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Colin Hinson
In the village of Blunham, Bedfordshire.

FOR OFFICIAL USE

A P 3302
(1st Reprint Nov. 1963)
(2nd Reprint Aug. 1965)

STANDARD TECHNICAL TRAINING NOTES

FOR THE

RADIO ENGINEERING TRADE GROUP

(FITTERS)

PART 1A

ELECTRICAL AND RADIO FUNDAMENTALS

(BOOK 2 — SECTIONS 8 to 12)

MINISTRY OF DEFENCE
August, 1965

DMD 302818 3350 10/65

STANDARD TECHNICAL TRAINING NOTES
FOR THE
RADIO ENGINEERING TRADE GROUP (FITTERS)

FOREWORD

1. These Notes are issued to assist airmen and apprentices under training as Fitters in the Radio Engineering Trade Group. Fitters in this trade group "require a thorough knowledge of the electrical and radio principles, and the elementary mathematics, appropriate to the theory of the specified equipment" (see A.P. 3282A, Vol. 2). It is with the intention of helping to attain this standard that these Notes are written. They are not intended to form a complete text-book, but are to be used as required in conjunction with lessons and demonstrations given at the radio schools. They may also be used to assist airmen on continuation training at other R.A.F. stations.

2. The Notes, which are based on the syllabuses of training for aircraft apprentices, are sub-divided as follows:—

Part 1A: Electrical and Radio Fundamentals.

This deals with the principles of electricity, electronics and radio at a level suitable for the upper technician ranks and for technician apprentices.

Because of its bulk Part 1A has been split into three separate books: Book 1 covers basic electricity; Book 2, basic electronics; and Book 3, basic radio.

Part 1B: Basic Electricity and Radio.

This deals with the principles of electricity, electronics and radio at a level suitable for the lower technician ranks and for craft apprentices.

Part 2: Communications.

This deals with the applications of the principles covered in Parts 1A and 1B to communication systems and is intended to be used as required by all fitters in the Radio Engineering Trade Group.

Part 3: Radar.

This deals with the applications of the principles covered in Parts 1A and 1B to radar and is intended to be used as required by all fitters in the Radio Engineering Trade Group.

3. In general, fitters employed on communications equipment will be interested mainly in Part 1A or 1B and Part 2 of these Notes. Similarly, radar fitters will be concerned mainly with Part 1A or 1B and Part 3. However it is difficult to draw a firm dividing line between the knowledge required by fitters engaged in communications and that required by radar fitters. There is considerable overlapping; much of what was once regarded as being exclusively in the province of the radar fitter is now a requirement for the communications fitter also, and *vice versa.* Therefore those under training in the radar trades may find much that is useful in Part 2, whilst those under training in the communications field may find much of interest in Part 3.

4. The Notes deal with the basic theory and the applied principles of electricity, electronics and radio in a general way. They do not cover specific details of equipment in use in the Service. Such details are to be found in the official Air Publication for the equipment and this should always be consulted during the servicing of the equipment.

5. No alteration to these Notes may be made without the authority of official Amendment Lists.

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(BOOK 2—SECTIONS 8 to 12)

First Promulgated by Command of The Air Council
By Command of The Defence Council

Henry Hardman

MINISTRY OF DEFENCE

August, 1965

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LIST OF AIR PUBLICATIONS ASSOCIATED WITH THE TRADE

Principles and Techniques

A.P. 1093	R.A.F. Signal Manual, Part 2 (Radio Communication)
A.P. 1093E	Interservices Radar Manual—Radar Techniques
A.P. 1093F	Radar Circuit Principles, with Aerials and Centimetric Techniques
A.P. 1093G	Radio Circuitry Supplement
A.P. 1093H	Suppressed Aerials
A.P. 1186V	C.V. Register of Electronic Valves
A.P. 2521A	V.H.F. Ground Station Aerial Systems
A.P. 2867	Interservices Standard Graphical Symbols
A.P. 2867A	Interservice Glossary of Terms used in Telecommunications
A.P. 2867B	Interservice Glossary of Terms used in Telecommunications (Radar)
A.P. 2878C	H.F. and M.F. Aerials for Ground Stations
A.P. 2900C	Handbook of Electronic Test Methods and Practices
A.P. 3158C	R.A.F. Technical Services Manual
A.P. 3214 (Series)	The Services Text book of Radio.

Equipment

Air Publications applicable to specific radio equipment are listed in:—

A.P. 2463	Index to Radio Publications

INSTRUCTIONAL FILMS

Title	Reference
Current of Electricity	14L/52
Nuts and Bolts	14L/178
Micrometer Calipers	14L/273
Vernier Scale	14L/413
Hammers, Chisels, Punches and Drifts	14L/1605
Files and Filing	14L/1606
Spanners, Screwdrivers and Pliers	14L/1636
Taps, Dies and Reamers	14L/1727
Hacksaws, Shears and Vice Clamps	14L/1728
Locking Devices	14L/1729
Measuring and Marking—Precision Instruments	14L/1730
Transmission Lines—Maintenance of Coaxial Cables	14L/3260
Transmission Lines and Waveguides	14L/3288

Title	Reference
Vacuum Tubes—Electronic Diode	14L/3953
Cathode Ray Tube	14L/4268
Electricity and Magnetism	14L/4708
Magnetism	14L/5557
Electrical Terms	14L/5607
What is Electricity?	14L/5609
Electricity and Heat	14L/5610
Electricity and Movement	14L/5611
Electrochemistry	14L/5612
Putting Free Electrons to Work	14L/5614
A.C. and D.C.	14L/5615
The Generation of Electricity	14L/5616
The Transmission of Electricity	14L/5617
Aircraft First Line Servicing	14L/5656
Audio Oscillator	14L/5666
Volts—Ohm Meter Operation	14L/5667
Radio Shop Technician	14L/5668
First Line Servicing, Fighter Aircraft	14L/5768
Radio Antennae Fundamentals, Parts 1 and 2	14L/5780–1
R.D.F. to Radar	14L/5826
Waveguides, Parts 1 to 5	14L/5958–5962
Tuned Circuits	14L/6037
Ground Handling of Aircraft	14L/6338
The Doppler Principle in Airborne Navigation Aids	14L/6388
Centimetric Oscillators, Parts 1 to 3	14L/6397
Servomechanisms	14L/6435
Radar Techniques, Part 1—Waveform Response of C.R. Circuits	14L/6500
Radar Techniques, Part 2—Multivibrator	14L/6502
Radar Techniques, Part 3—Miller Timebase	14L/6504
Radar Techniques, Part 4—Pulse Forming by Delay Lines	14L/6506
Radar Techniques, Part 5—Flip Flop	14L/6508
Problems of Radio and Electronic Fault Finding	14L/6594
Principles of the Transistor	14L/6620

INSTRUCTIONAL FILM STRIPS

Title	Reference
Primary Cells	14J /154
Time Constant	14J/155
Distribution of Electricity	14J/194
Electricity—its Production	14J/195
Uses of Electricity	14J/196
Radiation	14J/197
Thermionic Valve	14J/198
Electrical Measuring Instruments	14J/203
The D.C. Motor	14J/204
Basic Radio Trouble-shooting, Parts 1 to 5	14J/239–243
The Internal Combustion Engine	14J/369
Elementary Principles of Cathode Ray Oscillograph	14J/370
The Cathode Ray Tube	14J/404
Magnetism and Electricity	14J/407
Waveguide Theory	14J/495–511
Waveguide Theory	14J/512–517
Introduction to Control Engineering Theory	14J/578
Introduction to Electronics	14J/586
Electronic Devices—Electron Tubes	14J/587
Basic Valve Circuits, Parts 1 to 4	14J/588–9
The Meaning of Valve Characteristics	14J/590
Telecommunication Principles	14J/606

LIST OF SYMBOLS AND ABBREVIATIONS

TABLE 1

Greek Letters Used in the Text

Letter			Letter		
Small	**Capital**	**Name**	**Small**	**Capital**	**Name**
α	—	Alpha	λ	—	Lambda
β	—	Beta	μ	—	Mu
γ	—	Gamma	π	—	Pi
δ	Δ	Delta	ρ	—	Rho
ε	—	Epsilon	σ	—	Sigma
η	—	Eta	φ	Φ	Phi
θ	—	Theta	ω	Ω	Omega
ϰ	—	Kappa			

TABLE 2

Meaning of Symbols Used in the Text

Letter	Meaning	Letter	Meaning
A	Ampere, Amplification	j	Vector operator=$\sqrt{-1}$
B	Magnetic flux density, Susceptance, Bandwidth	k	Coupling coefficient, Kilo—(prefix), constant
C	Capacitance	L	Length
D	Electric flux density, Distance	m	Modulation factor, Metre, Mass, Milli—(prefix)
E	Electromotive force, Electric field strength	n	Number
F	Farad, Factor, Force	p	Pico—(prefix)
G	Conductance, Giga—(prefix)	q	Instantaneous charge
H	Magnetic field strength, Henry	r	Length (in polar co-ordinates)
I	Electric Current	r_a	Anode slope resistance

(*continued overleaf*)

J	Joule
L	Inductance
M	Mutual inductance, Mega—(prefix)
N	Number, Noise factor, Revs. per minute
P	Power
Q	Quantity or charge of electricity, Coil amplification factor
R	Resistance
S	Magnetic reluctance
T	Temperature (Abolute), Period, Transit Time, Transformation ratio
V	Potential difference, Volt, Volume
W	Energy or work, Watt
X	Reactance
Y	Admittance
Z	Impedance
a	Area
c	Velocity of light, Cycle
d	Distance
e	Instantaneous e.m.f., Electron charge
f	Frequency
g_0	Valve conversion conductance
g_m	Valve mutual conductance
i	Instantaneous current

(*continued overleaf*)

s	Second
t	Time, Temperature
u	Velocity
v	Instantaneous potential difference
x	Distance, Length
y	Length
α	Angle, Number
β	Number, Feedback factor
γ	Propagation constant
δ	Small increment, Loss angle
ε	Base of natural logs = 2·71828
η	Efficiency
θ	Angle
$\varkappa_r$	Dielectric constant
$\varkappa_o$	Permittivity of free space
λ	Wavelength
μ	Permeability, Valve amplification factor
μ_o	Permeability of free space
μ_r	Relative permeability
π	Ratio of circumference to diameter of a circle = 3·14159
ρ	Specific resistance
ϕ	Angle
Φ	Magnetic flux
ω	Angular velocity = $2\pi f$
Ω	Ohm
σ	Specific conductance

TABLE 3

Prefixes for Multiples and Sub-multiples

Multiple or sub-multiple	Name	Prefix	Multiple or sub-multiple	Name	Prefix
$1,000,000,000 = 10^9$	Giga-	G	$\frac{1}{1,000,000} = \frac{1}{10^6} = 10^{-6}$	Micro-	μ
$1,000,000 = 10^6$	Mega-	M			
$1,000 = 10^3$	Kilo-	k	$\frac{1}{10^{12}} = 10^{-12}$	Micro-micro- or Pico	$\mu\mu$ or p
$\frac{1}{1,000} = \frac{1}{10^3} = 10^{-3}$	Milli-	m			

TABLE 4

Abbreviations of Units

Unit	Abbreviation	Unit	Abbreviation
Ampere	A	Gramme	g
Ampere-hour	Ah	Henry	H
Ampere-turn	AT	Joule	J
Cycles per second	c/s	Metre	m
Decibel	db	Ohm	Ω
Degree	° Centigrade = C. Fahrenheit = F.	Second	s or sec.
		Volt	V
Electron-volt	eV	Watt	W
Farad	F	Weber	Wb

CONTENTS

SECTIONS

(A detailed contents list is given at the beginning of each Section and Chapter)

SECTION 8

FUNDAMENTAL ELECTRONIC DEVICES

SECTION 8

CHAPTER 1

THERMIONIC EMISSION AND THE DIODE VALVE

Paragraph

THERMIONIC EMISSION

DIODE VALVE

THERMIONIC EMISSION AND THE DIODE VALVE

THERMIONIC EMISSION

Introduction

1. An *electronic device* is defined as "a device which makes use of electrons free in vacuous or gaseous space, or in boundary layers between dissimilar conductors or semi-conductors". This definition includes valves, cathode-ray tubes, photocells, crystal detectors, metal rectifiers and transistors, all of which are dealt with in this Section. The definition excludes electrical devices dependent solely on ordinary metallic conduction such as meters, transformers, relays, etc., and all devices dependent on electrochemistry.

2. In this Chapter the principles of thermionic emission and the operation of the diode valve are considered. The diode is an electronic device which consists of an envelope, containing gas at a low (possibly negligible) pressure, provided with two *electrodes* between which conduction of electricity through the vacuum or the contained gas may take place as a result of thermionic emission from one of the electrodes (the *cathode*). The question of how thermionic emission occurs must first be considered.

Electron Emission

3 In Bk.1.Sect.1, Chap 1, it was stated that a conductor is a material in which a continual random movement of free electrons from atom to atom within the material is taking place. An electrostatic field super-imposed on the conductor by means of a potential difference between its ends causes the free electrons to 'drift' from the low potential to the high potential part of the electrostatic field.

4. The kinetic energy acquired by an electron in its passage through a conductor is usually expressed in *electron-volts*. One electron-volt is the energy acquired by an electron in falling through a potential difference of one volt.

5. Electrons cannot easily leave a conductor at normal room temperatures since their energy is usually insufficient to allow them to do so, and *additional* energy must be given to the electrons to enable them to overcome the forces acting at the surface of the conductor. The quantity of work which must be done to free an electron from a metal in order to produce *electron emission*, is usually given in electron-volts and is called the *work function* of the metal. The lower is the work function, the less the amount of additional energy which must be given to the electrons to cause electron emission. The work functions of some of the more common materials used in cathode construction are listed in Table 1.

Material	Work Function (*electron-volts*)
Barium oxide	1·12
Caesium	1·75
Potassium	1·85
Lithium	2·2
Thorium	3·4
Copper	4·2
Molybdenum	4·25
Tungsten	4·5
Silver	4·65

Table 1. WORK FUNCTIONS

6. Before an electron can leave the surface of a metal it must possess energy in excess of the work function of the metal. There are four principal ways in which the normal electron energy may be increased in order to produce electron emission:—

(*a*) **Thermionic emission.** By raising the temperature of the metal as in valves and cathode-ray tubes.

(*b*) **Secondary emission.** By bombarding the surface of the metal with electrons from another source; some of the energy of the bombarding electrons is transferred on collision to the electrons in the metal. This is utilized in certain specialised valves.

(*c*) **Photo-electric emission.** By the incidence of light on the surface of the material. This type of emission is used for photocells.

(*d*) **Field emission.** By the action of an external electric field in overcoming the surface barrier. Field emission has few deliberate applications although it does occur in some electronic devices.

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Thermionic Emission

7. Raising the temperature of a metal causes an increase in the movement of the atoms in the metal, and since the vibrating atoms are continually in collision with the free electrons, energy is transferred from the atoms to the electrons. If the metal is heated to a sufficiently high temperature the electrons may acquire sufficient additional energy to enable them to break through the surface barrier and escape from the metal, producing thermionic emission.

8. A metal with a low work function will release electrons at comparatively low temperatures and, if it also has a high melting point it will generally be suitable for use as a thermionic emitter. The two desirable properties of low work function and high melting point must both be taken into account. Thus tungsten, for which the work function is 4·5 and which has a melting point of 3,380°C, has a melting point emission of 450 A /sq. cm. Thorium, with a more desirable work function of 3·4, has a melting point of 1,830°C and an emission at that temperature of 2·9 A /sq. cm. So the lower work function, of thorium is more than off-set by its lower melting point.

Emission Current

9. If a length of wire is heated in a vacuum to a temperature sufficient to cause thermionic emission, and if a positively-charged plate is situated near the wire, the emitted electrons are attracted to the plate, giving rise to a thermionic current. This current increases very rapidly with the temperature of the wire as shown in Fig. 1 for three different types of emitter. The wire which is heated to produce thermionic emission is termed the *cathode*, and the positively-charged plate which collects the emitted electrons is called the *anode*.

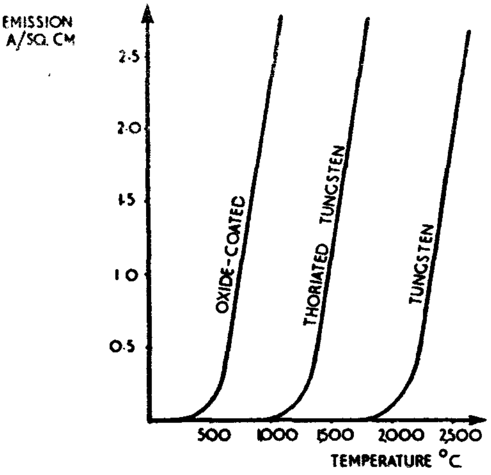


Fig. 1.—VARIATION OF EMISSION CURRENT WITH TEMPERATURE.

Practical Thermionic Emitters

10. The number of substances available for use as practical thermionic emitters is very few, because of the dual requirements of low work function and high melting point. At present, the only ones used commercially are:—

(*a*) Tungsten emitters.
(*b*) Thoriated-tungsten emitters.
(*c*) Oxide-coated emitters.

11. **Tungsten emitters.** Tungsten is superior to other emitters in its ability to operate under adverse conditions, particularly conditions associated with very high voltages on the anode. The emitter does not easily deteriorate even when supplying a large emission current for a long period of time. However, the operating temperature of a pure tungsten emitter is high (of the order of 2,500°C) and, since emitters are normally heated by passing a current through them the power required to raise the emitter to that temperature is large.

12. **Thoriated-tungsten emitters.** A filament of tungsten which has been impregnated with a small quantity of thorium oxide, is called a thoriated-tungsten emitter. Such emitters when properly activated, give thermionic emission at temperatures of the order of 1,500°C. Thus, the heating power necessary to produce emission in this type of emitter is less than that for a pure tungsten emitter. However, the thorium layer tends to be 'stripped off' under certain conditions when the valve is in operation, and for this reason the thoriated-tungsten emitter is restricted to medium-power transmitting valves.

13. **Oxide-coated emitters.** The oxide-coated emitter consists of a mixture of barium and strontium oxides coated on the surface of a suitable metal, commonly nickel. When properly prepared and activated, such a surface will emit electrons at temperatures of the order of 750°C. Thus, the heating power necessary to produce emission is low. Oxide-

coated emitters have the property of being able to give very high instantaneous electron emission for brief periods. This effect is utilized to advantage in valves designed to generate short pulses of high power, such as radar transmitting valves. The period of excessive emission cannot be prolonged without the danger of stripping the emitter. The oxide-coated emitter is employed in almost all small valves, and it is also replacing thoriated-tungsten in many of the larger valves.

14. **Emitter efficiency.** The efficiency of an emitter (or cathode) is generally specified in terms of its emission per unit area for a given heating power input. Typical figures are:—

(*a*) Pure tungsten wire	5mA/sq. cm./watt
(*b*) Thoriated-tungsten wire	50mA/sq. cm./watt
(*c*) Oxide-coated wire	500mA/sq. cm./watt

Cathode Construction

15. The construction of practical cathodes used to produce thermionic emission in valves may take either of two forms:—

(*a*) Directly-heated cathodes.

(*b*) Indirectly-heated cathodes.

16. **Directly-heated cathode.** This consists of a filament of wire which is heated to the desired operating temperature by a current passing through the filament itself. The material of the filament can be pure tungsten, thoriated-tungsten, or oxide-coated nickel depending on the requirement. Typical directly-heated cathode structures are shown in Fig. 2. Directly-heated cathodes are the most economical form so far as concerns the power necessary to heat the cathode. They are, therefore, used in most applications for operation from batteries, and for special applications in which very quick heating is required. Directly-heated cathodes of tungsten or thoriated-tungsten are also used in preference to indirectly-heated cathodes where high powers are involved.

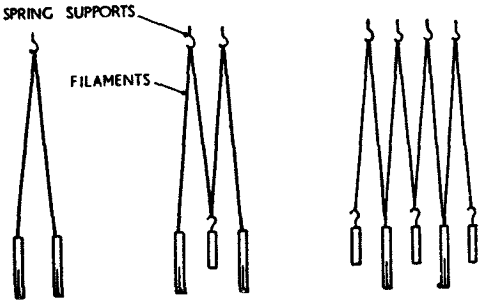


Fig. 2.—DIRECTLY-HEATED CATHODES.

17. **Indirectly-heated cathode.** This consists of an oxide-coated nickel tube surrounding a heater wire. The oxide-coated surface is heated to the desired operating temperature *indirectly*, by a current passing through the *heater*. The heater wire fits into the hollow cathode and is insulated from it by a coating of alumina, as shown in Fig. 3. Although this form of cathode may take up to a minute to reach its final working temperature it also has certain advantages over the directly-heated cathode:—

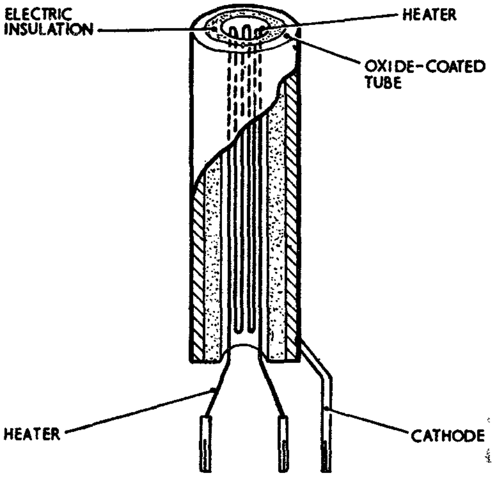


Fig. 3.—INDIRECTLY-HEATED CATHODE.

(*a*) The cathode can be given the best shape for its purpose.

(*b*) The thermal reservoir effect between heater and cathode ensures that the cathode remains at a constant temperature even when a.c. is used for the heater.

(*c*) Since no heating current is passed through the cathode itself, the whole of the latter is at the same potential. With directly-heated cathodes a p.d. is developed across the filament, and if a.c. is used as the heater current, the variations in p.d. produce fluctuations in the current through the valve.

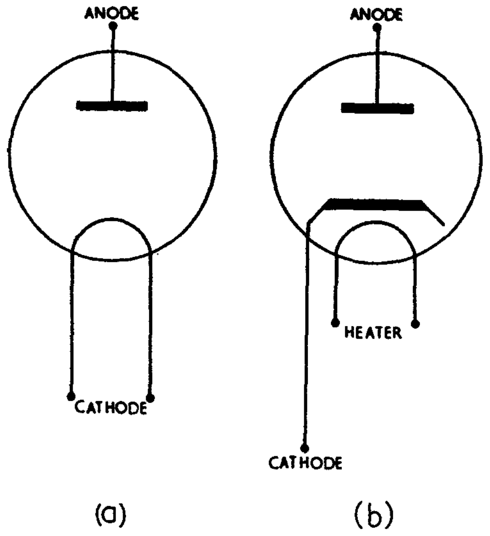


Fig. 4.—SYMBOLS FOR DIODE VALVES.

DIODE VALVE

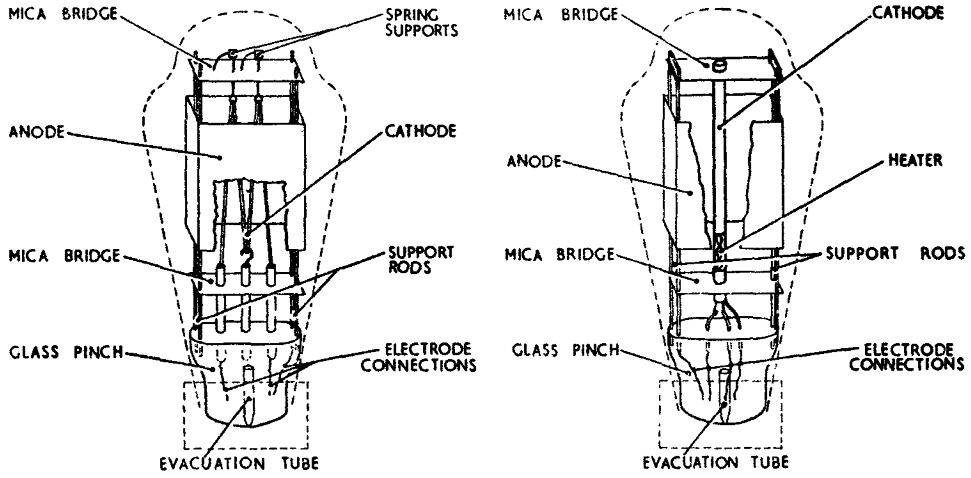
General Construction of Diode Valve

18. The motion of the free electrons that surround the cathode as the result of thermionic emission may be influenced by electrostatic fields applied by means of further electrodes. Valves are classified according to the *number* of electrodes they possess. A diode is therefore, a valve with two electrodes—a cathode and an anode. The conventional symbol for a diode valve is shown in Fig. 4, the directly-heated diode being given in (*a*), and the indirectly-heated diode in (b). It should be noted that in an indirectly-heated valve the heater is *not* regarded as one of the 'active' electrodes.

19. Fig. 5(a) shows the construction of a directly-heated diode, and Fig. 5(b) that of an indirectly-heated diode. In both instances the anode completely surrounds the cathode so that all electronic action is confined to the space between the electrodes. The cathode material will depend on the requirement, and the anode is made from nickel or molybdenum, metals which are easily worked and which have high melting points; they also dissipate heat readily.

20. During assembly, the support rods are sealed into a glass "pinch", and the electrode connections are taken through the pinch to the valve base, from which connection is made to the circuit. When the parts are assembled on the pinch, the remainder of the envelope is sealed to it. In order to create a vacuum inside the valve, connection is made from a pumping installation to a small evacuation tube in the pinch, and while the pumps are still operating to remove all gas from the valve:—

(*a*) The valve is 'baked' in an oven to



(a) DIRECTLY HEATED DIODE

(b) INDIRECTLY HEATED DIODE

Fig. 5.—CONSTRUCTION OF DIODE VALVES.

remove any absorbed water-vapour from the glass surface.

(*b*) The electrodes are heated by induction to remove the last traces of gas from the surface.

(*c*) A small pellet of magnesium (known as a '*getter*') is located inside the valve and fired by induction; the magnesium evaporates and deposits itself as a thin film on the wall of the tube, where it combines with any residual gas which appears during the operation.

(*d*) The evacuation tube in the valve pinch is sealed off.

The completed valve is then ready to be connected to its valve base and subsequently to be tested.

Current in a Diode

21. It was shown in Para. 9 that the rate of emission of electrons into the free space surrounding a cathode depends on the temperature of the cathode and its material. Thus, for a given cathode material at a given temperature, the emission is *constant*. Consider a directly-heated diode connected in the simple circuit of Fig. 6. A low tension (l.t.) supply is applied to the cathode and the resultant heater current raises the temperature of the cathode to the steady operating value, so that a constant electron emission occurs. A high tension (h.t.) supply is connected to apply a p.d. between anode and cathode, the anode being *positive* with respect to the cathode. This p.d. can be varied by means of a potentiometer, and the resultant '*anode voltage*' is indicated in a voltmeter. Finally, a milliammeter is connected in the anode circuit to read the '*anode current*'.

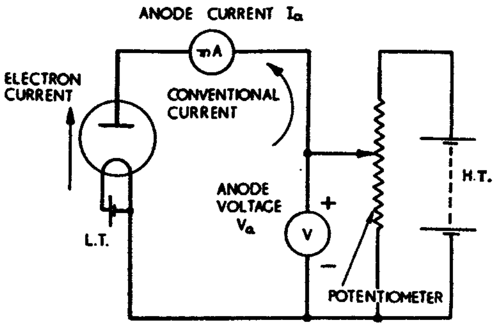


Fig. 6.—DIODE BASIC CIRCUIT.

22. **Space charge.** If the potentiometer is adjusted so that V=O, there is no electric force between the anode and the cathode to drive the emitted electrons to the anode. The velocity of emission of electrons is such that a few electrons may arrive at the anode, but these are so few that they may be neglected at this stage. For practical purposes then, the reading in the milliammeter will be zero, indicating zero anode current. The majority of the emitted electrons will distribute themselves throughout the space between the two electrodes, and a 'cloud' of negative electricity is formed in the vicinity of the cathode. This cloud of electrons is termed the *space charge*, and it tends to repel back to the cathode all electrons emitted with small energies. For zero anode voltage the space charge is *constant*, although its component electrons are continually changing. The cathode is left slightly positive by the emission of electrons so that it tends to attract the electrons back again. Thus, electrons are continually being emitted by, and attracted back towards the cathode and an equilibrium is set up which keeps the density of the space charge constant. If the anode voltage is now made slightly positive, some of the electrons in the space charge are attracted to the anode and an anode current is established, the loss of electrons to the anode causing the space charge to become less dense. The electron flow is from the point of low potential (the cathode) to the point of high potential (the anode), thence round the external circuit back to the cathode. It is more convenient in electronic circuits to consider electron flow rather than conventional current, which is in the *reverse* direction.

23. The *effect* of the space charge can be seen by considering the potential gradient of the electric field in the space between the two electrodes when the anode is positive with respect to the cathode (Fig. 7). Because of the existence of the space charge, the potential in this region will be lower than that of both anode and cathode, the cloud of electrons forming a '*virtual cathode*'. Only electrons with sufficient energy to surmount this negative potential barrier will reach the anode, and the anode current under such conditions is said to be *space-charge-limited.*

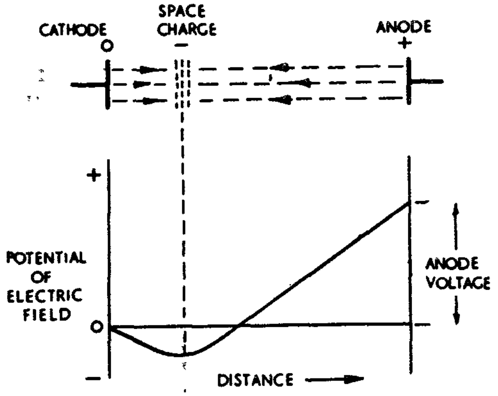


Fig. 7.—POTENTIAL GRADIENT IN A DIODE, SPACE-CHARGE-LIMITED.

24. Saturation current. As the anode potential is increased, electrons are drawn from the space charge at an increasing rate to give an increase in the anode current. The anode current is still space-charge-limited and this condition continues until electrons are lost by the space charge to the anode at a rate faster than the cathode emission can replace them. As the anode potential is increased, a point is eventually reached when *all the electrons* emitted by the cathode are carried directly across to the anode, and the space charge disappears. The graph of the potential gradient inside the diode under these conditions is shown in Fig. 8, from which it is seen that the negative potential barrier has disappeared.

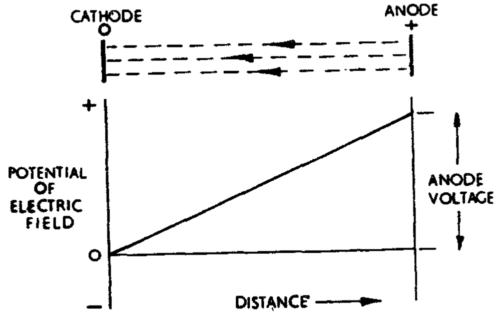


Fig. 8.—POTENTIAL GRADIENT IN A SATURATED DIODE.

25. When the anode potential is such that all the electrons emitted by the cathode are carried directly across to the anode, the anode current reaches its maximum value for a given cathode temperature. Thereafter it is impossible to increase the anode current by increasing the anode voltage, since the anode is already collecting all the emitted electrons. The limiting value of the anode current is called the ***saturation current***, and the valve in this condition is said to be ***saturated*** or *temperature-limited.* Fig. 9 shows a graph of anode current (I_a) plotted against positive values of anode voltage (V_a). For values of V_a less than the saturation value V_s, the current is space-charge-limited. When the anode potential is greater than V_s, the space charge has been dispersed and the anode current tends to its saturation value I_s.

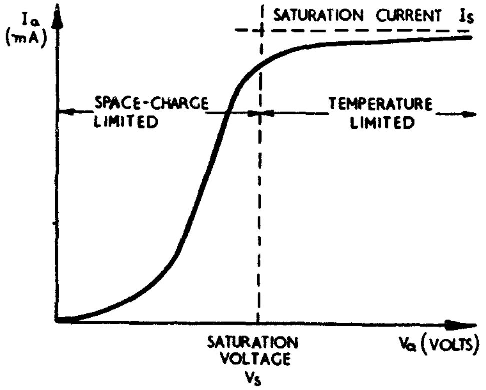


Fig. 9.—SPACE-CHARGE-LIMITED AND TEMPERATURE-LIMITED CONDITIONS.

26. Fig. 10 shows how the value of the saturation current is increased by raising the *temperature* of the cathode (by increasing the value of the l.t. supply). The significance of the term 'temperature-limited' will now

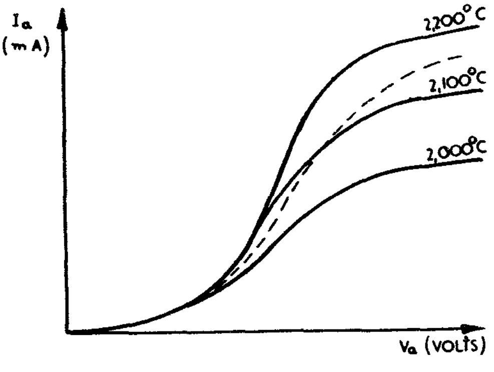


Fig. 10.—EFFECT OF INCREASING CATHODE TEMPERATURE.

be seen. The solid curves are plotted for a tungsten cathode, although thoriated-tungsten cathodes would give curves of the same general shape. In oxide-coated cathodes, saturation takes place much more gradually as shown by the dotted curve.

Anode Characteristic of Diode

27. Fig. 11 shows how the anode current I_a varies with changes in the anode voltage V_a, for negative as well as positive values of V_a. Such a curve is termed the *anode characteristic* of the diode. The lower part of the curve of Fig. 11, corresponding to space-charge-limitation of the anode current, can be represented by a "three-halves power law"; that is:—

$$I_a = kV_a^{3/2} \quad .. \quad .. \quad .. \quad (1)$$

where k is a constant depending on the construction of the valve.

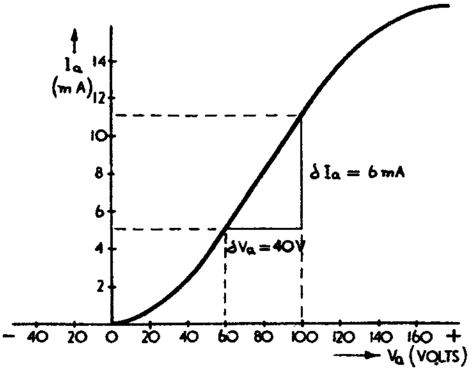


Fig. 11.—ANODE CHARACTERISTIC OF A DIODE.

28. Two important facts can be stated by examining the curve of Fig. 11:—

(*a*) A valve is a *non-linear* device. It does not obey Ohm's law since the current is not directly proportional to the voltage (see equation (1)).

(*b*) A valve is a *uni-directional* device. Electrons can pass only from the cathode to the anode, and that only when the anode is positive with respect to the cathode. It is this property of the diode that suggested the name 'valve'.

Valve Resistance

29. **D.C. Resistance.** This is the ratio of the steady voltage across the diode to the resultant steady current which is established in it. It is obviously not a constant since the valve does not obey Ohm's law.

30. **A.C. Resistance.** Of more importance than the d.c. resistance of a diode is the anode a.c. resistance or *slope resistance* (r_a). The slope resistance is defined as:—

$$r_a = \frac{\delta V_a}{\delta I_a} \text{ (ohms)} \quad .. \quad (2)$$

where δ is used to mean "a small change" in the quantity to which it is attached.

Example. Fig. 11 shows that by increasing the anode voltage from 60V to 100V (i.e. a change of 40V) the anode current increases from 5mA to 11mA (i.e. a change of 6mA). Hence:—

$$r_a = \frac{40\text{V}}{6\text{mA}}$$

$$\therefore\ r_a = 6{\cdot}66\ \text{k}\Omega.$$

This figure is true only if taken over the 'straight' part of the characteristic, but since in most modern valves the anode characteristic is almost a straight line over a large part of its useful range, it is permissible to assume that r_a is a constant over this range.

Anode Dissipation

31. When a valve is passing current, power is dissipated in the form of heat by the anode as result of bombardment by electrons from the cathode. The power so developed may be large. Each valve is given a *rated* anode dissipation by the manufacturer, and if the average values of anode current and anode voltage are such that their product exceeds this limit, damage to the valve may result. The construction of the valve depends on the power that the anode is called upon to dissipate:—

(*a*) *Low powers.* When the heat produced at the anode does not exceed 100 watts, it is possible to dissipate it inside the valve envelope by radiation from the anode to the tube walls.

(*b*) *Medium powers.* When the power to be dissipated at the anode is of the order of 100 to 750 watts, the valve is constructed in such a way that the anode extends outside the valve envelope and is fitted with cooling fins. The anode is then cooled by normal air circulation or by air blast from a 'blower' motor.

(*c*) *High powers.* For powers in excess of 750 watts, the anode again constitutes part of the valve envelope and is fitted with a jacket through which distilled water is pumped to give the necessary cooling.

32. In order to show the safe margin for operation of a valve it is often convenient to super-impose the rated dissipation curve for the valve on its anode characteristic. For example, the dissipation curve for a valve which has a rated anode dissipation of 10 watts, is shown in Fig. 12. Operation of the valve must be confined to the area underneath this curve if damage to the valve is to be avoided.

