at the valve grid and the stage gain.

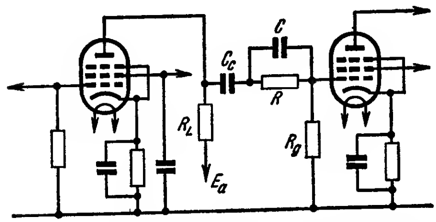


Fig. 15.15. *High-frequency compensation employing an RC network*

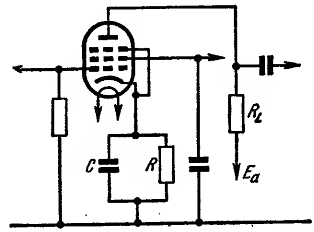


Fig. 15.16. *High-frequency compensation by negative feedback*

As the signal frequency increases, the capacitive reactance and, consequently, the impedance of the *RC* network, goes down. The negative feedback voltage decreases, and the stage gain increases. At the high frequencies, the increase in the stage gain compensates for the decrease in the frequency response caused by the usual reactive circuit components.

Figure 15.17 shows high-frequency series compensation. The operating principle of this arrangement can be understood from its equivalent circuit (Fig. 15.18). The inductance of the compensating coil L_g is so selected that the L_gC_{in} circuit will resonate at the high frequencies. As a result of series (voltage) resonance, the current flowing through C_{in} increases, and the voltage drop across the capacitance is also increased. This results in a virtual increase of gain. L_g is frequently shunted by R to decrease the resonance peak.

In a number of cases the stage simultaneously employs several methods of high-frequency compensation. Figure 15.19 shows an amplifier stage using shunt-inductive and series-inductive compensation.

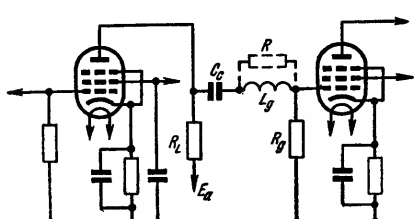


Fig. 15.17. *Series-compensated amplifier circuit*

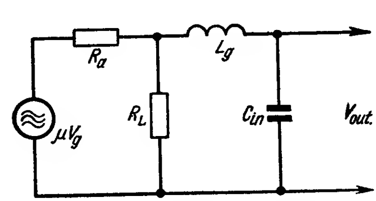


Fig. 15.18. *Equivalent circuit of series-compensated amplifier*

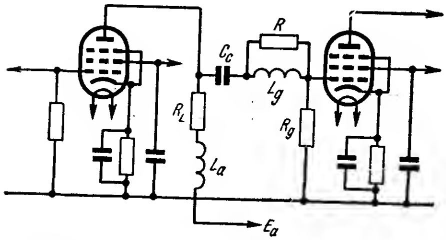


Fig. 15.19. Amplifier stage with shunt-inductive and series-inductive compensation

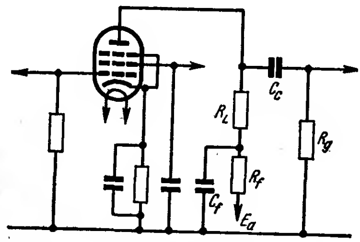


Fig. 15.20. Low-frequency compensation

Low-frequency Compensation. One of the most commonly used methods of low-frequency compensation is to insert a decoupling *RC* filter as shown in Fig. 15.20.

Let us see how this circuit operates. The stage gain of a stage employing a valve with a high a.c. anode resistance is decided by the mutual conductance and anode load impedance

$$K = g_m Z_l$$

In our case, the anode load impedance Z_L is made up of R_L and filter impedance Z_f

$$\dot{Z}_L = R_L + \dot{Z}_f$$

The value of C_f is so selected that its reactance is very low at the medium and high frequencies when the filter impedance may be neglected and the anode circuit may be regarded as being loaded only into R_L.

Then, the mid-band gain will be

$$K_0 = g_m R_L$$

At the low frequencies, Z_f increases noticeably and the stage gain goes up.

The increase in stage gain due to frequency compensation, usually called the *relative gain*, is defined as

$$m_l = \frac{\dot{K}_l}{K_0} = \frac{g_m Z_L}{g_m R_L} = \frac{Z_L}{R_L} \tag{15.34}$$

The value of R_f is usually specified in advance. It ought not to be too great because the anode voltage decreases as R_f is increased.

Usually

$$R_f = (0.2 \text{ to } 0.5)\, R_L$$

If the relative gain and the filter resistance are known, the capacitance of the filter capacitor may be determined from the following equation:

$$C_f = \frac{1}{\Omega_l R_f} \sqrt{\frac{\left(1+\frac{R_f}{R_l}\right)^2 - m^2}{m^2 - 1}} \qquad (15.35)$$

Apart from flattening the frequency response, the filter reduces parasitic coupling between the stages through the supply sources, which improves amplifier stability.

The overall frequency response, that is, one at the low and high frequencies, may be improved by means of negative feedback supplied by an *RC* network in the cathode lead. However, as we already know, negative feedback brings down the stage gain. Therefore, video amplifiers usually resort to compensating circuits in the anode and grid leads which all increase the stage gain.

Review Questions

1. Does the a.c. anode resistance of the valve affect the pulse shape?
2. Can h.f. compensation use a coil in parallel with the anode load resistance?
3. Why should video amplifiers use valves with high mutual conductance?
4. Can video amplifiers use transformer coupling?
5. Does the gain of video amplifiers depend on the pulse duration?

93. Transmission-line Amplifiers

Radio today is using pulses of progressively shorter duration. In some cases the pulse duration is several hundredths of a microsecond.

Shorter pulses improve the resolution of aircraft radars and make it possible to see on plan-position indicators (PPI) not only the signals that determine target coordinates, but also the outlines of streets, large ships and even of big aircraft.

Short-pulse operation calls for a very broad bandwidth of the video channel. For instance, with a pulse duration of 0.01 μsec, the upper limiting frequency should be

$$E_h = \frac{3\cdot 5}{\tau} = \frac{3\cdot 8}{0.01} = 350 \text{ megahertz}$$

It is impossible to obtain a uniform gain over such a frequency band, even when complex compensating circuits are used.

A detailed analysis of compensating circuits shows that the limit for the stage gain is determined by

$$K = \frac{g_m}{\pi F_h \sqrt{C_1 C_2}}$$

If $K \geqslant 1$, then

$$F_h \leqslant \frac{g_m}{\pi \sqrt{C_1 C_2}}$$

For a 6Ж1П valve (US equivalent, 6AK5) with $g_m = 5.2$ milliamperes/volt, $C_{in} = 4$ picofarads, $C_{out} = 2.1$ picofarads, and $C_w = 10$ picofarads,

$$C_1 = C_{out} + 0.5C_w = 2.1 + 0.5 \times 10 = 7.1 \text{ picofarads}$$
$$C_2 = C_{in} + 0.5C_w = 4 + 0.5 \times 10 = 9 \text{ picofarads}$$
$$F_h \leqslant \frac{g_m}{\pi \sqrt{C_1 C_2}} = \frac{5.2 \times 10^{-3}}{3.14 \sqrt{7.1 \times 10^{-12} \times 9 \times 10^{-12}}} = 207 \text{ megahertz}$$

With a gain of 2, the bandwidth is narrowed down to 100 megahertz.

Thus, in a stage employing a 6Ж1П valve it is practically impossible to obtain a gain of more than unity and a bandwidth of 200 megahertz; this is true even when a sophisticated form of frequency compensation is used.

A stage operating over such a broad bandwidth can produce amplification only if C_s shunting the load resistance is no longer a parasitic capacitance, but a parameter of the stage.

An example of such a circuit is the *transmission-line amplifier*, also called a *distributed amplifier*. For its operation such an amplifier depends on the propagation of travelling waves in transmission lines, or simply lines (Fig. 15.21).

When a voltage V_{in} is applied to the line input, current and voltage waves begin to propagate along the line. In an ideal line (the one without losses), the output voltage V_{out} is equal to the input voltage. If a.c. generators, operating in phase with

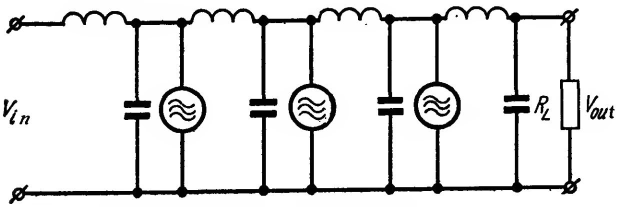


Fig. 15.21. Schematic of an artificial line

the incoming signal, are connected between the line sections, the energy transmitted from the beginning to the end of the line will gradually increase. This will lead to an increase in voltage across the load, i.e. the input signal will be amplified.

Transmission lines operating in a travelling-wave mode have a very broad pass-band. This is also true of artificial lines made up of lumped-constant sections. However, there is always an upper frequency limit (the *cuf-off frequency*) to artificial lines. Past this limit, the lines lose their broad-band properties.

The cut-off frequency is given by

$$f_c = \frac{1}{\pi \sqrt{L_0 C_0}}$$

where L_0 and C_0 are, respectively, the inductance and capacitance of a line section.