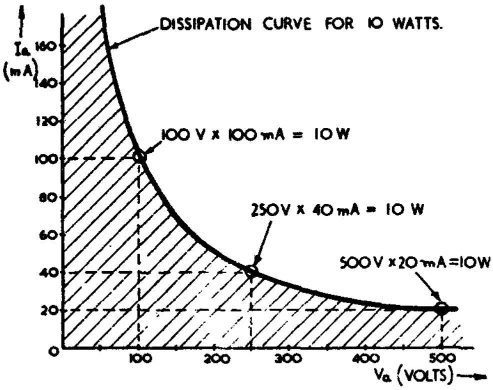


Fig. 12.—ANODE DISSIPATION CURVE.

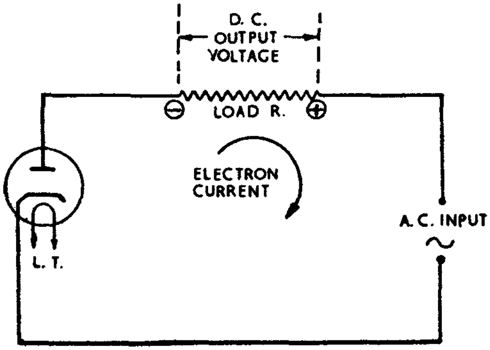


Fig. 13.—SIMPLE DIODE RECTIFIER CIRCUIT.

Applications of Diode

33. All the uses to which a diode valve may be put can be explained in terms of the non-linear and uni-directional properties of the valve. The uni-directional properties may be shown by considering Fig. 13. An alternating voltage is applied between the anode and the heated cathode of the diode valve in series with a load resistor R. Anode current is established only when the anode is positive with respect to the cathode. Thus, current flows during one half-cycle of an a.c. cycle as shown by the graph of Fig. 14. The voltage developed across the load resistor R by this pulsating current is, therefore, a *uni-directional* voltage. Hence, the diode valve has been used as a '*rectifier*' to convert an a.c. input to a d.c. output. This, and other applications of the diode valve, will be considered in detail in later Sections.

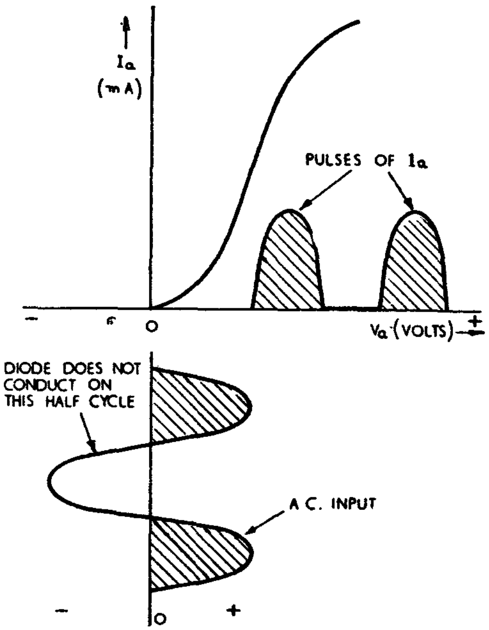


Fig. 14.—OPERATION OF DIODE RECTIFIER.

SECTION 8

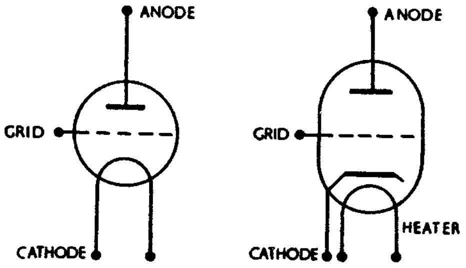
CHAPTER 2

THE TRIODE VALVE

THE TRIODE VALVE

Introduction

1. The triode valve contains *three* electrodes. As in the diode there is an electron emitter, the cathode, and a collector electrode, the anode. In addition, *between* these two electrodes there is a wire mesh or *grid* which *controls* the flow of electrons from the cathode to the anode. The triode is shown diagrammatically in Fig. 1. The introduction of the third electrode, generally referred to as the *control grid*, transforms the diode into a device which can control power over a wide range of frequencies.



(a) DIRECTLY-HEATED (b) INDIRECTLY-HEATED

Fig. 1.—SYMBOLS FOR TRIODE VALVES.

General Construction of Triode

2. The construction of a typical indirectly-heated triode is illustrated in Fig. 2. The general construction is the same as that for the diode considered in Chap. 1. The materials used for the cathode and the anode, and the method of assembly are also similar. The grid, in the form of an open mesh of molybdenum wire, is interposed between the anode and the cathode as shown in Fig. 2.

Effect of Grid

3. The number of electrons that reach the anode in a triode valve under space-charge-limited conditions is determined mainly by the electrostatic field in the cathode-grid space and this is determined by the potential of the grid relative to the cathode. Assuming that the anode potential has a steady positive value and the grid potential is variable, the following conditions apply:—

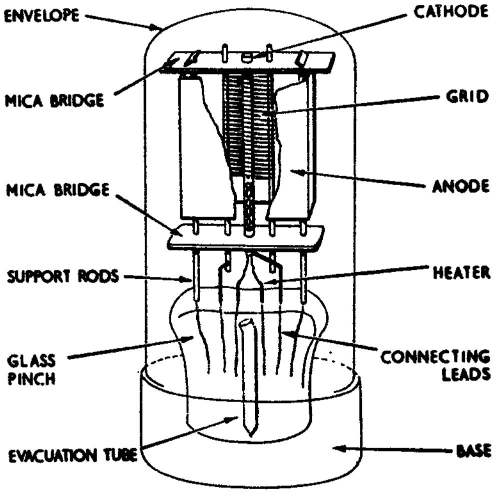


Fig. 2.—CONSTRUCTION OF A TYPICAL TRIODE.

(*a*) When the potential on the grid is sufficiently *negative* with respect to the cathode the electrostatic field is such that the emitted electrons are repelled back to the cathode and the anode current is said to be '*cut off*'.

(*b*) By progressively reducing the negative potential on the grid a point is reached where the anode current 'cuts on', the electrostatic field being such that a few of the emitted electrons are attracted to the anode. Further reduction in the negative value of grid potential causes an increase in the number of electrons reaching the anode.

(*c*) When the potential on the grid is *positive* with respect to the cathode, electrons are attracted to the grid as well as to the anode. Thus, as well as an increase in anode current I_a being produced by a positive potential on the grid, a grid current I_g is established. This latter current is generally small. The total valve current, known as the *space current* I_k is given by:—

$$I_k = I_a + I_g \quad \ldots \quad \ldots \quad (1)$$

Current in a Triode

4. Consider a triode valve connected in the simple circuit of Fig. 3. A l.t. supply is applied to the heater to give a constant

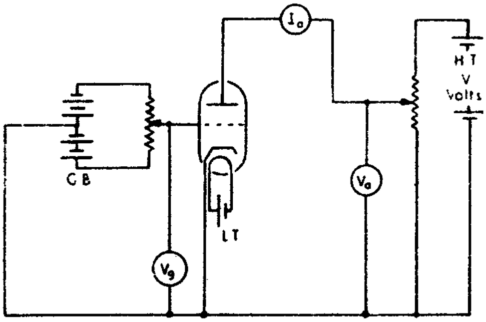


Fig. 3.—SIMPLE TRIODE CIRCUIT.

electron emission from the cathode. A h.t. supply is applied to the anode via a potentiometer by means of which the anode voltage V_a can be varied between zero and +V volts. A grid bias (g.b.) supply is connected to the grid via a potentiometer. This supply is centre-tapped to the cathode so that the grid voltage V_g may be made positive or negative with respect to the cathode. Finally, a milliammeter is connected in the anode circuit to read the anode current I_a.

5. In such a circuit there are three variables, assuming constant electron emission, V_a, V_g and I_a. The circuit may be used to show how I_a varies with changes in V_g for several *fixed* values of V_a. Fig. 4 shows a typical example of such curves plotted from the readings given in Table 1.

V_g (volts) ..	+2	0	−2	−4	−6	−8	−10
I_a (mA) .. (V_a = 100V)	11	6·5	3	0·5	—	—	—
I_a (mA) .. (V_a = 140V)	15	11	7	3·5	1	—	—
I_a (mA) .. (V_a = 180V)	20	16	12	8	4	1·5	—

Table 1. READINGS FOR FIG. 4.

6. Inspection of Fig. 4 will show that:—

(*a*) The anode potential V_a influences the current through the valve.

(*b*) The grid potential V_g influences the current to a *greater* extent. This is because the grid is much closer to the cathode than is the anode.

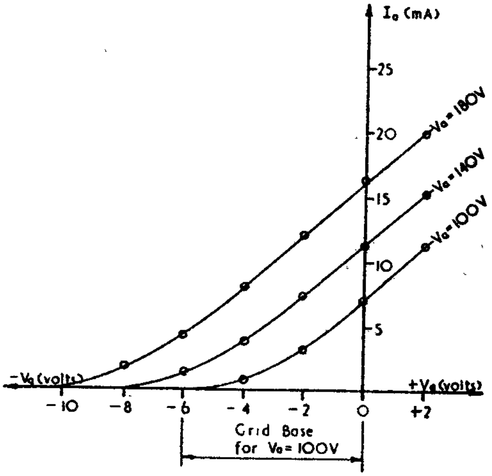


Fig. 4.—TYPICAL Ia—Vg CURVES FOR A SMALL TRIODE.

(*c*) Current can flow through the valve when the grid is negative.

(*d*) To every anode potential there corresponds a grid potential for which the current is just cut off. This value of grid potential gives the '*cut-off-bias*' in the valve and determines the '*grid base*' of the valve as shown in Fig. 4.

7. In the diode valve the relation between the anode current and the anode voltage is of the form:—

$$I_a = kV_a^{3/2}$$

The current in a triode is given by a similar relationship:—

$$I_a = k\left(V_g + \frac{V_a}{\mu}\right)^{3/2} \quad .. \quad (2),$$

where k is a constant determined by the dimensions of the valve, and μ is a constant known as the *amplification factor* of the valve (see Para. 14).

This expression assumes that the grid is negative, so that the grid current is zero, and the total number of electrons drawn from the cathode (the space current) reaches the anode.

Triode Characteristic Curves

8. The characteristics of triode valves are commonly expressed by families of curves, plotted by using a circuit similar to that of Fig. 3. They show the relationship between:

(*a*) I_a and V_g, with constant values of V_a. Such curves are termed *mutual characteristics*.

(*b*) I_a and V_a, with constant values of V_g. Such curves are termed *anode characteristics.*

9. **Mutual characteristics.** These are the curves shown in Fig. 4. They have not been plotted for sufficiently high values of anode current and anode voltage to show the effect of saturation. From equation (2), the anode current is zero when:—

$$(V_g + \frac{V_a}{\mu}) = 0$$

$$\text{i.e., when } V_g = -\frac{V_a}{\mu} \quad .. \quad (3)$$

This gives the value of the cut-off bias or the grid base of a triode for a given value of V_a.

10. **Anode characteristics.** In the circuit of Fig. 3 if V_g is set to a given value and V_a increased from zero, I_a increases. By noting the values for V_a and I_a, an $I_a - V_a$ curve can be plotted. If this is repeated for other fixed values of V_g a family of anode characteristic curves is obtained, as shown in Fig. 5.

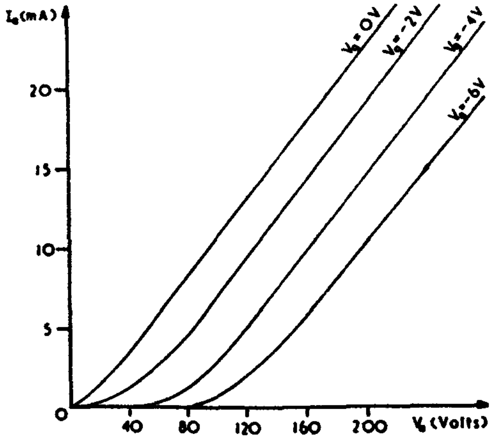


Fig. 5.—ANODE CHARACTERISTICS OF A TRIODE.

Valve Constants

11. The behaviour of a triode valve in the vicinity of a particular 'operating point' can be expressed in terms of three *constants:*—

(*a*) Slope resistance, r_a.
(*b*) Mutual conductance, g_m.
(*c*) Amplification factor, μ.

Note. These factors are constant only if taken over the 'straight' part of the characteristics.

12. **Slope resistance.** This constant has already been defined for a diode valve. For a triode, the slope resistance r_a is defined as the ratio of the change δV_a in the anode voltage to the resulting change δI_a in the anode current, V_g remaining constant. Thus:

$$r_a = \frac{\delta V_a}{\delta I_a} \ (V_g \text{ constant}) \quad .. \quad (4)$$

The slope resistance is expressed in ohms and it is equal to the *reciprocal* of the slope of the *anode* characteristics for the valve.

13. **Mutual conductance.** This is the ratio of the change in anode current δI_a to the change in grid voltage δV_g producing it, the anode voltage V_a being kept constant. Thus:—

$$g_m = \frac{\delta I_a}{\delta V_g} \ (V_a \text{ constant}) \quad .. \quad (5)$$

Mutual conductance is expressed either in micromhos, or more commonly in *milliamps per volt*, and it is equal to the *slope* of the *mutual* characteristic for the valve.

14. **Amplification factor.** The amplification factor μ, already noted in equation (2), is a measure of the relative effectiveness of the grid and anode voltages in controlling the anode current. It is defined in terms of the voltage changes δV_a and δV_g that cause identical changes of anode current. Thus, if δV_a changes I_a to some new value and δV_g returns I_a to its original value:—

$$\mu = \frac{\delta V_a}{\delta V_g} \ (I_a \text{ constant}) \quad .. \quad (6)$$

A value of 20 for μ indicates that the grid is 20 times more effective than the anode in controlling the anode current, and this is due to the closer proximity of the grid to the cathode.

15. **Relation between r_a, g_m and μ.** The three valve constants are not independent, but as may be seen from equations (4), (5) and (6) they are connected by the relation:—

$$\begin{array}{ccccc} \mu & = & r_a & \times & g_m \\ \downarrow & & \downarrow & & \downarrow \\ \dfrac{\delta V_a}{\delta V_g} & = & \dfrac{\delta V_a}{\delta I_a} & \times & \dfrac{\delta I_a}{\delta V_g} \end{array} \quad .. \quad (7)$$

Derivation of Valve Constants from Characteristics

16. The valve constants μ, g_m and r_a may be derived under given operating conditions

from a study of either a set of mutual characteristics or a set of anode characteristics for the valve.

17. Fig. 6 shows a family of *mutual* characteristic curves for a low-impedance triode. Suppose the operating conditions, are given as $V_a = 100V$, $V_g = -5V$. From the curve corresponding to $V_a = 100V$, $V_g = -5V$ gives $I_a = 5$ mA (point P). Keeping V_a constant at 100V (i.e. remaining on the same curve) a change of 2V in grid potential to $V_g = -3V$, gives an increase in I_a to 8 mA (point Q). Thus, from the definition for g_m:—

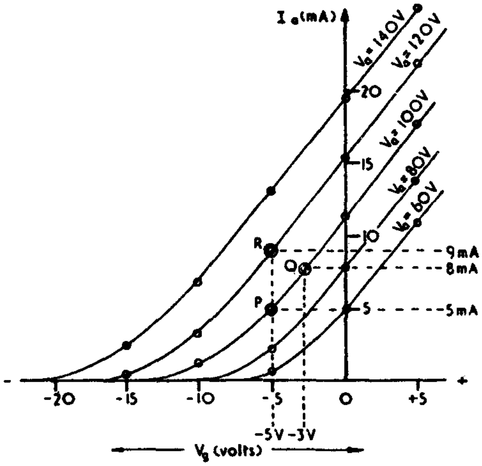


Fig. 6.—DERIVATION OF CONSTANTS FROM MUTUAL CHARACTERISTICS.

$$g_m = \frac{3mA}{2V}$$

$$\therefore g_m = 1 \cdot 5 \text{ mA/V}.$$

Going back to the initial operating point P, and keeping V_g constant at $-5V$, a change in V_a means leaving the curve for $V_a = 100V$. Considering the curve corresponding to $V_a = 120V$, the anode current for $V_a = 120V$ and $V_g = -5V$ is 9 mA (point R). Thus, an increase of 20V in V_a gives an increase of 4 mA in I_a, provided V_g is kept constant at $-5V$.

Thus, from the definition for r_a :—

$$r_a = \frac{20V}{4mA}$$

$$\therefore r_a = 5{,}000 \ \Omega.$$

Finally, since $\mu = g_m \times r_a$:—

$$\mu = \frac{1 \cdot 5}{1{,}000} \times 5{,}000$$

$$\therefore \mu = 7 \cdot 5.$$

18. Fig. 7 shows a family of *anode* characteristic curves for the same valve as that considered in Para. 17. Suppose the operating point is given as $V_g = -5V$, $V_a = 100V$ when $I_a = 5$ mA (point S). Keeping V_g constant at $-5V$ (i.e. remaining on the same curve) an increase in V_a from 100V to 120V gives an increase in I_a from 5 mA to 9 mA (point T). Thus from the definition for r_a :—

$$r_a = \frac{20V}{4mA}$$

$$\therefore r_a = 5{,}000 \ \Omega.$$

Going back to the initial operating point S, and keeping V_a constant at 100V, an increase in V_g from $-5V$ to $-3V$ gives an increase in I_a from 5 mA to 8 mA (point U). Thus, from the definition for g_m :—

$$g_m = \frac{3mA}{2V}$$

$$\therefore g_m = 1 \cdot 5 \text{ mA/V}.$$

Finally, since $\mu = g_m \times r_a$:—

$$\mu = \frac{1 \cdot 5}{1{,}000} \times 5{,}000$$

$$\therefore \mu = 7 \cdot 5.$$

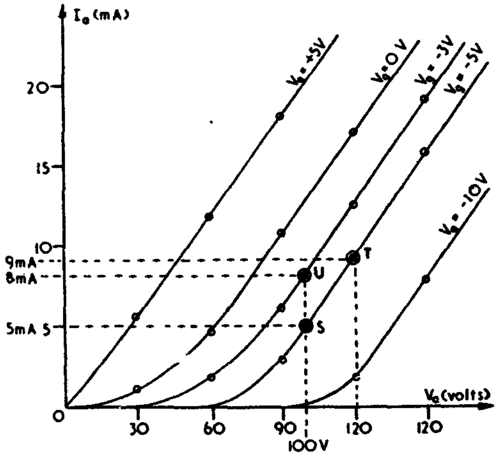


Fig. 7.—DERIVATION OF CONSTANTS FROM ANODE CHARACTERISTICS.

19. The example given in the preceding paras. emphasises that the mutual characteristics and the anode characteristics for a valve are merely two different ways of imparting the same information. One or other of these sets of characteristics (sometimes both) is given in valve data sheets issued by the manufacturers in order that suitable operating conditions may be chosen to suit the purpose for which the valve is to be used.

Grid Bias

20. It has been shown that a variation in V_g will result in a variation in I_a, V_a remaining constant. Thus, if an alternating voltage is applied between grid and cathode of the valve shown in Fig. 8(a), a corresponding variation about a mean value occurs in I_a (Fig. 8(b)). The amount by which I_a varies depends on the mutual conductance of the valve, since:—

$$\delta I_a = \delta V_g \times g_m.$$

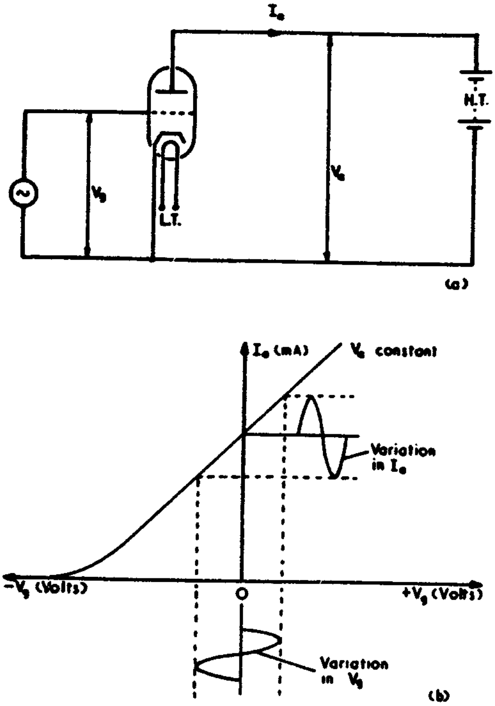


Fig. 8.—ALTERNATING VOLTAGE APPLIED TO THE GRID.

21. It is normal when using a triode valve as an amplifier to apply a *bias voltage* to the grid. This bias consists of a d.c. voltage connected between grid and cathode, usually in such a way that the grid is *negative* with respect to the cathode. The input alternating voltage is then applied *in series* with the bias voltage to the valve grid as shown in Fig. 9. The main purpose of grid bias is to adjust the operating point of the valve to suit the condition for which the valve is required. Where a negative bias is used, grid current will not be established. If current is allowed to flow in the grid circuit, three effects are apparent:—

(*a*) Distortion in waveform occurs on the positive half-cycle of the alternating grid input voltage.

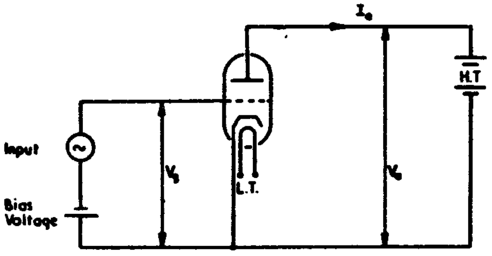


Fig. 9.—GRID BIAS.

(*b*) Power is dissipated in the grid circuit.

(*c*) The division of space current between grid and anode means that the anode current does not rise to as high a value as that to be expected.

22. **Classes of bias.** There are four main 'classes' of bias used to denote the operating conditions for a valve:—

(*a*) *Class A.* Anode current flows at all times during the entire input cycle. The normal biasing point to achieve this is midway between zero grid volts and the bottom bend (Fig. 10(a)).

(*b*) *Class AB.* Anode current flows for less than the entire input cycle but for appreciably more than half the cycle (Fig. 10(b)).

(*c*) *Class B.* The grid bias voltage is approximately equal to the cut-off value so that the anode current is approximately zero when no alternating grid voltage is applied. Anode current flows for approximately half of each cycle when an alternating grid voltage is applied (Fig. 10(c)).

(*d*) *Class C.* The grid bias voltage is appreciably greater than the cut-off value so that the anode current is zero when no alternating grid voltage is applied. Anode current flows for appreciably less than half of each cycle when an alternating grid voltage is applied (Fig. 10(d)).

Note. To denote that grid current does *not* flow during any part of the input cycle, the suffix 1 may be added to the letter of the class identification, e.g., Class AB_1.

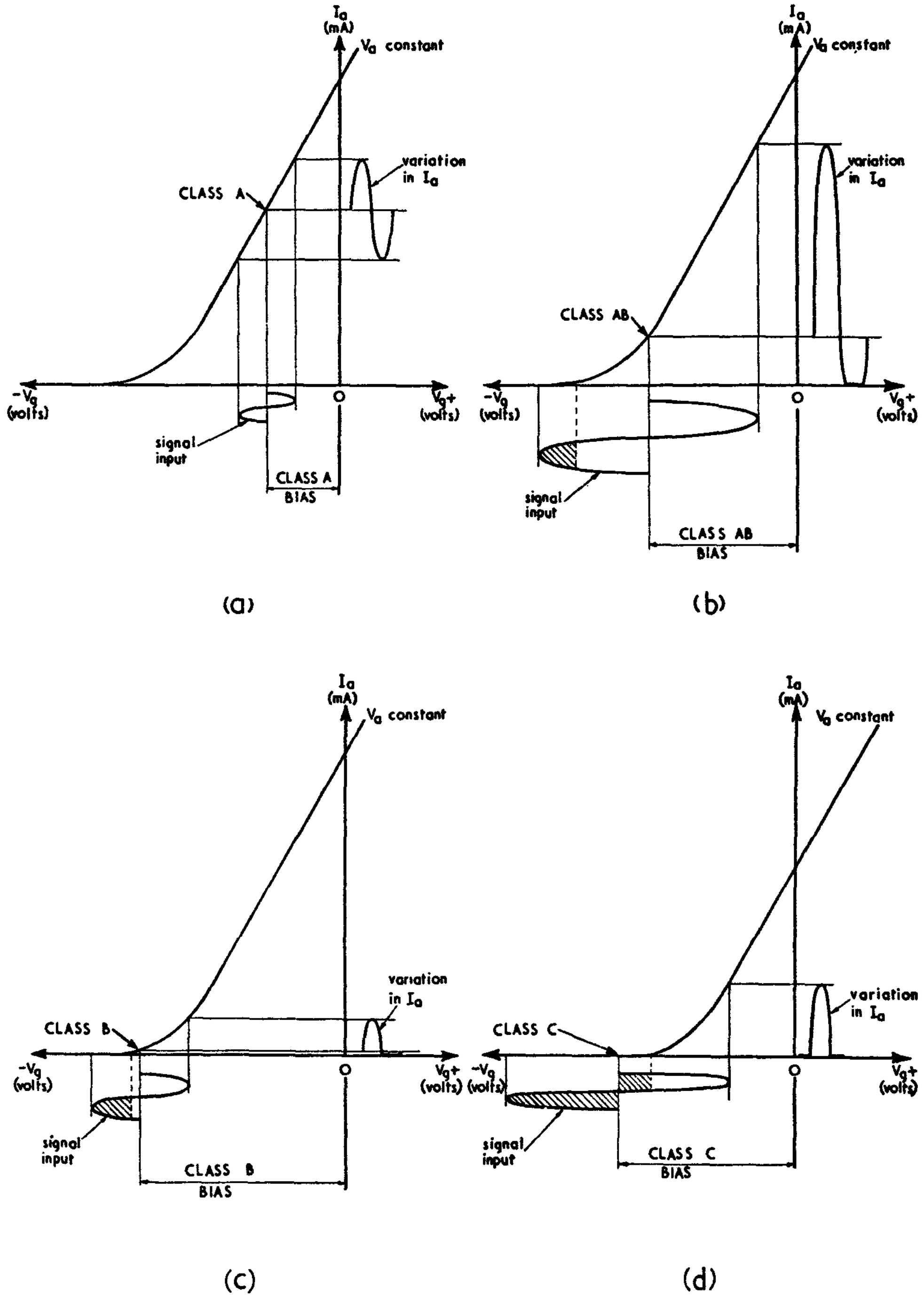


Fig. 10.—CLASSES OF BIAS.

The suffix 2 may be used to denote that grid current flows during some part of the cycle.

23. **Methods of biasing.** There are various ways in which negative bias may be applied to the grid of a valve. Some of the more common methods are described below:—

(*a*) *Battery bias.* This is used in battery-operated equipment, the bias voltage being provided by a separate battery as shown in Fig. 11.

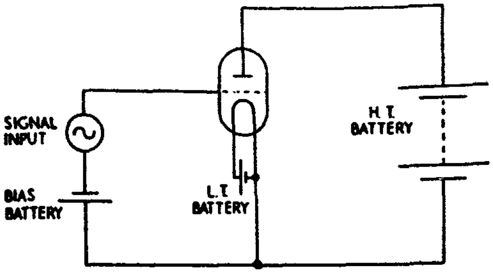


Fig. 11.—BATTERY BIAS.

(*b*) *Cathode bias.* This is used extensively in all types of radio equipment. As shown in Fig. 12, the total space current flowing through the cathode bias resistor R develops a p.d. across it in the polarity shown, so that the grid becomes negative with respect to the cathode. The magnitude of the bias depends on the d.c. value of the space current and on the value of the resistor R. The capacitor C ensures a steady d.c. bias, since the value of C will be such that to any alternating component in the current it offers negligible reactance. No alternating p.d. will then be developed across C or R. Typical values for an a.f. amplifier are R = 750Ω, C = 25 μ F. For a mean space current of 4mA the bias voltage then is :—

$$V = \frac{4}{1{,}000} \times 750$$

$$\therefore V = 3V.$$

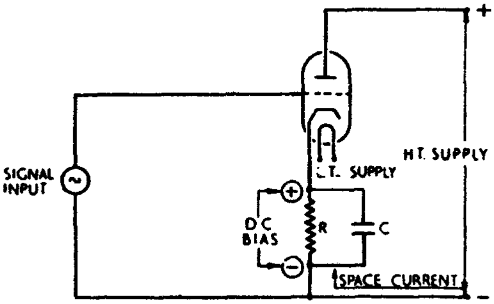


Fig. 12.—CATHODE BIAS.

(*c*) *Automatic grid bias.* This depends for its action on the flow of *grid* current. Initially, the grid is at the same potential as the cathode. On applying an alternating input voltage, grid current flows on the positive half-cycle. The flow of electrons from the grid charges the capacitor C in the polarity shown in Fig. 13(a) so that the valve operating point is moved further back. During the period of the input cycle

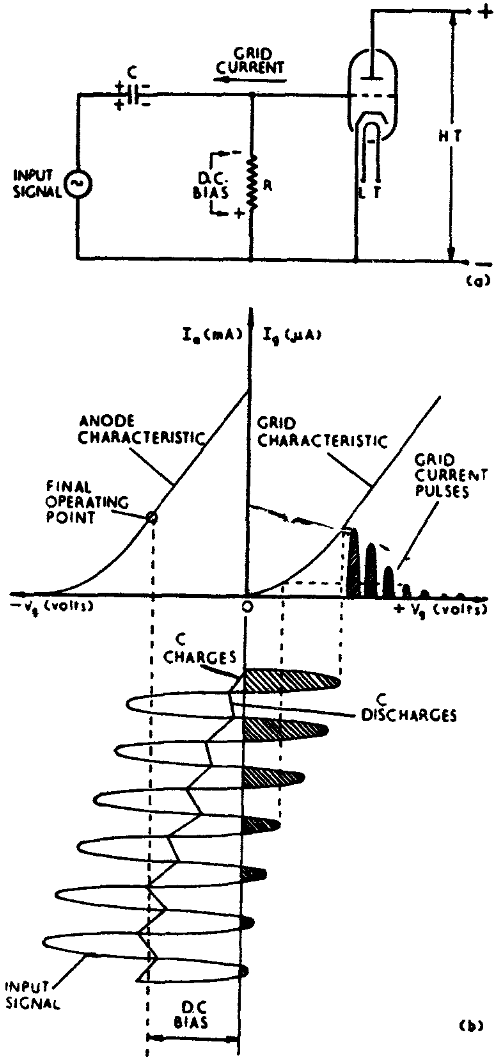


Fig. 13.—AUTOMATIC GRID LEAK BIAS.

when there is no grid current, part of the charge on C 'leaks away' through the grid leak resistor R at a rate determined by the time constant CR. On the next positive half-cycle of input, a further pulse of grid current recharges C, part of this charge leaking away through R when grid current ceases. In this way the charge on C is built up until a point of equilibrium is reached where only sufficient grid current flows to supply a charge to C equal to that leaking away through R. At this point the d.c. bias is virtually constant. The bias is automatic since an increase in the amplitude of the input produces an increase in grid current and the bias will build up to a larger value. The bias is also dependent on the time constant CR, and the values must be chosen in relation to the frequency of the input signal and to the value of bias required. A small value of R causes C to discharge more rapidly and the bias voltage is correspondingly reduced. Similar results are obtained with a small value of C. *Vice versa* for large values of C and R. Due regard must also be paid to the reactance of C at the input frequency.

The Triode as an Amplifier

24. Consider a triode valve in which the grid bias voltage is such that the valve operates under Class A conditions. When a d.c. voltage is applied from the h.t. supply to make the anode positive with respect to the cathode, the anode current assumes a steady d.c. value, known as its 'pre-signal value'. On applying an alternating signal voltage to the grid in series with the bias, the anode current will vary in a manner similar to that of the signal, the *change* in I_a depending on the g_m of the valve. If a suitable load is connected in series in the anode circuit, a p.d. is developed across it by the anode current. The variations in p.d. across this load will be similar to the signal variations but will be greater in magnitude so that *voltage amplification* has taken place. A basic form of triode amplifier having a resistor as the anode load is shown in Fig. 14. The output is taken across the anode-earth line. The potential on the anode is the h.t. potential *minus* the voltage drop across the anode load. Thus, the anode voltage (and hence the output) varies in a manner similar to that of the signal voltage to give an amplified output. It should be noted that as V_g rises, I_a rises, the p.d. across R_L rises and V_a *falls*. Thus V_g and I_a are in phase with each other, but both are 180° out of phase with V_a as shown in Fig. 15.

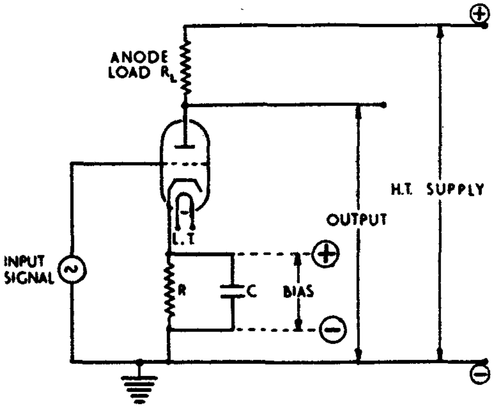


Fig. 14.—BASIC TRIODE AMPLIFIER.

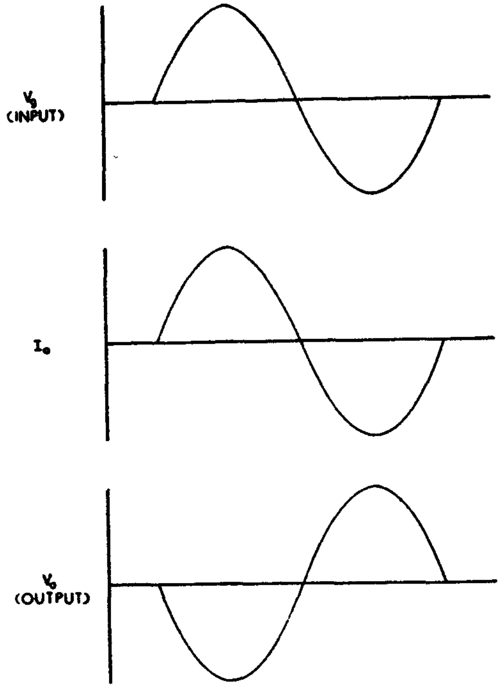


Fig. 15.—WAVEFORMS IN A CLASS 'A' TRIODE AMPLIFIER.

Equivalent Circuits

25. The equivalent circuit of an amplifier is a representation in which direct currents and voltages are neglected, and alternating components only are considered. Such circuits are useful in calculating the gain of an amplifier, and in showing the correct

operating conditions. Fig. 16(a) shows a basic amplifier circuit. Neglecting the d.c. components the equivalent circuit of Fig. 16(b) may be obtained. From the relationship for μ, a signal v_g at the grid is equivalent to a voltage $-\mu v_g$ acting at the anode, the negative sign indicating that the output from the anode is in *antiphase* with the input. The alternating component of current i_a is then due to a voltage $-\mu v_g$ acting through two resistors r_a and R_L in series so that:—

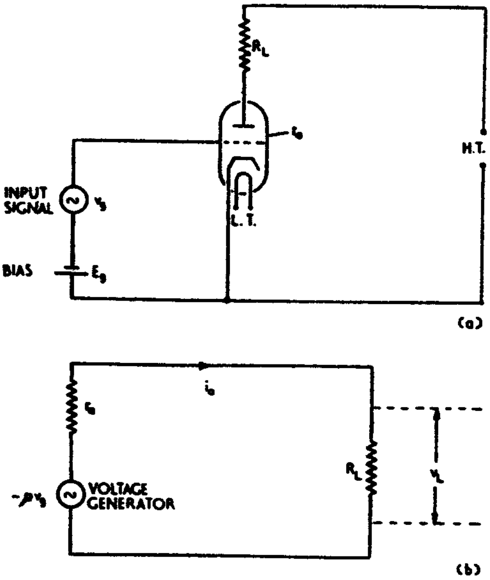


Fig. 16.—CONSTANT-VOLTAGE GENERATOR TYPE OF EQUIVALENT CIRCUIT.

$$i_a = \frac{-\mu v_g}{(r_a + R_L)} \quad .. \quad .. \quad (8)$$

The voltage v_L developed across the load R_L is $i_a R_L$. Thus:—

$$v_L = \frac{-\mu v_g R_L}{(r_a + R_L)} \quad .. \quad (9)$$

The ratio of output voltage v_L to input voltage v_g is termed the *voltage amplification factor* (v.a.f.) of the circuit. Thus:—

$$\text{V.A.F.} = \frac{\mu R_L}{(r_a + R_L)} \quad .. \quad (10)$$

The v.a.f. is always less than μ but by making R_L large in relation to r_a it can approach μ. In practice R_L is about 3 or 4 times r_a.

26. The equivalent circuit of Fig. 16 is known as a *constant voltage generator* type of circuit, and is generally used where the impedance of the source (i.e., the r_a of the valve) is relatively small. For valves with a high value of r_a it is preferable to use the *constant current generator* type of equivalent circuit. This is constructed as follows:—

In equation (9), substitute for μ in terms of g_m and r_a.

$$\therefore \; v_L = -g_m v_g \frac{r_a R_L}{(r_a + R_L)} \quad .. \quad (11)$$

From equation (11) it is seen that the output voltage v_L is produced by a current $-g_m v_g$ acting in a circuit consisting of two resistors r_a and R_L *in parallel.* The equivalent circuit for this is shown in Fig. 17. Now:—

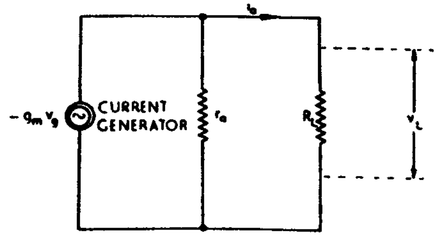


Fig. 17.—CONSTANT-CURRENT GENERATOR TYPE OF EQUIVALENT CIRCUIT.

$$\text{V.A.F.} = \frac{\text{Output } v_L}{\text{Input } v_g}$$

$$\therefore \; \text{V.A.F.} = \frac{g_m r_a R_L}{(r_a + R_L)} \quad .. \quad (12)$$

Equation (12) is identical with (10) since $g_m r_a = \mu$. This shows the equivalence of the two circuits.

Dynamic Characteristics

27. The mutual characteristic curves considered in Para. 9 are referred to as '*static*' curves since, in plotting them, V_a was assumed to remain *constant.* These static characteristics give information about the valve itself. However, when a valve is connected in a circuit with an anode load, V_a no longer remains constant. If V_g rises, the p.d. across the load rises and V_a falls. Thus, with an anode load, any variation in V_g results in a simultaneous (antiphase) variation in V_a. To take account of this, a '*dynamic*' mutual characteristic is required which is characteristic not of the valve itself, but of the valve with a particular value of anode load. It can be plotted directly from a family of static mutual curves issued by the manufacturers.

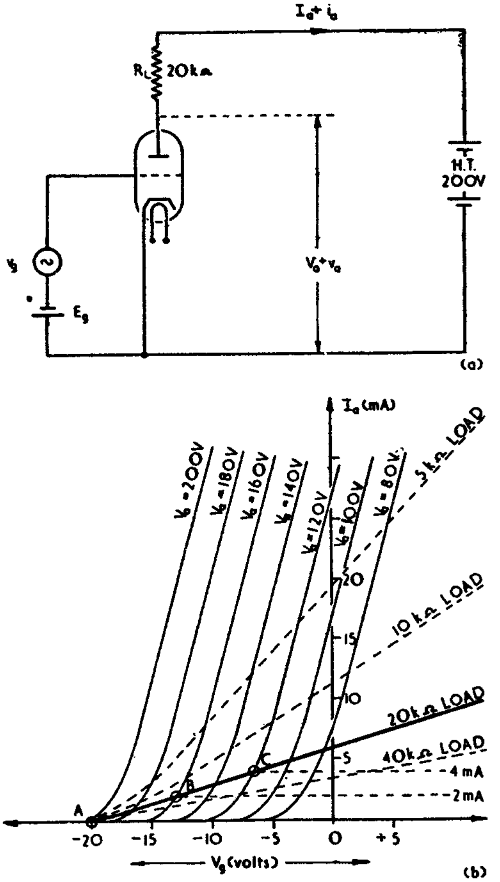


Fig. 18.—DYNAMIC MUTUAL CHARACTERISTICS

28. Fig. 18(a) shows the circuit of a triode amplifier with a purely resistive load of 20 kΩ and a h.t. supply of 200V. The static mutual characteristics for this valve are shown in Fig. 18(b). If V_g is sufficiently negative to make $I_a = 0$, the p.d. across the load R_L is zero and V_a = H.T. = 200V. This occurs at $V_g = -20V$, so that $V_g = -20V$, $I_a = 0$, $V_a = 200V$ is point A on the dynamic curve. When $I_a = 2mA$, the p.d. across R_L is 40V and $V_a = 160V$. The point B corresponding to $I_a = 2mA$, $V_a = 160V$ is another point on the dynamic curve. When $I_a = 4mA$, the p.d. across R_L is 80V and $V_a = 120V$. Thus, $I_a = 4mA$, $V_a = 120V$ is point C on the dynamic curve. The full line joining the points A, B and C gives the dynamic mutual characteristic for the valve with a load of 20 kΩ. The dynamic characteristic for any value of load resistance may be plotted in a similar manner, and curves for 5 kΩ, 10 kΩ, and 40 kΩ loads are shown dotted in Fig. 18(b).

29. Examination of Fig. 18(b) shows that:—
(*a*) The dynamic characteristics are almost straight throughout.
(*b*) The *higher* is the load resistance, the *lower* the slope of the dynamic curve.
(*c*) The smaller is the load resistance the more nearly does the dynamic characteristic coincide with the static characteristic for $V_a = 200V$, and the greater is the curvature near cut-off.

30. Suppose the operating point chosen for the circuit of Fig. 18(a) is point Q in Fig. 19, corresponding to $V_g = -7{\cdot}5V$, $I_a = 3{\cdot}5$ mA. The application of a signal of peak value 2·5V in series with the bias causes V_g to vary between −5V and −10V, and I_a to vary between 2·6mA and 4·4mA about its pre-signal value of 3·5mA. The alternating anode current of peak value 0·9mA flows in the load resistance of 20 kΩ, developing a peak voltage of 18V across this load. The v.a.f. of the circuit considered is, therefore, $\frac{18}{2{\cdot}5} = 7{\cdot}2$. It depends on the value of the load resistance as shown by equation (10). An undistorted output always results if the dynamic characteristic is straight over the entire grid swing, and the load resistance and operating point are chosen to ensure this.

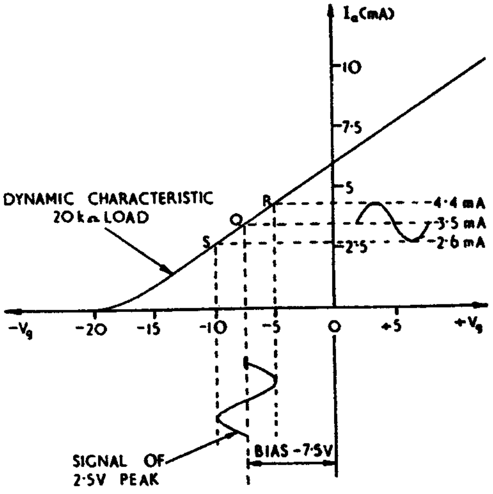


Fig. 19.—USE OF DYNAMIC CURVES.

The Load Line

31. A load line plotted on the *anode* characteristics for a valve corresponds exactly to the dynamic mutual characteristics for the

same valve and load resistance. Under no-signal conditions, the d.c. value of anode voltage is given by:—

$$V_a = \text{H.T.} - I_aR_L$$

$$\therefore I_a = \frac{\text{H.T.} - V_a}{R_L} \quad .. \quad (13)$$

This is the equation of the load line which is plotted in the form of a graph on the anode characteristics. For a purely resistive load this graph is a straight line. From equation (13) it is seen that when $I_a = 0$, $V_a = \text{H.T.}$; when $V_a = 0$, $I_a = \frac{\text{H.T.}}{R_L}$. This gives two points on the anode characteristic which enable the load line to be plotted.

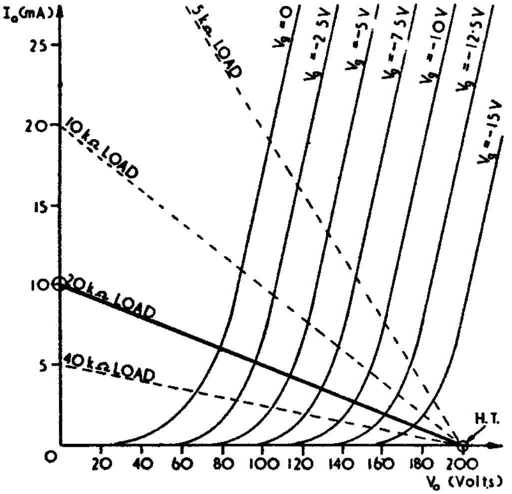


Fig. 20.—LOAD LINES.

32. Fig. 20 shows the static anode characteristics for the same valve and h.t. supply (200V) as that considered in Para. 28. The load line for a 20kΩ load is plotted from the points $I_a = 0$, $V_a = 200V$; and $V_a = 0$, $I_a = \frac{200V}{20{,}000\Omega} = 10mA$.

Load lines for 5kΩ, 10kΩ and 40kΩ may be plotted in a similar manner as shown dotted in Fig. 20.

33. In determining the operating conditions for a triode amplifier several factors are important:—

(*a*) The value of load resistance and the operating point on the resultant load line must be so chosen that distortion is reduced to a minimum.

(*b*) Grid current must not be allowed to flow if distortion and power dissipation at the grid are to be prevented.

(*c*) The rated dissipation for the valve must not be exceeded.

34. Fig. 21 repeats the 20 kΩ load line for the valve and circuit previously considered. The rated dissipation curve for 1 watt has also been inserted. Suppose the operating point chosen is point Q, which corresponds to $V_g = -7 \cdot 5V$, $I_a = 3 \cdot 5$ mA, $V_a = 130V$. The application of a signal of 2·5V peak in series with the bias causes V_g to vary between −5V and −10V *along the load line.* This causes I_a to vary between 2·6mA and 4·4mA about its pre-signal value of 3·5mA; V_a varies between 112V and 148V about its pre-signal value of 130V. Thus, for equal swings of grid voltage about the standing bias, equal swings in the values of anode current and anode voltage are obtained. This implies no distortion, since the load line makes equal intercepts on the anode characteristics. This is equivalent to selecting a straight dynamic mutual characteristic.

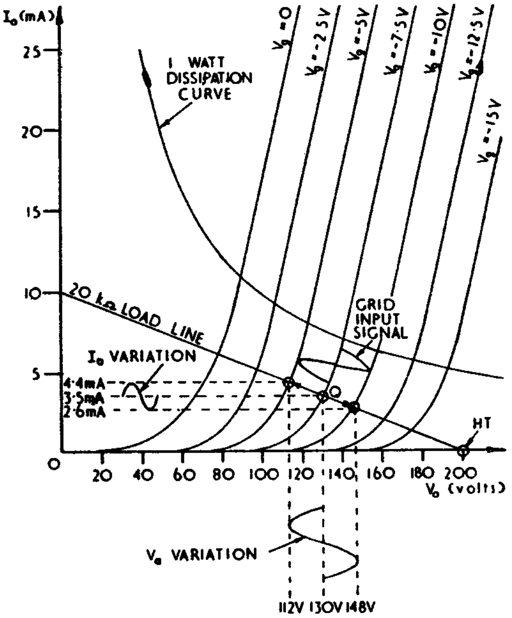


Fig. 21.—USE OF LOAD LINES.

35. With a signal of peak voltage 2·5V on the grid, an alternating anode current of 0·9mA peak flows in the load resistance of

20kΩ, thus developing a peak voltage of 18V across this load. The v.a.f. of the circuit is, therefore, $\frac{18}{2\cdot 5} = 7\cdot 2$. This is the same value as that given in Para. 29, so that the dynamic mutual characteristic and the load line are equivalent ways of expressing the behaviour of the valve with a given resistive anode load. The load line method is more usual since distortion is easier to detect, and all the waveforms may be shown simultaneously.

Practical Triodes

36. Triode valves are used for a large number of purposes in radio equipment. Such applications include voltage and power amplifiers, oscillators and trigger devices in control systems, all of which are considered in later Sections. To give this wide range of activity, triode valves are available from subminiature types up to valves capable of developing several hundred kilowatts of power output:—

(*a*) The small general purpose types of triode, such as those used in radio receivers, are air-cooled, employ either glass or metal envelopes, and use oxide-coated cathodes. In general, r_a is high (several thousand ohms), g_m is small (about 3mA/V) and μ is high (up to 100). Anode voltages rarely exceed 250V, grid bias voltages are low (−1V to −15V), and anode currents are of the order of 2mA to 20mA.

(*b*) In small power triodes capable of supplying up to 50W power output, r_a is small (a few hundred ohms), g_m is high (about 10mA/V) and μ is small (up to 10). Grid bias voltages are much higher (up to −100V), and anode currents up to 150 mA are obtained.

(*c*) In high-power water-cooled triodes used in ground transmitting equipment, the valve constants are similar to those given in (*b*), but the anode voltage may be several thousand volts, the grid bias several hundred volts and the anode current of the order of amperes.

Triode Inter-electrode Capacitances

37. The electrodes in a triode valve are conductors with small spacing between them, so that between every pair of electrodes there exists a *capacitance*. The three capacitances are:—

(*a*) C_{ga}, between grid and anode;

(*b*) C_{gk}, between grid and cathode;

(*c*) C_{ak}, between anode and cathode.

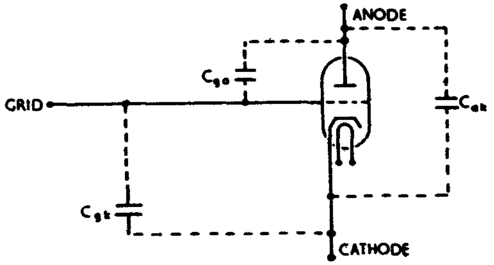


Fig. 22.—TRIODE INTER-ELECTRODE CAPACITANCES.

These are shown in Fig. 22. C_{ak} is normally so small as to be neglected, but C_{ga} and C_{gk} can have a considerable effect on the operation of the valve.

38. At low frequencies the inter-electrode capacitances may be neglected since the capacitances are very small (a few picofarads) and their reactance at those frequencies very high. However, at higher frequencies the effect (especially that of C_{ga}) may be considerable, resulting in a reduction in gain, increased distortion, instability and even o lation. The exact effect depends on the nature of the load. The triode valve has therefore, certain limitations when used at high frequencies. The effect which C_{ga} has on the valve and its associated circuits is known as the '*Miller effect*' and will be considered in detail in a later Section.

SECTION 8

CHAPTER 3

MULTI-GRID AND MULTI-UNIT VALVES

Paragraph

TETRODE VALVES

PENTODE VALVES

FREQUENCY-CHANGER VALVES

MULTI-UNIT VALVES

HIGH FREQUENCY EFFECTS

VALVE NOISE

VALVE BASES

COMPARISON OF VALVE CONSTANTS

MULTI-GRID AND MULTI-UNIT VALVES

Introduction

1. The first multi-grid valve was the *tetrode*, which has two grids. The second grid was introduced as a 'screen' between the first grid and the anode to reduce C_{ga}, thereby reducing Miller effect. Subsequently, valves with up to six grids have been produced for various purposes. In addition, there are multi-unit types consisting of more than one valve within a single envelope.

TETRODE VALVES

Tetrode Construction

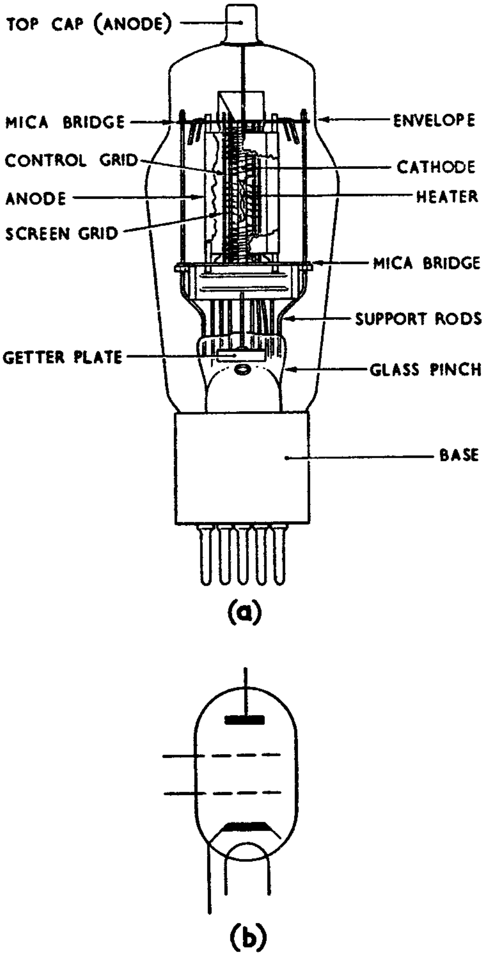


Fig. 1.—TETRODE VALVE.

2. The construction of the tetrode valve is as shown in Fig. 1 (a) and the symbol in Fig. 1(b). If the screen grid were maintained at a d.c. potential approaching that of the anode, most lines of electric force leaving the cathode and control grid would terminate on the screen, and the electric field in the space between anode and screen would be very weak. The effective capacitance between anode and screen, and hence between control grid and anode is then considerably reduced. The C_{ga} for a tetrode is of the order of 0·005 pF as compared with about 2 pF for a triode, and Miller effect becomes negligible.

Screen Grid Potential

3. If the screen grid were maintained at earth potential, the number of electric lines of force reaching the control grid from the anode would be very few because of the screening effect of the screen grid. The resultant anode current would be negligible. It is therefore essential that the screen grid be maintained at a positive potential with respect to the cathode so that an electric field is established between the control grid and the screen grid. Electrons, accelerated by this field, will then pass through the screen grid and reach the anode to establish anode current. The electric field reaching the control grid from the anode is virtually unaffected by the *anode* potential and consequently the anode current is only in a limited degree dependent upon anode voltage.

4. The screen potential of a normal tetrode valve is maintained at about two-thirds that of the anode potential. This is obtained by connecting the screen to the h.t. positive line via a dropping resistor R_s in which there is a certain volts drop due to the screen grid current I_s (Fig. 2(a)). The screen potential V_s is then given by:—

$$V_s = \text{H.T.} - I_s R_s \quad .. \quad (1)$$

Alternatively, the screen is taken to a tapping on a potential divider connected across the h.t. supply, the values of the two resistors R_{s1} and R_{s2} being adjusted in relation to the h.t. supply and the current I_s to give

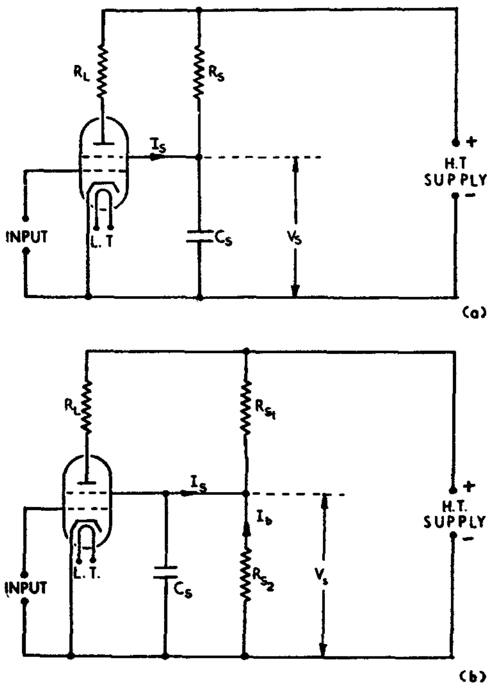


Fig. 2.—SCREEN GRID POTENTIAL.

the required screen potential V_s (Fig. 2(b)). Hence:—

$$V_s = \text{H.T.} - (I_b + I_s)\,R_{s_1} \quad .. \quad (2)$$

Alternatively:—

$$V_s = I_b R_{s_2} \quad .. \quad .. \quad .. \quad (3)$$

5. In both cases a *de-coupling* capacitor C_s is essential, and it is chosen to be of a sufficiently large value to have a very low reactance at the input signal frequencies. Thus, although the voltage swing at the control grid varies the current flowing through the valve to both the screen and the anode, the capacitance C_s maintains the screen at a constant d.c. potential. In fact, *so far as the frequencies in the signal voltage are concerned*, the screen is at the same potential as the cathode.

Secondary Emission Effects

6. "Secondary" electrons are emitted from the anode of a valve when the surface of the anode is bombarded with high-velocity "primary" electrons from the cathode. This phenomenon of *secondary emission* occurs in *all* valves, but in diodes and triodes the secondary electrons are attracted back into the anode and they have no effect on the behaviour of the valve. In the tetrode however, the fast-moving primary electrons pass through the screen grid to the anode and produce secondary emission. The slow-moving secondary electrons in the anode-screen space are then attracted either to the anode or to the screen grid, whichever is at the higher potential relative to the cathode. The resultant anode characteristic for a typical tetrode having a fixed screen potential of V_s = 100V is as shown in Fig. 3. When the anode potential is zero, the emitted primary electrons are attracted to the screen, giving a high screen current I_s and zero anode current I_a.

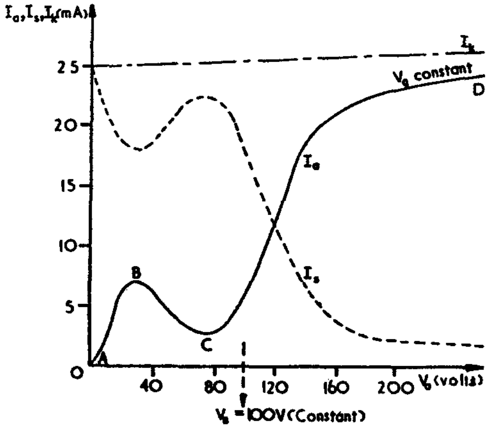


Fig. 3.—ANODE CHARACTERISTICS OF A TETRODE.

7. Over the portion AB, as the anode potential is increased, some of the electrons passing through the meshes of the screen are carried on by their momentum to the anode giving an anode current which increases with increased anode voltage. Over this portion of the curve the primary electrons are not travelling sufficiently fast to produce secondary emission. Because of the screen grid, the potential of the anode will have very little effect on the electric field in the vicinity of the cathode and an increase in anode potential will *not* appreciably increase the total space current I_k, which remains virtually *constant* throughout. Any increase in anode current will therefore be at the expense of the screen current as shown in Fig. 3.

8. Over the portion BC of Fig. 3, the velocity of the primary electrons is such that secondary emission is occurring. Since

the potential of the anode is lower than that of the screen, the slow-moving secondary electrons from the anode are attracted to the screen, thereby causing an increase in screen current at the expense of anode current. Thus, over this portion of the graph the anode current *decreases* with increased anode potential, and the valve then behaves as a *negative resistance.*

9. Over the portion CD of Fig. 3 secondary emission is still occurring but as the anode potential is increased, more secondary electrons will be drawn back into the anode and the anode current will once more increase at the expense of a decreasing screen current. Thus, to prevent the effects of secondary emission in a tetrode the anode potential must always be greater than that of the screen.

10. To prevent operating over the 'kink' portion of the characteristic, the load and the operating point on the resultant load line must be correctly adjusted. If the load is such that the load line is lower than that shown in Fig. 4 distortion will occur. Over the operating portion the anode characteristic becomes practically a horizontal line. This does *not* indicate saturation but shows what little effect the anode voltage has on the anode current. Thus, the anode slope resistance r_a is much higher than that in a triode. The mutual conductance g_m is of the same order as that for a triode. Hence, from the relationship between the valve constants the amplification factor μ is higher than that in a triode. Due to the restriction on the working part of the characteristic imposed by secondary emission, the basic screen grid tetrode is of little use as an amplifier and is now almost obsolete.

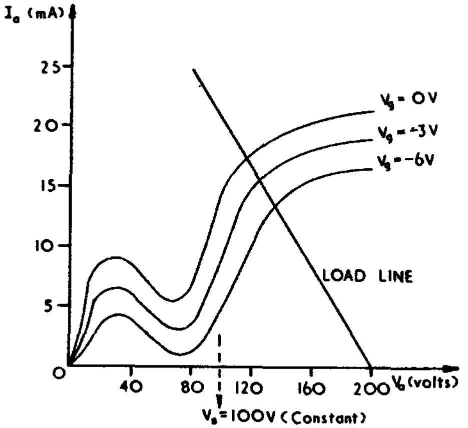


Fig. 4.—TETRODE AS AN AMPLIFIER.

Critical Distance Tetrodes

11. The essential features of the tetrode are:—

(*a*) Low grid-anode capacitance.

(*b*) High amplification factor.

To retain these features and at the same time increase the usefulness of the valve it is necessary to suppress the effects of secondary emission. One way of achieving this is to design the valve with a small control grid/screen grid separation, and a large screen grid/anode separation combined with the use of an open-meshed screen. The screen potential will still largely govern the space current, and the open-meshed screen will allow most of the electrons to pass through

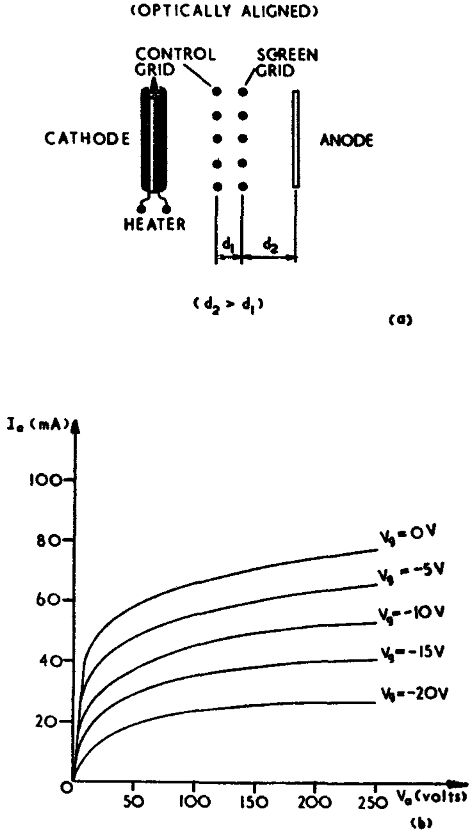


Fig. 5.—CRITICAL DISTANCE TETRODE.

and come under the influence of the anode. The screen current is kept small by *optical alignment* of the wires of the control grid and of the screen, so that very few electrons are intercepted by the screen.

12. Even if the anode potential is lower than that of the screen, secondary electrons emitted at the anode will have to travel a considerable distance before the attraction of the screen is greater than that of the anode. The result is that the secondary electrons are attracted back into the anode and the effects of secondary emission are thus eliminated. Fig. 5(a) shows the essential features of the construction, while Fig. 5(b) illustrates the anode characteristics of a typical critical-distance tetrode.

Beam Tetrodes

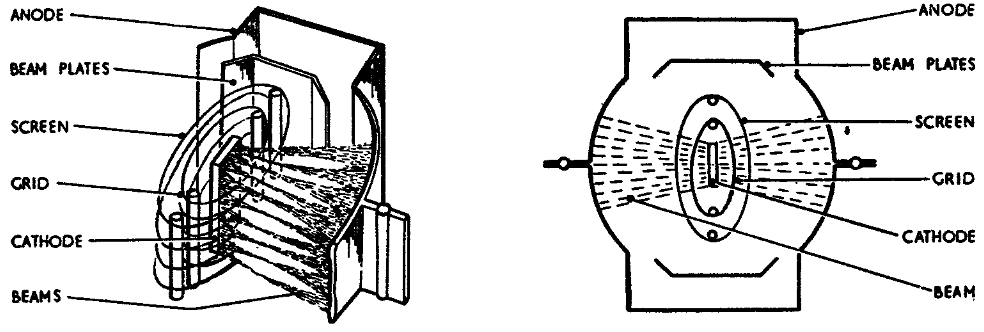
13. To reduce further the effects of secondary emission, additional elements are inserted. The electrode arrangements of a *beam tetrode* are illustrated in Fig. 6. The distinctive features of this valve are as follows:—

(*a*) The control grid and screen grid wires are of equal pitch and are so aligned that the screen grid wires lie in the 'shadow' of the control grid wires. Very few electrons are intercepted by the screen and the resultant screen current is small. At the same time, the electrons escaping through the grids form beams of relatively high current density.

(*b*) Beam-forming plates which are internally connected to the cathode are inserted to give a beam with small lateral spread.

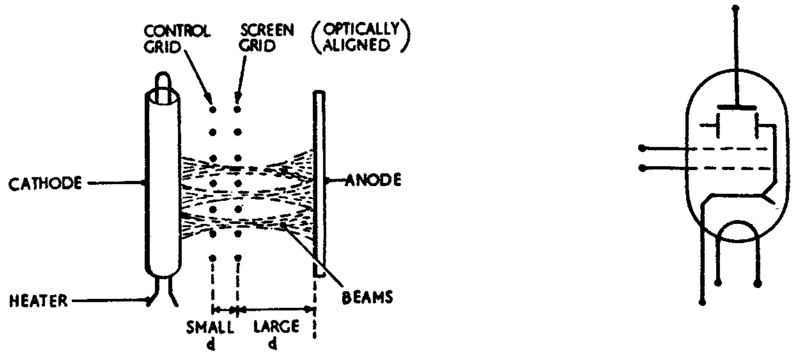
(*c*) A flat cathode is used to assist in beaming.

(*d*) There is a large anode-screen separation.



(a) SECTIONAL VIEW

(b) PLAN VIEW



(c) SIDE VIEW

(d) SYMBOL

Fig. 6.—BEAM TETRODE.

14. The concentration of the electrons into a beam, combined with a large distance between anode and screen, gives an intensified space-charge effect in the anode-screen space. The potential gradient of the electric field in this region for given values of anode and screen voltages is as shown in Fig. 7. The potential minimum near the anode prevents the slow-moving secondary electrons from the anode reaching the screen and the effects of secondary emission are eliminated. The effect on the high velocity primary electrons is negligible.

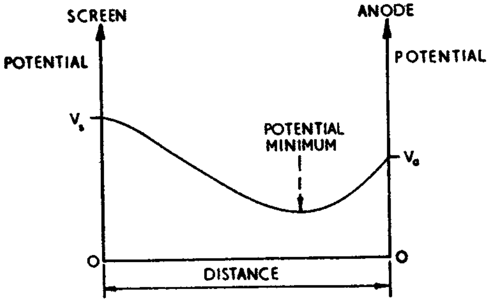


Fig. 7.—ELECTRIC FIELD IN ANODE-SCREEN SPACE.

15. Characteristic curves giving the variation of anode current with anode voltage for a constant value of screen voltage are illustrated in Fig. 8 for a typical beam tetrode. The kink evident in the screen-grid tetrode characteristic has been eliminated, and for this reason critical-distance tetrodes and beam tetrodes are often referred to as 'kinkless' tetrodes.

16. In beam tetrodes the forming of the electrons into beams produces a relatively high anode current. For this reason beam tetrodes are suitable for use mainly as *power amplifiers* at both a.f. and r.f. and in this connection are referred to as 'output tetrodes' or 'beam power valves'.

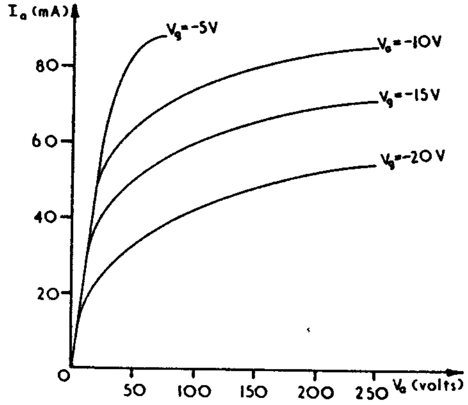


Fig. 8.—ANODE CHARACTERISTICS OF A BEAM TETRODE.

PENTODE VALVES

Pentode Characteristics

17. The pentode, as its name implies, is a five-electrode valve. It comprises a cathode, a control grid, a screen grid and a coarse mesh grid between the screen and the anode, called the *suppressor grid* (Fig. 9). The function of the suppressor grid is to prevent the effects of secondary emission from the anode. The suppressor grid is normally connected to the cathode (sometimes internally) and the resultant potential distribution in the screen-anode space shows a potential minimum at the suppressor. This has very little effect on the fast-moving primary

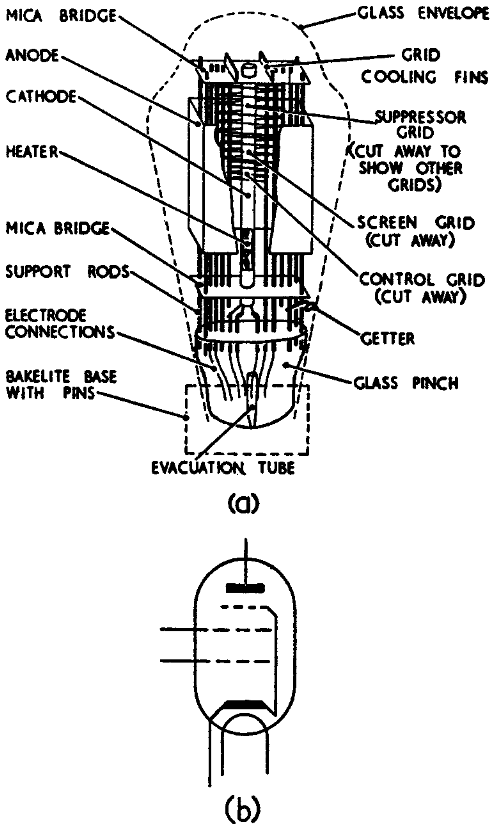
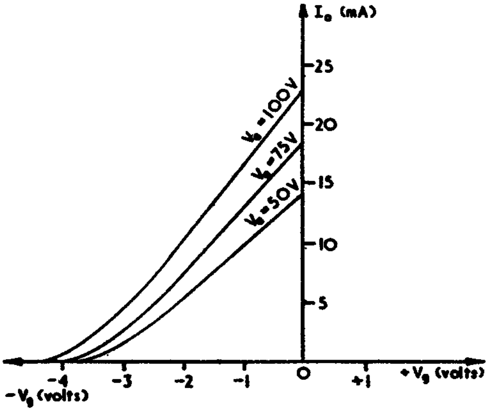


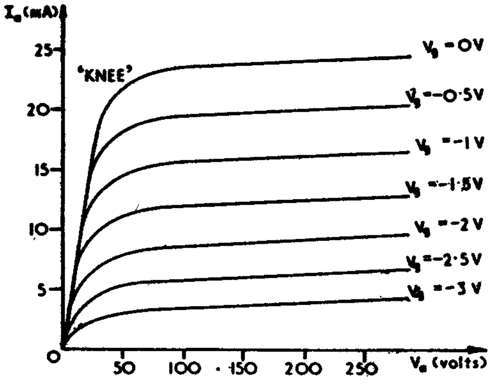
Fig. 9.—PENTODE VALVE.

electrons, but it is effective in preventing the slow-moving secondary electrons from the anode reaching the screen.

18. The mutual characteristics of a typical pentode are shown in Fig. 10(a) and the anode characteristics in Fig. 10(b). These curves have a shape similar to those for a beam tetrode but the 'knee' in the anode characteristics is sharper in a pentode. Below the knee the anode voltage has considerable effect on the anode current. The reason for this is that the primary electrons are slowed down by the potential minimum at the suppressor grid, and at zero anode voltage (when there is no electric field from the anode) the primary electrons actually stop and produce a *virtual cathode* on the screen side of the suppressor grid. Any slight increase in anode voltage reduces the negative space charge to give a rapid increase in anode current. This continues until the virtual cathode disappears at the knee of the anode characteristics. Thereafter, the anode voltage has practically no effect on the anode current since the anode is screened from the true cathode by three grids. The screen voltage, on the other hand, has an effect comparable to that of the anode of a triode.



(a) MUTUAL CHARACTERISTICS



(b) ANODE CHARACTERISTICS

Fig. 10.—PENTODE CHARACTERISTICS.

Pentode as an Amplifier

19. Fig. 11(a) shows a pentode connected as a voltage amplifier and Fig. 11(b) shows the constant-current generator form of equivalent circuit. When using a *triode* as a voltage amplifier, the anode load R_L is typically of the order of four times the anode slope resistance r_a. With a pentode, r_a is high and the value of R_L is normally small in comparison. The constant-current form of equivalent circuit is used since the impedance of the source is high in relation to that of the load.

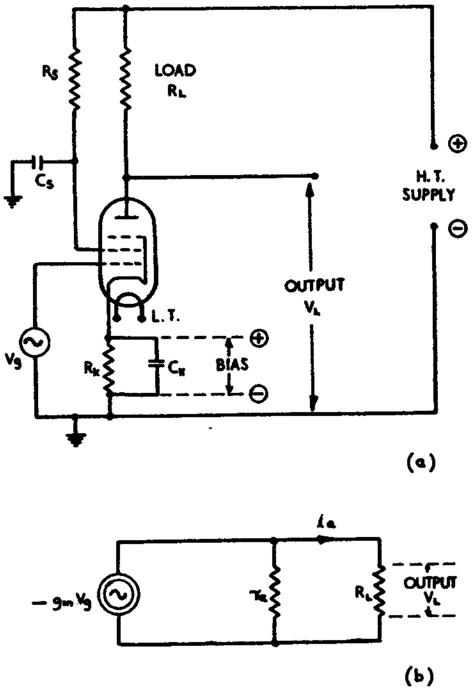


Fig. 11.—PENTODE AS AN AMPLIFIER.