The propagation time through one line section depends upon the inductance and capacitance of the section:

$$t_1 = \sqrt{L_0 C_0}$$

The transient time, or the time required for the output to reach its steady-state value, depends upon the parameters and the number of line sections:

$$t_{st} = 1.1 n^{1/3} \sqrt{L_0 C_0}$$

With low values of section parameters it is possible to obtain a short transient time and a relatively wide pass-band. The number of sections has a limited effect on the transient time.

Let us examine in greater detail the operation of the distributed amplifier shown in Fig. 15.22. As is seen, there are two lumped (artificial) transmission lines, and a number of valves with their grids connected into one, and with their anodes into

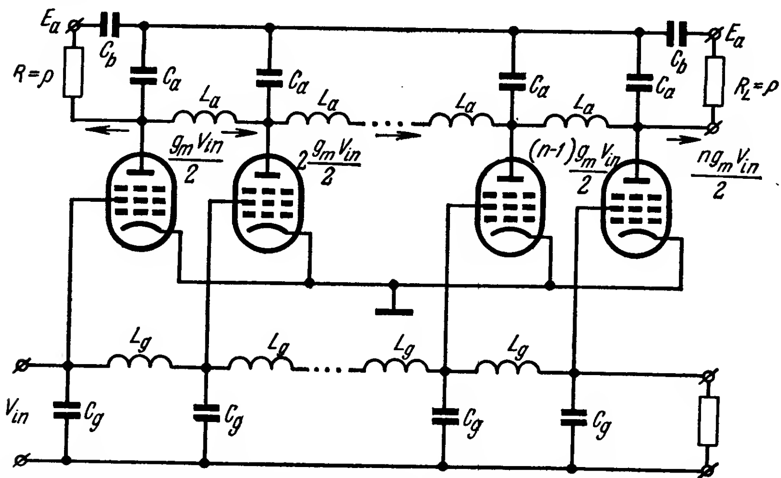


Fig. 15.22. Circuit diagram of a transmission-line (distributed) amplifier

the other line at appropriate node points. When a signal voltage V_{in} is applied to the input, a voltage wave begins to propagate along the grid line. If the propagation time through one section is t_{g1}, the energy will be transferred to the end of the line in time $t=(n-1)t_{g1}$ to be absorbed by the load resistance R equal to the characteristic impedance of the grid line:

$$R=\rho_g=\sqrt{\frac{L_g}{C_g}}$$

With V_{in} applied to the grid line, the current in the anode circuit will change by

$$I_a=V_{in}g_m$$

Each section to the left and right of each anode has the resistance equal to the characteristic impedance of the anode line:

$$\rho_a=\sqrt{\frac{L_a}{C_a}}$$

Therefore, the change of anode current will propagate to the left and right of each anode and will only be half as great as the change in the valve current

$$I'_a=\frac{1}{2}I_a=\frac{1}{2}V_{in}g_m$$

If the propagation time through a section of the anode line, t_{a1}, is selected to be equal to the propagation time through a section of the grid line, i.e.

$$t_{a1}=t_{g1}$$

or

$$\sqrt{L_aC_a}=\sqrt{L_gC_g}$$

then on arriving at the input of the second section the change of current caused by the anode circuit of the first valve will be added to the current change of the second valve, then of the third and, finally, of the nth valve. The total change of anode current at the output of the anode line will be n times the change of anode current of one valve

$$I_{a\,out}=n\frac{V_{in}}{2}g_m$$

The output voltage due to this current across the load resistance of the anode line will be

$$V_{out}=n\frac{g_m}{2}V_{in}\rho_a$$

The gain of a distributed-amplifier stage is

$$K_d=\frac{V_{out}}{V_{in}}=\frac{n}{2}g_m\rho_a \tag{15.36}$$

Let us compare the gain K of the conventional stage and the gain K_d of a distributed-amplifier stage. At the highest frequency of the band, the gain of the conventional stage may be expressed in terms of the reactance of the capacitance that shunts the anode load

$$K=g_mX_a=\frac{g_m}{2\pi F_hC_a} \tag{15.37}$$

Let us assume that the highest frequency F_h that can pass through the line without noticeable distortion is equal to the cut-off frequency f_c (usually $F_h<f_c$). Writing the characteristic impedance of the anode line ρ_a in terms of the cut-off frequency, we have

$$\rho_a=\frac{1}{\pi f_cC_a}$$

Substituting this expression into Eq. (15.36) gives

$$K_d=\frac{n}{2}g_m\rho_a=n\frac{g_m}{2\pi f_cC_a}$$

Noting that $F_h = f_c$, we obtain

$$K_d = n\frac{g_m}{2\pi F_h C_a}$$

Thus, all other conditions being equal, the gain of a distributed-amplifier stage is several times that of the conventional stage.

The performance of a single-stage amplifier is often compared in terms of the *gain-bandwidth product*, $K(F_h - F_l)$, or, for a broad-band stage, KF_h.

For a conventional stage

$$KF_h = \frac{g_m}{2\pi C_a} \tag{15.38}$$

while for a distributed-amplifier stage

$$K_d F_h = n\frac{g_m}{2\pi C_a} \tag{15.39}$$

As is seen, in a distributed-amplifier stage the product KF_h is inversely proportional to C_a. Therefore, an increase in gain in a distributed-amplifier stage will always lead to a narrower bandwidth. This is why the gain of a distributed amplifier ought not to be increased by increasing the load impedance.

A detailed analysis of distributed amplifiers shows that an increase in gain can be obtained not only by employing a larger number of line sections, but also by using the conventional cascading when the total gain is equal to the product of the stage gains

$$K_d = K_{d1} K_{d2}$$

Review Questions

1. What is the principle of bandwidth widening in distributed amplifiers?
2. Does it pay to use a distributed amplifier for a bandwidth of a few megahertz?
3. What kind of valve should be used in a distributed amplifier?
4. List methods for broadening the bandwidth of a distributed amplifier.

94. Video Amplifier Design

Given:

1. Input voltage, V_{in}.
2. Maximum pulse duration, τ_{max}.
3. Minimum pulse rise time t_r or upper cut-off frequency f_c.
4. Equivalent shunting capacitance C_s.

To Find:

1. Type of valve.
2. Upper cut-off frequency f_c or pulse rise time t_r.
3. Stage circuit parameters.
4. Stage gain K and output voltage V_{out}.
5. Operating conditions of the valve and associated circuit parameters.

Design Procedure:

1. Select the valve. The main thing to remember when selecting the type of valve for a video amplifier is to secure maximum gain.

The stage gain is determined by the same equation for any of the discussed circuits

$$K = g_m R_L$$

In the simplest case, the anode load resistance is decided by the equivalent shunt capacitance C_s and the upper cut-off frequency:

$$R_L = \frac{1}{\omega_c C_s}$$

The value of C_s chiefly depends upon the input and output capacitances of the valve. Therefore, when selecting the type of valve, make sure that the mutual conductance is as high as possible, while the input and output capacitances are a minimum.

2. For calculation of circuit parameters it is necessary to know the upper cut-off frequency or the minimum pulse rise time. In the simplest case, these values are related to each other as

$$f_c = \frac{0.35}{t_r}$$

When the upper cut-off frequency f_c is specified in advance, the pulse rise time will be

$$t_r = \frac{0.35}{f_c}$$

When the pulse rise time is specified in advance, the upper cut-off frequency can be found from the same relation.

3. The order in which the circuit parameters should be calculated depends on the selected circuit configuration. For a comparatively narrow signal frequency band, or for a comparatively long pulse rise time, an *RC*-coupled circuit may be used. For a frequency band several megahertz wide and for a rise time of several hundredths of a microsecond, a frequency-compensated circuit should be used. After the circuit has been chosen, its parameters are found from the equations given in the preceding sections of the present chapter.

4. The midband stage gain is given by

$$K_0 = g_m R_L$$

At the upper frequency of the range it is given by

$$K_h = K_0 \frac{1}{\sqrt{1 + (\omega_c C_s R_L)^2}}$$

The output voltage is

$$V_{out} = K V_{in}$$

5. The operating conditions of the valve are selected according to the signal to be amplified and the requirements that the stage is to meet. When positive pulses are to be amplified, the operating point should be positioned in the lower part of the linear section of the characteristic. The bias voltage must provide for operation of the valve without grid current:

$$E_g + V_{in} < 0$$

However, if the stage is to amplify negative pulses, the position of the operating point will be determined by the maximum anode dissipation

$$P_a = I_{a0} V_{a0}$$

and by the length of the characteristic tail. If the stage is to operate so that the signal wave is limited, the bias voltage should, above all, be such that

$$E_g + V_{in} < E'_g$$

and the anode dissipation

$$I_{a0} V_{a0} < P_{asafe}$$

The characteristic $i_a = F(e_g)$ can be shifted by varying the screen-grid voltage.

Having selected the operating conditions of the valve, the circuit parameters may be calculated.