20. From the equivalent circuit it is seen that the v.a.f. is given by:—

$$\text{V.A.F.} = \frac{\text{Output}}{\text{Input}}$$

$$= \frac{V_L}{V_g}$$

$$= g_m \frac{r_a R_L}{r_a + R_L}$$

$$\therefore \text{V.A.F.} = g_m \frac{R_L}{1 + R_L/r_a}$$

If, as is usually the case, r_a is very much larger than R_L, the v.a.f. for a pentode may be written as:—

$$\text{V.A.F.} \simeq g_m R_L \quad .. \quad .. \quad (4)$$

Load Lines for a Pentode

21. The anode characteristics shown in Fig. 10(b) for a typical pentode are repeated in Fig. 12. With an available h.t. supply of 300 V and a load resistance of 10 kΩ the resultant load line can be plotted and the operating point Q chosen. From the equal intercepts it is seen that the grid input can vary between O V and —3 V about the standing bias of —1·5 V without producing appreciable distortion. The peak value of anode current corresponding to a peak grid voltage of 1·5 V is 9·5 mA, giving a peak voltage of 95 V across the anode load. The v.a.f. of the stage is:—

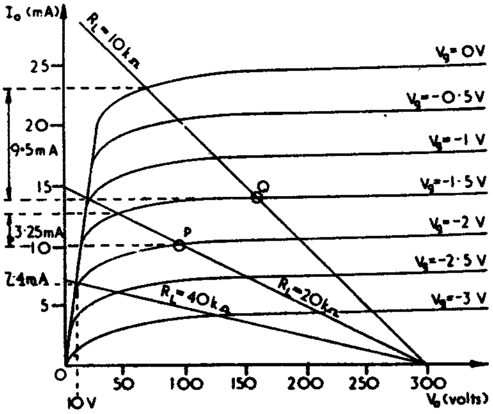


Fig. 12.—PENTODE LOAD LINES.

$$\text{V.A.F.} = \frac{95}{1 \cdot 5} = 63.$$

Alternatively, since the g_m for this valve is 6·5 mA/V, the approximate v.a.f. can be obtained from equation (4):—

$$\text{V.A.F.} = g_m R_L$$
$$= \frac{6 \cdot 5}{10^3} \times 10^4$$
$$= 65.$$

22. If an anode load of 20 kΩ is chosen, another load line is obtained with the operating point P. This time the intercepts indicate that a peak signal of 0·5 V only may be applied without producing undue distortion. The peak value of anode current corresponding to a peak signal of 0·5 V is 3·25 mA, giving a peak voltage of 65 V across the anode load. The v.a.f. is:—

$$\text{V.A.F.} = \frac{65}{0 \cdot 5} = 130.$$

Alternatively:—

$$\text{V.A.F.} = g_m R_L$$
$$= \frac{6 \cdot 5}{10^3} \times 20 \times 10^3$$
$$= 130.$$

Care must be taken in choosing the value of anode load for the available input signal, if distortion is to be prevented.

23. **Anode bottoming.** If a high anode load of 40 kΩ is chosen and the load line for this is plotted, an interesting result is obtained. The characteristics for:—

$V_g = 0$ V, $- 0 \cdot 5$ V, $- 1$ V and $-1 \cdot 5$ V all cut the load line at the same point, namely $I_a = 7 \cdot 4$ mA, $V_a = 10$ V. Thus, as the value of V_g is increased from —3 V along the load line, I_a rises and V_a falls until V_g reaches a value of $-1 \cdot 5$ V. Thereafter, any further increase in V_g produces no further rise in I_a or *fall* in V_a. The valve is said to have *'bottomed'* at a grid voltage

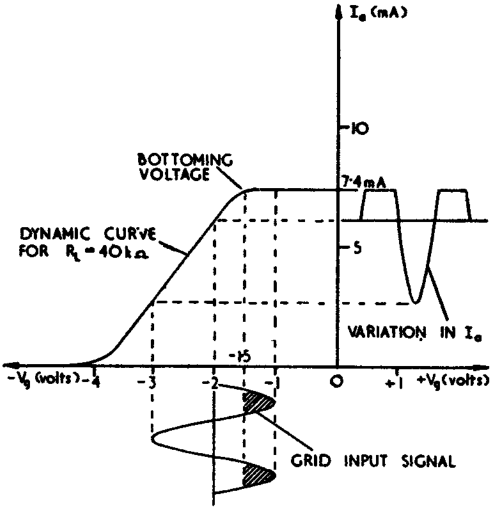


Fig. 13.—ANODE BOTTOMING.

of $-1{\cdot}5$ V (Fig. 13). This arrangement is used in some circuits on radio equipment. The cause of bottoming is the formation of the virtual cathode on the screen side of the suppressor grid at low values of anode voltage (see Para. 18). The control grid voltage has then no effect on the anode current.

Variable-mu Pentode

24. This is a particular type of pentode valve which is used considerably in the early stages of a receiver. The valve is so constructed that the anode current diminishes more and more slowly as the grid is made increasingly negative. The value of g_m (and hence of μ) therefore varies considerably over the operating range, being high at low values of negative bias and *vice versa.* The mutual characteristics of such a valve are shown in Fig, 14(a), the dotted curve representing the more normal 'sharp cut-off' pentode. The corresponding anode characteristics are shown in Fig. 14(b).

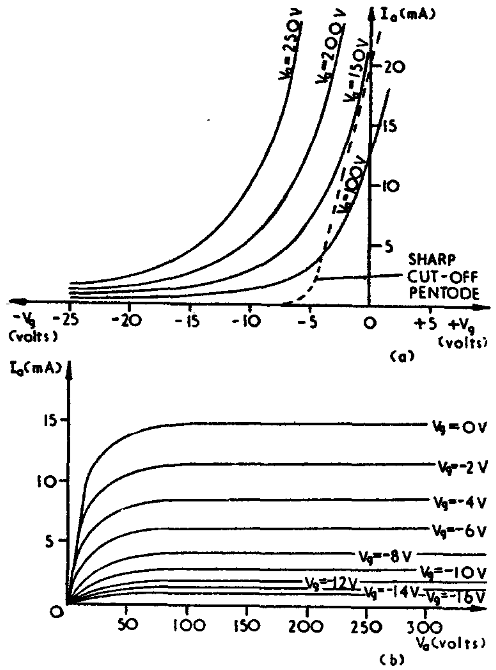


Fig. 14.—VARIABLE-MU PENTODE CHARACTERISTICS.

25. The variable-mu pentode was introduced in an attempt to keep the output from a receiver constant despite wide variations in the amplitude of the signal input, caused for example by fading. When the input signal is small, the valve operates with a small negative bias, whence the g_m and hence the gain is high (since v.a.f. = g_mR_L). When the signal amplitude increases for some reason, the negative bias is caused to increase, so that the gain is reduced and the output remains sensibly constant. This is illustrated in Fig. 15. To avoid distortion due to the curvature of the mutual characteristics the input signal should not be too large.

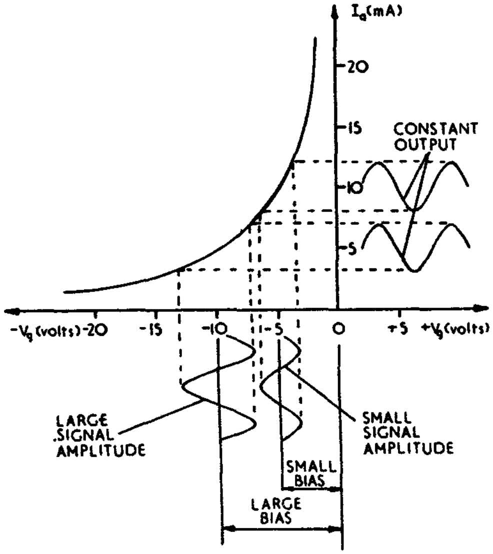


Fig. 15.—USE OF VARIABLE-MU PENTODE.

26. The variable-mu characteristic is obtained by winding the control grid so that the

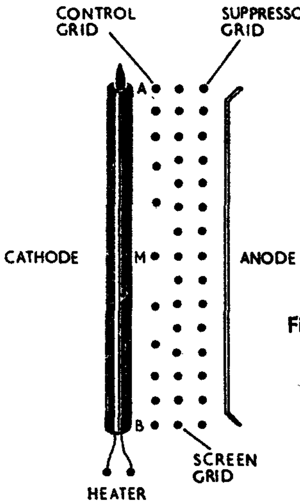


Fig. 16.—CONSTRUCTION OF VARIABLE-MU PENTODE.

wires are close together near the ends of the electrode structure (A and B of Fig. 16) but are much further apart at the middle (M). As the grid is made more negative the space current is cut off quickly at A and B and more slowly at M. A uniform variation in g_m and μ, and a long 'tail' to the mutual characteristic, may be obtained in this way. The term '*remote cut-off pentode*' is sometimes used to describe a variable-mu pentode since a very large negative bias is required to reduce the anode current to zero.

Applications of a Pentode

27. Pentode valves are available from sub-miniature types up to valves capable of handling 2 kW of power. They are used extensively in radio equipment for a wide variety of purposes but their main use is as voltage amplifiers and power amplifiers at both a.f. and r.f. Distortion is more difficult to prevent than in a triode and the '*valve noise*' of a pentode is greater (see Para. 48).

28. For most purposes, the suppressor grid is held at cathode potential in order to prevent the effects of secondary emission. However, in certain specialized circuits, it may be required to vary the suppressor grid potential (e.g., in suppressor grid modulation). The suppressor is then biased negatively and its negative potential is varied by an applied signal. This affects the potential distribution in the screen-anode space in such a way that the primary electrons passing through the screen are 'controlled' by the suppressor potential and the anode current is thus varied. Therefore, the control grid and the suppressor grid both affect the anode current and signal inputs may be applied to each grid.

FREQUENCY-CHANGER VALVES

Purpose

29. There is a group of valves designed especially for use as *frequency-changers* (see Sect. 14). In these, two separate grids exert dual control of anode current, while additional grids remove the effects of secondary emission and prevent unwanted coupling between the circuits connected to the grids and the anode.

Hexodes

30. The symbol for a hexode (six electrodes) is shown in Fig. 17. The valve is a modified pentode having an extra screen interposed between the third grid and the anode. This additional screen G_4 is usually connected internally to the existing screen G_2 and is maintained at a steady positive potential so that it functions in the same manner as the screen grid in a tetrode to reduce inter-electrode capacitance. One signal input is applied to G_1 and the other to G_3 to give dual control of anode current.

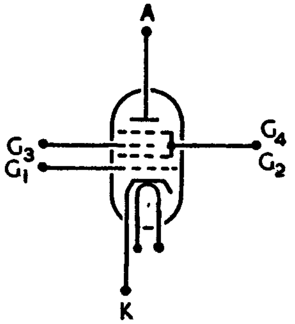
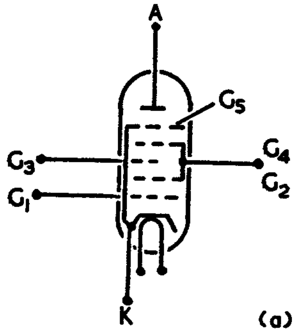


Fig. 17.—HEXODE VALVE.

Heptodes

31. In the hexode, the screen G_4 at a positive potential is adjacent to the anode and



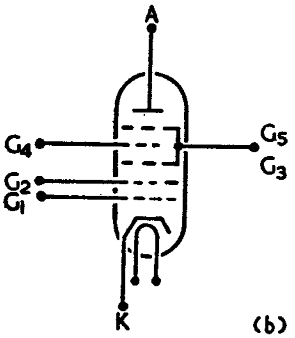


Fig. 18.—HEPTODE ARRANGEMENTS.

secondary emission effects may occur as in a tetrode. This difficulty is overcome by inserting a suppressor grid G_5 between G_4 and the anode, as illustrated in Fig. 18(a). An alternative form of heptode connection is shown in Fig. 18(b), where the cathode, G_1 and G_2 form an effective 'triode' which is used to *produce* one of the signals. The grids G_3 and G_5 are connected together and maintained at a positive potential so that they form a screen round the grid G_4 to which the second signal is applied.

Octodes

32. The heptode arrangement shown in Fig. 18(b) has a positive screen adjacent to the anode, and to prevent the effects of secondary emission a suppressor grid G_6 is inserted between the screen G_5 and the anode as shown in Fig. 19. Again, the cathode, G_1 and G_2, form a 'triode oscillator' producing one of the signals. The grids G_3 and G_5 form a screen round G_4 to which the second signal is applied, and G_6 is the suppressor grid.

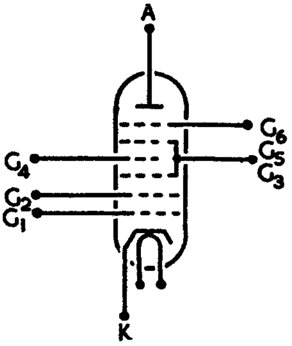


Fig. 19.—OCTODE VALVE.

MULTI-UNIT VALVES

Types

33. To save space and to economise in wiring, valves have been developed which enclose in one envelope two or more complete sets of electrodes, each set behaving as a single valve. The *triode-hexode* is a typical example and is one of the most commonly-used frequency-changer valves. The arrangement is shown in Fig. 20. The triode portion produces one of the signals which is applied to the 'injector' grid G_3 of the hexode portion. The other signal is applied to the hexode grid G_1 and the hexode grids G_2 and G_4 form a screen round G_3.

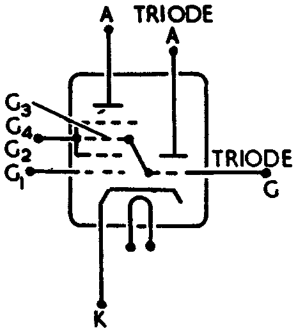
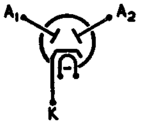


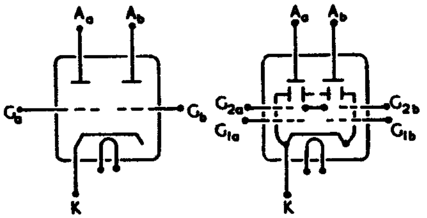
Fig. 20.—TRIODE-HEXODE VALVE

34. Other multi-unit valves are illustrated in Fig. 21. They are:—

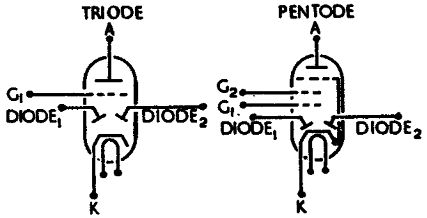
(*a*) Double-diode.
(*b*) Double-triode.
(*c*) Double-beam-tetrode.
(*d*) Double-diode-triode.
(*e*) Double-diode-pentode.



(a) DOUBLE DIODE



(b) DOUBLE TRIODE (c) DOUBLE BEAM-TETRODE



(d) DOUBLE DIODE-TRIODE (e) DOUBLE DIODE-PENTODE

Fig. 21.—MULTI-UNIT VALVES.

HIGH FREQUENCY EFFECTS

Factors Affecting Use of Valves at Higher Frequencies

35. The operation of valves is affected considerably by several factors as the frequency at which they are required to

operate is increased. The factors limiting the use of the valves so far discussed at very high frequencies are effects caused by:—

(*a*) Valve inter-electrode capacitance.

(*b*) The inductance of the connecting leads to the valve electrodes.

(*c*) The time taken for an electron to travel from the cathode to the anode of the valve (the '*transit time*' of an electron).

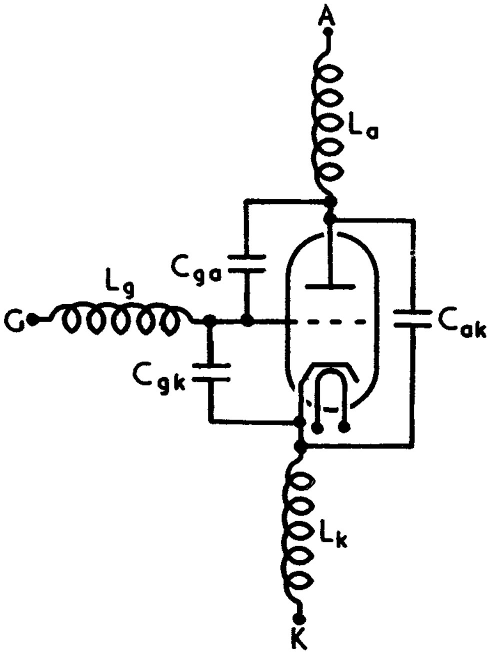


Fig. 22.—VALVE CAPACITANCES AND INDUCTANCES.

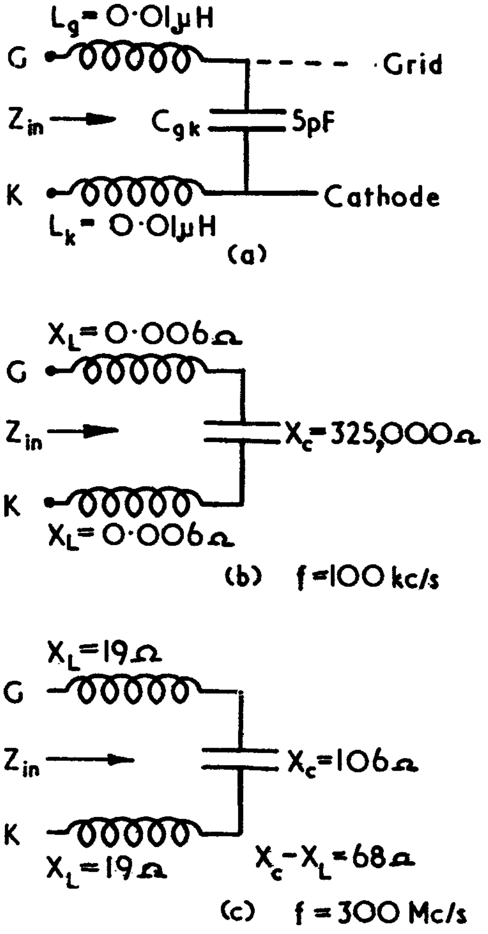


Fig. 23.—EFFECT OF VALVE CAPACITANCES AND INDUCTANCES.

Inter-electrode Capacitance and Lead Inductance

36. Fig. 22 shows a triode valve in which the valve inter-electrode capacitances and the inductance of the leads connected to the electrodes have been annotated. The anode lead inductance L_a can usually be neglected since it can be included as part of the tuned LC circuit in the anode. Similarly, C_{ak} is so small as to be negligible and C_{ga} can be considered to augment C_{gk} by virtue of the Miller effect. The important elements are then C_{gk}, L_g, and L_k.

37. Fig. 23(a) shows L_g, C_{gk}, and L_k connected in series between the grid and cathode pins of the valve. Assuming these elements to have the values indicated, it is seen that at a low frequency of 100 kc/s the input impedance Z_{in} will have a high value of 325,000 ohms (Fig. 23(b)). However, as the frequency is increased, X_L increases and X_c decreases. Thus, at a high frequency of 300 Mc/s, Z_{in} falls to the low value of 68 ohms (Fig. 23(c)). Further, at the resonant frequency of this circuit (approx. 500 Mc/s), the input impedance is zero. Thus, one effect of the inter-electrode capacitances and the lead inductances is to cause a reduction in the input impedance of the valve as the frequency increases. This results in *damping* of the valve input circuit so that the power required to supply a given voltage at the grid will be high.

Transit Time

38. In all the valve actions so far discussed it has been assumed that electrons move from the cathode to the anode in a negligibly short time. This assumption of zero transit time is satisfactory up to frequencies of about 100 Mc/s and for practical purposes, up to those frequencies, a rise in grid voltage gives an *instantaneous* rise in anode current. For example, in a typical pentode, the time taken by an electron to cross from the cathode to the anode is approximately 3×10^{-10} sec. If the valve is operating at a frequency of 1 Mc/s, one period of oscillation is 10^{-6} sec., and this time is long *in relation* to the transit time. Thus, transit time can be neglected. However, at a frequency of 300 Mc/s one period of oscillation is $\frac{1}{3 \times 10^8} = 3 \times 10^{-9}$ sec. The time taken by the electron to cross the valve is now comparable to the time taken by one cycle of the input and transit time becomes a limiting factor at such high frequencies.

Effects of Transit Time

39. When the time taken by an electron to traverse a valve is of the same order as the period of the high frequency input, several effects appear:—

(*a*) The grid-cathode input impedance falls to a low value and the power required to supply a given input voltage will be high.

(*b*) The amplification factor μ of the valve falls as the input frequency is increased.

(*c*) The phase difference between anode and grid potentials is disturbed.

40. **Effect on input impedance.** Electrons approaching the grid from the cathode induce a charge in the grid and since this charge is changing, an induced grid current is established. However, electrons which have passed through the grid and are moving away to the anode will induce an *opposite* charge on the grid and since, at low frequencies, the number of electrons approaching the grid from the cathode is approximately equal to the number leaving the grid for the anode at any given instant of time, the average grid current is zero. At high frequencies, the effect of transit time is such that more electrons may be approaching the grid than are leaving it, or *vice versa*. The two induced charges do not now cancel out and an average induced current is established, indicating that the effective input impedance Z_{in} has fallen. This effect increases with frequency. For example, in a typical pentode:—

At 10 Mc/s, $Z_{in} = 100$ kΩ
At 50 Mc/s, $Z_{in} = 4$ kΩ
At 300 Mc/s, $Z_{in} = 250$Ω

41. **Effect on μ.** A reduction in the negative potential applied to the control grid of a valve causes the anode current to increase. At very high frequencies, the transit time is such that before the electrons are able to pass through the grid, the grid potential will become more negative repelling some of the electrons back to the cathode. Thus, the anode current does not rise to the extent indicated by the valve characteristics and the effective mutual conductance g_m (and hence μ) falls as the frequency is increased.

42. **Effect on phasing.** Because of the transit time of the electron, a variation in the signal voltage at the grid does *not* cause an instantaneous variation in the anode current. This effect increases with frequency with the result that at very high frequencies the anode current may *lag* the grid voltage by a considerable angle. The phase difference between grid and anode potentials then becomes less than 180°. This effect is important in oscillator circuits.

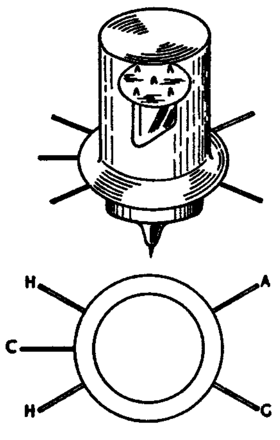
Reduction of H.F. Effects

43. To combat the above noted effects valves of special construction are used:—

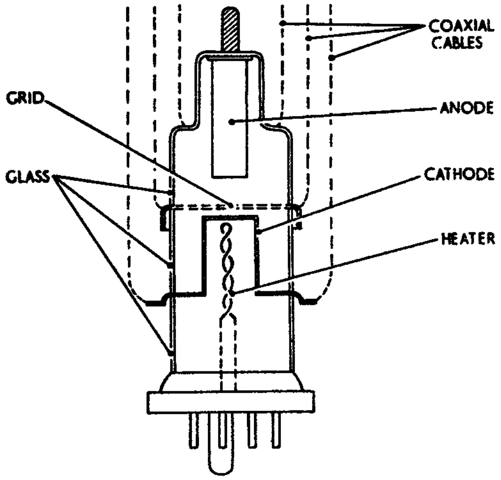
(*a*) The spacing between the electrodes is reduced and the h.t. applied to the anode is increased in order to reduce transit time.

(*b*) The reduction in spacing gives an *increase* in inter-electrode capacitances, and in order to compensate for this it is necessary to reduce the *area* of the electrodes. The two factors of close spacing and small area of the electrodes give the reason for the introduction of *miniature* and *sub-miniature* valves used at very high frequencies. Their small size however, sets a limit on the power they are capable of handling.

(*c*) Lead inductances may be minimized by the use of short leads of large diameter or by multiple leads. Alternatively, the connecting lead may be included as an integral part of the external circuit.



(a) ACORN



(b) DISC SEAL

Fig. 24.—VALVES FOR V.H.F.

44. Fig. 24 shows two examples of valves incorporating the features given above. Fig. 24(a) shows an *acorn valve* capable of operation up to 1,000 Mc/s at a power output of 500 mW. Fig. 24(b) shows a *disc-seal* (*lighthouse*) *valve* which is capable of operation at frequencies of the order of 1,000 Mc/s at a power output of 100 W. In the disc-seal valve the electrodes are connected directly to the external circuit elements (coaxial cables) so that lead inductance is eliminated.

45. Even when every possible step has been taken to reduce h.f. effects, the ultimate frequency limit in order to give a reasonable output from valves previously described is of the order of 1,000 Mc/s. Transit time is the limiting factor and when the use of other valves becomes impossible, types functioning on a different principle are used. These valves actually *make use* of transit time and include klystrons and magnetrons. They are described in Part 3 (Radar). ech-

VALVE NOISE

Noise

46. 'Noise' in a radio receiver may be defined as any unwanted electrical disturbance, usually of a random character, due to any causes. The effect of such noise is to tend to mask or obscure a desired signal, and if the ratio of the amplitude of the desired signal to that of the unwanted noise is too low, the signal will be completely submerged in the noise and will become indistinguishable. Noise in the output of a receiver is derived from a large number of sources, some external to the receiver and some internal. At this stage it is desired to discuss noise caused by valves, all other sources of noise being considered in Sect. 14.

Main Sources of Noise in Valves

47. **Shot effect.** This is due to random variations in the rate of emission of electrons from the cathode. The number and the velocity of the electrons arriving at the anode therefore varies from instant to instant, and the corresponding variations in anode current develop a random noise voltage across the anode load. A valve which is operating under space-charge-limited conditions produces a relatively low noise voltage since the electrons crossing the valve from the space charge are more constant in number and velocity.

48. **Partition noise.** This occurs in valves with more than one grid and is caused by random variations in the *division* of valve current between the anode and any other positive electrode(s). The resultant variations in anode current produce a noise voltage across the load in addition to that produced by shot effect. Thus, multi-grid valves are noisier than triodes. To reduce partition noise, optically-aligned grid structures are used so that the number of electrons intercepted by the grids is small.

49. **Induced grid noise.** This is only effective at very high frequencies where transit time effects are becoming apparent. It was explained in Para. 40 that at such frequencies a current will be induced in the control grid and since this current is random in nature, a resultant noise voltage appears across the anode load.

Equivalent Noise Resistance

50. It is convenient to express the *total* noise produced by a valve in terms of an equivalent 'noisy' resistor which, when placed in the grid circuit of an ideal (noiseless) valve, would produce the same noise voltage across the anode load as that produced by the noisy valve itself. For a triode valve, the equivalent noise resistance R_n is given approximately by the relationship:—

$$R_n = \frac{2 \cdot 5}{g_m} \qquad \ldots \quad \ldots \quad (5)$$

Multi-grid valves introduce additional (partition) noise so that the equivalent noise resistance of pentode valves is of the order of three to ten times as great as that of triode valves. Typical values are given in Table 1.

Valve Reference	Type	Equivalent Noise Resistance
CV417 (EC91)	Triode	220 Ω
CV1933 (6J5)	Triode	920 Ω
CV1091 (EF50)	Pentode	2,500 Ω
CV1981 (6SK7)	Pentode	10,500 Ω
CV1966 (6SA7)	Heptode	240,000 Ω
CV1945 (6K8)	Triode-Hexode	290,000 Ω

Table 1. VALVE NOISE RESISTANCE.

VALVE BASES

Common Types

51. The electrodes in a valve are connected via leads through the glass pinch to pins on the valve base, as noted in Chap. 1, or directly to a 'top cap'. The valve is then connected to its external circuit by inserting the valve base into the appropriate valve holder mounted on the chassis of the equipment. A.P. 1186V gives a complete list of Service valves, together with the name of the manufacturer and the valve civilian equivalent. The type of base fitted to a valve varies considerably, and depends on a number of factors. It is not, therefore, possible to give a complete list of valve bases in this Publication. However, Fig. 25 shows a few typical Service valves and the corresponding valve bases as viewed from underneath. This is in no way comprehensive and the appropriate equipment Air Publication should be consulted for the valve base type and the base pin connections used in the equipment.

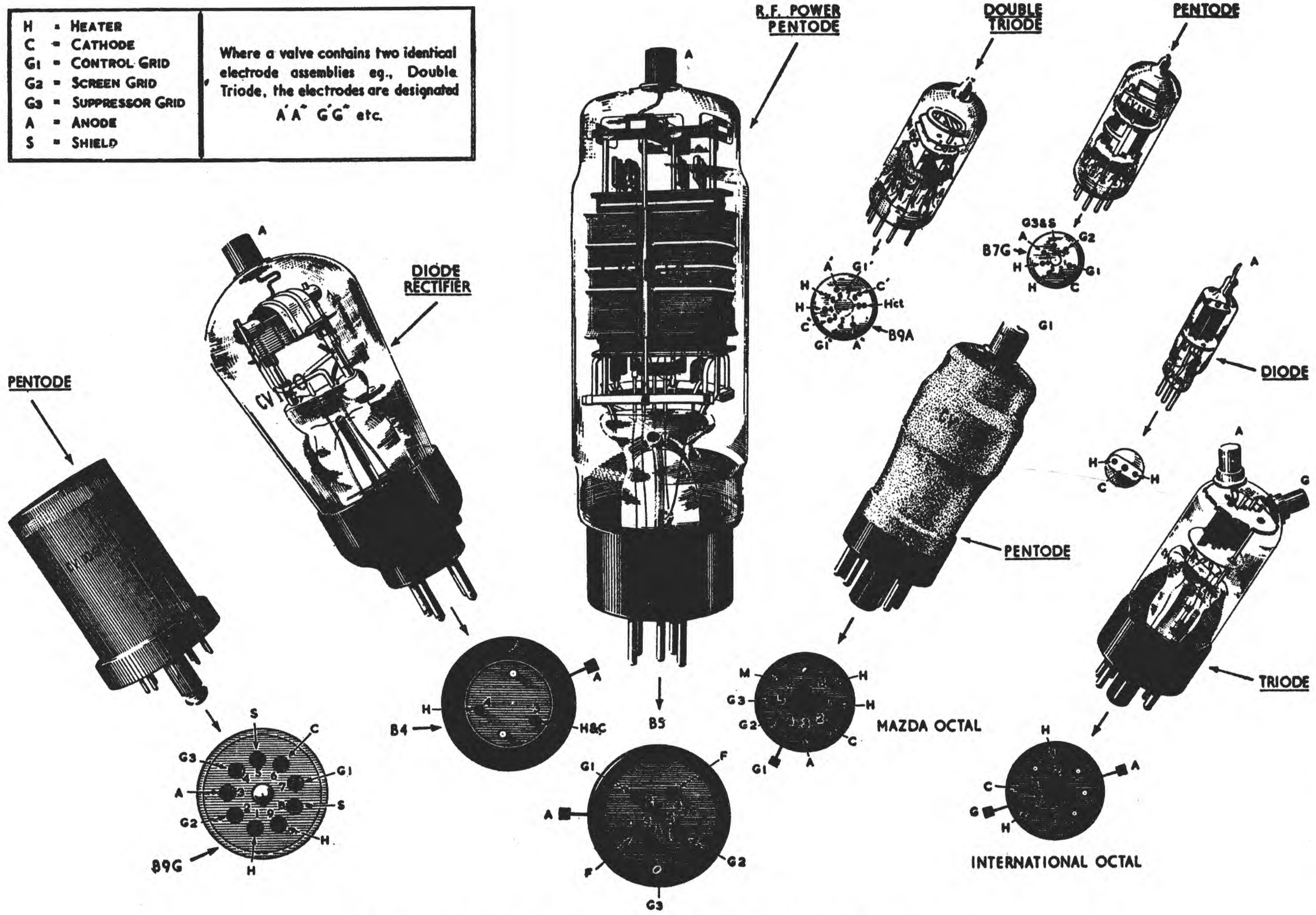
COMPARISON OF VALVE CONSTANTS

Table of Comparisons

52. Table 2 shows values of the valve constants for a few typical Service valves.

Valve reference	Valve type	r_a (ohms)	g_m (mA/V)	μ	Application
CV133(6C4)	Triode	7·7k	2·2	17	Power amplifier
CV417(EC91)	Triode	12k	8·5	102	A.F. voltage amplifier
CV1066(P61)	Triode	3·7k	4·5	17	A.F. voltage amplifier
CV510(6V6)	Beam tetrode	77k	3·75	288	Power amplifier
CV1948(6L6)	Beam tetrode	33k	5·2	171	Power amplifier
CV138(EF91)	Pentode	1 M	7·65	7,650	R.F. voltage amplifier
CV1053(EF39)	Pentode	1·25M	2·2	2,750	R.F. voltage amplifier
CV1056(EF36)	Pentode	2·5M	1·8	4,500	R.F. voltage amplifier
CV1966 (6SA7)	Heptode	1 M	—	—	Frequency-changer
CV1426(EK2)	Octode	2 M	—	—	Frequency-changer
CV1347(ECH35)	Triode-hexode	1·3M	—	—	Frequency-changer
CV859(6JEG)	Triode-heptode	4 M	—	—	Frequency-changer

Table 2. COMPARISON OF VALVES.



H = Heater
C = Cathode
G1 = Control Grid
G2 = Screen Grid
G3 = Suppressor Grid
A = Anode
S = Shield

Where a valve contains two identical electrode assemblies eg., Double Triode, the electrodes are designated A′ A″ G′ G″ etc.

Fig. 25.—TYPICAL SERVICE VALVES AND VALVE BASES.

SECTION 8

CHAPTER 4

GAS-FILLED VALVES

GAS-FILLED VALVES

Introduction

1. The valves described in the previous chapters are termed '*hard-vacuum*' or '*hard*' valves. That is, the valve envelopes have been evacuated to such an extent that the effect of any remaining gas on the operation of the valve is negligible, and for practical purposes the valve electrodes function in a vacuum. It is possible for a hard-vacuum valve to become '*soft*' with use. That is, gas occluded in the inner surface of the valve walls or in the electrodes may become active, or the valve seal may not be perfect so that the pressure inside the valve rises. The result is that the operation of the valve is affected appreciably by the gas in an indeterminate manner, and the valve is no longer reliable. One indication of this is a glow inside the valve at high values of anode voltage.

2. '*Gas-filled*' valves are considered in this Chapter. These are valves in which gas is included *intentionally* inside the envelope at a pressure less than about 10^3 newtons/sq. metre. Normal atmospheric pressure is about 10^5 newtons/sq. metre. This amount of gas is enough to determine entirely the characteristics of the valve when ionization takes place. Gas-filled valves have many uses ranging from power rectification by hot-cathode mercury-vapour diodes to extremely rapid switching by cold-cathode discharge valves in certain radar equipment.

Ionization in Gases

3. If an electron is accelerated from rest by a potential difference V between two electrodes it gains kinetic energy.

If such an electron collides with a gas atom, ionization will occur, providing that the kinetic energy of the electron (dependent on V) is greater than a certain critical value. This value varies from one gas to another and is termed the *ionization potential* of the gas concerned; it is expressed in electron-volts. Table 1 gives the ionization potentials of the gases commonly used in valves.

Gas	Ionization Potential (electron-volts)
Helium	24.5
Neon	21.5
Argon	15.7
Krypton	14.0
Hydrogen ..	13.5
Xenon	12.1
Mercury-vapour ..	10.4

Table 1. IONIZATION POTENTIALS OF GASES

Hot-cathode Mercury-vapour Diode

4. **Action.** The hot-cathode mercury-vapour diode is a diode valve containing mercury at a low pressure of the order of 1 newton/sq. metre. Assuming that the cathode is heated so that electron emission is taking place, Fig. 1 shows that as the anode voltage is increased from zero, the anode current at first increases in the same manner as that in a hard-vacuum valve. However, when the voltage becomes slightly greater than the ionization potential of mercury vapour (i.e. about 15V), ionization by collision

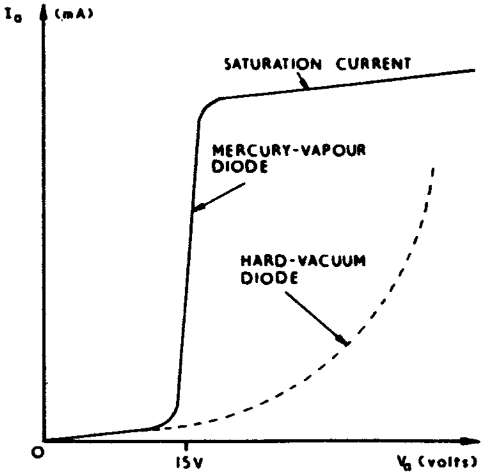


Fig. 1—ANODE CHARACTERISTIC OF A MERCURY-VAPOUR DIODE

begins. If external circuit conditions permit, the anode current will then increase to the full electron emission of the cathode with only a very small increase in the anode voltage.

5. This behaviour results from the fact that the ionization of the gas causes some of the atoms to split up into electrons and relatively heavy positive ions. The released electrons join the stream of primary electrons and travel rapidly towards the anode making a slight contribution to the anode current. The heavier positive ions move towards the cathode where they encounter and neutralize the electron space charge, so that the anode current is no longer space-charge limited and saturation level is rapidly reached.

6. In order to obtain adequate emission with this type of valve, it is normal to use oxide-coated or thoriated-tungsten cathodes. When the valve is operated correctly, the voltage drop across it when it is conducting is about 15V. The velocity of the positive ions moving towards the cathode in potential gradients of this order is such that the cathode is not adversely affected. However, if the valve is over-run, the voltage drop across the valve when it is conducting increases, and at voltages in excess of 22V, the velocity of the positive ions striking the cathode is sufficient to cause progressive disintegration of the cathode emitting surface. It is, therefore, important that the circuit external to a gas-filled diode be so arranged that the anode current is limited to a value that is safely below the full electron emission of the cathode.

7. Mercury-vapour valves do not start at their correct operating point if the ambient temperature is too low. It is only when the effective temperature of the mercury is about 40°C that the vapour pressure is adequate. If the pressure is too low, the intensity of ionization is reduced to the point where the space charge neutralization action does not occur. The voltage drop in the valve is then excessive and cathode disintegration results. The l.t. supply should therefore be applied before the h.t. in order to allow the valve time to reach the required temperature before anode current is established. This normally necessitates the use of a '*time delay*' circuit in the h.t. supply (see Sect. 9).