The self-biasing resistor is

$$R_k = \frac{|E_g|}{I_{g2} + I_{a0}}$$

The bypass capacitor is

$$C_k = \frac{\tau_{max}}{\delta_c} g_m$$

The dropping resistor in the screen-grid circuit is

$$R_{dr} = \frac{E_a - E_{g2}}{I_{g2}}$$

The value of I_{g2} is found from the valve characteristics. The capacitance of the screen-grid blocking capacitor is

$$C_{g2} = \frac{\tau_{max}}{\delta_g R_s}$$

where

$$R_s = \frac{E_{g2}}{I_{g2}}$$

Example 15.3. Calculate a wide-band amplifier if the input voltage is 0.5 volt, pulse duration $\tau = 2$ μseconds, pulse rise time $t_r = 0.35\ t_{r\,tot}$. Equivalent shunt capacitance of the stage $C_s = 30$ picofarads.

To Find:

1. Type of valve.
2. Upper cut-off frequency, f_c.
3. Anode load resistance, R_L.
4. Inductance of compensating coil, L.
5. Midband gain K_0 and high-frequency gain K_h.

6. Resistance R_g and capacitance C_c of the coupling connection.

7. Operating conditions of the valve and associated circuit parameters.

Solution. 1. Select a type 6Ж9П valve.

2. Determine the upper cut-off frequency

$$f_c = \frac{0.35}{t_r} = \frac{0.35}{0.35\times 0.1\,\tau} = \frac{10}{2\times 10^{-6}} = 5 \text{ megahertz}$$

3. To determine the anode load resistance, from the curve of Fig. 15.11 find that

$$\Omega_h \tau_L = 0.707$$

whence

$$\tau_L = \frac{\Omega_h \tau_L}{\Omega_h} = \frac{0.707}{6.28\times 5\times 10^6} = 2.25\times 10^{-8} \text{ second}$$

Determine the time constant τ_a

$$\tau_a = (1+\sqrt{2})\,\tau_L = 2.41\times 2.25\times 10^{-8} = 5.43\times 10^{-8} \text{ second}$$

Find the anode load resistance

$$R_L = \frac{\tau_a}{C_s} = \frac{5.43\times 10^{-8}}{30\times 10^{-12}} \cong 1{,}800 \text{ ohms}$$

4. Find the inductance of the compensating coil

$$L = \tau_L R_a = 2.25\times 10^{-8}\times 1{,}800 = 40.8 \text{ microhenrys}$$

5. Determine the midband and high-frequency gain of the stage

$$K_0 = g_m R_L = 17.5\times 10^{-3}\times 1{,}800 = 31.5$$

$$K_h = \frac{K_0}{M_h} = \frac{31.5}{\sqrt{2}} = 22.2$$

6. Calculate the parameters of the coupling connection:

$$R_g = 30 R_L = 30\times 1{,}800 = 54.0 \text{ kilohms}$$

$$C_c = \frac{\tau_g}{R_g} = \frac{10\tau}{R_g} = \frac{20\times 10^{-6}}{54.3\times 10^3} = 370 \text{ picofarads}$$

7. Calculate the operating conditions of the valve for an anode supply voltage of 180 volts. For a typical case: $E_a = 150$ volts, $E_{g2} = 150$ volts.

Bias voltage

$$-E_g = V_{in} + 0.5 = 0.5 + 0.5 = 1.0 \text{ volt}$$

From the valve characteristic find the quiescent anode current

$$I_{a0} = 15 \text{ milliamperes}$$

The anode quiescent voltage

$$V_{a0}=E_{a\,sup}-I_{a0}R_L=180-15\times10^{-3}\times1.8\times10^{3}=150\ \text{volts}$$

Screen-grid current

$$I_{g2}=4.5\ \text{milliamperes}$$

Hence,

$$R_{g2}=\frac{E_{g2}}{I_{g2}}=\frac{150}{4.5\times10^{-3}}=33\ \text{kilohms}$$

Find the value of the self-bias resistor

$$R_k=\frac{|E_g|}{I_{a0}+I_{g2}}=\frac{1.0}{(15+4.5)\times10^{-3}}=51\ \text{ohms}$$

Determine the value of the cathode bypass capacitor

$$C_k=\frac{\tau_{max}}{\delta_k}g_m=\frac{2\times10^{-6}}{0.01}17.5\times10^{-3}=3.5\ \text{microfarads}$$

Find the value of the dropping resistor in the screen-grid circuit and the capacitance of the blocking capacitor:

$$R_{dr}=\frac{E_a-E_{g2}}{I_{g2}}=\frac{180-150}{4.5\times10^{-3}}\cong6.8\ \text{kilohms}$$

$$C_b=\frac{\tau_{max}}{R_{g2}\delta_s}=\frac{2\times10^{-6}}{33\times10^{3}\times0.01}=6{,}000\ \text{picofarads}$$

95. The Final Stage of a Video Amplifier

The output, or final, stage of a video amplifier above all serves to build up the signal voltage to a value necessary to control the indicator tube. This voltage, as already mentioned, is 20 to 100 volts.

The output voltage of an amplifier is decided by the voltage applied to the valve grid, the mutual conductance, and the anode load resistance

$$V_{out}=V_g g_m R_L$$

The anode load resistance R_L is determined by the limits of frequency and phase distortion. Usually, R_L does not exceed 1 to 3 kilohms.

The driving voltage is limited by how far the foot of the valve characteristic extends into the region of negative grid voltage, i. e. by the E_g' value. Therefore, in order to obtain the necessary voltage, the final stages should employ power remote-

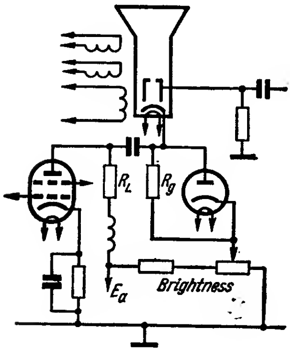


Fig. 15.23. Circuit of a final stage loaded into a beam intensity-modulated tube

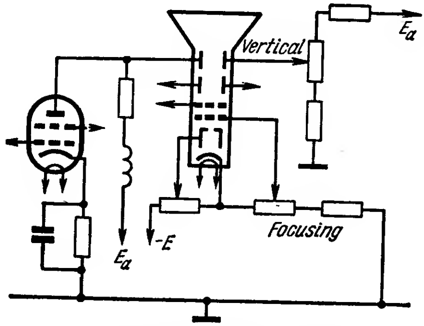


Fig. 15.24. Final stage loaded into a beam-deflecting tube

cutoff valves with a large emission current and considerable mutual conductance. The final stages of video amplifiers in TV and radar receivers use power pentodes and beam tetrodes.

For higher gain, the output stage should be frequency-compensated. Apart from reducing signal distortion, frequency compensation considerably increases the anode load resistance and, consequently, the stage gain.

The signal voltage developed at the detector output always contains, apart from the a. c. component, a slowly varying or d. c. component which determines the level of the signal. In the process of signal amplification, the d. c. component is lost, and the average signal level might be changed. In order to restore the original signal shape, the final stage incorporates suitable circuits to restore the d. c. component.

Operation of the final stage depends upon the relative position of the amplifier and cathode-ray tube (CRT). If the output stage and the CRT are close to each other, the voltage from the anode load is applied directly to the electrodes or deflecting plates of the CRT (Figs. 15.23 and 15.24). When the CRT and

Fig. 15.25. Block diagram of a final stage with a separately located cathode-ray tube

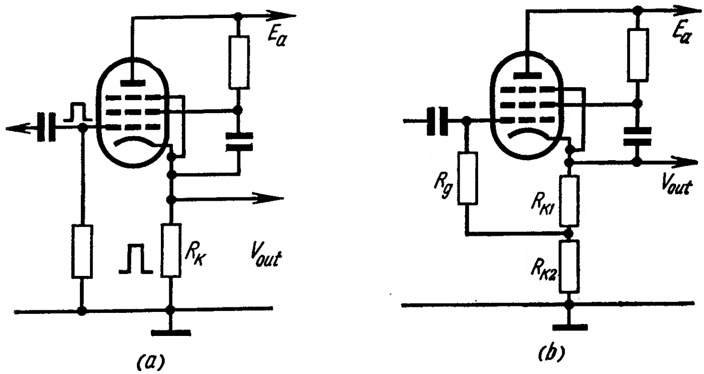


Fig. 15.26. Cathode follower

the final stage are placed at a distance from each other, an auxiliary stage is incorporated in the video amplifier, connected to the final stage by an r. f. cable (Fig. 15.25). The characteristic impedance of the cable is 50 to 100 ohms, as a rule. Therefore, connecting the cable directly to the anode load resistance may considerably lower the total resistance of the anode load and the stage gain. To avoid this, a *cathode follower* is incorporated in the circuit between the amplifier and the cable. The cathode follower (Fig. 15.26*a*) has a low output impedance and secures a perfect match between the video amplifier and the r. f. cable, thereby keeping the pulse polarity unchanged.

Consider operation of the cathode follower. A cathode follower is an amplifier stage which has the load connected in the cathode circuit while the anode is decoupled to earth. Therefore, the whole of the voltage developed across R_k is applied to the grid circuit of the same valve in phase opposition, while the output voltage is in phase with the input voltage. Thus, the cathode follower can be thought of as a single stage of negative voltage feedback with $\beta = 1$.