8. **Construction.** The space-charge neutralization that occurs in a mercury-vapour diode permits the use of electrode structures that would not be practical in hard-vacuum valves. Anode structures can be made small in proportion to current rating because the low voltage drop of the valve makes the power dissipation at the anode small. Relatively large spacing between anode and cathode is also permissible. The symbol for a hot-cathode, gas-filled diode is shown in Fig. 2(a), where the cross-hatching indicates that it contains gas, and Fig. 2(b) illustrates the general construction of the valve.

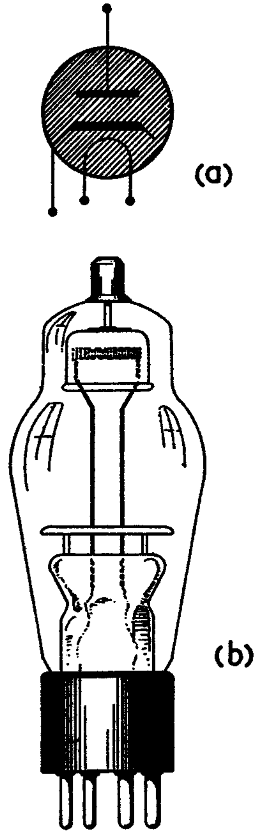
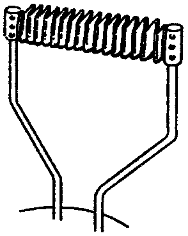


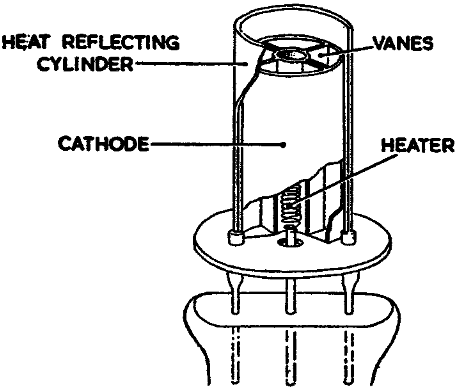
Fig. 2—MERCURY-VAPOUR DIODE

9. In hard-vacuum valves the cathode emitting surface must face the other electrodes if the space charge is not to be too large. In a gas-filled valve the cathode can be designed to have a high thermal efficiency by enclosing or shielding the cathode since the positive ions penetrate into the cathode and neutralize the space charge in any case. For a directly-heated

cathode, a wide ribbon that is crimped and folded to give a large emitting surface in a small volume is used as shown in Fig. 3(a). Indirectly-heated cathodes are commonly constructed as shown in Fig. 3(b). An oxide-coated inner cylinder with attached vanes is placed round the heater and acts as the cathode. This provides a large emitting surface in relation to the heat-radiating area. Surrounding the cathode is another nickel cylinder of high reflecting power to reduce the radiation of heat from the cathode. This type of cathode is known as a '*heat-shielded*' cathode.



(a) DIRECTLY HEATED



(b) INDIRECTLY HEATED

Fig. 3—CATHODE CONSTRUCTION FOR A GAS-FILLED DIODE

10. **Use.** Mercury-vapour diodes are widely used for rectification of alternating current supplies where the power requirements are high. The current they are capable of supplying is much greater than that supplied by a hard-vacuum diode rectifier. Table 2 gives a comparison between the two types.

Note. Mercury vapour is not the only gas which may be used in hot-cathode diodes. Xenon-filled and argon-filled diodes have been successfully developed and although the ionization potentials and the voltage drops across such valves are greater than those of a mercury-vapour diode, the danger of failure due to very low or very high temperatures is elimated.

	Peak Current	Rated Average Current	Filament	
			Volts	Amps
Typical hard-vacuum diode ...	0.4A	0.075A	5	2
Typical mercury-vapour diode ...	5.0A	1.25A	5	7.5

Table 2. COMPARISON BETWEEN HARD-VACUUM AND MERCURY-VAPOUR DIODES

Hot-cathode Gas-filled Triodes

11. **Action.** The most common valve of this type is termed a '*thyratron*', which is essentially a triode valve containing gas at low pressure. The contained gas may be mercury-vapour, helium, argon, hydrogen or a mixture of inert gases depending on the requirements. The thyratron grid controls the flow of electrons from the cathode to the anode, but only in the sense of an 'on-off' switch. If the grid potential is considerably more negative than the cut-off bias value, the anode current is zero. By keeping the anode voltage constant and gradually reducing the value of negative bias it is found that, at the point where anode current would just start to flow in a normal hard-vacuum triode, the current jumps from zero to a high value. If external circuit conditions permit, the anode current can reach saturation value with anode voltages as low as 15 V. Once anode current has been established by the action of the grid this electrode *loses control* and can be made much more negative than the cut-off value of bias without affecting the anode current. To cut off the anode current it is then necessary to reduce the *anode* voltage below the ionizing potential of the contained gas. The control grid potential merely determines the anode voltage required to 'strike' the valve. The graph showing how the striking voltage is determined by the grid potential is shown in Fig. 4(a). Fig 4(b) shows the anode characteristic for a typical thyratron.

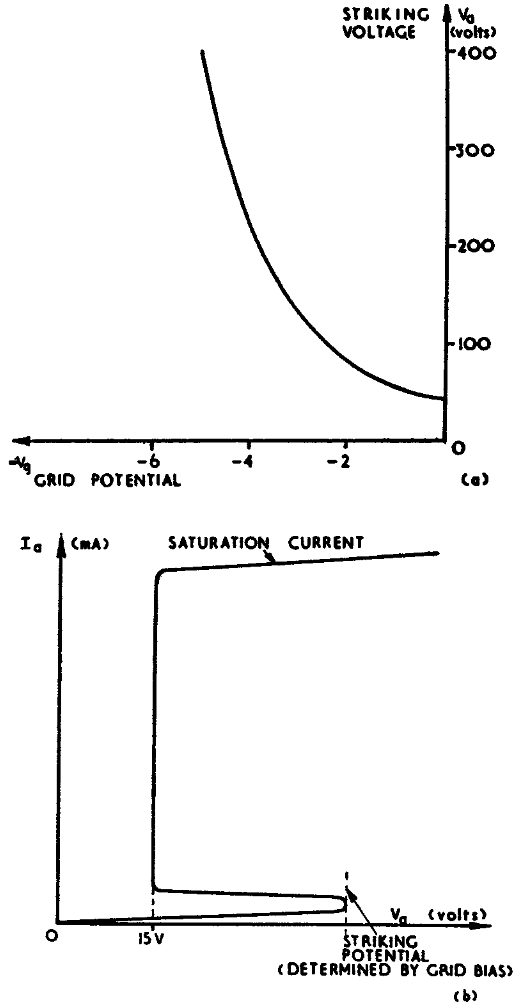


Fig. 4—THYRATRON CHARACTERISTICS

12. The thyratron behaviour is caused by the fact that, as soon as the control grid bias allows anode current to flow, positive ions are produced as a result of ionization by collision. Some ions are attracted towards the grid, surrounding it with a sheath of positive ions that neutralize the bias, so that the grid loses control. At the same time, other positive ions are attracted towards the cathode and neutralize the space charge. Hence, once ionization has started, there is no space charge to limit the anode current, and the control action of the grid has been lost.

13. **Construction.** A thyratron is not constructed in the same manner as a conventional hard-vacuum triode. The symbol for a thyratron is shown in Fig. 5(a), and Fig. 5(b) illustrates a typical construction. The cathode is a heat-shielded type as described in Para. 9. The grid is a cylinder surrounding the cathode and containing an annular disc, through the aperture of which the electrons pass to the cup-shaped anode. This construction allows the grid to have a very effective control over the anode current before ionization.

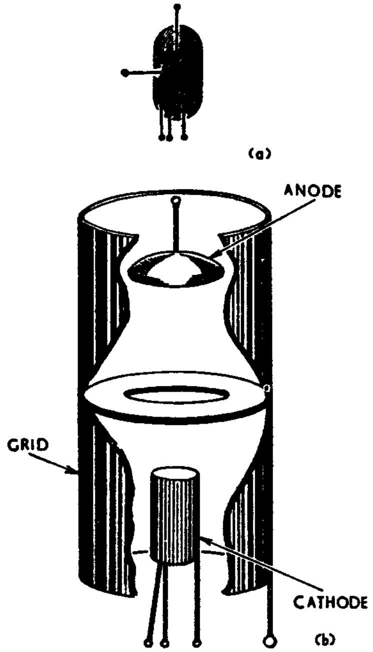


Fig. 5—THYRATRON CONSTRUCTION

14. **Gas used.** In a mercury-vapour triode the time lag between the application of a switching voltage to the control grid and the flow of current to the anode (the *ionization time*) may be a few micro-seconds. The *de-ionization time* (i.e. the time between cutting off the anode voltage and the cessation of anode current) will generally be several hundred micro-seconds. These time lags are due to the inertia of the gas molecules and they set a limit to the usefulness of a thyratron for rapid switching. To reduce the ionization and de-ionization times, thyratrons with argon or helium gas are used. Where very rapid switching is required for the production of pulses, hydrogen thyratrons may be used, the hydrogen atom having a high relative mobility. The ionization potential and the voltage drop across the valve are greater with these gases than

with mercury vapour but these disadvantages are offset by the reduced ionization and de-ionization times.

15. **Uses.** The thyratron may be used as a 'trigger' device that has numerous important practical applications in control work. It takes little energy at the control grid to initiate the discharge and at the same time the resulting energy available at the anode may be very large. The thyratron is also used in basic 'time-base' circuits for cathode-ray tubes.

Cold-cathode Gas-filled Diode

16. This consists of two electrodes in a bulb containing an inert gas, usually neon or argon, at low pressure. The symbol is shown in Fig. 6(a). The cathode is unheated so that no thermionic emission takes place. However, in any volume of gas there are always a few free electrons. As the p.d. between anode and cathode is increased the kinetic energy of the free electrons rises until a level, sufficient to ionize the gas, is reached. A glow then appears inside the valve; if the gas is neon, the glow is pink; if argon, blue. The *striking potential* depends on the distance between the electrodes and on the pressure of the contained gas, but a typical figure is 120V. Once ionization commences, the positive ions bombard the cold cathode to produce new electrons and a large current can then be established in the valve. The voltage required to maintain this discharge is less than the striking voltage, so that once the valve has struck the voltage across it falls to a value known as the *extinction voltage* (typically about 80V) and then remains constant over a wide range of current. A graph relating current and voltage for a cold-cathode discharge diode is shown in Fig. 6(b).

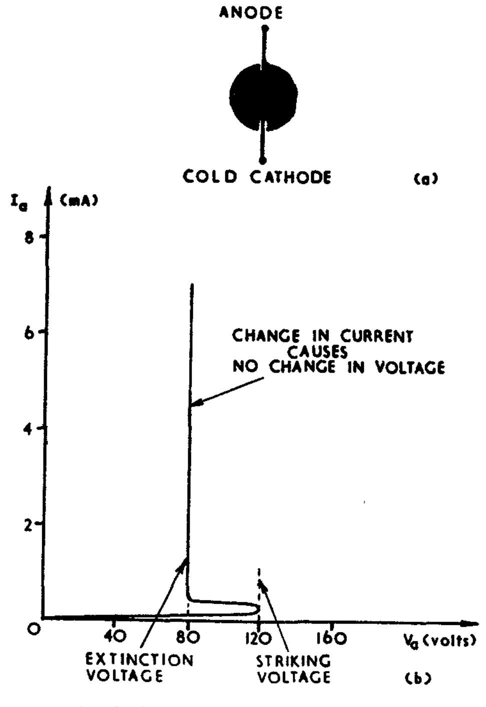


Fig. 6—COLD-CATHODE DISCHARGE DIODE

17. A valve of this type is used as a *voltage stabilizer* since, once it has struck, its anode-cathode voltage is stabilized at the value of the extinction voltage, provided the current through it is of sufficient magnitude. The arrangement of the valve as a voltage stabilizer is shown in Fig. 7. Any change in the input voltage across AB produces a change in current through R and V, but no change in the voltage across CD nor in the current in the load resistance R_L (see Fig. 6(b)). Similarly, any variation in the load resistance R_L causes a change in the current in R and V but no change in the voltage across CD. Thus the output voltage across CD is stabilized by the valve V.

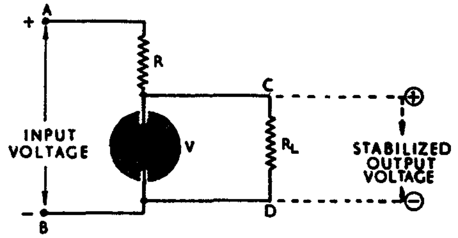


Fig. 7—COLD CATHODE DIODE AS A VOLTAGE STABILIZER

18. Cold-cathode gas-filled diodes are used to stabilize voltages of the order of 100V, at different current ranges from 1mA to 200mA. For supplies of several hundreds of volts, three or four valves may be used in series. To prevent instability it is usual to connect each unused electrode to earth through a suitable resistance (Fig. 8(a)), to prevent the valves 'flashing'. Flashing is caused by the current falling to such a low value that the valve ceases to conduct.

The anode voltage then rises to the striking value at which the valve reconducts and flashing occurs. The shunting resistances tend to maintain the p.d. across the valve at its stabilized value. Instead of several valves in series, a multi-electrode valve, known as a '*stabilovolt*' (trade name), may be used. This contains, besides the anode and the cold cathode, several intermediate gridded electrodes as shown in Fig. 8(b). The whole envelope is filled with gas at low pressure and the effect is similar to that of having four separate diodes in series. The stabilovolt is generally constructed with all five electrodes brought out to terminals so that the intermediate stabilized values, as well as the full output, may be used.

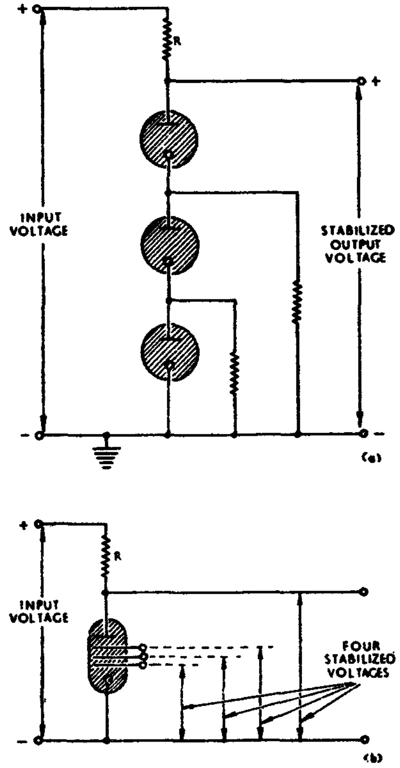
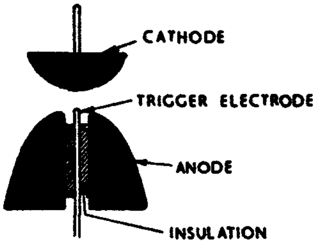


Fig. 8—STABILOVOLT

Cold-cathode Gas-filled Triodes

19. It is sometimes necessary to arrange a circuit so that a 'switch' is closed when the potential across its terminals reaches a certain high value. This process is conveniently carried out by putting a cold-cathode gas-filled valve between the terminals. When the p.d. across the valve reaches the value of the striking voltage the valve conducts, and while it is conducting its resistance is so low as to be considered a short-circuit for practical purposes. When the potential difference is removed the spark is extinguished after a period of de-ionization. In radar circuits, the spark in the gap between cathode and anode should occur at precise regular intervals. In the cold-cathode diode, the striking and extinction potentials are rather indeterminate and variable so that it is preferable to initiate the ionization by the use of a control electrode whose function is similar to that of the grid of a thyratron. This is obtained by the use of the cold-cathode gas-filled triode (sometimes called a triggered spark-gap) which consists of a cold cathode, an anode and a third electrode termed the trigger electrode.

20. In the cold-cathode gas-filled triode the p.d. between anode and cathode necessary to spark the gap is much lower and the discharge is much more precise when ionization is already present. A common construction of a triggered spark-gap, sometimes known by its trade name of '*trigatron*', is illustrated in Fig. 9. The cathode and anode take the form of hemispherical caps, and the trigger electrode is a wire the end of which protrudes through an aperture in the anode. The spacing between the anode and cathode is such that ionization does not occur even when the maximum



(a) CONSTRUCTION

(b) SYMBOL

Fig. 9—TRIGATRON

available p.d. is applied. Ionization is initiated by the application of a pulse of high voltage to the trigger electrode. A resistance in series with the trigger electrode limits the current in this circuit so that little power is required for the trigger pulse. On the application of the trigger pulse a spark occurs between the trigger electrode and one of the main electrodes. Once the gas has been ionized in this way, the p.d. between the main electrodes is sufficient to spark the main gap and the valve 'fires', its resistance falling virtually to a short-circuit value. The spark is extinguished, and the valve ceases to conduct when the p.d. between the two main electrodes falls below the ionization potential of the gas in the gap. This action is normally made repetitive and is used in the production of high-frequency pulses. Such systems are considered in Part 3 (Radar)

SECTION 8

CHAPTER 5

CATHODE-RAY TOBES

ELECTROSTATIC C.R.T.

MAGNETIC C.R.T.

CATHODE-RAY TUBES

Introduction

1. The cathode-ray tube (c.r.t.) is primarily an indicating device with a very light 'pointer' possessing no appreciable inertia; it can, therefore, be moved at high velocities. The 'pointer' is a beam of electrons which may be deflected by means of electric or magnetic fields. Where the electron beam impinges on the screen of the tube, it produces a spot of light and as the beam is moved by the deflector system, a pattern is traced on the screen showing graphically the variations of the electrical input. Alternatively, the brightness of the spot may be varied to provide additional information. The c.r.t. has many applications, e.g. displaying current and voltage waveforms in the osilloscope, as the indicator of a radar set (see Part 3) or for presenting a television picture. In the latter case the brightness of the spot is varied as well as its position on the screen.

2. The essential elements of a c.r.t. may be listed as follows:—

(*a*) An *electron gun* which 'fires' the electrons at a fluorescent screen. The gun comprises:—

(i) a *cathode* which emits the electrons;

(ii) a *modulator electrode* (sometimes called the control electrode or grid) which controls the intensity of the beam and, therefore, the brightness of the spot on the screen;

(iii) an *accelerator* (anode) system which accelerates the electrons.

(*b*) A means of *focusing* the electrons so that they arrive at the screen at a single point. Electrostatic or magnetic focusing may be used.

(*c*) A means of *deflecting* the electron beam. Electrostatic or magnetic deflection may be used.

(*d*) A *screen* which becomes luminescent under the impact of the electron beam.

All these elements are contained inside an evacuated glass tube which is suitably shaped.

3. Cathode-ray tubes are of two main types:—

(*a*) *Electrostatic tubes*, in which focusing and deflection are due to the action of electric fields.

(*b*) *Magnetic tubes*, in which focusing and deflection are carried out by the action of magnetic fields.

Both these types are in common use in the Service. However, tubes are not rigidly classified in this way since there are variations for different purposes. For instance, tubes using electric focusing and magnetic deflection may be met (see Para. 38).

ELECTROSTATIC C.R.T.

Construction

4. Fig. 1 shows the schematic layout of a typical two-anode electrostatic c.r.t. The electrons are emitted from the indirectly-heated cathode K, the filament power being supplied to the heater H. The intensity of the beam is controlled by varying the negative bias on the modulator electrode M. The

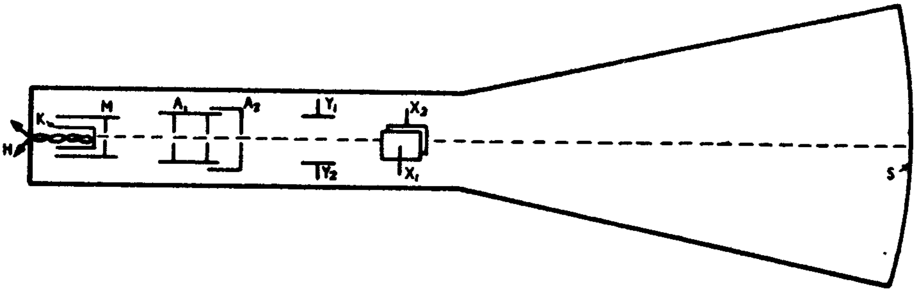


Fig. 1—ELEMENTS OF A TWO-ANODE ELECTROSTATIC C.R.T.

electrode A_1 is a metal cylinder that is provided with centre baffles to confine the beam. Coaxial with this first accelerator or anode is a second cylinder A_2, the anode that gives the electrons their final velocity in the direction of the screen. Both A_1 and A_2 are positive with respect to the cathode. Focusing of the electron beam is accomplished by varying the potential of A_1. Because of this, A_1 is called the *focusing* electrode. The combination of H, K, M, A_1 and A_2 forms the *electron gun*. The plates X_1 X_2 and Y_1 Y_2 deflect the electron beam when a p.d. exists between the plates of a pair. The X plates produce horizontal deflection and the Y plates vertical deflection. The screen S is deposited on the inside surface of the tube face and is made of a material which glows when excited by electrons striking it. These elements of the electrostatic c.r.t. will now be considered in detail.

5. **Cathode.** Fig. 2 shows a typical cathode designed for use in a c.r.t. It consists of a small capped cylinder C of sheet nickel. The cap is coated with emitting material E (barium or strontium oxide) and the cathode is heated by a non-inductive insulated filament wire H inside the cylinder. The cathode is sometimes surrounded by a cylinder S which acts as a heat shield.

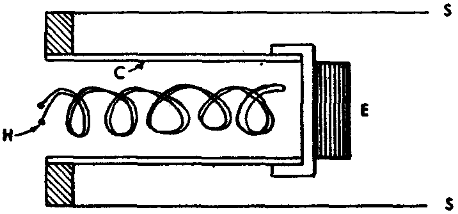


Fig. 2—TYPICAL C.R.T. CATHODE

6. **Modulator electrode.** This consists of a cylinder concentric with the cathode as shown in Fig. 3, one end having a small hole through which the electrons pass. A variable potential, negative with respect to the cathode, is applied to the modulator electrode and this has two functions:—

(*a*) It tends to concentrate the electrons leaving the cathode into a narrow beam, thus assisting in the focusing action.

(*b*) It controls the magnitude of the beam current (of the order of 30 to 300μA) and thus the brilliance of the spot on the screen. The potentiometer used to adjust the negative potential on the modulator electrode is termed the "*brillance*" control (also called "brightness" or "intensity" control). If the negative bias voltage is increased sufficiently, the beam current will reach its cut-off value and the spot on the screen will "black out". Some c.r.t.s use "intensity modulation" where the signal to be examined is applied to the modulator electrode to control the brilliance of the spot.

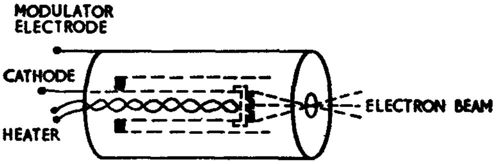


Fig. 3—MODULATOR ELECTRODE STRUCTURE

7. **Anode structure.** Fig. 4 shows the arrangement of a typical two-anode structure for a c.r.t. The second anode is normally held at a high positive potential with respect to the cathode, its function being to accelerate the electrons to a high velocity. The potential of the first anode is normally considerably lower and may be varied by the "focus" control. The resultant variation in the electrostatic field between the first and second anodes, and between the first anode and the modulator electrode gives a variation in the focus of the beam at the screen (see Para. 14). The two-anode arrangement suffers from the disadvantage that the focus and brilliance controls are *not* independent. The voltage on the first anode affects the beam current and hence the brilliance as well as the focus, while adjustment of the modulator electrode potential affects the focus as well as the beam current. This disadvantage is overcome in the three-anode structure.

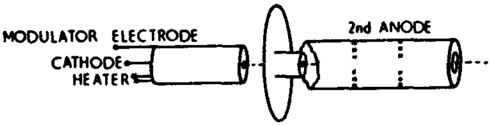


Fig. 4—TYPICAL TWO-ANODE STRUCTURE

8. Fig. 5 shows the arrangement of a typical three-anode structure. In this arrangement the first and third anode potentials are *fixed* and the focus is adjusted by altering

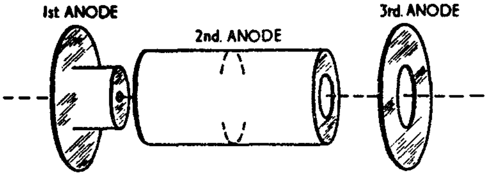


Fig. 5—TYPICAL THREE-ANODE STRUCTURE

the potential on the *second* anode. Owing to the screening effect of the first anode at a fixed potential, interaction of the focus and brilliance controls is practically negligible.

9. **Deflector plates.** These are the plates X_1 X_2 and Y_1 Y_2 mutually at right angles as shown in Fig. 1. Deflection is caused by applying suitable voltages to the respective pairs of plates (see Para. 16). The pair of plates nearer the final anode is always termed the Y plates irrespective of the orientation of the tube. But in general the Y plates deflect the beam *vertically*, while *horizontal* deflection is given by the X plates. The deflector plates are normally shaped as shown in Fig. 6, so that the beam does not strike them when the deflection is large.

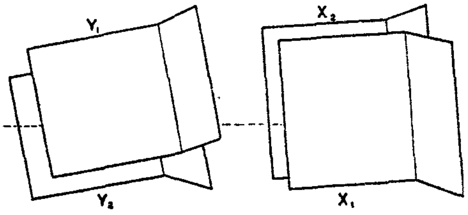


Fig. 6—DEFLECTOR PLATES

10. **The screen.** Many substances are fluorescent; that is, they have the property of emitting light when subjected to electron bombardment. However, the intensity and the colour of the light emitted varies from substance to substance. All fluorescent materials continue to emit light for some time after the electron bombardment has ceased; this is known as '*after-glow*', and its duration varies with different substances from a few micro-seconds to many seconds. In some applications, it is an advantage for a c.r.t. to have a long after-glow (e.g. plan position indicators in some radar systems). The material used for the screen is sprayed on to the inside of the tube face. Normally, a material emitting a high intensity of light has a short after-glow. To combine long after-glow with brightness, two fluorescent layers are used; the first gives an intense response of short duration; the second, a long after-glow. Suitable light filters may then be used to mask the instantaneous response.

11. Some modern c.r.t.s have a very thin film of alumium (4×10^{-6} inches) deposited over the inside surface of the fluorescent screen. The advantages obtained by 'aluminizing' the screen are:—

(*a*) Light that would otherwise be emitted by the screen backwards into the tube, is reflected by the aluminium film to increase the light output of the tube.

(*b*) Negative ions, released by electrons bombarding any gas atoms remaining inside the valve, are accelerated towards the screen along with the electrons. In view of their large mass in relation to that of an electron, considerable damage may be caused to the fluorescent material by ion bombardment and an '*ion burn*' results. This is a bare patch on the screen from which no light can be emitted. The aluminium film prevents this from occurring, since the ions cannot penetrate the film, although the electrons continue to pass through easily and reach the fluorescent screen.

12. **Aquadag.** This is a conductive graphite coating extending up inside the neck of the tube almost to the screen. It is connected to the final anode by a spring contact and attracts the secondary electrons emitted from the screen due to bombardment by the beam. The secondary electrons are then returned to the cathode via the power supply, so that a complete circuit for the beam current is established. If the aquadag were absent the screen would become negatively charged and would tend to repel the electron beam. The aquadag coating also acts as an electrostatic screen which shields the beam from stray electric fields. Cathode-ray tubes are often shielded from stray magnetic fields by an external mu-metal shield surrounding the whole tube. The constructional details of a typical three-anode electrostatic c.r.t. are shown in Fig. 7.

13. **Ion trap.** As noted in Para. 11(b), negative ions produced by collision between the primary electrons and gas atoms may cause considerable damage to the screen,

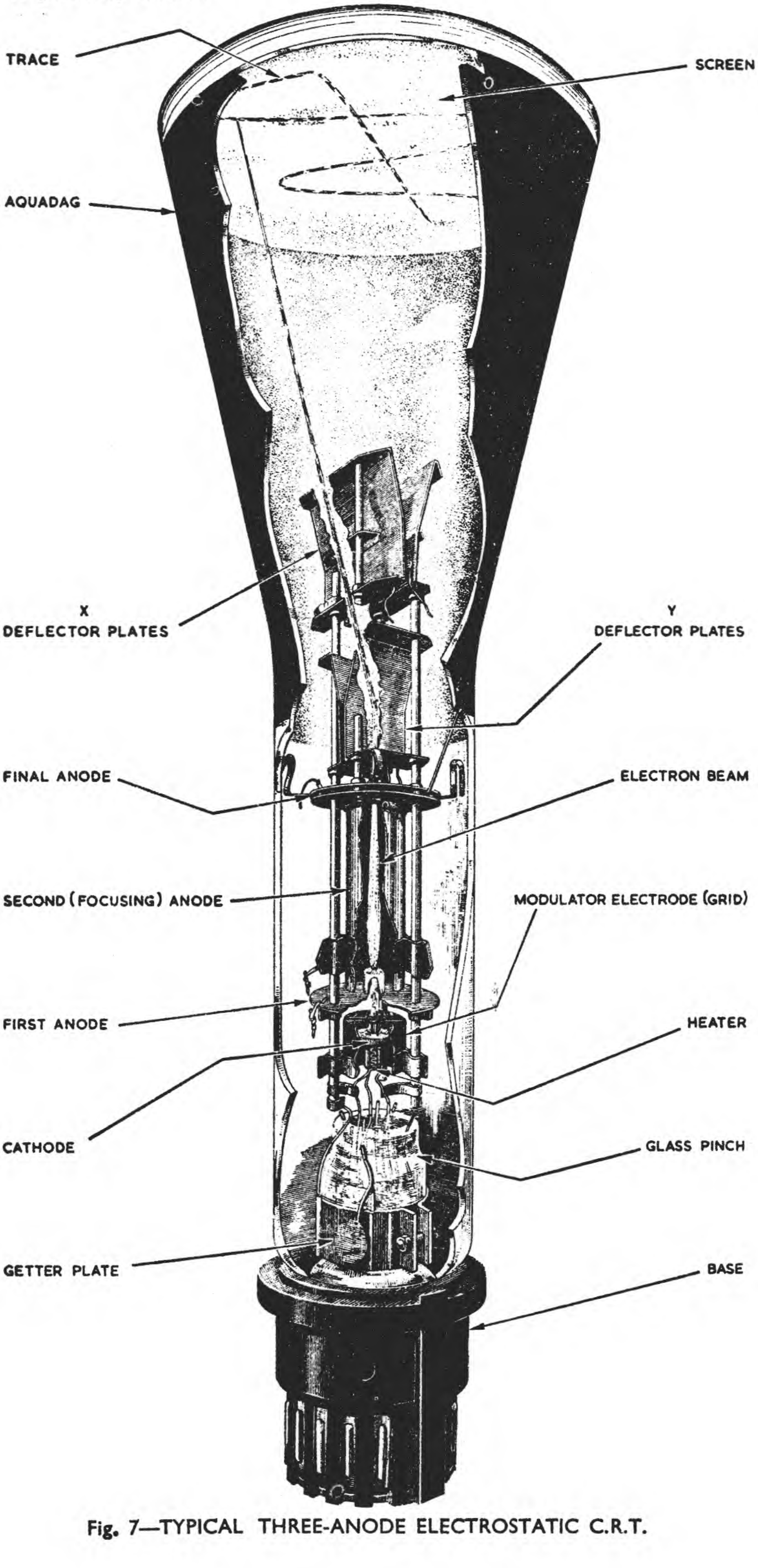


Fig. 7—TYPICAL THREE-ANODE ELECTROSTATIC C.R.T.

resulting in an ion burn. An aluminized screen will reduce the danger to a limited extent, but a much more effective method is the use of an ion trap. The electron gun is bent as shown in Fig. 8 and a small magnetic deflecting coil M is placed round the neck of the tube. The magnetic field produced by the current in this coil results in considerable deflection of the electrons which accelerate to strike the screen. The negative ions, of greater mass, are not deflected to the same extent and they strike the anodes A_2 and A_3 where they are neutralized. In some tubes a permanent magnet is used in lieu of the small deflecting coil M of Fig. 8.

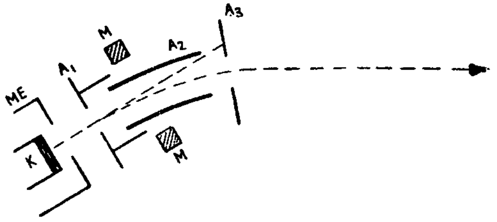


Fig. 8—SIMPLE ION TRAP

Electrostatic Focusing

14. In Fig. 9 the broken lines represent the lines of electric force between the ends of two cylindrical anodes A_1 and A_2, when A_2 is at a higher potential than A_1. An electron emitted from the centre of the cathode and accelerated along the path AB is unaffected by the electric field between the anodes. However, an electron which deviates from the central beam and moves along the path CD comes under the influence of the electric field. The electron at D is acted on by a force tending to move it *towards* the axis, and this *converging* force continues until the electron has crossed the plane XY midway between the two anodes, so that the electron follows the path CDE. On the right-hand side of the plane XY the force acting on the electron tends to move it *away* from the axis and the path *diverges* along EFG. The deflection actually produced depends upon the time spent by the electron in the electric field. Since the potential of A_2 is greater than that of A_1 the electron velocity is continually increasing, so that the time spent by the electron in the converging field is greater than that in the diverging field. Hence, the overall effect is to cause the beam to converge on the axis of the tube so condensing the beam. By suitable adjustment of the potentials of A_1 and A_2 the beam can be made to converge at a single point on the screen and correct focusing has been achieved.

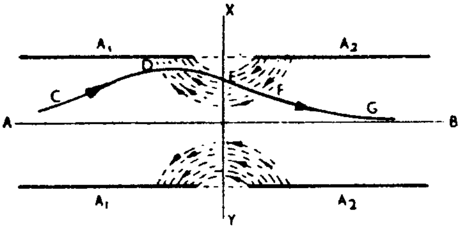


Fig. 9—TWO-ANODE ELECTROSTATIC FOCUSING

15. Fig. 10 shows the complete focusing system in a *three-anode* electrostatic tube. The electric field between the modulator electrode and the first anode is such that the electron beam converges and meets at a point, known as the '*cross-over*' point between these two electrodes. The beam is diverging as it enters the anode structure where focusing action occurs as explained in the preceding paragraph. The potential of the second anode is made variable so that the complex electric field in the anode structure may be adjusted to cause the beam to focus at the screen.

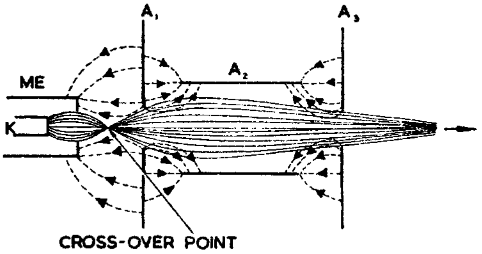


Fig. 10—THREE-ANODE ELECTROSTATIC FOCUSING

Electrostatic Deflection

16. In the electrostatic method of deflection, two pairs of plates are arranged as shown in Fig. 11. To produce *vertical*

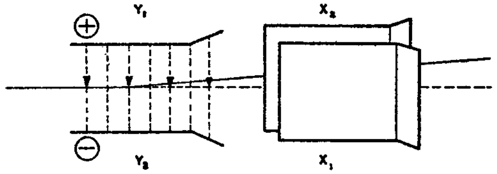


Fig. 11—ELECTROSTATIC DEFLECTION

deflection of the beam a p.d. is applied across the Y plates. The beam is then acted upon by the electric field between the plates so that it is attracted towards that plate which is positive and repelled from the negative plate. Thus, the spot of light on the screen will change its position in a vertical plane. Horizontal deflection is obtained by applying a voltage across the X plates.

17. From Fig. 12 it may be seen that the deflection of the spot on the screen *increases* as:—

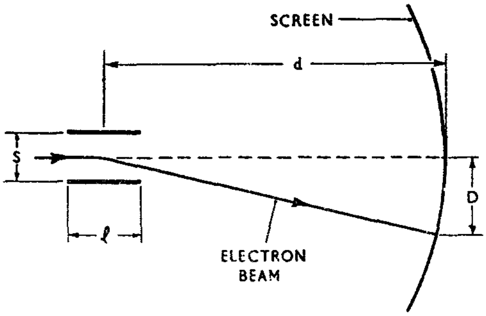


Fig. 12—FACTORS AFFECTING DEFLECTION SENSITIVITY

(*a*) the distance *d* between the midpoint of the deflector plates and the screen is *increased;*

(*b*) the length *l* of the deflector plates is *increased* (since the electrons then spend a longer time in the deflecting field);

(*c*) the spacing *s* between the plates is *decreased* (since the field strength between the plates is then increased);

(*d*) the deflection voltage V_d is *increased;*

(*e*) the potential V_a of the final anode is *decreased* (since the resultant decrease in electron velocity causes the electron to spend a longer time in the deflecting field).

The deflection D at the screen is, in fact, given by:—$D = \frac{d\ l\ V_d}{2\ s\ V_a}$ mm.—(1)

18. **Deflection sensitivity.** This is the deflection produced at the screen by a p.d. of one volt between the deflector plates for a given potential of the final anode. It is normally quoted as:—

$$\frac{x}{V_a} \text{ mm./deflection volt,}$$

where x is a factor depending on the tube, and V_a is the voltage of the final anode. For a typical c.r.t. x may be 730. For a final anode voltage of 2,000 V, the deflection sensitivity then is:—

$$\frac{730}{2{,}000} = 0.365 \text{ mm./deflection volt.}$$

Typical values range from 0.1 mm. to 0.5 mm./deflection volt.