As we already know, the gain of a stage using negative feedback is

$$K_{fb} = \frac{K_0}{1+\beta K_0}$$

At $\beta = 1$

$$K_{fb} = \frac{K_0}{1+K_0} < 1 \qquad (15.40)$$

Thus, in a cathode follower the gain is always less than unity.
In Eq. (15.40), K_0 is the gain of an RC-coupled stage without feedback, equal to

$$K_0 = \frac{\mu R_L}{R_a + R_L}$$

Substituting this expression into Eq. (15.40) and noting that in our case $R_L = R_k$ we obtain

$$K_{fb} = \frac{\mu \dfrac{R_k}{R_a + R_k}}{1 + \dfrac{\mu R_k}{R_a + R_k}} = \frac{\mu R_k}{R_a + (1+\mu) R_k}$$

$$= \frac{\mu}{1+\mu} \frac{R_k}{\dfrac{R_a}{1+\mu} + R_k} = \mu_e \frac{R_k}{R_{ae} + R_k} \qquad (15.41)$$

where $\mu_e = \dfrac{\mu}{1+\mu}$ = equivalent amplification factor of the valve

$R_{ae} = \dfrac{R_a}{1+\mu}$ = equivalent a. c. resistance of the valve.

Thus, the change in amplifying properties of a stage with negative feedback may be thought of as being due to the effect of feedback on R_a and μ.

Writing μ_e in Eq. (15.41) in terms of g_m and R_{ae} gives

$$K_{fb} = g_m R_e \qquad (15.42)$$

where $R_e = \dfrac{R_{ae} R_k}{R_k + R_{ae}}$ is the equivalent output resistance of the cathode follower.

The fact that the equivalent resistance R_e of the stage is reduced to a few tens of ohms brings down the time constant $\tau_a = C_s R_e$, which determines the upper limit of the bandwidth. The bandwidth of the cathode follower, with C_s held constant, is increased just as many times as the stage gain is reduced by feedback. Practically, bandwidth is increased 10 to 20 times, with no compensation used. The cathode follower also broadens the bandwidth of the preceding stage in the video amplifier.

The capacitance C_s of the preceding stage, as we know, is the sum of the output capacitance of the stage, distributed wiring capacitance, and input capacitance of the valve in the following stage.

A detailed analysis of cathode-follower operation shows that its input capacitance C_{in} is approximately equal to the anode-grid capacitance C_{ag}. As a proof, the input capacitance of the cathode follower is

$$C_{in}=C_{ag}+C_{gk}(1-K_{fb})$$

Since

$$K_{fb}\cong 0.8 \text{ to } 0.9$$

it follows that

$$C_{in}=C_{ag}+C_{gk}(0.1 \text{ to } 0.2)\cong C_{ag}$$

In an RC-coupled amplifier

$$C'_{in}=C_{gk}+(1+K)C_{ag}>C_{ag}$$

The voltage drop across R_k may turn out to be greater than is required to supply negative voltage feedback. Therefore, practical cathode followers use the arrangement of Fig. 15.26*b*.

In this arrangement, a fixed potential difference between grid and cathode is equal to the voltage drop across R_{k1}, and not across the whole of the cathode load.

As compared with the simple cathode follower, the one shown in Fig. 15.26*b* has a higher input resistance

$$R_{in}=\frac{V_{in}}{I_{in}}$$

where I_{in} is defined as

$$I_{in}=V_g/R_g=\frac{V_{in}-KV_{in}\dfrac{R_{k2}}{R_{k1}+R_{k2}}}{R_g}=\frac{V_{in}\left(1-K\dfrac{R_{k2}}{R_{k1}+R_{k2}}\right)}{R_g}$$

Substituting this expression in that for the input resistance gives

$$R_{in}=V_{in}/I_{in}=\frac{R_g}{1-K\dfrac{R_{k2}}{R_{k1}+R_{k2}}}$$

Thus, the input resistance of a cathode follower may be several times that of a conventional stage.

D.C. Restoration. This consists in maintaining a definite average signal voltage produced at the detector output. Signals obtained at the output of a TV or radar receiver are, as a rule, one-sided (Fig. 15.27). Such signals, apart from a.c. components, also have a d.c. component.

In an RC-coupled circuit, which is the basic configuration of video amplifiers, the d.c. component is lost in the coupling

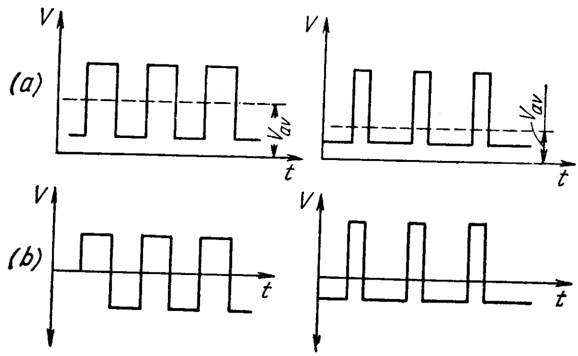


Fig. 15.27. Diagrams showing changes in the signal level with loss of d.c. component

capacitor C_c which is an open-circuit to direct current. As a result, there is a considerable change in the maximum or minimum level of the signal.

Figure 15.27*a* shows two signals with a different ratio of pulse duration to pulse spacing, so that they have the same peak but different average values V_{av}. Figure 15.27*b* shows the same signals after they have passed through an amplifier. The change in the maximum to minimum level ratio causes a change in the background brightness of the tube and the accuracy of signal reproduction on the CRT screen.

In order to restore the d.c. component in the signal, an additional diode is connected into the grid circuit of the final-stage valve (Fig. 15.28). In the no-signal state, the grid bias is E_{go},

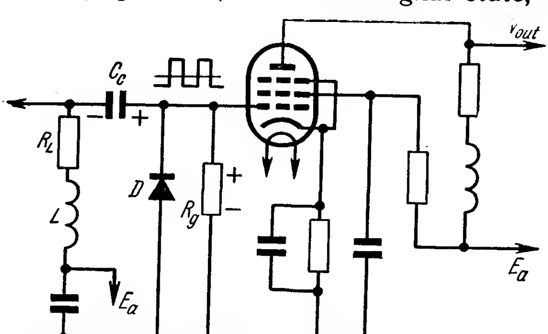


Fig. 15.28. D.c. restoration circuit

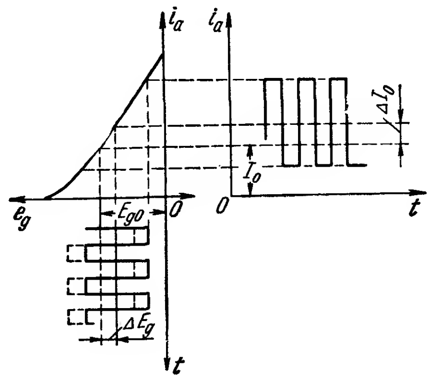


Fig. 15.29. Diagram of operation of a d.c. restoration circuit

and the anode voltage of the diode is very low, being negative to the cathode. A positive pulse will not change the condition of the diode, and the initial bias at the amplifier valve will remain unchanged. A negative pulse causes the diode to conduct, and a current begins to flow through the diode and charges up C_c to the reference level. The next positive pulse cuts off the diode, and C_c discharges through R_g, producing a positive voltage drop across this resistor. The average value of this voltage is proportional to the average value of the incoming pulses.

This additional positive voltage shifts the operating point of the amplifier valve to the right into the region of a higher quiescent current (Fig. 15.29). That is, the voltage generated across C_c may be regarded as being added to the pulse appearing across R_L, thereby producing a d.c. restored signal across the grid resistance R_g.

Review Questions

1. Which characteristics of the video amplifier control the choice of circuit configuration?
2. What are the requirements for a cathode-follower valve?
3. Why is the gain of a cathode follower less than unity?
4. Why is the screen-grid in a cathode-follower valve decoupled to the cathode and not earth?
5. How does the polarity of the output pulse control the choice of the valve for the final stage?

96. Calculation of the Final Stage for a Video Amplifier

Given:

1. Maximum output voltage, V_{max}.
2. Maximum pulse duration, τ_{max}.
3. Minimum pulse rise time, t_r.

To Find:

1. Type of valve.
2. Circuit parameters.
3. Alternating anode current, $\Delta I_{a\,max}$.
4. Bias voltage, E_g.
5. Input voltage, V_{in}.

Design Procedure:

1. The valve is chosen for the maximum voltage to be obtained at the stage output. The maximum voltage that can develop across the anode load resistance depends upon the value of this resistance and the incremental anode current

$$V_{max} = \Delta I_{a\,max} R_L \tag{15.43}$$

In order to utilize the valve as fully as possible, the incremental anode current $\Delta I_{a\,max}$ should be 90 per cent of the maximum anode current at the anode dissipation quoted in the valve manual

$$\Delta I_{a\,max} = 0.9 I_{a\,max}$$

The anode load resistance depends, as we already know, on the circuit configuration and the shape of the amplified signals.

In the simplest case, that is, an uncompensated amplifier, R_L may be defined in terms of signal rise time and the distributed capacitance of wiring as follows

$$R_L = \frac{t_r}{2.2C_s}$$

Substituting the values of $\Delta I_{a\,max}$ and R_L into Eq. (15.43), we obtain:

$$V_{max} = 0.9 I_{a\,max} \frac{t_r}{2.2C_s} = D_0 t_r$$

where $D_0 = \frac{0.9 I_{a\,max}}{2.2 C_s}$ is the grid-through, a coefficient characterising the valve employed in the final stage.

In compensated stages, R_L may be made greater than in uncompensated ones; this will increase the coefficient D.

In a stage with simple shunt inductive compensation and with an overshoot of 6 per cent, the coefficient D is

$$D = 1.8 D_0$$

In the same stage, with an overshoot of 1 per cent, the coefficient D will be

$$D = 1.67 D_0$$

In stages with modified shunt compensation, the value of D is

$$D = \frac{2.2}{q} D_0$$

where the value of q is taken from Table 15.1.

In the general case, the maximum anode voltage of the valve is given by

$$V_{max} = D t_r \tag{15.44}$$

Table 15.3 lists approximate values of $I_{a\,max}$, C_s and D_0, proposed by A. P. Sivers for selected valves.

When V_{max} and t_r are specified in advance, it is possible, applying Eq. (15.44), to determine the necessary Q of the valve and select the type of valve from Table 15.3.

TABLE 15.3

Valves	6Ж3П	6Ж9П	6Ж11П	6Ж20П-1	6Ж20П-2	6П9	6П14П	6П18П	6П6С	6П3С
$I_{a\,max}$, mA	10	15.6	25	16	28	30	48	53	45	75
$C_{s\,min}$, pf	20	25	25	20	25	28	29	27	27.5	31.5
D_0, V/μ sec	180	255	410	330	460	440	670	800	600	775
I_g, mA	2	4.5	8	5	9	6.5	5	8	5	5

2. The order in which circuit parameters should be calculated depends on the circuit configuration of the stage and its operating conditions. According to the expected performance, the final

stage of a video amplifier may be the conventional RC-coupled circuit or a circuit with frequency compensation. The latter may be employed in cases where the selected valve cannot provide the specified output voltage when connected into an uncompensated amplifier.

After the circuit configuration has been selected, the circuit parameters are calculated just as in a video voltage amplifier.

3. The incremental anode current is determined from the maximum output voltage V_{max} and anode load resistance R_L

$$\Delta I_{a\,max} = \frac{V_{max}}{R_L}$$

4. The bias voltage E_g is so selected that the anode dissipation will not exceed the maximum safe value

$$P_a = I_0 V_{a0} < P_{a\,safe}$$

5. The input voltage is usually found graphically from the valve characteristic, as shown in Fig. 15.30.

If the final stage is loaded into a cable transmission line with a low wave impedance, it is essential to use a cathode follower. Without it, the low wave impedance of the line would shunt the anode load of the earthed-cathode stage, and the stage gain would be too low. Besides, in order to retain the flat top of the pulse, this circuit would need a high-value blocking capacitor.

To prevent reflections from the end of the cable, its output is terminated in a resistance R_{k2}, equal to its wave impedance

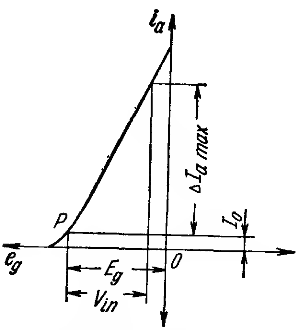


Fig. 15.30. Graphic calculation of the final stage

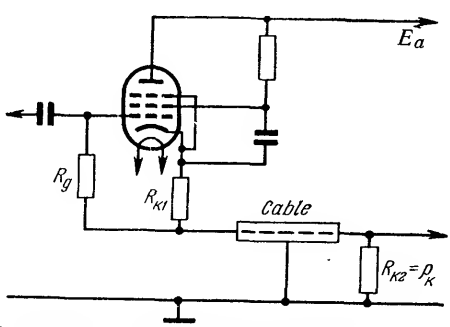


Fig. 15.31. Cathode-coupled final stage of a video amplifier

ρ_k (Fig. 15.31). The incremental cathode current in this circuit is found from the maximum output voltage and R_{k2}:

$$\Delta I_{k\,max} = V_{max}/R_{k2} = V_{max}/\rho_k$$

The bias voltage and the signal voltage, V_{in}, are found from the cathode-grid characteristic, which is the overall anode and screen-grid characteristic

$$i_k = i_a + i_{g2} = F(e_g)$$

The bias voltage is chosen such that the power dissipation in the quiescent state will not exceed the safe value

$$P_a = I_{k0} E_a < P_{a\,safe}$$

where

$$I_{k0} = F(E_g) < \frac{P_{a\,safe}}{E_a}$$

A graphical method for determining the bias voltage is shown in the plot of Fig. 15.32. The quiescent cathode current is laid off as ordinate (segment OA'). From point A', a perpendicular is raised until it intersects the cathode-grid curve at point A. From point A, a perpendicular is dropped onto the axis of abscissas at point A''. The requisite bias voltage will be represented by the segment $A''O$. Now segment $A'B'$, representing the requisite incremental current, $\Delta I_{k\,max}$, is laid off along the cathode-current axis from point A', and a perpendicular is erected from point B' until it cuts the curve at point B. Now a perpendicular is dropped from point B onto the grid-voltage

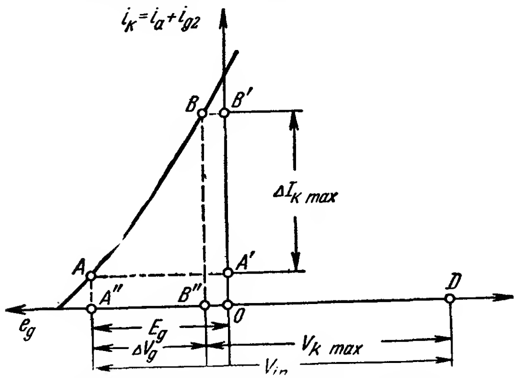


Fig. 15.32. Graphic calculation of input voltage for a cathode follower

axis so that it cuts the curve at point B''. Segment $A''B''$ will represent the requisite incremental grid voltage, ΔV_g.

The self-bias resistor R_{k1} is found by the equation

$$R_{k1}=\frac{|E_g|}{I_{k0}}$$

Before finding the input-signal voltage, it is necessary to determine the maximum voltage across the cathode load

$$V_{k\,max}=R_k\Delta I_{k\,max}=(R_{k1}+R_{k2})\,\Delta I_{k\,max}$$

The incremental cathode current, $\Delta I_{k\,max}$, necessary to generate this voltage, is due to the application to the grid of a voltage ΔV_g equal to the difference between the input voltage and the voltage drop $V_{k\,max}$:

$$\Delta V_g=V_{in}-V_{k\,max}$$

Hence,

$$V_{in}=\Delta V_g+V_{k\,max}$$

V_{in} can also be found graphically by laying off segment $B''D$, representing $V_{k\,max}$, from point B''. Segment $A''D$ will represent the sought input voltage.

SUMMARY

1. A video amplifier is a broad-band amplifier intended to amplify pulses of various duration and within a large range of frequencies.
2. Amplification over a large range of frequencies can only be provided by an RC-coupled circuit.
3. The bandwidth of an RC-coupled amplifier can be broadened by reducing the anode load resistance and by use of frequency compensation.
4. In the high-frequency range the best method of compensation is to place coils in the anode and load circuits of the valve. These coils both improve the frequency response and increase the stage gain.
5. High-frequency compensation with an RC network connected into the load or cathode circuit of the valve decreases the total stage gain.
6. Low-frequency compensation is provided by a decoupling filter connected in the anode circuit.
7. Simultaneous low- and high-frequency compensation can be obtained by use of negative feedback.

8. The final stage of a video amplifier is intended to ensure normal operation of the cathode-ray tube.

9. The final stage may be connected to the CRT electrodes either directly or by cable.

10. When the final stage and the tube are located separately from each other, a cathode follower must be connected between them.

11. The cathode follower is used to maintain the polarity of the signal and to match the output stage to a cable transmission line of low input impedance.

12. The final stage uses a d.c. restorer, an electronic circuit which re-inserts the d.c. component of the signal lost during amplification.

■ Problems

15.1. Determine the pulse rise time t_r in a stage with $R_L = 2{,}500$ ohms, and $C_s = 40$ picofarads.

Answer: $t_r = 0.23$ μsecond.

15.2. Pulse duration $\tau = 100$ μsec; $R_g = 250$ kilohms; $C_c = 400$ picofarads; $V_g = 100$ volts.

Find the droop of the pulse top.

Answer: $\Delta V_g = 9.5$ volts.

15.3. From the data of Problem 15.2 find the capacitance of C_c that will limit the droop to 5 volts.

Answer: $C_c = 8{,}000$ picofarads.

15.4. An amplifier has a bandwidth of 2 megahertz. The limit of frequency distortion is $M = 1.05$. The distributed capacitance $C_s = 30$ picofarads. Determine the anode load resistance R_L and inductance of the compensating coil L.