19. Since the waveform to be examined is normally applied to the Y plates, these are further from the screen than the X plates. From equation (1) it is seen that the increase in *d* then gives a larger deflection at the screen for a given deflection voltage than that given by the X plates and this is the required condition.

Post-deflection Acceleration

20. To improve deflection sensitivity, tubes are sometimes used in which a large part of the acceleration of the electrons takes place *after* they have been deflected. The final anode is held at a relatively low potential (positive with respect to the cathode) so that deflection occurs when the electrons are travelling at a low velocity, and deflection sensitivity is thereby improved. The necessary acceleration is provided by using an additional *accelerator*. This is in the form of a narrow band of graphite placed round the inside of the tube between the screen and the main aquadag coating (Fig. 13). The post-deflection accelerator, as it is called, is connected to an external pin on the glass envelope and the required potential is applied to this pin. In a typical system, the cathode is at −1.5 kV, the final anode is at earth, and the post-deflection accelerator is at + 1.5 kV.

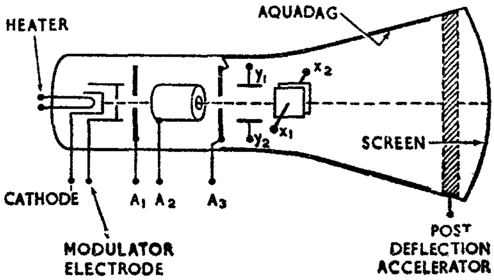


Fig. 13—POST-DEFLECTION ACCELERATOR

Double-beam Tubes

21. This is a c.r.t. of special construction in which the beam, after leaving the final anode A_3, impinges on the edge of a splitting plate C placed midway between the Y plates (Fig. 14). The beam is thus split into *two* sections, each of which may be deflected almost *independently* in the Y-plane by voltages applied between Y_1, and C and between Y_2 and C. Both beams are then equally affected by the X plates. This type of c.r.t. is of particular use in an '*oscilloscope*' when it is desired to compare two waveforms simultaneously. It is possible by the use of external circuits to produce a double trace on an ordinary c.r.t. The spot is switched rapidly from one trace to the other by a switching voltage applied to conventional plates, and the two signals to be examined are switched on to the traces in synchronism.

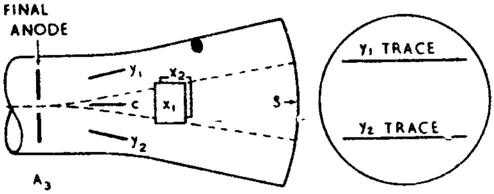


Fig. 14—DOUBLE-BEAM C.R.T.

Power Supplies

22. Fig. 15 shows a typical arrangement for applying the required potentials to the electrodes of an electrostatic c.r.t. The voltage supply is normally obtained from a half-wave rectifier circuit (see Sect. 9), and since the supply is greater than the usual h.t. supply to an equipment, it is termed '*extra high tension*' (e.h.t.). For oscilloscope work it is common practice to earth the final anode. Since the potential of the deflector plates is of the same order as that of the final anode, the external connections to the X and Y plates may then be made without danger of electric shock. These potentials are, of course, still positive with respect to the cathode, which in Fig. 15, is held at a potential of−1.48 kV.

Shift Voltages

23. Shift controls are provided to enable the spot to be moved anywhere on the screen. This is done by applying balanced shift voltages to the deflector plates by means of potentiometers which are inserted across a

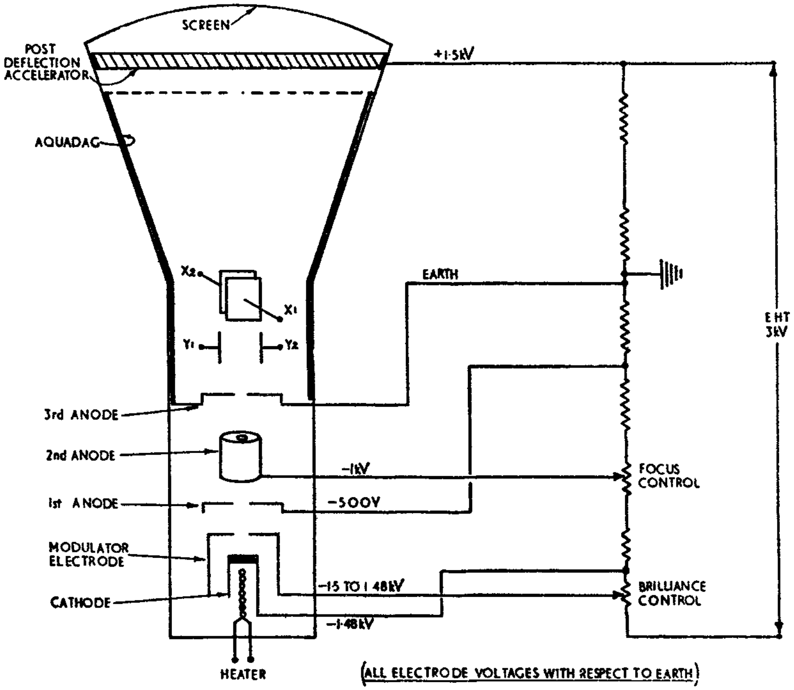


Fig. 15—C.R.T. E.H.T. NETWORK

section of the supply bleeder network as shown in Fig. 16. R_1 and R_2 are potentiometers of equal value, so ganged that when R_1 increases in value, R_2 decreases and *vice versa*. R_3 and R_4 are equal resistors of high value inserted in the bleeder network. Their junction is earthed to ensure that the shift voltages are balanced with respect to earth. The *mean* potential of the deflector plates is then the same as that of the final anode irrespective of shift. C_1 and C_2 are by-pass capacitors which ensure a steady d.c. shift voltage.

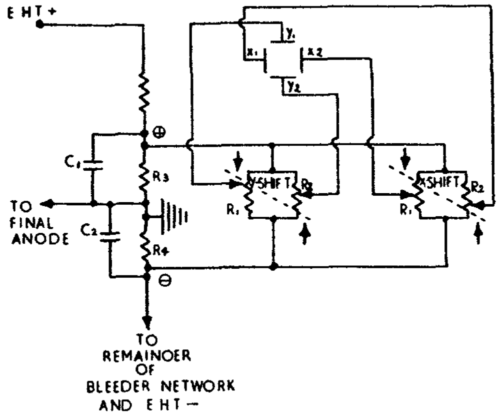


Fig. 16—SHIFT CONTROLS

Time-bases

24. To study voltage waveforms on a c.r.t. a uniformly-changing deflection voltage is applied to the X plates, while the waveform to be studied is applied to the Y plates. The voltage applied to the X plates should change linearly and at the end of the 'sweep' the original conditions (determined by the shift controls) should be quickly restored; that is, the 'fly-back' should be rapid. Such a variation, known as a *sawtooth* waveform, is shown in Fig. 17. The effect of applying this voltage to the X plates is to cause the electron spot on the screen to move horizontally at a constant speed and to fly back rapidly to its starting point, repeating the process at anything from a few c/s to several hundred kc/s. Provided the movement of the spot across the screen is sufficiently fast it appears as a continuous line of light which is variously described as the *time-base*, the *scan* or the *trace*. The apparatus for producing the necessary voltage or current for establishing the time-base is termed the *time-base generator*, circuits for which are discussed in Section 18 BK.3.(Radio Measurements). The voltage to be examined is applied to the Y plates so that vertical and horizontal deflection occur simultaneously. If the time-base frequency is *synchronized* with that of the signal (i.e., at the same frequency as, or an integral part of, the signal frequency) the signal waveform will appear to be stationary on the screen.

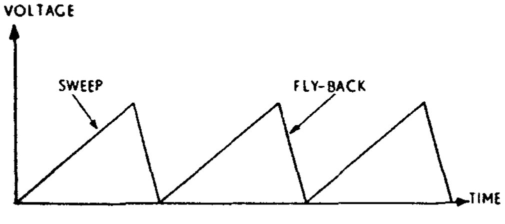


Fig. 17—SAWTOOTH VOLTAGE WAVEFORM

Distortion

25. Distortion in an electrostatic c.r.t. arises from mechanical or electrical causes, and may take several forms. The three main types of distortion are discussed in the following paragraphs.

26. **Trapezium distortion.** This form of distortion occurs when *unbalanced* (asymmetrical) deflection is used. The deflection produced by one pair of plates is affected by an alteration in the *mean* potential of the other. The mean potential of both pairs of plates should be the same as that of the final anode. If, however, X_1 is earthed (i.e. held at the final anode potential) and X_2 made positive with respect to it, the beam will be deflected to the right. At the same time the mean potential of the plates rises *above* that of the final anode. Thus, the velocity of the electrons increases so that the deflection produced by the Y plates for a given deflection voltage *decreases*, and a short trace CD results (Fig. 18). Conversely, if X_2 is made negative with respect to X_1, the decrease in electron velocity *increases* the Y plate deflection sensitivity, and a long trace EF results. Trapezium distortion is so named because of its effect on what would otherwise be a square or rectangular frame on the screen. The Y plates have a similar effect on the deflection sensitivity of the X plates. If symmetrical deflection voltages are applied, trapezium distortion will be reduced. For

example, instead of applying +10V on one plate of the pair with the other earthed, +5V may be applied to one and −5V to the other. The deflection will be the same, but the mean potential of the plates will be constant at that of the final anode. Other methods of minimizing trapezium distortion include:—

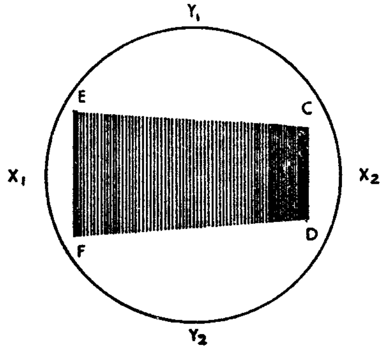


Fig. 18—TRAPEZIUM DISTORTION

(*a*) Curving the plates to produce counter-distortion (Fig. 19(a)).

(*b*) Cutting part of the X plates away and inserting an earthed screen S between the X and the Y plates. (Fig. 19(b)); the electric field is then such that trapezium distortion is neutralized.

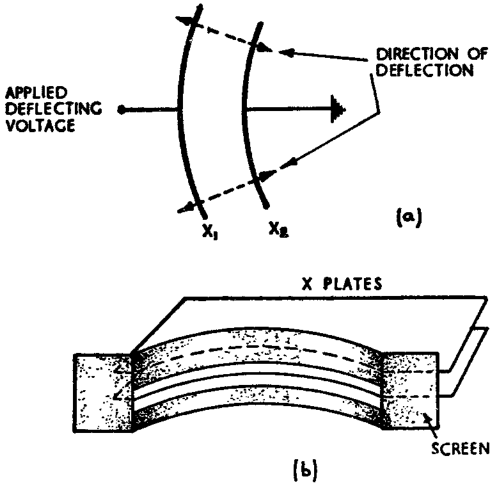


Fig. 19—COUNTERING TRAPEZIUM DISTORTION

27. **Deflection defocusing.** This form of distortion occurs when *asymmetrical* deflection is used. The velocity of the electrons in the region of the deflection system depends upon the p.d. between the deflection system and the cathode. If this p.d. varies, the variation in the velocity of the electrons will move the focal point of the beam at the screen. If asymmetrical voltages are applied to the deflector plates, the mean potential of the deflection system does *not* remain constant and defocusing may occur. Thus, if the focus is correctly adjusted at the centre of the screen when there is no deflection voltage, the application of a deflection voltage will cause the spot to move and at the same time alter the velocity of the electrons in the beam. The result is that the focus deteriorates as the deflection voltage and the distance from the centre of the screen increase, as shown in Fig. 20.

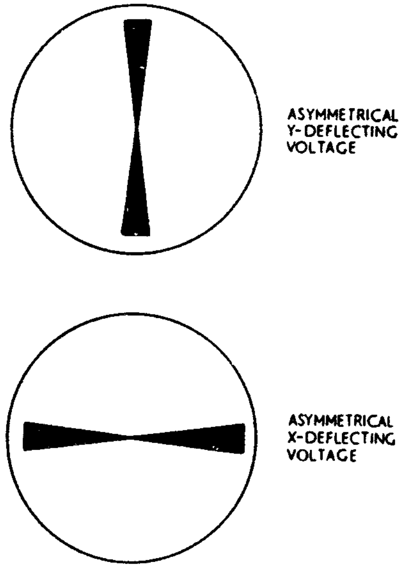


Fig. 20—DEFLECTION DEFOCUS

The term 'deflection defocus' is obvious from the figure. It may be avoided by the use of *symmetrical* deflection voltages.

28. **Astigmatism.** This is a defect in a c.r.t. which is the result of the production of an electron beam having different focal points in the horizontal and vertical planes, so that a well defined spot of light cannot be obtained. The most likely causes are interaction between the deflecting and focusing fields and misalignment of the electrode system. If the screen is placed at position 1 in Fig. 21, the display is a horizontal line as shown at (a). If placed at position 3, a vertical line results as at (c). At position 2, an ill-defined patch of light, known as the

'circle of least confusion' is produced as at (b). Alternatively if the position of the screen is fixed and the focus control of the tube is altered the display on the screen will pass from (a) through (b) to (c). Where a c.r.t. suffers from astigmatism it is usual to adjust the focus control to give (b). Astigmatism may be corrected by having the mean potential of each pair of deflector plates separately adjustable to a value different from that of the final anode by means of a '*stig*' control which must be independent of shift controls as shown in Fig. 22. The 'stig' controls are ganged so that one resistance increases in value as the other decreases. The p.d. across the shift network thus remains constant although its mean value about earth is altered. Such an arrangement introduces deliberate distortion in the electric field between the electrodes, thereby neutralizing astigmatism.

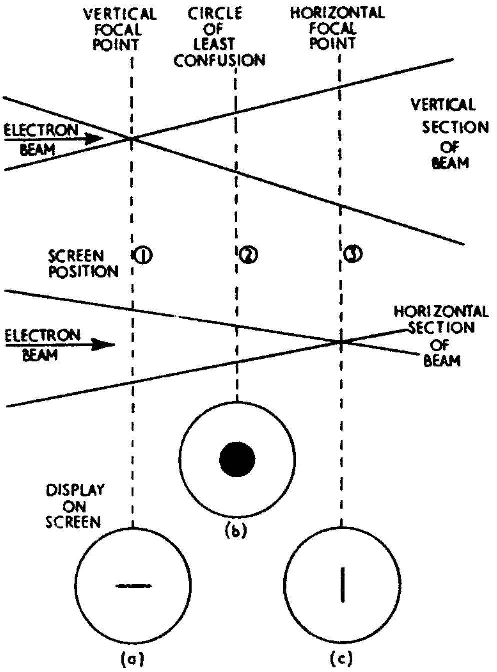


Fig. 21—ASTIGMATISM

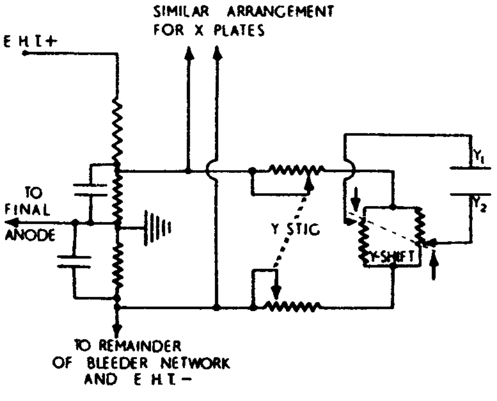


Fig. 22—'STIG' CONTROL

MAGNETIC C.R.T.

Construction

29. The magnetic c.r.t. differs from the electrostatic type in that the focusing and deflection of the beam are accomplished by means of magnetic fields and so the construction of the tube is simpler. The electron gun has one anode only, in some cases the anode being the aquadag coating. The neck of the tube is narrow since the focusing and deflector coils (which take the place of the focusing anode structure and the deflector plates respectively of an electrostatic tube) are mounted outside the tube round the neck. The arrangement is shown in Fig. 23.

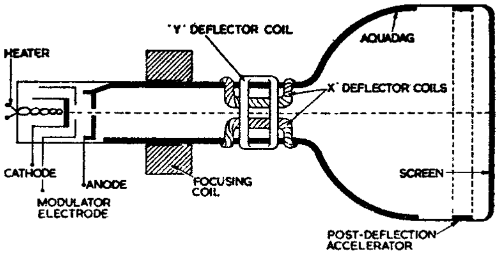


Fig. 23—MAGNETIC C.R.T.

Magnetic Focusing

30. This is carried out by a coil wound coaxial with the neck of the tube in the position shown approximately in Fig. 23. The coil is enclosed by a case of magnetic material which has an annular ring on the inside, and is so shaped that when direct current is passed through the coil, the resultant magnetic field has its main component *parallel* with the axis of the tube as shown in Fig. 24(a). Consequently, only those electrons which are deviating from the axis will be affected by the field.

31. Electron *a* in Fig. 24(a) passes through a field which is at all times parallel to its path PQ and it is therefore unaffected by

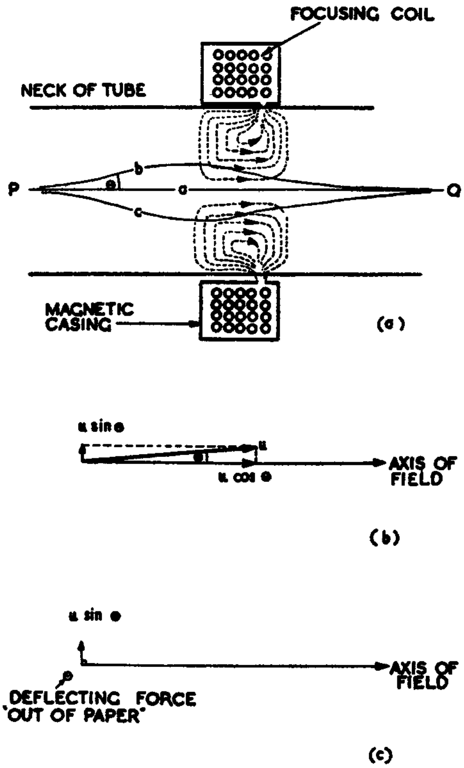


Fig. 24—MAGNETIC FOCUSING

the field. Electron *b*, travelling at velocity u, enters the magnetic field at an angle θ to the axis. By vector resolution, this electron can be shown to have two components of velocity; one, u cos θ parallel to the axis, and the other, u sin θ at right angles to the axis as shown in Fig. 24(b). The magnetic field has no effect on the component u cos θ parallel to its axis but by Fleming's left-hand rule (and remembering that a moving electron is equivalent to a current in the *reverse* direction), it is seen that a force acts on the electron at right angles to both the component u sin θ and the axis of the field (Fig. 24(c)). Thus electron *b* in Fig. 24(a) is deflected initially in a direction 'out of the paper' and given a *radial* motion. At the same time it has a high forward velocity u cos θ. The radial motion of the electron in conjunction with the axial velocity causes a combination of forces to act on the electron in such a manner that it is given a converging *spiral* path towards the axis. Similarly, electron *c* is given a radial motion in a direction initially 'into the paper' and follows a similar path towards the axis.

32. In practice, only a short focus coil is used and it is arranged that the electrons, after describing a portion of a spiral within the magnetic field, leave the field moving inwards towards the axis again. The angle at which each electron converges is proportional to the angle at which it was originally diverging. Consequently all the electrons tend to come together at one point on the axis. By correctly adjusting the magnitude of the current in the focus coil and also the position of the coil on the neck of the tube, the focal point may be made to lie on the screen, giving a sharp spot of light. Fig. 25(a) shows the paths taken by the electrons, looking from the screen towards the cathode, when the focus is correctly adjusted. Fig. 25(b) shows the condition when the focusing field is too weak, and Fig. 25(c) when the field is too strong.

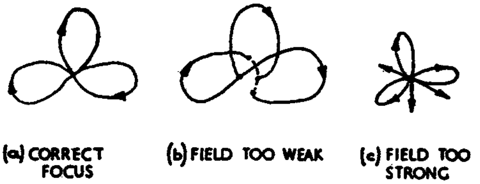


Fig. 25—EFFECT OF VARYING FOCUSING FIELD

Magnetic Deflection

33. Fig. 26 shows two coils connected in series and placed between the focus coil and the screen with their axes perpendicular to the axis of the tube. A current passed through these coils will produce a magnetic field across the neck of the tube

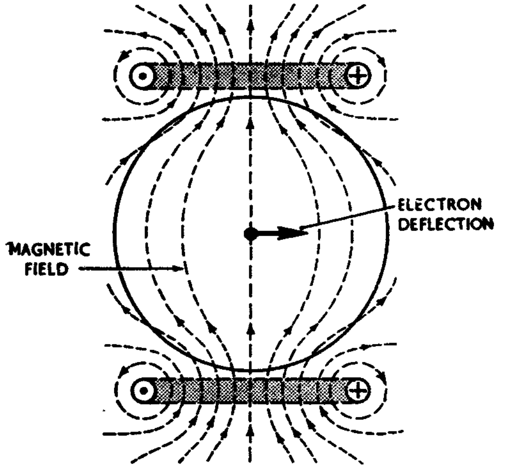


Fig. 26—MAGNETIC DEFLECTION

as shown. A beam of electrons travelling up the tube in a direction 'out of the paper' will, by Fleming's left-hand rule, be deflected to the *right* by an amount proportional to the intensity of the field. By reversing the direction of the deflection current in the coils, the direction of deflection is reversed. Thus, if any *current* waveform is passed through the coils the deflection of the spot will follow this waveform.

34. The coils used in practice to produce the deflecting magnetic field are generally wound on a rectangular former and then bent round the neck of the tube so that the horizontal sides of the coils almost touch. In this way a more uniform magnetic field is obtained. Fig. 27 shows a pair of coils used for horizontal (X) deflection. The coils for vertical deflection will be similar, placed on top of the X coils and at an angle of 90° to them.

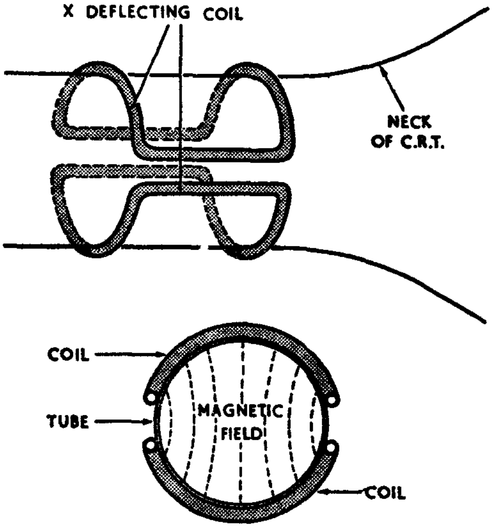


Fig. 27—DEFLECTOR COILS

35. From Fig. 28 it may be seen that, with magnetic deflection, the deflection of the spot on the screen *increases* as:—

(*a*) the distance *d* between the midpoint of the deflector coils and the screen is *increased*;

(*b*) the length *l* of the deflector coils is *increased* (since the electrons will then spend a longer time in the deflecting field);

(*c*) the value of the magnetic flux density B is *increased*, by increasing the current in the coils;

(*d*) the final anode voltage V_a is *decreased* (since the resultant decrease in electron velocity causes the electrons to spend a longer time in the deflecting field).

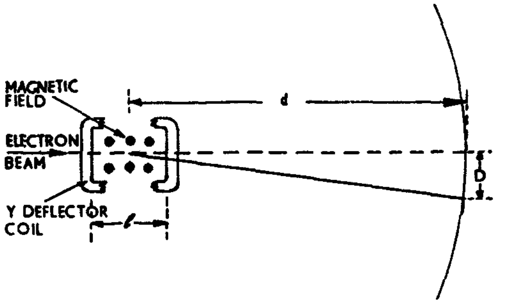


Fig. 28—DEFLECTION SENSITIVITY

The deflection D at the screen, is in fact, given by:—

$$D = \frac{d\,l\,B}{3\cdot37\sqrt{V_a}}\ \text{mm.} \quad \ldots \quad \ldots \qquad (2)$$

The *deflection sensitivity* of a magnetic tube is the deflection produced at the screen by a current of 1 mA in the deflector coils.

Comparison of Magnetic and Electrostatic Tubes

36. Advantages of magnetic c.r.t.

(*a*) The construction is simpler than that of an electrostatic tube.

(*b*) The beam current may be higher, resulting in a brighter spot on the screen, since:—

(i) the cross-sectional area of the beam and (hence the number of electrons in the beam) is not confined by anode structures or deflector plates as in an electrostatic tube:

(ii) the deflection sensitivity is inversely proportional to $\sqrt{V_a}$ and not V_a as in an electrostatic tube (compare equations (1) and (2)). Thus much higher anode voltages may be applied without loss of sensitivity so that the display on the screen will be brighter.

(*c*) Magnetic deflection is usually more suitable for radial deflection and polar representations on the screen. The coils may be made to rotate about the axis of the tube where the frequency of rotation required is not too high.

37. **Disadvantages of magnetic c.r.t.**

(*a*) Considerable power is dissipated in the resistance of the focusing and deflector coils.

(*b*) Because of the inductance of the deflector coils it is more difficult for associated valve circuits to produce the necessary *current* waveforms for magnetic deflection than it is to produce voltage waveforms for electrostatic deflection.

(*c*) The practical upper frequency limit for magnetic deflection is of the order of 10 kc/s. Electrostatic deflection may be used up to frequencies as high as several hundred kc/s.

Electrostatic Focusing with Magnetic Deflection

38. This system uses a tube which resembles an electrostatic c.r.t., but it has a deflector coil assembly mounted round the neck between the focusing anodes and the screen instead of deflector plates. With this system, advantages (b) and (c) are retained, while disadvantage (a) is reduced.

Distortion in Magnetic Tubes

39. The magnetic c.r.t. does not suffer from trapezium distortion or deflection defocus, and astigmatism can usually be avoided by correctly adjusting the position of the focus coil. Distortion may arise however, if the deflecting magnetic field is not uniform over the area through which the beam passes. This necessitates careful coil design.

Shift Controls in Magnetic Tubes

40. These are provided to enable the spot to be moved anywhere on the screen. It is done by passing a direct current through the deflector coils from a shift voltage supply as shown in Fig. 29. The current is varied by the shift control, and since a similar arrangement applies to the other pair of deflector coils, the requirement for shift is met.

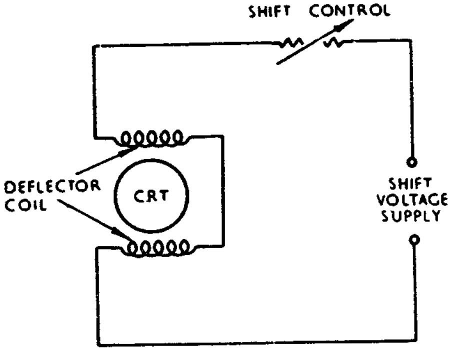


Fig. 29—MAGNETIC C.R.T. SHIFT CONTROL

Time-bases for Magnetic Tubes

41. As for the electrostatic c.r.t., the magnetic type may be required to give a horizontal trace, known as the time-base. Para. 23 shows that to provide this time-base on an electrostatic tube a sawtooth voltage waveform is applied to the X deflector plates. Since in a magnetic tube, the deflecting field is proportional to the *current* in the deflector coils, a sawtooth *current* waveform is applied to the X deflector coils. The methods used for producing such current waveforms are dealt with in Part 3 (Radar).

SECTION 8

CHAPTER 6

PHOTO-ELECTRIC DEVICES

PHOTO-ELECTRIC DEVICES

Introduction

1. Photo-electric devices or *photocells* are devices the electrical properties of which undergo a change when they are exposed to light. There are three types of photocell, which differ according to the effect that light has on their properties:—

(*a*) **Photo-emissive cell.** Light causes the *emission* of electrons from the prepared surface of a cathode.

(*b*) **Photo-conductive cell.** Light alters the *resistance* of the material on which it falls.

(*c*) **Photo-voltaic cell.** Light produces an *e.m.f.* across the cell.

Vacuum Photo-emissive Cells

2. **Construction.** Fig. 1(a) shows the symbol for a photo-emissive cell, while (b) shows a method of construction used for a typical cell. In (b), the cathode consists of a semi-cylindrical metal plate the concave surface of which is coated with the photo-emissive material. The anode, which must occupy little space if it is not to obstruct the light falling on the cathode, is a thin vertical rod mounted at the centre as shown.

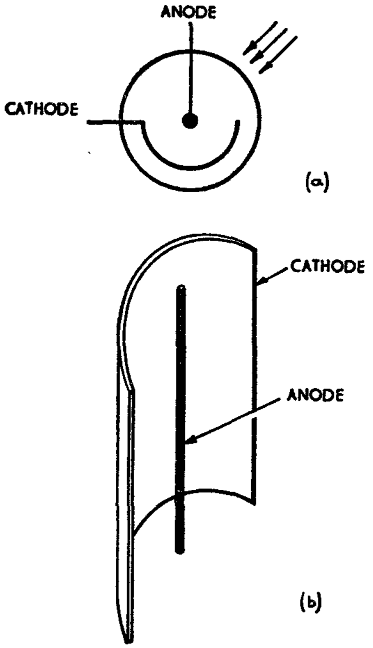


Fig. 1—PHOTO-EMISSIVE CELL

3. **Photo-emissive materials.** Light is electromagnetic in nature and the frequency or wavelength of the light determines its *colour*. The wavelength of light is measured in *Angstrom units* (Å), where 1Å = 10^{-8} cm. The range of light wavelengths over which a photocell gives maximum response is determined by the material of which the cathode is constructed. Fig. 2 compares the response of typical materials to different wavelengths, the *intensity* of the light remaining constant. The caesium-antimony type of cathode has its maximum sensitivity in the visible portion of the spectrum and is most sensitive to blue light. The caesium-silver cathode has maximum response in the near infra-red portion of the spectrum, while the calcium-sulphide cathode covers the entire visible spectrum and extends into the near infra-red. The type of cathode used in the photocell depends on the wavelength of the light at which it is required to be used.

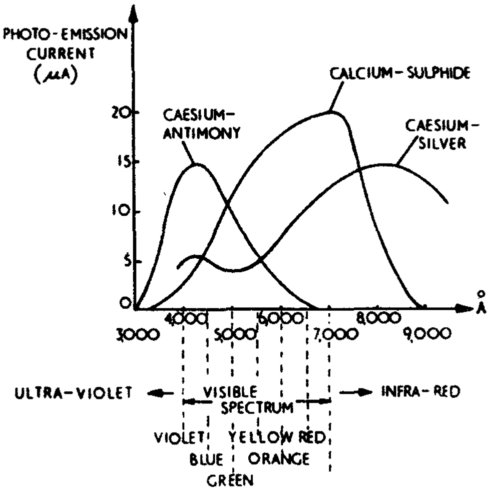


Fig. 2—RESPONSES OF PHOTO-EMISSIVE CATHODES

4. **Action.** When a photo-emissive cell is exposed to light of sufficient intensity and of a wavelength suitable for the cathode material used, the energy imparted to the electrons by the light is sufficient to enable them to overcome the surface barrier at the cathode and electron emission is achieved. By making the anode positive with respect to the cathode the emitted electrons are drawn towards the anode and a photo-electric current is established. The characteristic curve showing the variation of the anode current I_a with the anode voltage V_a is as shown in Fig. 3; in this graph, both the intensity and the wavelength of the light are assumed *constant*. At a small positive value of anode voltage (generally around 20V) all the electrons being emitted by the cathode reach the anode and the valve saturates.

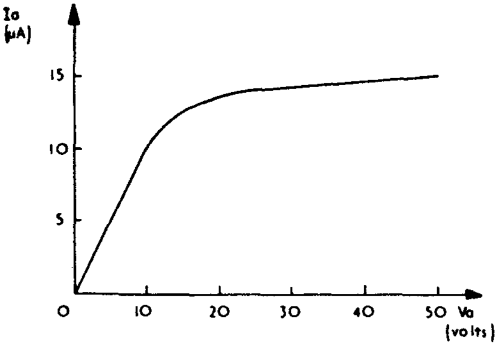


Fig. 3—PHOTOCELL CHARACTERISTIC

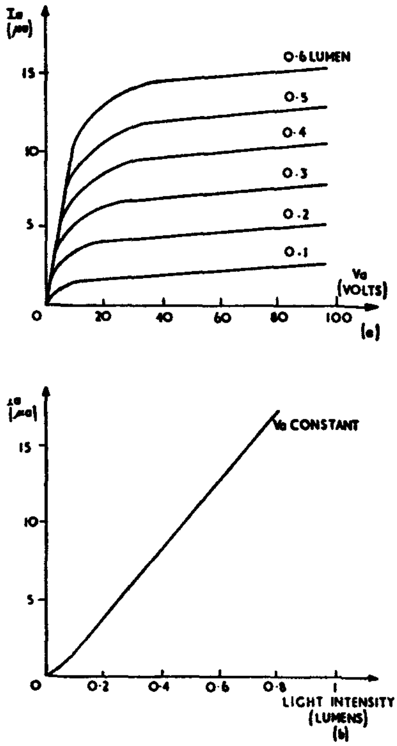


Fig. 4—VARIATION OF PHOTO-CURRENT WITH LIGHT INTENSITY

5. The saturation current of a photocell is a measure of the light intensity. Intensity is measured in '*lumens*', where one lumen is the light energy falling on one square inch of surface at a distance of one inch from a standard candle. Fig. 4(a) shows how the saturation current increases with the intensity of the light. Fig. 4(b) shows the relationship between the anode current and the light intensity for a constant (saturation) value of anode voltage. This gives the condition under which the photo-emissive cell is normally used. That is, the anode voltage is maintained at a constant value of, say, 20V and the anode current is then proportional to the light intensity. A basic *photometer* may be obtained by arranging to measure this current. Photo-emissive cells are also used to control the operation of relays in certain radio equipment. Consider the circuit shown in Fig. 5. The photocell, when not illuminated, is in effect an open circuit so that no current is established in R_1. The grid of the triode is thus held at h.t. negative potential and the cathode potential is adjusted so that the bias between grid and cathode is sufficient to cut off the triode. No current flows through the triode and the relay is, therefore, de-energised. When the photocell is illuminated, the resultant photo-current establishes a voltage drop in R_1, raising the poten-

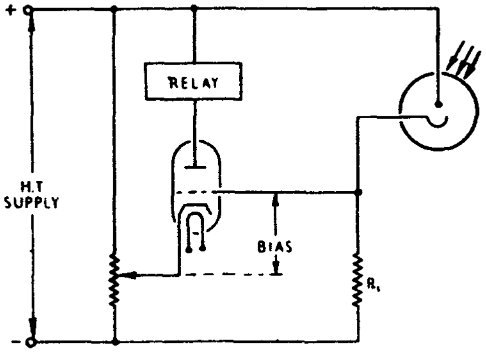


Fig. 5—USE OF PHOTOCELL TO OPERATE RELAY

tial, of the grid and causing the triode to cut on, energising the relay.

Gas-filled Photo-emissive Cells

6. The maximum photo-electric current obtainable from a vacuum photocell is of the order of $50\mu A$/lumen. For many purposes this small current is inconvenient. To increase the sensitivity, cells are made which contain a small quantity of argon gas. The voltage between anode and cathode is then adjusted so that the photo-electrons, emitted by the action of the light, produce ionization by collision with the gas atoms. The total current arriving at the anode may then be many times greater than the primary photo-electric current. The value of the current for a given light intensity is, however, influenced by the anode potential to a far greater extent than that in the vacuum photocell. This is shown in Fig. 6. When the anode voltage exceeds the ionization potential of the gas, the current increases rapidly, and an amplication factor of 10 may be obtained by using the cell at a sufficiently high voltage. There is a limit to the amplification which may be obtained, for if the voltage is increased sufficiently a self-sustained cold-cathode discharge occurs and the current is then independent of the light intensity. Therefore in practice, voltages between 90V and 200V are generally used in gas-filled photocells. In addition the upper frequency limit is of the order of 10 kc/s because of the inertia of the gas ions.

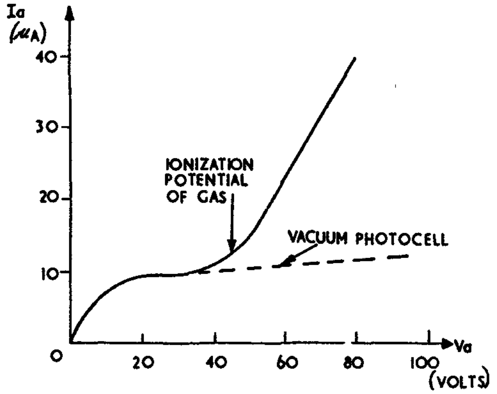


Fig. 6—GAS-FILLED PHOTOCELL CHARACTERISTIC

Photo-multiplier Cells

7. An alternative method of directly amplifying the photo-electric emission from a cathode is based upon the secondary emission effect, which permits amplification of the electron current within the cell itself by electronic multiplication. Such a device is known as a *photo-multiplier.*

8. When primary electrons strike a target, secondary electrons may be liberated. The average number of secondary electrons released by each primary electron is called the *secondary emission coefficient* and may be as high as 4 or 5. It depends on:—

(*a*) the energy of the primary electrons;

(*b*) the material of which the target (or '*dynode*') is constructed; a dynode is an electrode, the chief function of which is to emit secondary electrons;

(*c*) the angle of incidence of the primary electrons.