Answer: $R_L = 2{,}400$ ohms, $L = 72$ microhenrys.

15.5. Calculate the parameters of a stage employing modified shunt inductive frequency compensation and a 6Ж9П valve. Pulse duration $\tau = 2$ μsec; distributed capacitance $C_s = 25$ picofarads; grid-leak resistance of the following stage $R_g = 200$ kilohms; droop $\delta_g = 0.05$.

Answer: $R_L = 7$ kilohms; $L = 180$ microhenrys; $L_a = 880$ microhenrys; $R_{sh} = 14$ kilohms; $C_c = 200$ picofarads.

15.6. Find the gain of a cathode follower using a 6П3С valve and loaded into a cable with a characteristic impedance of 100 ohms.

Answer: $K = 0.375$.

CHAPTER XVI

FUNDAMENTALS OF PRACTICAL DESIGN OF PULSED-RADAR RECEIVERS IN THE SHF BAND

97. General

The design of pulsed-radar receivers for the SHF band is more complicated than that of radio receivers operating on long, medium and short waves. To begin with, they call for a broad bandwidth, running into many megahertz and a very high sensitivity of the order of 10^{-9} to 10^{-10} milliwatt.

This sensitivity, which is comparable with internal (receiver) noise, coupled with a broad bandwidth, is obtained by use of low-noise circuits in the early stages of the receiver and multistage intermediate-frequency amplifiers. Besides, radar receivers incorporate many auxiliary circuits, such as AFC, different kinds of gain control, etc. All this complicates the circuitry and design of the receiver.

The design of a radar receiver consists of preliminary and final calculations. During the preliminary stage the functional units of the receiver are selected and data necessary for the calculation of the individual circuits and stages are clarified.

The design of radar receivers is carried out according to specifications which stipulate the following:

1. Application of the receiver.
2. Wavelength λ or frequency f.
3. Pulse duration, τ_p.
4. Pulse repetition frequency, F.
5. Pulse rise time, t_r.
6. Signal-to-noise power ratio, α_p.
7. Indicator type (the specificatons must show whether the indicator and receiver are to be separated from each other).
8. Receiver output voltage, V_{out}.
9. Control requirements.

98. Preliminary Calculations

Break-down of Pulse Rise Time among the Receiver Units. The bandwidth of a radar receiver is wholly decided by the pulse rise time t_r. Whether the specification can be met with a minimum number of stages depends to a great measure on the correct break-down of rise time among the receiver units.

Using the notation:

$t_{r\,v}$ = pulse rise time in the video amplifier

$t_{r\,d}$ = pulse rise time in the detector

$t_{r\,rf}$ = pulse rise time in the r.f. section, the total pulse rise time t_r in the receiver may be defined as

$$t_r = \sqrt{t_{r\,v}^2 + t_{r\,d}^2 + t_{r\,rf}^2} \tag{16.1}$$

The values of $t_{r\,v}$ and $t_{r\,rf}$, $t_{r\,d}$ and $t_{r\,rf}$ are related as

$$\frac{t_{r\,v}}{t_{r\,rf}} \cong 0.4; \quad \frac{t_{r\,d}}{t_{r\,rf}} \cong 0.3$$

After substitutions, we obtain

$$\left.\begin{aligned} t_{r\,rf} &\cong 0.9t_r \\ t_{r\,d} &\cong 0.27t_r \\ t_{r\,v} &\cong 0.36t_r \end{aligned}\right\} \tag{16.2}$$

The relations thus obtained are the basis for the selection of functional units for the video and r.f. sections.

Choice of Functional Units for the R.F. Section. The functional set-up of the r.f. section in a radar receiver depends only on the frequency at which the radar is to operate. The number of stages in one unit or another and their circuit configuration are decided by a series of factors which must be determined in the process of design. An SHF receiver has no r.f. amplifier, and includes only a frequency changer and i.f. amplifier (see Fig. 14.2). We shall leave out the use of a travelling-wave tube for r.f. amplification in the SHF band.

To finalize the choice of functional units, the following factors should be determined:

(1) bandwidth of the whole receiver;

(2) minimum bandwidths of the individual units and circuits of the receiver;

(3) receiver sensitivity;

(4) the gain of the r.f. section and break-down of the gain among receiver units;

(5) the intermediate frequency.

Determining the Receiver Bandwidth. The bandwidth of a receiver depends upon the expected use and performance of the radar equipment. In cases where the distance of radar operation is of prime importance and the ranging accuracy is not critical, the bandwidth of the receiver may be taken as

$$2\Delta F'_r = \frac{1 \text{ to } 1.37}{\tau}$$

The ranging accuracy depends upon the steepness of the leading edge of the pulse, which is in turn determined by the rise time t_r stipulated in the specifications. Therefore, if high ranging accuracy is essential, the bandwidth is chosen to be much wider than the optimum one

$$2\Delta F'_r(\text{MHz}) \cong \frac{0.75}{t_{r\,rf}\,(\mu\,\text{sec})} \tag{16.3}$$

The obtained bandwidth $2\Delta F'_r$ should be increased by Δf_v to provide for possible variations in the frequency of the transmitter and the local oscillator

$$2\Delta F_r = 2\,(\Delta F'_r + \Delta f_v) \tag{16.4}$$

For receivers without AFC, Δf_v should be learned from an authority in radar design. For receivers with AFC, Δf_v, representing the AFC error, may be taken equal to 0.2 to 0.4 megahertz.

Determining the Minimum Bandwidth of the Individual Units and Circuits of the Receiver. In the SHF range, the bandwidth of the receiver and of its circuits are usually determined for the same value of frequency distortion, $M=0.7$. With this method, each circuit in the receiver has a bandwidth necessarily wider than the total receiver bandwidth $2\Delta F_r$. To establish the relation between the bandwidth of the whole system and those of its elementary circuits we shall use the following approximate equality:

$$t_r \cong \sqrt{t_{r1}^2 + t_{r2}^2 + \ldots + t_{rn}^2} \tag{16.5}$$

If the individual units and the receiver as a whole have single-humped resonance curves, then

$$t_{r1} = \frac{0.75}{2\Delta F_1}; \quad t_{r2} = \frac{0.75}{2\Delta F_2}; \quad t_{r3} = \frac{0.75}{2\Delta F_3}, \quad \ldots, \quad t_{rn} = \frac{0.75}{2\Delta F_n}$$

After substitutions, we obtain

$$\frac{1}{(2\Delta F)^2}=\frac{1}{(2\Delta F_1)^2}+\frac{1}{(2\Delta F_2)^2}+\ldots+\frac{1}{(2\Delta F_n)^2}$$

Hence, the total bandwidth of the system is

$$2\Delta F=\frac{2\Delta F_1}{\sqrt{1+\left(\frac{2\Delta F_1}{2\Delta F_2}\right)^2+\left(\frac{2\Delta F_1}{2\Delta F_3}\right)^2+\ldots+\left(\frac{2\Delta F_1}{2\Delta F_n}\right)^2}} \tag{16.6}$$

For a special case, where all n circuits have the same bandwidth

$$2\Delta F_1=2\Delta F_2=\ldots=2\Delta F_n$$

we obtain

$$2\Delta F=\frac{2\Delta F_1}{\sqrt{n}} \tag{16.7}$$

In receiver design, it is usually necessary to find the bandwidth of the elementary circuits from the specified overall bandwidth of the receiver, $2\Delta F_r$. An "elementary circuit of the receiver" may be a valve loaded into a tuned circuit, the aerial-input circuit of the receiver, the coupling network of the i.f. amplifier, or a TR-switch. In preliminary design, the number of elementary circuits, n, may be assumed to be from 6 to 12. Then the bandwidth of a single elementary circuit will be

$$2\Delta F_1=(2.5 \text{ to } 3.5)\,2\Delta F_r \tag{16.8}$$

In preliminary calculations it is necessary to determine the bandwidths of the receiver circuits that will be subsequently calculated separately.

For the SHF range, the following bandwidths should be determined: for the TR switch, i.f. amplifier coupling network, and main i.f. amplifier (the effect of the mixer upon the bandwidth is usually neglected).

The bandwidth of the TR switch is given by

$$2\Delta F_{sw}=\frac{f_s}{Q_{sw}} \tag{16.9}$$

where f_s is the signal frequency, and Q_{sw} is the figure-of-merit of the switch circuit which can be taken from Table 14.4.

The bandwidth of the i.f. amplifier coupling network is determined from Eq. (16.8

$$2\Delta F_{in\,if}=3\times 2\Delta F_r \tag{16.10}$$

After this, the overall bandwidth of the switch and i.f. amplifier coupling network, $2\Delta F_{sw\ in\ if}$ is found from Eq. (16.6), with the subscripts properly changed

$$2\Delta F_{sw\ in\ if}=\frac{2\Delta F_{sw}}{\sqrt{1+\left(\frac{2\Delta F_{sw}}{2\Delta F_{in\ if}}\right)^{2}}} \qquad (16.11)$$

Changing the subscripts in (16.6) gives

$$2\Delta F_{r}=\frac{2\Delta F_{sw\ in\ if}}{\sqrt{1+\left(\frac{2\Delta F_{sw\ in\ if}}{2\Delta F_{if}}\right)^{2}}}$$

and the bandwidth of the i.f. amplifier

$$2\Delta F_{if}=\sqrt{\frac{(2\Delta F_{r})^{2}\,(2\Delta F_{sw\ in\ if})^{2}}{(2\Delta F_{sw\ in\ if})^{2}-(2\Delta F_{r})^{2}}} \qquad (16.12)$$

Choice of the Intermediate Frequency. When choosing the value of f_i, one should be guided by the following basic considerations:

1. If the intermediate frequency is to be reliably filtered out past the video detector, the upper modulating frequency F_h should be one-fifth to one-tenth of the intermediate frequency, so that $F_h=\Delta F_r$.

2. For better reproduction of the pulse envelope, the period of the intermediate frequency should be 0.05 to 0.1 of pulse duration τ.

3. A low intermediate frequency secures better stability, reduces the effect of valve replacement, and lowers the noise factor. At the same time, it calls for a more sophisticated AFC system.

4. To standardize the units, the intermediate frequency is usually taken to be 30 or 60 megahertz.

Determining Receiver Sensitivity. As established in Sec. 69, the effective sensitivity $P_{s\,ef}$ of radar receivers depends on the signal power developed in the aerial:

$$P_{s\,ef}=kTN\Delta f\alpha_{p}$$

In preliminary calculation, the noise factor N may be taken equal to 17 decibels for a wavelength of 3 centimetres and 15 decibels for 10 centimetres. The noise bandwidth Δf of the receiver is assumed to be equal to the receiver bandwidth, $2\Delta F_r$. If the value of N is expressed numerically, and the bandwidth Δf is in megacycles, and if we note that the product kT equals 4×10^{-15} watts/megahertz, then $P_{s\,ef}$ will be in watts.

Determining the Necessary Gain of the R.F. Section. In the SHF range, the signal is fed from the aerial to the detector over two different paths. One is from the aerial to the i.f. amplifier coupling network, and the other from the i.f. amplifier coupling network to the detector. In the former, the transmission of the signal is more conveniently characterized by the power transfer ratio. In the latter, this is best done in terms of the voltage transfer ratio.

To begin with, it is necessary to determine the power transfer ratio of the first path comprising the coaxial or waveguide transmission line, TR switch and crystal mixer. If we denote as respectively the power transfer ratios of these components K_{pf}, K_{pp}, and K_{pm}, the total transfer ratio of the path will be

$$K_p = K_{pf}K_{pp}K_{pm}$$

The values of K_{pp} and K_{pm} may be taken from Tables 14.4 and 14.5. We shall assume that the transfer ratio of the transmission line is $K_{pf}=0.9$. The signal power at the input of the i.f. amplifier is

$$P_{in\,if} = P_{s\,ef}K_p$$

Noting that for a matched load

$$P = \frac{V^2}{4R}$$

the signal voltage at the i.f. preamplifier will be

$$V_{p\,if} = 2\sqrt{P_{s\,ef}K_pR} \tag{16.13}$$

where R is input resistance of the i.f. amplifier coupling network. When the i.f. amplifier is perfectly matched to the frequency changer, R is equal to the output resistance of the changer R_{if} at the intermediate frequency. R_{if} may be found in Table 14.4. The coupling network of the i.f. amplifier may be a double-tuned transformer in which the coupling between the primary and secondary is chosen to comply with the matching criterion just stated.

To determine the signal voltage to be applied to the grid of the first valve in the i.f. amplifier, V_{in1}, it is necessary to find the transformation ratio of the coupling network. Using the same notation as in Sec. 86, the transformation ratio is

$$m=\frac{V_1}{V^2}$$

Noting that $V_1=V_{in\,if}$ and $V_2=V_{in1}$, we obtain

$$m=\frac{V_{in\,if}}{V_{in1}}$$

The signal voltage to be applied to the grid of the first valve is

$$V_{in1}=\frac{V_{in\,if}}{m}$$

If the coupling network of the i.f. amplifier is to secure both a perfect match and a minimum noise factor in the i.f. amplifier, the transformation ratio m should be such as given by Eq. (14.44). Then

$$V_{in1}=\sqrt{\frac{G_{if}}{G_{in}+G_{k2opt}}}V_{in\,if} \tag{16.14}$$

The method for determining G_{if} and G_{in} is explained in Sec. 86; the value of G_{k2opt} is determined from Eq. (14.43).

Determine the necessary voltage gain up to the detector input:

$$K'_{if}=\frac{V_{in}}{V_{in\,if}} \tag{16.15}$$

V_{in} at the detector input should be one or two volts.

With a gain margin of 2 or 3, we finally obtain

$$K_{if}=(2\text{ to }3)\,K'_{if} \tag{16.16}$$

Choice of Functional Units for the Receiver Video Section. When we say "the video section of a radar receiver", we mean the stages following the i.f. amplifier, namely the video detector, limiting amplifier, and video amplifier.

The output indicator of a radar receiver is usually a cathode-ray tube with intensity modulation or deflection of the beam The voltage applied to the CRT is stated in the specifications. The indicator and video amplifier may be either combined in a single unit, or may be separate. In the latter case, the indicator is connected to the video amplifier by an r.f. cable, and the output stage of the amplifier is a cathode follower.

The video detector in a radar receiver is generally a diode detector.

The functional set-up for the video section is chosen from considerations of the desired gain. Since the input voltage of the video detector is always known ($V_{in} = 1$ or 2 volts) and the transfer ratio of the detector may be assumed to be $K_d \cong 0.5$, the detector output voltage corresponding to the actual sensitivity of the receiver can be determined as follows:

$$V_{d\,out} = K_d V_{in} \tag{16.17}$$

where V_{in} and $V_{d\,out}$ are the peak voltage values.

Since the signal at the receiver input can possibly exceed the receiver sensitivity, the input voltage to the limiting amplifier, V_0, can be determined as advised in Sec. 90.

For a CRT with intensity modulation

$$V_0 = 2V_{d\,out}$$

For a CRT with a beam deflection

$$V_0 = (2 \text{ to } 5)\, V_{d\,out}$$

Now determine the necessary gain of the video section

$$K_v = \frac{V_{out\,max}}{V_0} \tag{16.18}$$

The value of K_v, as obtained from Eq. (16.18), usually exceeds the gain offered by a single stage. Therefore, several, usually three, stages are used between the video detector and indicator. When a remote indicator is used, a cathode follower is incorporated in the final stage. The transfer ratio of the follower may be taken equal to 0.8.

The circuit configuration of the video detector and the polarity of the voltage at its output depend upon the way the CRT is controlled. The necessary gain of the video section usually does not exceed 80 to 100. Therefore each stage should have a gain of nearly ten, which fact helps to obtain the required bandwidth or the required rise time of the video pulse. If the video amplifier consists of two stages, the video pulse rise time per stage is given by

$$t_{rv} = \sqrt{t_{r\,v1}^2 + t_{r\,v2}^2} \tag{16.19}$$

99. Final Calculations

The starting point for final calculations is provided by the specification and the preliminary calculations.

Since the method for the calculation of the basic circuits of a radar receiver has been presented in Chapter XIV, we shall only give recommendations as to the order of final calculations and some additional notes.

The final calculations cover:

1. The noise factor of the i.f. amplifier.
2. The i.f. amplifier coupling network.
3. The i.f. amplifier.
4. The overall frequency response of the receiver which involves knowledge of the frequency responses of the TR switch, i.f. amplifier coupling network, and i.f. amplifier.

The frequency response of the TR switch is found from Eq. (5.4), which holds for a single resonant circuit. The Q of the TR switch is found from Table 14.5, according to the selected type of switch.

The frequency response of the double-tuned coupling network can be found from the equation

$$Y = \frac{1}{\sqrt{1+4x^4}}$$

where

$$x \cong \frac{2\Delta f}{f_0} \frac{Q_{ef1} Q_{ef2}}{Q_{ef1}+Q_{ef2}}$$

The frequency response of the i.f. amplifier is plotted on the basis of bandwidth ratios (relative bandwidth) $K_{20\,db}$ and $K_{40\,db}$ and the bandwidth $2\Delta F_{if}$, obtained by calculation.

The frequency responses should be plotted on a common graph. The overall frequency response of the receiver is then obtained by multiplying the ordinates of each curve for the same amount, Δf, from resonance.

5. The video detector.
6. The video amplifier.
7. The basic characteristics of the receiver:

(a) receiver noise, using Eq. (14.38);

(b) the bandwidth between the 3-db points on the overall frequency response curve;

(c) effective sensitivity, using Eq. (14.6) and the noise factor and bandwidth of the receiver in (a) and (b) above.

8. A complete circuit diagram of the radar receiver.

100. Practical SHF Radar Receiver

Figures 16.1 and 16.2 show circuit diagrams of an SHF receiver made up of the stages discussed in the preceding sections. The receiver uses a twin mixer, a separate AFC section, and an IAGC circuit.

Figure 16.1 shows the twin mixer, the i.f. preamplifier of the receiver, and the separate automatic frequency control (AFC) section. The main i.f. amplifier, video detector, video amplifier and the IAGC are shown in Fig. 16.2.