9. The principles involved in the operation of a basic photo-multiplier are illustrated in Fig. 7. The dynodes A, B, C and D, and the anode E, are at successively higher potentials, positive with respect to the cathode. The primary electron stream is directed from the cathode to A, where secondary emission occurs. The increased electron stream from A is then accelerated towards B, where further secondary emission takes place. The process is repeated between B and C, and between C and D, the electron stream increasing at each stage. Finally, the secondary electrons produced at D are collected by the anode E to establish the anode current. The amplification of such an arrangement is high. If the secondary emission coefficient is K and the number of dynodes n, the overall amplification is given by K^n.

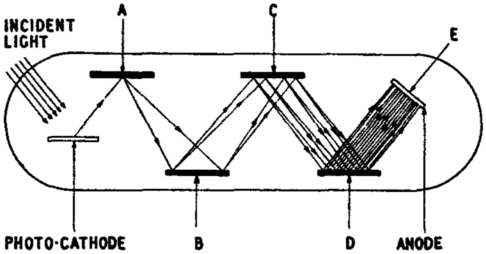


Fig. 7.—PRINCIPLE OF PHOTO-MULTIPLIER

10. The simplified arrangement shown in Fig. 7 would be of little practical use since the dynodes must not only have a high

secondary emission coefficient, but some means must be provided to *focus* the electron stream on to each dynode in turn. A variety of methods, employing magnetic and electrostatic focusing are in general use. Fig. 8 shows a typical nine-stage photo-multiplier with electrostatic focusing, sometimes known as a 'zig-zag' multiplier. Focusing of the electron stream is achieved by a geometrical arrangement of specially shaped dynodes, their potentials increasing in equal steps above the cathode potential. The effect of the electrode shape and the resultant electrostatic field is to guide the electrons emitted from the photo-cathode 0 to the first dynode 1, so that they impinge at the optimum angle for maximum secondary emission. The electrons emitted from dynode 1 are then guided to dynode 2, the process being repeated in a zig-zag manner until the electrons reach the anode 10. This system gives a gain of 100,000 using a final anode-cathode potential of 1,000V.

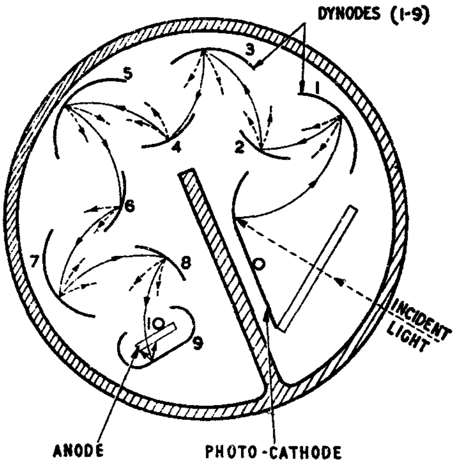


Fig. 8—CONSTRUCTION OF PHOTO-MULTIPLIER

Photo-conductive Cells

11. When light falls on certain semi-conductors (see Chap. 7) electrons are released from their orbits inside the material because of the additional energy imparted to them and the number of 'free' electrons available is increased, thus reducing the resistance of the surface layer. This effect is negligible in conductor materials because of the very large number of free electrons already available. The photo-conductive effect may be used to detect and measure light by embodying the semi-conductor in a suitable circuit.

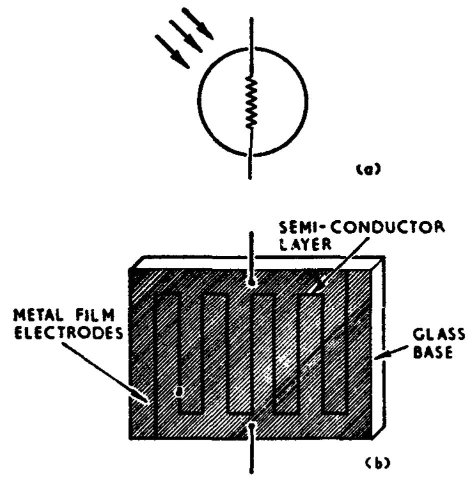


Fig. 9—PHOTO-CONDUCTIVE CELL

12. Fig. 9(a) shows the symbol for a photo-conductive cell. The elements of a typical cell are shown in Fig. 9(b). A thin coating of a chemically inactive metal (e.g., gold) is deposited on a glass base and is divided into two separate electrodes by etching a narrow gap through the metal film to the glass. A thin layer of semi-conductor material is then deposited in the gap. A transparent envelope is generally used to enclose the cell, so that light falling on the envelope will alter the resistance of the semi-conductor layer. Various semi-conductors are used, depending on the requirement. They include selenium, which has a peak response at about 7,000Å and is therefore sensitive to the red end of the visible spectrum; and compounds of lead and thallium with sulphur, selenium and tellurium. These compounds are sensitive to various regions of the infra-red band and are finding increasing use in radio equipment. For instance, infra-red detectors are used in guided weapons to detect and home on to a target.

Photo-voltaic Cells

13. This class of photocell is a completely solid arrangement of various layers of metal and semi-conductor materials and forms a battery, activated by light. The symbol for a photo-voltaic cell is shown in Fig. 10(a), and Fig. 10(b) shows the

elements of a typical cell. A base plate A carries a thin layer of semi-conductor material E which forms one electrode of the cell. The semi-conductor is separated from a conductor layer C (the counter-electrode) by a barrier layer B, and connection is made to C by a metal grid, Because of the transparency of the arrangement some of the light falling on C is transmitted through B to the surface of the semi-conductor where it is responsible for the release of photo-electrons. These electrons pass through the barrier B and accumulate on the counter-electrode C, giving it a negative potential with respect to the semi-conductor. A current is then established when the base-plate A and the grid of the counter-electrode C are connected through an external circuit. In the photo-voltaic effect, radiation incident on the cell causes an e.m.f. to be set up at the terminals of the cell.

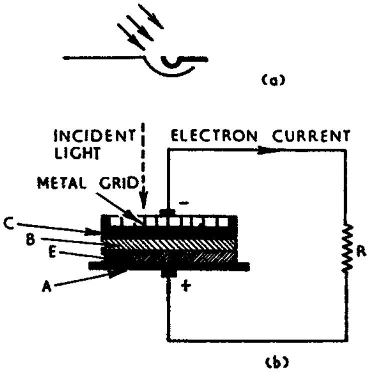


Fig. 10—PHOTO-VOLTAIC CELL

14. Various semi-conductor materials may be used, including selenium. A typical modern cell uses silicon with boron as the conducting layer. Such a cell is sensitive to light in the visible spectrum and its applications include photometry and instrument power supplies. One such application is the provision of the necessary power supplies for the radio equipment in earth satellites by photo-voltaic cells, activated by sunlight.

SECTION 8

CHAPTER 7

SEMI-CONDUCTOR DEVICES

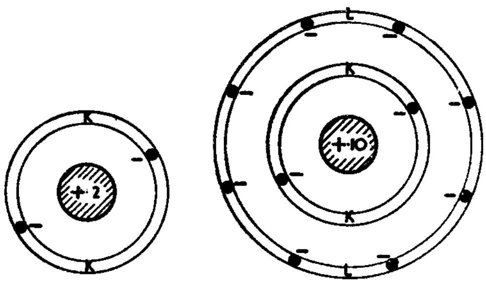
SEMI-CONDUCTOR DEVICES

Introduction

1. The term 'semi-conductor' denotes solid material having a resistivity considerably less than that of good insulator and much more than that of a good conductor. The resistivity of a semi-conductor is about 1 ohm per cm. cube compared to about 10^6 ohms per cm. cube for an insulator and about 10^{-6} ohm per cm. cube for a conductor. In addition, the resistivity of a semi-conductor *decreases* with rise of temperature in contrast with normal metallic conductors.

Atomic Structure

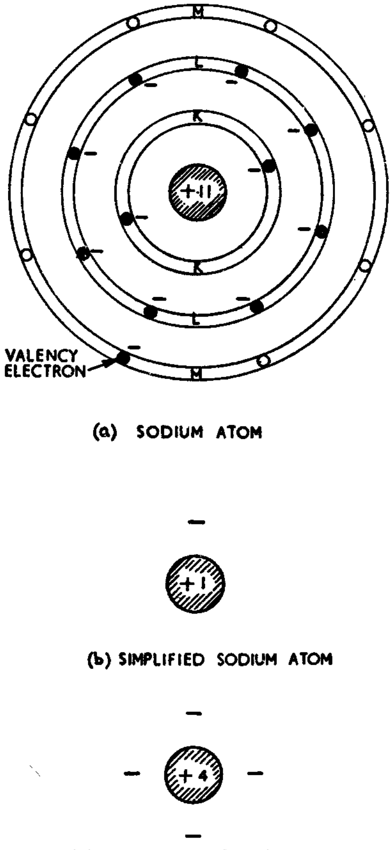
2. The behaviour of a material, chemically and electrically, depends on the disposition of the electrons in relation to the nucleus of the atom of the material. The electrons rotate round the nucleus in a definite and orderly manner. They can exist only in certain '*shells*' round the nucleus, and the number of electrons that each shell is capable of accommodating is fixed. Each shell is 'filled' in order of atomic number, commencing at the innermost shell. Working outwards from the nucleus, the first (K) shell may contain one or two electrons but not more than two, so this shell is filled when the atom is helium which has two electrons (Fig. 1(a)). The second (L) shell, further away from the nucleus, may contain anything up to eight electrons and the first *two* shells are completely filled by the atom of atomic number 10, which is neon (Fig. 1(b)). An atom which has just sufficient orbital electrons to fill completely one or more shells has no 'free' electrons and is completely *inert* (e.g. the inert gases helium and neon). An inert gas is one which shows practically no tendency to combine with other elements. Further shells, of greater radii and having accommodation for a fixed number of electrons, exist for elements of high atomic number.



(a) HELIUM ATOM (b) NEON ATOM

Fig. 1—ATOM SHELL STRUCTURE

3. For the great majority of elements, the number of electrons in the outermost shell is less than the maximum number that this shell is capable of accommodating. For example, the sodium atom (atomic number 11) has the K and L shells filled and has one 'odd' electron in the M shell (Fig. 2(a)). The electron in the M shell is free to enter into external relationships with other atoms



(a) SODIUM ATOM

(b) SIMPLIFIED SODIUM ATOM

(c) SIMPLIFIED GERMANIUM OR SILICON ATOM

Fig. 2—SIMPLIFIED ATOM DIAGRAMS

and because of this the sodium atom is extremely active. The electrons in an incomplete shell that tend to enter into external relationships are termed '*valency electrons*'; thus sodium has one valency electron. For many purposes the atom diagrams can be simplified by combining the filled shells with the nucleus and showing only the net electric charge inside a circle. Each of the electrons left over (the valency electrons) is represented by a minus sign outside the circle, the atom as a whole being electrically neutral. Fig. 2(b) replaces the complete sodium atom diagram of Fig. 2(a) in this notation. The semi-conductors germanium and silicon may be represented as in Fig. 2(c), from which it is seen that there are *four* valency electrons.

'Pure' Semi-conductors

4. The most important group of semi-conductors includes germanium and silicon, each atom of which has four valency electrons in its outer shell. In this Chapter, attention will be almost exclusively directed to germanium, although the action in silicon and other semi-conductors is similar. The atoms of chemically pure germanium are arranged in a crystal lattice as shown in Fig. 3. Each of the four valency electrons is 'shared' by a valency electron in an adjacent atom. The link between two atoms via the shared pair of electrons is called a *co-valent bond*, and in germanium all the valency electrons are so occupied. It is this co-valent bond created by electron pairs which holds the atoms together in the crystal. In such a configuration the only 'free' electrons available are those released by the partial breakdown of the lattice due to the application of thermal or photo-electric energy. Owing to the lattice construction, therefore, the conductivity is very low under normal conditions.

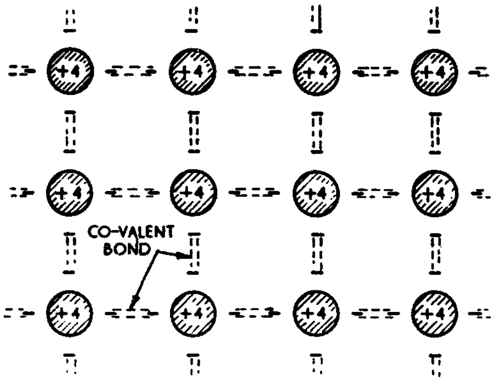


Fig. 3—GERMANIUM CRYSTAL LATTICE

5. If however, the temperature of the germanium is raised or if it is subjected to irradiation, some electrons receive sufficient energy to enable them to break away from the lattice. The release of an electron in this way leaves a '*hole*' in the otherwise regular array of electrons. Little energy is required for a valency electron in a neighbouring atom to jump into this vacant hole, and the hole is now transferred to the adjacent atom where it may be filled by yet another electron. In this way the hole *diffuses* through the material. Thus, in chemically pure germanium, the release of an electron from its parent nucleus by thermal or photo-electric energy increases the conductivity of the material not only by the creation of such free electrons, but by the creation of 'positive holes'. The increase in conductivity in this way is said to be due to '*intrinsic semi-conduction*' where electrons and holes are created in *equal numbers*. By applying a potential difference across the material under such conditions electrons drift towards the positive terminal causing holes to move towards the negative terminal. An increase in energy results in more electron-hole pairs being created and the conductivity increases. Reducing the energy to its original level causes the free electrons and holes to combine and thus disappear.

6. As the temperature is increased, more and more electrons receive sufficient energy to enable them to break away from the lattice. At temperatures above about 100°C this effect becomes so great that current control is not possible, the heating due to increasing current ionising still more atoms and allowing a still larger current to flow. This effect is cumulative and eventually results in the destruction of the crystal lattice. It is therefore necessary to maintain germanium at comparatively low temperatures.

N-type Germanium

7. By the introduction of a controlled amount of 'impurity', a higher conductivity can be obtained while at the same time preserving the lattice configuration. For

instance, if an atom containing *five* valency electrons (e.g. antimony or arsenic) is introduced into the pure germanium lattice, one free electron becomes available as a charge carrier. In the process of bonding with the germanium, since only four valency electrons are needed, one of the electrons of the impurity atom is liberated as shown in Fig. 4. Even at room temperatures this surplus electron has sufficient energy to leave the lattice and to migrate through the germanium as a negative charge carrier. The impurity atom will exhibit a positive charge, having lost one of its constituent electrons, and such an atom is called a '*donor*' since it donates one electron to the current. Electron-hole pairs may still be produced by intrinsic action as stated in Para. 5, but the introduction of impurity atoms of the type described causes the free electrons to be in the majority and this type of semi-conductor is termed *n-type* since the majority carriers (electrons) are negative (n) particles. For practical purposes, n-type germanium can be considered to be constructed of immobile impurity *positive* atoms called donors which provide a number of *negative* charge carriers or electrons. This construction is shown in Fig. 5. The material as a whole is electrically neutral.

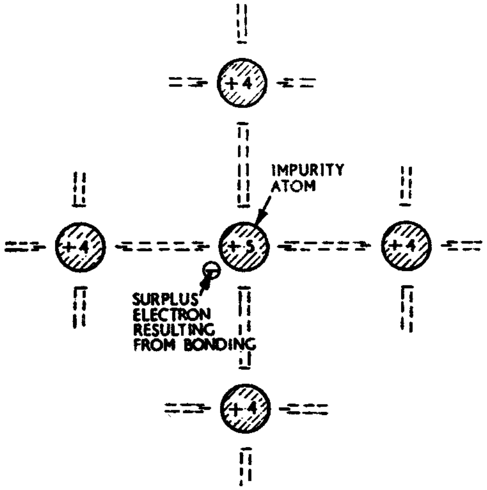


Fig. 4—N-TYPE GERMANIUM

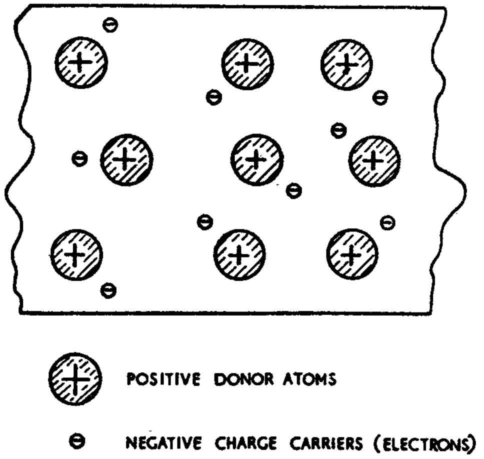


Fig. 5—DONORS AND ELECTRONS IN N-TYPE GERMANIUM

P-type Germanium

8. By introducing into the pure germanium lattice impurity atoms having *three* valency electrons (e.g. aluminium or indium) a different type of semi-conductor is obtained. Since such an atom possesses only three electrons available for bonding, a *hole* will be left within the lattice as shown in Fig. 6. This hole exerts an attractive force on, and will 'capture', any electron which, having been liberated from a germanium atom by thermal energy, happens to pass near the hole. This results in the formation of another hole which, in turn, is filled by yet another free electron. This process is continuous throughout the germanium so constituting a 'diffusion' of *positive* charge carriers. The impurity atom, which loses a hole in this way, gains an electron and so exhibits a *negative* charge. Impurity atoms of this type are thus termed

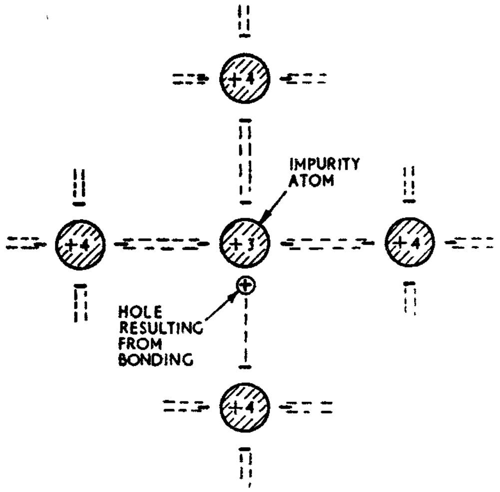


Fig. 6—P-TYPE GERMANIUM

'*acceptor*' atoms since they accept an electron to neutralise the hole in the lattice. Electron-hole pairs may still be produced by intrinsic action as stated in Para. 5, but the introduction of impurity atoms of the type described causes the holes to be in the majority, and this type of semi-conductor is termed *p-type* since the majority carriers (holes) are effectively positive (p) particles. For practical purposes, p-type germanium can be considered to be constructed of immobile impurity *negative* atoms called acceptors which provide a number of *positive* charge carriers or holes. This construction is shown in Fig. 7. The material as a whole is electrically neutral.

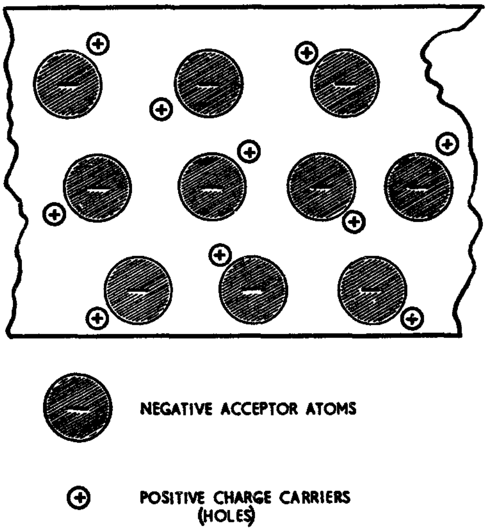


Fig. 7—ACCEPTORS AND HOLES IN P-TYPE GERMANIUM

Metal-to-Semi-conductor Junctions

9. When a semi-conductor is placed in contact with a metal, a barrier layer is set up at the contact and this layer has a higher resistance in one direction than in the other. The asymmetrical resistance means that current can flow much easier in one direction through the layer than in the other, and such a device may be used for *rectification.*

10. When a p-type semi-conductor is in contact with a metal plate, as in Fig. 8, electrons migrate from the metal to fill the positive holes in the semi-conductor, a process which continues until the transference of charge has led to a p.d. sufficient to stop it. A narrow strip at the junction has then been cleared of positive holes and so has no means of conduction; it is thus a very good insulator for very small applied p.d.s and is termed a *barrier layer*. However, when a p.d. in excess of about 1 volt is applied, the barrier layer is subject to infiltration from one side or both. If the p.d. is such that the semi-conductor is *positive* with respect to the metal, positive holes will migrate from the body of the semi-conductor into the layer, thereby restoring its conductivity. If the p.d. is reversed, more electrons are drawn from the metal simply to fill more positive holes, and the barrier layer is reinforced. Thus, the 'forward' resistance of the layer is low if the semi-conductor is positive with respect to the metal, and the 'back' resistance is high if the direction of the p.d. is reversed. Since the layer is very thin, a p.d. of several volts applied in the high resistance direction suffices to break down the insulation of the layer.

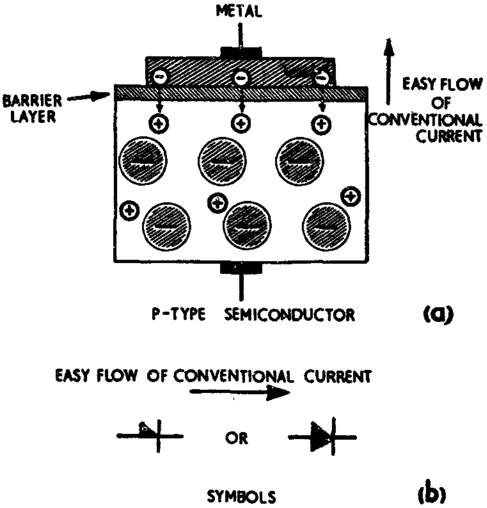


Fig. 8—METAL-TO-SEMI-CONDUCTOR JUNCTION

11. For the boundary of a p-type semi-conductor and a metal, the forward direction of easy flow of conventional current is *towards* the metal, and the current flows in the reverse direction only when the 'back voltage' exceeds a certain value. Similar behaviour, with the operating directions reversed, is observed for an n-type semi-conductor such as germanium with antimony

as added impurity. The forward direction of easy flow of conventional current in this case is *away* from the metal.

12. **Plate-type rectifiers.** Rectifiers using the principles described in Para. 10 have been constructed and are usually referred to as '*metal rectifiers*'. Fig. 9 shows rectifier elements using (a) cuprous oxide and (b) selenium, as the semi-conductors. The elements are thin flat discs with central axial holes so that several can be mounted on a spindle. In the copper-oxide rectifier the barrier layer is formed between the cuprous oxide and the copper plate; in the selenium rectifier the junction is between a layer of tin alloy and the selenium. In both cases the semi-conductors are p-type, so the direction of easy conventional current flow is from the semi-conductor to the metal. A graph relating current and voltage in a selenium rectifier element is shown in Fig. 9(c). Metal rectifiers are considered in greater detail in Sect. 9, Chap. 3.

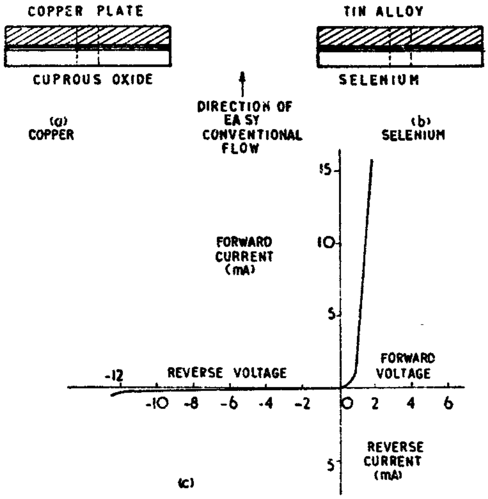


Fig. 9—PLATE TYPE METAL RECTIFIERS

13. **Point contact rectifiers.** Fig. 10 shows a germanium rectifier, in which a small piece of n-type germanium A, with its outer surface polished and etched, is held in contact with a tungsten '*cat's whisker*' B. The easy direction of flow of conventional current at the tungsten-germanium contact is from B to A, and the arrangement will pass appreciable current in this direction, and only in this direction, if an alternating voltage of peak value less than the back voltage for the contact is applied across the rectifier. The purpose of the point contact is to concentrate the electric field near the surface of the germanium to ensure as large a current flow as possible. Silicon is also used in rectifiers of this type. One great advantage of the point contact rectifier is its very small self-capacitance, which means that it can be used at very high frequencies; in this respect it is superior to the diode valve. In addition, no heater voltage is required. The point contact crystal diode has extremely small dimensions and is easily wired into a circuit with such components as resistors and capacitors. The relative size of the component is illustrated in Fig. 11. Point contact crystal diodes are considered in greater detail in Part 3 (Radar).

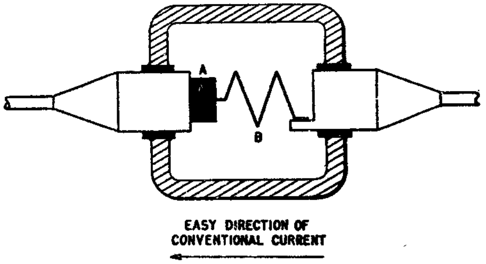


Fig. 10—POINT CONTACT GERMANIUM RECTIFIER

Fig. 11—DIMENSIONS OF POINT CONTACT RECTIFIER

P-N Junctions

14. Consider a piece of germanium consisting of a p-type zone and an n-type zone, with a defined line of demarcation between them as depicted in Fig. 12(a). On one side of this line, acceptors and holes are in the majority (p-type), and on the other side donors and electrons are in the majority (n-type). Initially, both the n-type and p-type are electrically neutral as previously explained. However, when the two types of material meet at a junction, electrons diffuse from the n-type to the p-type material,

and holes diffuse from the p-type to the n-type material. As a result of these movements, the n-type material becomes *positively* charged, and the p-type material becomes *negatively* charged as shown in Fig. 12(b). This prevents any further movement of charge carriers across the junction. This effect may be considered as the building up of a 'potential barrier' across the junction. The mobile current carriers (electrons and holes) are repelled from the vicinity of the junction leaving only the fixed acceptor atoms on one side and the donors on the other side.

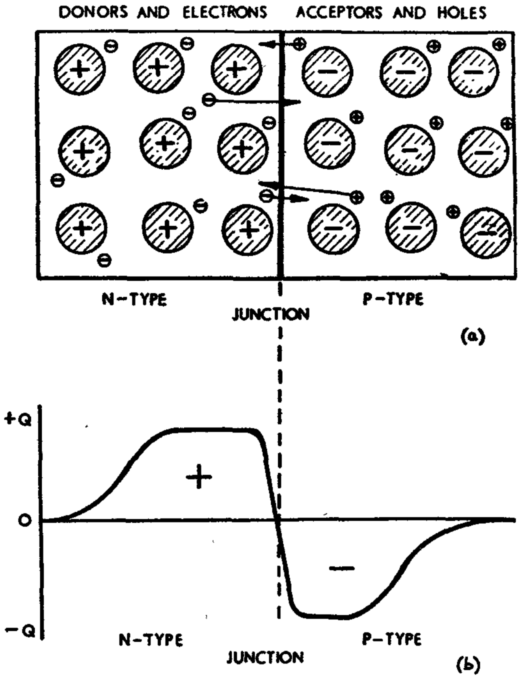


Fig. 12—P-N JUNCTION

15. Suppose now that a battery is connected across the p-n junction in such a sense that the barrier potential is reduced as shown in Fig. 13. The junction is then said to be '*forward biased*' and it is easier for the holes in the p-type to cross to the n-type. Similarly, the electrons in the n-type cross easily to the p-type. The result is that several milliamperes of current can flow. A further increase in forward bias removes the reverse potential gradient at the junction entirely and a rapid increase in current results.

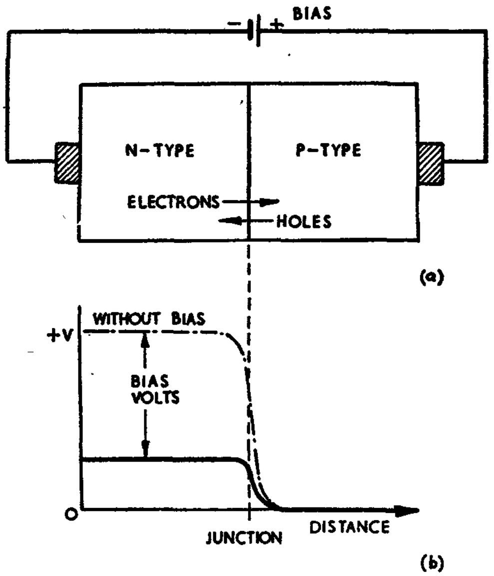


Fig. 13—FORWARD-BIASED JUNCTION

16. If however, the battery is connected across the junction in the opposite polarity so that the barrier potential is increased, as shown in Fig. 14, the reverse potential gradient at the junction is reinforced and very little current flows. The junction is said to be '*reverse or back biased*'. At a critical value of reverse voltage the co-valent

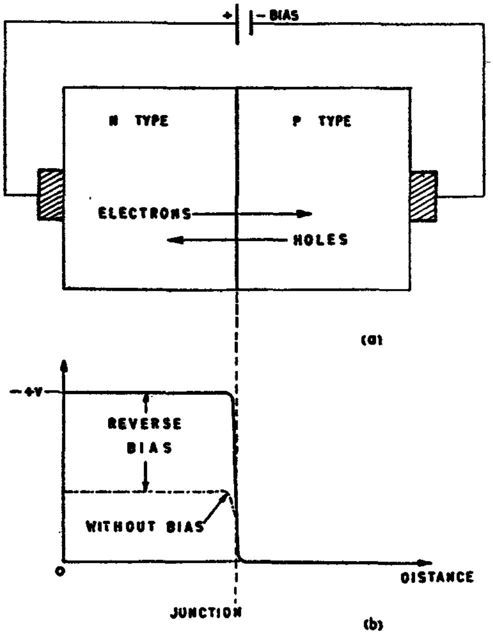


Fig. 14—REVERSE-BIASED JUNCTION

bonds in the semi-conductor break up and a high reverse current flows.

17. If an alternating voltage of peak value less than the critical reverse voltage is applied to a p-n junction diode, as in Fig. 15(a), the potential barrier will be alternately strengthened and weakened; that is the resistance of the junction will vary with the value and polarity of the applied voltage. If the current is plotted against various values of voltage, both positive and negative, it will be seen that the p-n junction acts in a manner similar to that of a metal rectifier. A typical characteristic curve is shown in Fig. 15(b).

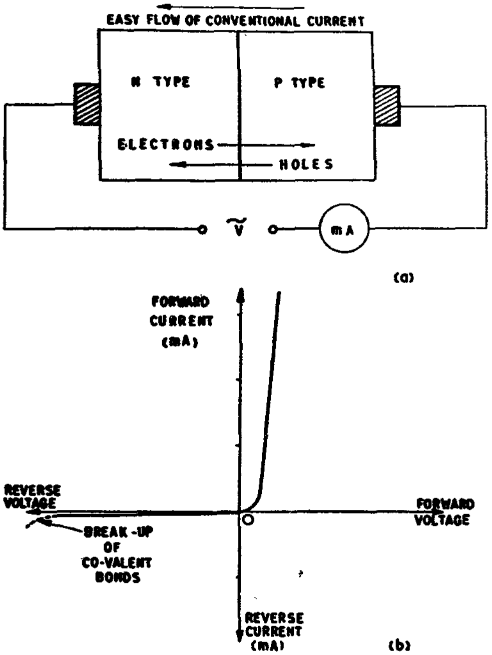


Fig. 15—JUNCTION DIODE CHARACTERISTICS

18. Semi-conductor junction diodes of the germanium and silicon type are now used for a wide variety of purposes including high power rectification. They compare favourably with thermionic diode valves since:—

(*a*) They are robust and will withstand severe vibration.

(*b*) No heater voltage is required.

(*c*) They are relatively small in size.

Point Contact Transistors

19. The transistor is a more complex semi-conductor device than the crystal diode, as it involves two contacts or two junctions. It may be considered analogous to the thermionic triode valve, and it can be employed for many purposes including amplification.

20. Fig. 16(a) shows the functional diagram of a type of transistor known as the *point contact* type. It consists of a small pellet of n-type or p-type germanium having three external connections. These connections are named the *collector*, the *emitter* and the *base*. The collector and emitter connections are cat's whisker point contacts and the base is a soldered contact. The transistor symbol is shown in Fig. 16(b).

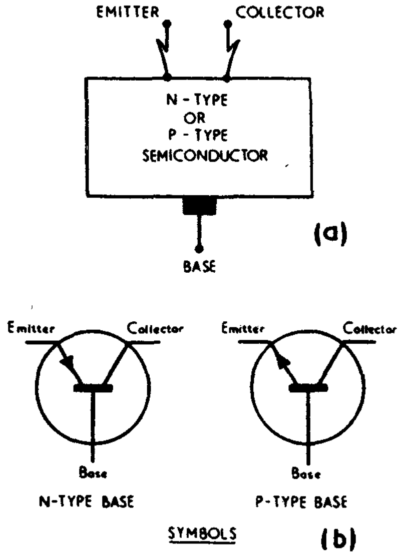


Fig. 16—POINT CONTACT TRANSISTOR

21. To start the transistor action it is necessary to ensure that all the connections are biased correctly as follows:—

(*a*) The emitter is biased in the *forward* direction with respect to the base. For an n-type base the emitter is made *positive* with respect to the base, the polarity being reversed for a p-type base.

(*b*) The collector is biased in the *reverse* direction with respect to the base. For an n-type base the collector is made *negative* with respect to the base, the polarity being reversed for a p-type base.

22. In a point contact transistor employing n-type germanium as the base material (Fig. 17), when the emitter bias is applied in the forward direction, the emitter point contact causes a field of sufficiently high intensity for the valency electrons to break their co-valent bonds and flow out through the emitter giving rise to an emitter current. This action causes holes to be formed in the base, a process that is called 'hole injection'. These holes will migrate towards the collector under the influence of the electric field, since the collector is biased negatively, thereby establishing a collector current. In a correctly designed point contact transistor, an injected current at the emitter causes an *amplified* version to flow in the collector. Further, a change of current in the emitter circuit causes a corresponding change of current in the collector circuit. This is known as transistor action and in a correctly designed and adjusted circuit it can be used to produce amplification.

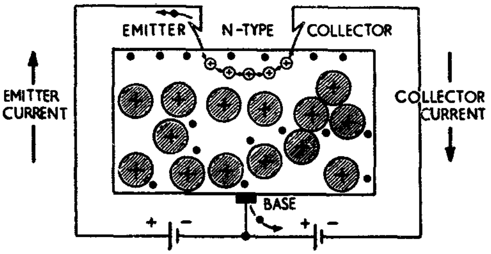
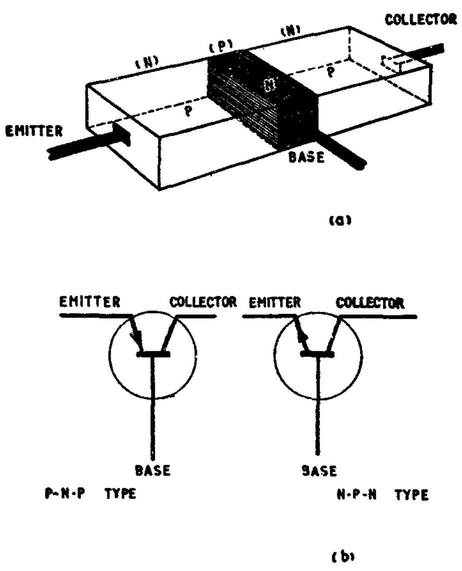


Fig. 17—POINT CONTACT TRANSISTOR ACTION

Junction Transistors

23. The junction transistor has virtually replaced the point contact type. The junction transistor consists of three regions of germanium arranged in sandwich form; that is, p-n-p or n-p-n as shown in Fig. 18. There are three external connections; one is soldered to the emitter, one to the base and one to the collector, each connection being placed in one of the impurity regions with the base in the middle. Whether of the p-n-p or the n-p-n type, the emitter-base junction is biased in the *forward* direction and the collector-base junction is biased in the *reverse* direction.



SYMBOLS

Fig. 18—JUNCTION TRANSISTOR

24. A p-n-p junction transistor is depicted in Fig. 19(a). Under conditions of thermal equilibrium, with no external voltages applied to the transistor, the potentials of the p and n layers are different as explained in Para. 14 and as shown in Fig. 19(b). If now the junctions are biased so that the emitter-base junction has forward bias (emitter positive with respect to base) and the collector-base junction has reverse bias (collector negative with respect to base) the potential levels are altered as shown in Fig. 19(c).

25. When a positive potential is applied to the emitter terminal, the holes in the p material are repelled from the region of the emitter and drift towards the p-n junction under the influence of the applied field. The junction potential barrier has been lowered by the forward bias and therefore the holes are able to cross to the n region (base). While passing through the base region a few holes will re-combine with the electrons present there, and it is therefore important that the width of the n region is as small as possible (about 0·05mm.) Many of the holes, however, continue to drift towards the collector region and are finally able to cross the n-p junction since the barrier potential is in favour of hole conduction from base to collector (although not in the reverse direction). Holes which cross into the collector p region represent the collector current since each hole on reaching the collector requires one electron

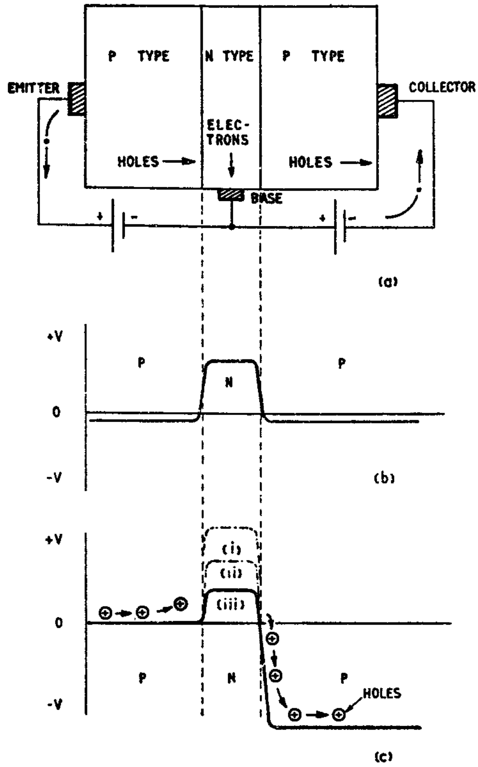


Fig. 19—JUNCTION TRANSISTOR ACTION

from the battery to neutralize it. The base current is small compared with the emitter current, most of which goes to the collector. Thus, a change of base potential as at (i), (ii) and (iii) in Fig. 19(c) varies the diffusion current of holes from the emitter to the base (i,e., the emitter current is varied) thereby causing proportionate changes in the collector current.