The twin mixer is made up of two crystal mixers located in a common waveguide cavity. The aerial signal is fed to the receiver mixer through the TR switch *1*, while the signal from the magnetron is applied to the AFC mixer via an attenuator *2*. The twin mixer is coupled to a common local oscillator through a jack *3*. The local oscillator is a mechanically tuned klystron.

The i.f. signals from the mixers are fed over r.f. cables to the i.f. preamplifier of the receiver and to the AFC mixer. The i.f. preamplifier consists of a coupling network and three amplifying stages. The coupling network is a double-tuned transformer. To decrease the undesirable capacitive coupling between the coils, the primary is split up into two sections, L_1 and L_2. Of these two sections, only L_2 is inductively coupled to the secondary L_3. The capacitance of the primary is that of the wiring, and the capacitance of the secondary is the input capacitance of the valve.

The first two stages of the i.f. preamplifier (V_1 and V_2) make up a cascode circuit ensuring a low noise level at the input of the receiver. In the first stage the cathode is earthed and series feed is used; in the second stage, the grid is earthed and the stage uses parallel feed. The tuned-circuit inductances are L_5 and L_6. L_4 serves to neutralize C_{ag} of V_1.

The intermediate-frequency signal from the i.f. preamplifier is conveyed to the main i.f. amplifier by a stage, V_3, using a coaxial transmission line K_2.

The i.f. amplifier consists of a coupling network and five single-tuned stages using pentodes V_4-V_8. The i.f. signal is applied to the tuned circuit formed by L_8 and the input capa-

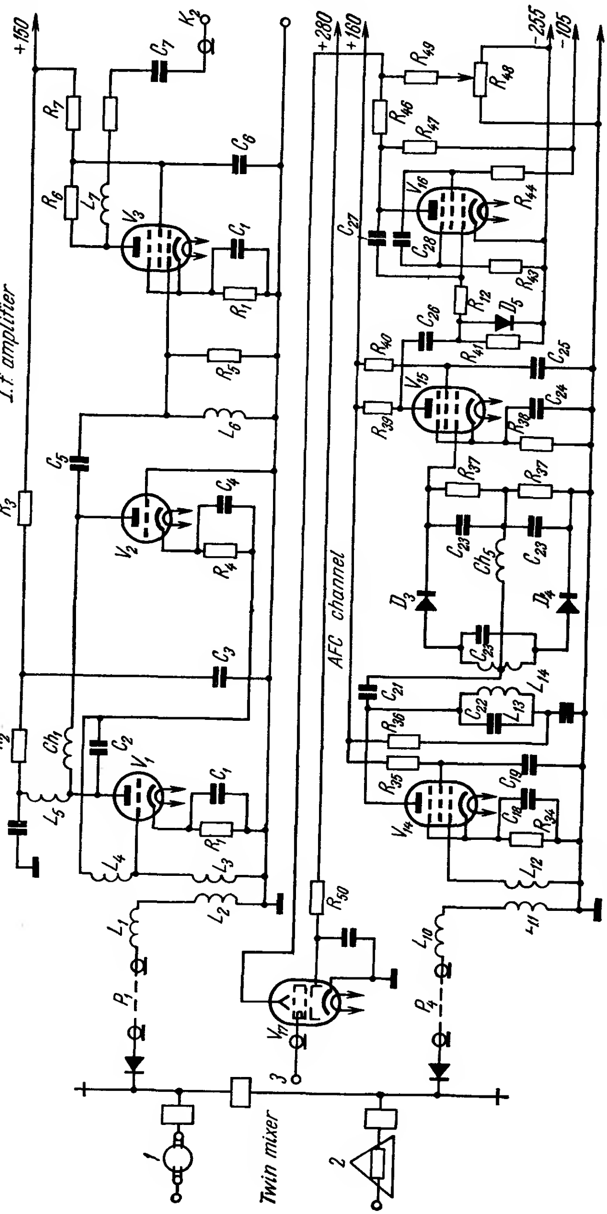


Fig. 16.1. Twin mixer, i.f. preamplifier and AFC channel

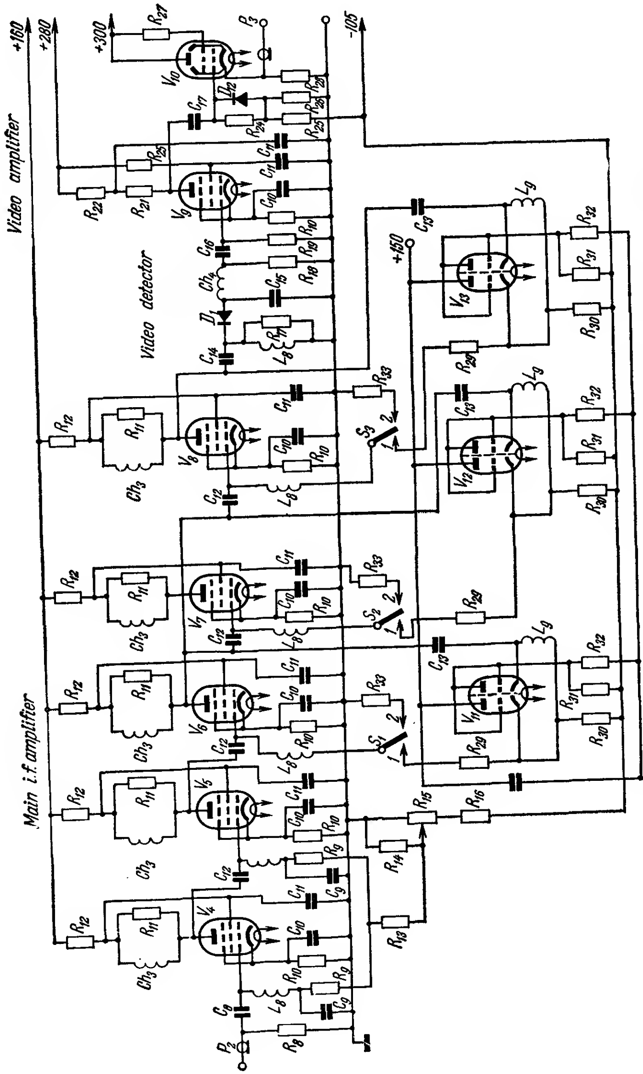


Fig. 16.2. *Main i.f. amplifier, IAGC, video detector, and video amplifier* citance of valve V_4 over the coaxial cable K_2 and through a coupling capacitor C_8. R_8 at the input of the i.f. amplifier serves to match this amplifier to the coaxial line.

All five stages of the i.f. amplifier employ parallel feed, have their tuned circuits connected in the grid leads, and are of similar type. The chokes Ch_3 connected in parallel with the anode resistor R_{11} of each stage minimize the d.c. voltage drop across these resistors. To broaden the bandwidth, the tuned circuit of the last stage is shunted by R_{17}.

Manual gain control in the receiver is effected in the first two i.f. stages with a potentiometer R_{15}, from which an adjustable negative bias may be applied to the control grids of valves V_4 and V_5, through the decoupling filters R_9C_9.

The video detector D_1 uses a semiconductor diode and is arranged into the usual circuit described in Sec. 88.

The video amplifier consists of a pentode stage (valve V_9) and a final stage, V_{10}, which is a cathode follower. The crystal diode D_2 connected in the grid circuit of the final stage operates as a d.c. restorer.

The video signals taken from the load resistor R_{28} of the cathode follower are fed over an r.f. cable to the indicator of the radar station.

The IAGC circuit consists of three similar stages, V_{11} through V_{13}, using double triodes. Each IAGC stage automatically controls the gain of one i.f. stage (V_6-V_8). The right-hand triode of each valve is used as a diode detector with a negative output voltage. The left-hand triode works as a cathode follower. The negative voltage taken from the cathode loads R_{30} is fed via contact *1* of switches S_1, S_2 and S_3 to the grids of the corresponding valves. In position *2*, the IAGC circuit is disabled.

The AFC section contains a coupling circuit, an AFC amplifier, a discriminator, and control circuit. The coupling circuit of the AFC section does not differ in any respect from that of the receiver section.

The i.f. amplifier of the AFC section employs a pentode; from its anode load the i.f. voltage is applied to the stagger-tuned discriminator. The non-linear elements used in the discriminator are crystal diodes D_3 and D_4.

The control circuit includes an amplifier V_{15} intended to amplify the discriminator signal and a sawtooth sweeping system consisting of a crystal diode D_5 and a pentode V_{16}, connected in a phantastron circuit. The sawtooth voltage from the anode of the phantastron is fed to the repeller electrode of the klystron V_{17} and controls its frequency.

The receiver is powered by two rectifiers. One develops positive voltages (150, 160, 280, and 300 volts); the other one negative voltages (—105 and —255 volts).

The positive voltages go to the anode and screen-grid circuits of the various stages. —255 volts is applied to the repeller electrode of the klystron V_{17}, through a divider R_{48}, by means of which the local oscillator can be tuned.

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Mir Publishers would be grateful for your comments on the content, translation and design of this book. We would also be pleased to receive any other suggestions you may wish to make.

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RADIO RECEIVERS