26. The general form of the collector current-collector voltage characteristic of the junction transistor is shown in Fig. 20. The section OA is linear, but above A a given increase in the collector voltage causes a much smaller increase in the collector current. This phenomenon is a 'saturation' effect due to the fact that at a voltage corresponding to A, practically all the holes made available as charge carriers (p-n-p) are reaching the collector. Beyond this point, therefore, the collector resistance rises steeply as indicated by the slope of the graph. The transistor is normally operated beyond the knee at A to make use of this high resistance effect.

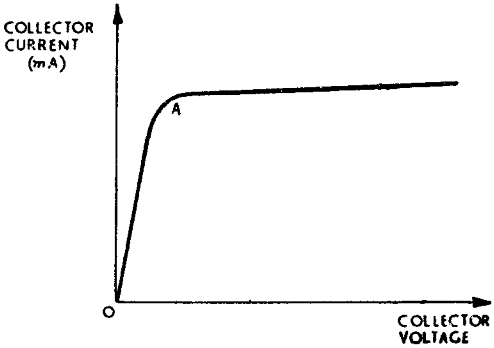


Fig. 20—JUNCTION TRANSISTOR CHARACTERISTIC

27. Not all the holes which represent the emitter current will be received in the collector circuit for some will re-combine with electrons in the base as explained in Para. 25. Hence, the current gain of a junction transistor is less than unity and is usually of the order of 0·98. However, since the collector resistance is so high, considerable voltage and power gains may be achieved. Typical values of emitter-to-base and collector-to-base resistances are 500 Ω and 1 MΩ respectively. This represents a resistance 'gain' of 2,000, a voltage gain of about 1,960 and a power gain of about 1,900. The frequency response of a junction transistor is limited because of the relatively low velocity of electrons and holes within the germanium. A typical junction transistor is shown in Fig. 21.

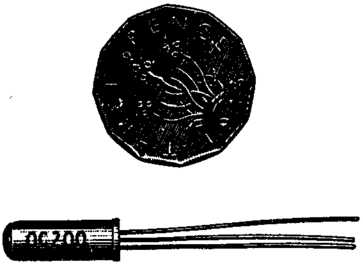


Fig. 21—TYPICAL JUNCTION TRANSISTOR

28. The n-p-n junction transistor is similar to the p-n-p type except that the polarities of the applied voltages are reversed, as shown in Fig. 22, and that it is electrons and not holes which are 'injected' at the emitter and collected at the collector. The amplifying action may be summed up generally by imagining the base layer to be a barrier

separating the two different levels of emitter and collector, the height of this barrier being varied by the signal. The holes (in p-n-p type) or electrons (in n-p-n type) from the emitter 'spill over' this barrier to reach the collector at a rate determined by the height of the barrier. The signal, therefore, in controlling the height of the barrier, effectively controls the collector current, and in so doing consumes only a fraction of the power which can be dissipated by the collector circuit. Thus there is a power gain and amplification is then possible.

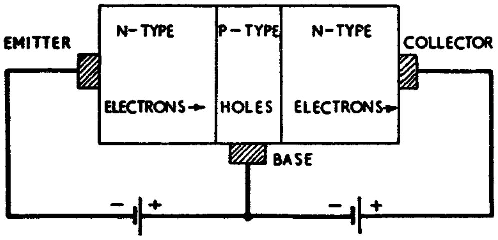
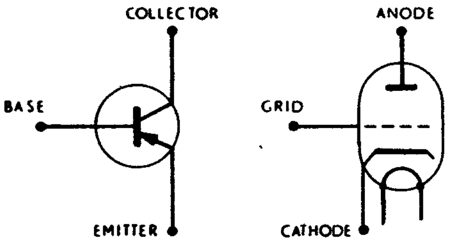


Fig. 22—N-P-N JUNCTION TRANSISTOR

Comparison of Transistors with Valves

29. The emitter, base and collector of a transistor are the equivalents of the cathode, grid and anode respectively of a triode valve as shown in Fig. 23, and any form of valve circuit may be used in transistor circuits provided suitable adjustment is made to power supplies. The power that a transistor is capable of handling however, is limited by the heat produced during operation, and at the present stage of development the maximum power handling capacity for a single transistor is of the order of a few watts.



(a) N-TYPE BASE TRANSISTOR (b) INDIRECTLY-HEATED TRIODE

Fig. 23—COMPARISON OF TRANSISTORS WITH VALVES

30. **Advantages of transistors.**

(*a*) Very small and light.

(*b*) Robust and reliable.

(*c*) High efficiency.

(*d*) Very long life.

(*e*) No heater supply required.

(*f*) The supply voltage required is small.

31. **Limitations of transistors.**

(*a*) Power handling capacity is limited to a few watts.

(*b*) Frequency response is limited to under 100 Mc/s.

(*c*) Noise factor is higher than that in valves.

The Transistor as an Amplifier

32. The transistor, like the valve, has many applications. One application which is now in use in various electronic devices is the transistor amplifier. The circuit diagrams of devices using transistors bear a marked resemblance to conventional valve circuits. For instance, a basic triode amplifier circuit is illustrated in Fig. 24(a). This circuit, which is discussed in Chap. 2, Para. 24, has its transistor 'equivalent' in Fig. 24(b). In the latter, R_b acts as a self-bias resistor to the base of the transistor in the following manner. The base of the transistor draws a small amount of current from the supply. This current causes a potential to be developed across the bias resistor, thus maintaining the base of the transistor at a small positive potential with respect to the collector. By the introduction of R_b into the circuit, one source only of e.m.f. (a 4·5 volt battery) is necessary. With such a circuit the amplification obtained is similar to that obtained from the triode amplifier, but the power requirements are very much smaller for the transistor circuit.

Other Semi-conductor Devices

33. **Photodiode.** This is the semi-conductor equivalent of the photocell discussed in Chap. 6. Two types are used:—

(*a*) *The p-n junction type.* This consists of a piece of germanium with two distinct regions, one p-type and the other n-type. The whole is enclosed in an insulated container, light being allowed to fall on the p-n junction. As shown in Fig. 25(a) the cell is biased in the reverse direction and with no light incident at the junction, the current is practically negligible. When light strikes the junction however, hole-electron pairs are formed at this junction by intrinsic action as

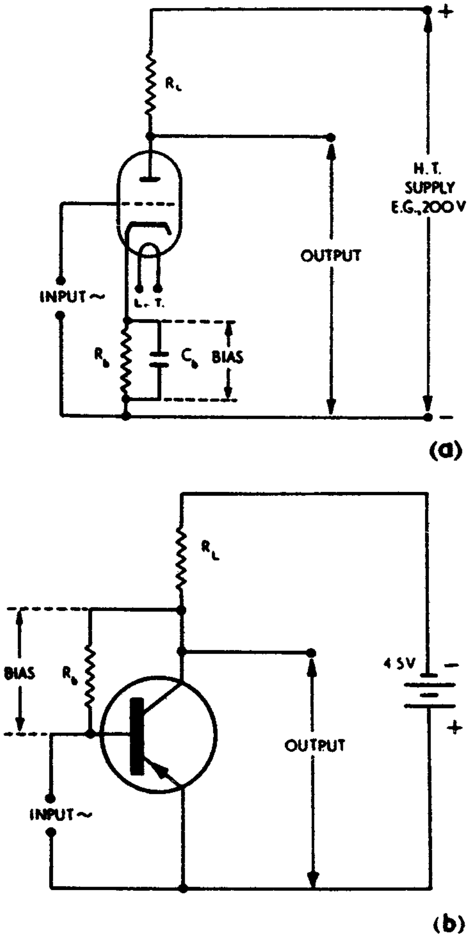
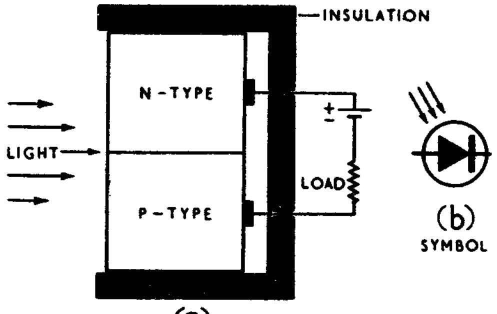


Fig. 24—TRANSISTOR AMPLIFIER

explained in Para. 5, and the current increases. Typical characteristics are reproduced in Fig. 25(c).

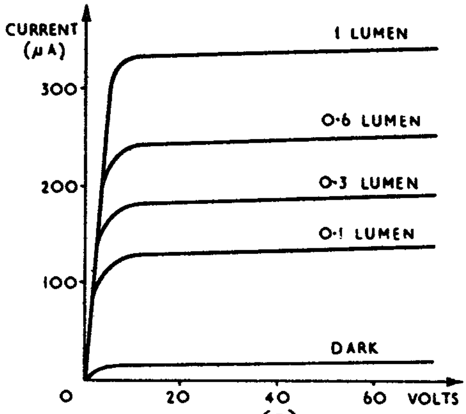
(*b*) *The point contact type.* This consists of a piece of p or n type germanium with one point contact terminal as shown in Fig. 26. The characteristics of this type of photodiode are similar to those of the junction type photodiode.

34. **Photo-transistors.** This development is effectively a junction transistor with a light sensitive base region. The advantage of such a device over the photodiode or photocell is that it is not only light sensitive but is also an amplifier. This means that photo-transistors can be connected in circuit directly to a relay without an intermediate amplifying stage. The symbol for a photo-transistor is shown in Fig. 27.



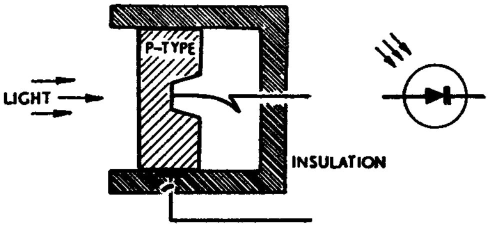
(a)
CONSTRUCTION

(b)
SYMBOL



(c)
CHARACTERISTICS

Fig. 25—JUNCTION PHOTODIODE



(a) CONSTRUCTION (b) SYMBOL

Fig. 26—POINT CONTACT PHOTODIODE

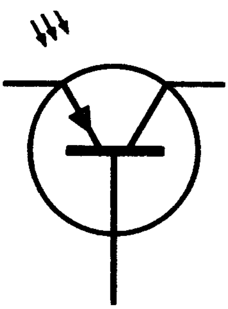


Fig. 27—SYMBOL FOR A PHOTO-TRANSISTOR

SECTION 9

POWER SUPPLIES

SECTION 9

POWER SUPPLIES

SECTION 9

CHAPTER 1

ELECTROCHEMICAL POWER SUPPLIES

Paragraph

PRIMARY CELLS

LEAD-ACID SECONDARY CELLS

ALKALINE CELLS

ELECTROCHEMICAL POWER SUPPLIES

Introduction

1. There are, in general, three different sources of electrical energy required for a radio equipment:—

(*a*) *Low tension* (l.t.) supply of a few volts for application to valve heaters.

(*b*) *Grid bias* (g.b.) supply up to the order of 100 volts for application to control grids.

(*c*) *High tension* (h.t.) supply of several hundreds of volts for application to valve anodes and screen grids.

In addition, some high power communication equipment and most radar equipment require an *extra high tension* (e.h.t.) supply of several thousands of volts.

2. The methods used for obtaining the required power supplies include:—

(*a*) *Electrochemical*, from primary or secondary batteries. This method, which is used mainly for low power mobile equipment, is considered in this Chapter.

(*b*) *Mechanical*, from rotating machines and vibrators. This method, commonly used in airborne communication equipment, is considered in Chapter 2.

(*c*) *Electronic*, from valve and metal rectifiers. Electronic power supply systems are used almost invariably on radar and ground communication equipment, and are considered in Chapter 3.

Electrolysis

3. When certain soluble substances are dissolved in water they ionise spontaneously, producing positive and negative ions. Such substances are termed *electrolytes*. For example, sulphuric acid (H_2SO_4) becomes, in solution, positive hydrogen ions (H+) and negative sulphate ions (SO_4^{-}).

4. If an e.m.f. is applied to two metal plates (electrodes) immersed in an electrolyte, the positive ions travel to the negative electrode (cathode) and the negative ions to the positive electrode (anode). This process is termed *electrolysis*, and is used in electroplating. In silver plating, the anode is silver, the electrolyte is silver nitrate and the article to be plated is the cathode. When an external e.m.f. is applied such as to make the anode positive and the cathode negative the migration of ions results in silver being deposited on the article while the anode is gradually eaten away.

5. The exchange between electrical and chemical energy can be made to work both ways. In electrolysis, electrical energy is being converted into chemical energy. The reverse process may be obtained as in the simple primary cell.

PRIMARY CELLS

The Simple Cell

6. Fig. 1 shows a zinc (Zn) rod and a copper (Cu) rod immersed in a solution of dilute sulphuric acid (H_2SO_4). Atoms of zinc are dissolved in the solution as positive ions, leaving the zinc rod *negative* with respect to the electrolyte. Positive hydrogen ions from the electrolyte are attracted to the copper, making the copper rod *positive* with

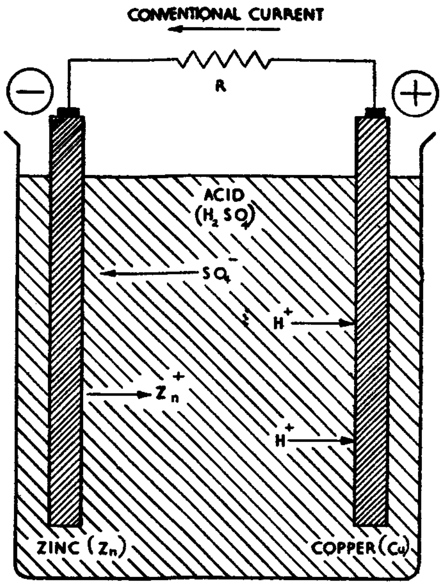


Fig. 1.—SIMPLE PRIMARY CELL

respect to the electrolyte. Thus, an e.m.f. (of approximately one volt) is established between the two rods.

7. On connecting an external circuit between the two rods, an electric current is established. This causes more zinc to dissolve in the acid and more hydrogen to be liberated at the copper electrode so as to maintain a p.d. between the terminals. Current continues to flow until the zinc is dissolved, until the acid is exhausted, or until 'polarisation' renders the cell inoperative (see Para. 8).

Polarisation

8. The formation of hydrogen bubbles at the positive (copper) electrode when the cell is connected to an external circuit is termed *polarisation*. The film of hydrogen tends to insulate the copper electrode from the electrolyte, thereby increasing the internal resistance of the cell and reducing the chemical action.

9. It is possible to remedy the effects of polarisation by dispersing the film of hydrogen; this process is known as *depolarising*. A second chemical, the depolariser, is included in the cell and this combines with the hydrogen to form a compound which is harmless in its action. The depolariser does not combine with the hydrogen instantaneously, and, since the release of hydrogen is proportional to the current taken from the cell, primary cells are suitable only for intermittent current demands or for very low continuous current work.

Local Action

10. Impurities present in the zinc result in the formation of small 'local' cells between the pure zinc and the impurities. The local currents resulting cause the zinc to be dissolved without supplying power to an external circuit. This factor determines the 'shelf life' of a cell.

Internal Resistance

11. Every cell has a certain internal resistance, depending on the construction of the cell. Thus, when current is being supplied to an external circuit, a voltage drop is developed across the internal resistance, and the p.d. at the terminals is then the *difference* between the open-circuit e.m.f. and the internal voltage drop. It is therefore necessary to test the terminal voltage of a cell *on normal load.*

Types of Primary Cell

12. Several types of primary cell, differing in their construction and choice of materials have been produced for various purposes. Table 1 lists a few of the better known types.

Type	Positive Electrode	Negative Electrode	Electrolyte	Depolariser	Terminal P.D.	Remarks
Daniell	Copper	Zinc	Sulphuric Acid	Copper Sulphate	1·1V	Low currents only
Weston	Mercury	Cadmium	Cadmium Sulphate	Mercurous Sulphate	1·02V	Standard of e.m.f.
Leclanche	Carbon	Zinc	Ammonium Chloride	Manganese Dioxide	1·5V	Intermittent low currents
Mallory	Mercury	Zinc	Potassium Hydroxide	Mercuric Oxide	1·34V	Fairly large currents

Table 1. TYPES OF PRIMARY CELL.

Dry Cells

13. The majority of primary cells used in the Service are of the '*dry*' Leclanche type, in which the electrolyte (ammonium chloride) is in the form of a paste containing sufficient moisture to permit the chemical action. The construction of a dry cell of this type is shown in Fig. 2. The positive electrode is a carbon rod and is surrounded by a mixture of powdered manganese dioxide and carbon which forms the depolariser. The reaction between the depolariser and the hydrogen formed round the carbon rod during discharge produces additional water. The carbon rod and depolariser rest on an insulating washer fitted into the base of the

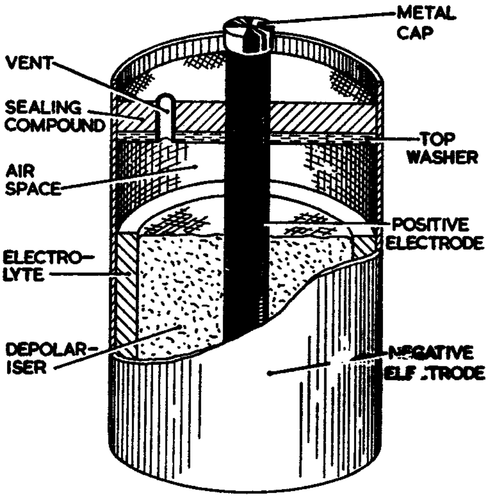


Fig. 2.—LECLANCHE DRY CELL

zinc cylinder, the latter serving as the negative electrode. The space between the depolariser and the zinc cylinder is filled with paste electrolyte of ammonium chloride. The e.m.f. of such a cell is about 1·5V and its internal resistance is between 0·5Ω and 4Ω.

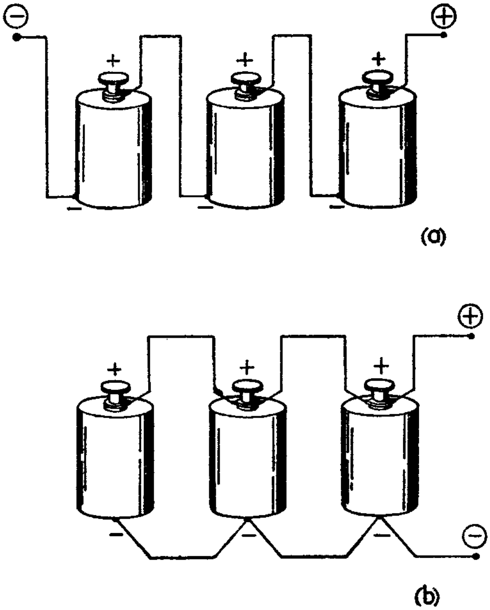
14. Mercury type dry batteries are used in the Service to provide power for certain types of radio equipment. The '*Kalium*' cell, based on the Mallory type (see Table 1), has potassium hydroxide (caustic potash) as the electrolyte and mercuric oxide as the depolariser. The zinc negative electrode is placed inside a zinc container and the positive electrode of mercury surrounds a central carbon rod, so that the external contacts are the same as those for the Leclanche dry cell. The Kalium cell can supply comparatively large currents (up to 1 amp) at a constant voltage of approximately 1·3V. This cell has curious discharge characteristics; the voltage drops suddenly when it is first put on load, and then rises slowly to its steady load value. The shelf life is longer than that of the Leclanche dry cell, but it should not be used at temperatures much below O°C or much above 40°C.

Batteries

15. A battery consists of two or more cells connected in series or in parallel. Alternatively, a battery may consist of several parallel banks of series-connected cells. For cells in series the positive terminal of one cell is connected to the negative terminal of the next as shown in Fig. 3(a), the total voltage being the *sum* of the cell voltages. H.T. batteries giving an open-circuit voltage of 120V are obtained in this way. For cells in parallel, all the positive terminals are joined together, and all the negative terminals as shown in Fig. 3(b). The total voltage is that of *one* cell, but the current that the battery is capable of supplying is increased (i.e., its '*capacity*' has been increased).

Life of Primary Cells

16. Primary cells, either singly or in batteries, may be used to provide the necessary power supplies in certain low power mobile radio equipment and test equipment. The dry type deteriorates in storage due to local action and its shelf life is therefore limited. In the '*inert*' cell, the electrolyte is in the form of dry crystals to which water must be added to make the cell active. Thus, no local action takes place in storage, and the shelf life of the inert cell is indefinite. When current is being drawn from a primary cell, the negative electrode is consumed and the electrolyte and depolariser become exhausted, after which the cell becomes useless. *Primary cells cannot be re-charged.*



(a)

(b)

Fig. 3.—BATTERIES

LEAD-ACID SECONDARY CELLS

Introduction

17. Primary batteries are capable of supplying small currents for limited periods, but their high internal resistance and their liability to polarisation make them unsuitable when a continuous supply of heavy current is necessary. Furthermore, the active material in a discharged primary cell cannot be regenerated and dry batteries must then be discarded. The *secondary cell* overcomes these disadvantages.

18. Secondary cells operate on the same principle as primary cells; that is, an e.m.f. is developed between dissimilar plates which are partially immersed in an electrolyte, and current flows when an external circuit is connected between the plates, the action continuing until the chemical energy of the cell is exhausted. However, in a secondary cell the chemical energy can be restored by passing a 'charging' current through the cell in the *reverse* direction to that of the discharge current. This charge is continued until the electrolyte and plates have been restored to their original condition. The action of the secondary cell is therefore reversible, unlike that of the primary cell, and its life consists of a large number of cycles of charge and discharge. The secondary cell does *not* 'store electricity'. It converts electrical energy into chemical energy during the charging process, and retains the chemical energy until an external circuit is connected to the terminals.

Definitions

19. (*a*) **Voltage.** The terminal p.d. of a secondary cell depends on the materials of which the plates are composed, the nature of the electrolyte and state of charge. For a fully charged lead-acid cell, the voltage is approximately 2·1 V on normal load. During discharge, the voltage of the cell gradually falls to 1·8 V, beyond which the discharge should not be continued.

(*b*) **Capacity.** This is the *amount* of charge that can be taken from a cell while it is discharging from the fully-charged to the fully-discharged condition. Since the coulomb is too small a unit for this purpose capacity is measured in *ampere-hours* (Ah); it is the product of discharge current (in amperes) and time of discharge (in hours), one Ah being equal to 3,600 coulombs.

(*c*) **Ten-hour rate.** The capacity of a cell varies with the *rate* of discharge. Thus, a cell which gives 8 amperes for 10 hours (i.e., capacity=80 Ah) will *not* give 80 amperes for 1 hour. For this reason, capacity is normally measured at the rate which discharges the cell in 10 hours. A cell marked '2 V, 14 Ah' should give 1·4 amperes over a period of 10 hours, the maximum discharge current for any cell being given by:—

$$\text{Current} = \frac{\text{Nominal Capacity}}{10} \text{ (amps).}$$

This is called the '*ten-hour rate*' and unless otherwise specified, it is also the normal rate at which the cell should be *charged.*

(*d*) **Specific gravity.** This is the ratio of the weight of a given volume of the electrolyte to the weight of an equal volume of water at the same temperature. The specific gravity (S.G.) of the electrolyte in a lead-acid cell when fully charged should be between 1·24 and 1·27; when the cell is discharged the S.G. falls to between 1·15 and 1·18. The exact figures should accord with the maker's instructions.

(*e*) **Batteries.** Secondary cells may be grouped to form batteries in the same manner as primary cells and with similar results. Thus, a battery might comprise six 2 V 40 Ah cells in series; the battery voltage would be 12 V and its capacity 40 Ah. Two such batteries connected in series would give 24 V with the *same* capacity (40 Ah), while two such batteries connected *in parallel* would give 12 V, but their combined capacity would be 80 Ah. Grouping should be restricted to cells or batteries of the same capacity.

Construction of Lead-acid Cell

20. A lead-acid cell consists essentially of two plates immersed in an electrolyte of dilute sulphuric acid. It is constructed as shown in Fig. 4:—

(*a*) **Plates.** These are grids of a lead-antimony alloy serving as supports for the active material, which is in a paste form. The negative plate carries spongy lead (Pb) and the positive plate, lead peroxide (PbO_2). Each cell contains several negative and positive plates, interleaved alternately in such a way that both outer plates are *negative.* All plates of the same polarity are attached to a lug which forms both a mechanical support and an electrical connection to the terminal. The plates are further supported by ribs on the inside of the container and are raised clear of the bottom of the cell.

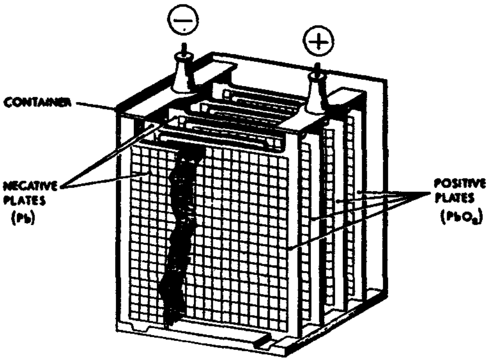


Fig. 4.—LEAD-ACID SECONDARY CELL

(*b*) **Separators.** Spacers are provided within the cell to prevent contact between positive and negative plates and to prevent buckling of the plates. They are designed to permit free circulation of the electrolyte, and are made of specially treated wood or perforated rubber or plastic materials.

(*c*) **Containers.** The containers used for portable batteries are made of black moulded insulating material which is impervious to the action of the acid. Containers for stationary batteries may be of glass.

(*d*) **Vent plugs.** Each cell is fitted with a vent plug to permit the escape of gas generated by the chemical action. Aircraft batteries are fitted with special *non-spill* vent plugs, and, to allow the gas to escape while at the same time preventing leakage of the electrolyte, the level of the electrolyte must be as given in the maker's instructions.

(*e*) **Electrolyte.** Only pure sulphuric acid (H_2SO_4) diluted with distilled water is to be used as an electrolyte and it must be of the correct S.G. as indicated on the maker's instruction label. Sulphuric acid of the correct S.G. may be supplied ready for use in batteries, or it may be supplied in concentrated form (S.G. 1·84). In the latter case the acid must be 'broken down' to the required S.G. by diluting with distilled water. Great heat is developed during mixing, which must be done gradually and with care. *Never pour water into acid; always add acid to water* until the correct S.G. is obtained.

Chemical Action

21. **Discharge.** When an external circuit is connected across the positive and negative terminals of a charged cell, a current is established. The conventional current is from the positive lead peroxide (PbO_2) plate through the external circuit to the negative spongy lead (Pb) plate and thence through electrolyte (H_2SO_4). The discharge current through the plates and electrolyte sets up a chemical action in the cell which:—

(*a*) Results in the formation of lead sulphate ($PbSO_4$) at *both* plates.

(*b*) Produces additional water in the electrolyte.

As the discharge continues, the terminal p.d. and the S.G. of the electrolyte both fall. In practice, the discharge should be discontinued before the voltage of the cell has fallen to 1·8 V on load, otherwise the lead sulphate on the plates is removed only with difficulty.

22. **Charge.** The plates of a discharged cell can be cleared of lead sulphate and the S.G. of the electrolyte restored by passing current through the electrolyte in the reverse direction. This is done by connecting the cell to a d.c. supply which has a voltage slightly higher than that of a fully charged cell. For charging, the positive terminal of the supply is connected to the positive terminal of the cell, and the charging action is continued until all the $PbSO_4$ has been reconverted to Pb or PbO_2 and the S.G. of the electrolyte has risen to its original level. The chemical energy which was expended during discharge has thus been restored and the cell is again available as a source of electrical energy.

Indications of Charge and Discharge

23. The state of charge of a cell is indicated in three ways as shown in Table 2.

Hydrometer

24. The S.G. of an electrolyte is measured with a hydrometer (Fig. 5). This consists of a graduated float contained in a glass

State	Terminal P.D.	Specific Gravity	Plates
Discharged	Not less than 1·8V	According to maker's instructions (commonly S.G. 1·18).	Both plates $PbSO_4$ whitish-grey lead sulphate.
Charged (on load).	Approx. 2·1V	According to maker's instructions (commonly S.G. 1·27).	Positive = PbO_2, brown in colour. Negative = Pb, slate grey in colour.
Charged (whilst on charge)	Approx. 2·7V	ditto	ditto

Table 2.—INDICATIONS OF STATE OF CHARGE

syringe into which the electrolyte is drawn by means of a rubber bulb. Sufficient liquid is drawn into the syringe to float the hydrometer and the S.G. is read off the scale graduation at the point where the surface of the electrolyte intercepts it.

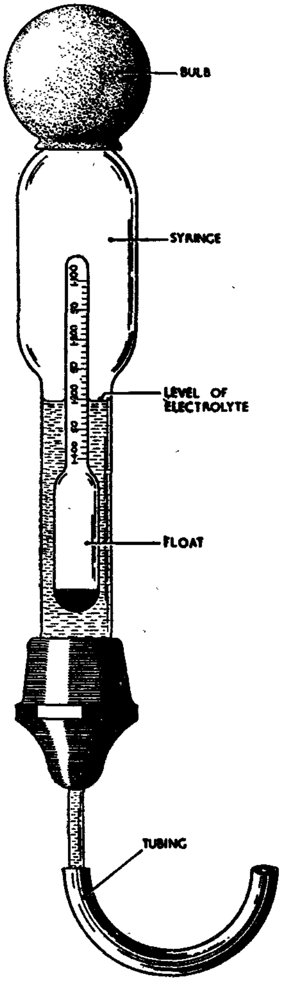


Fig. 5.—HYDROMETER

Faults in Lead-acid Cells

25. **Sulphation.** A sulphated cell is one in which the lead sulphate formed on the plates during discharge is allowed to become hard and insoluble so that it becomes difficult to remove by normal charging, and the cell suffers loss of capacity. This fault is caused by persistent undercharging, persistent discharge below 1·8 V and by leaving in a discharged state. It is indicated by a high terminal voltage during charging (because of the increased internal resistance), by a low S.G. at the end of the charge and by the formation of a white deposit on the plates. There is no remedy for this if matters have gone very far, but if the sulphating is not too deep-seated a cure may be effected by carrying out the procedure detailed in R.A.F. Poster No. 20. This Poster, which provides information on the treatment of lead-acid batteries, should be prominently displayed in all R.A.F. battery charging rooms.

26. **Internal short-circuits.** A cell whose plates are short-circuited will show little or no voltage across its terminals. Causes are as follows:—

(*a*) **Sediment.** This is due to active material being dislodged from the plates, and occurs to some extent with normal usage. It is, however, accelerated by persistent overcharging or charging at too high a rate. The sediment can be removed by washing out the cell with weak acid, refilling with electrolyte of the correct S.G., and re-charging; but the loss of active material reduces the capacity of the cell.

(*b*) **Buckling of the plates.** If the formation of lead sulphate at the plates during discharge occurs too suddenly, a serious mechanical strain is thrown on the framework, which may tend to buckle and warp. As a result, the separators may split and an internal short-circuit occur. It is caused by excessive discharge currents, and no local treatment is possible.

Charging Boards

27. The charging board is the means whereby the batteries are connected to the d.c. supply for charging. It contains a main switch, fuses, an ammeter and a rheostat, all connected in series with the terminals to which the batteries are connected (Fig. 6). A reverse-current automatic cut-out is sometimes fitted to prevent the batteries discharging back through the supply should the latter fail.

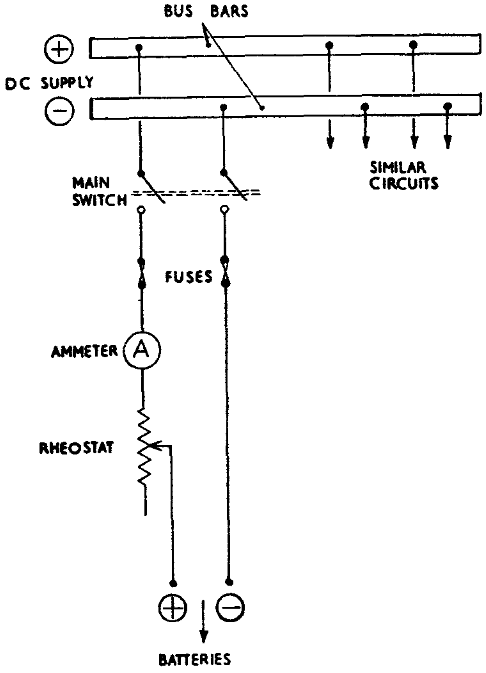


Fig. 6.—CHARGING BOARD

Initial Charge

28. The majority of lead-acid batteries require initial filling with electrolyte, followed by an initial charge of up to 24 hours or more, depending on the type of battery, before being ready for Service use. The capacity and useful life of a battery depend on the treatment it receives both before and after it is put into service. Full manufacturer's instructions for preparation and initial charge are usually issued with each battery; these must be followed in detail. Where instructions are not given, the procedure for initial charge detailed in R.A.F. Poster No. 20 should be followed.

Routine Charging

29. When arranging batteries for charging, they are placed on insulating grids laid on the charging bench and connected in series in a single line, with the supply leads from the charging board terminals connected to opposite ends of the line, positive to positive and negative to negative. A voltage of 2·8 V must be allowed for each cell so that the maximum number of lead-acid cells which can be connected in series for charging is found by dividing the charging voltage by 2·8. Batteries of the *same capacity* should be connected to one pair of charging terminals and the charging rate adjusted by the rheostat. The charging rate may be given by the manufacturers; where this is not done, the batteries should be charged at the ten-hour rate (i.e., 40 Ah batteries should be charged at 4 amps.). The charge is continued at the prescribed rate until the voltage of each cell on charge rises to 2·7 V and the S.G. rises to the level given by the manufacturer (commonly 1·27). These two figures should remain constant for at least one hour before the battery is taken off charge. Further, when charging is complete it is indicated by the liberation of gas bubbles at the plates. *Free gassing* is, therefore, a third sign of the completion of the charge. The detailed procedure for routine charging of lead-acid batteries is given in R.A.F. Poster No. 20.

ALKALINE CELLS

Principle

30. Lead-acid batteries require a considerable amount of careful servicing to maintain them in good condition. They are adversely affected by high rates of charge and discharge, and they suffer from sulphation if allowed to stand in a discharged condition for considerable periods. Where a battery is required to withstand such

conditions it is sometimes possible to make use of another form of secondary cell, the *alkaline cell.* An alkaline battery differs from a lead-acid battery in that it uses an alkaline instead of an acid electrolyte, and different electrode materials. The fundamental principle of this type of battery is the oxidation of metals in a solution of potassium hydroxide (caustic potash). The electrolyte does not combine with, nor does it dissolve, the metals.

Construction

31. A typical alkaline cell is constructed as shown in Fig. 7.

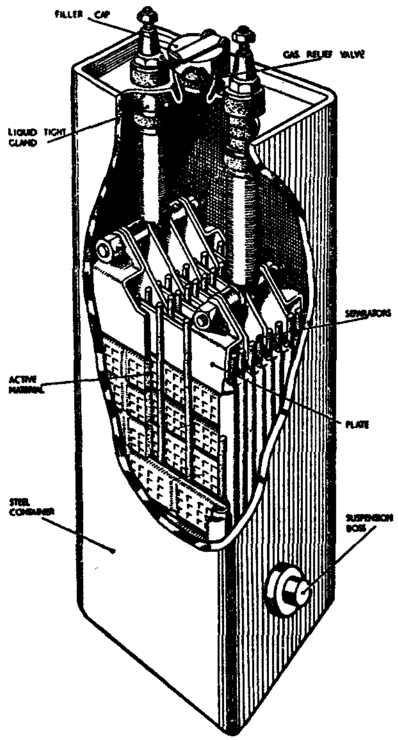


Fig. 7.—CONSTRUCTION OF ALKALINE CELL

(*a*) **Plates.** The active material in the positive plates is nickel hydroxide mixed with graphite, while that in the negative plates is a mixture of cadmium oxide and iron oxide. The active materials are compressed into briquettes under high pressure and held in steel 'pockets' in nickel plated steel frames. This method prevents shedding of the active material.

(*b*) **Separators.** Ebonite rods or perforated sheet ebonite are used as separators.

(*c*) **Containers.** These are made of welded steel, plated on the outside to prevent rusting. Since the container is in contact with the electrolyte it is 'alive', and care must be taken to avoid accidental short circuit to the steel case. The container is closed at the top and capable of being made airtight.

(*d*) **Vent plugs.** Cells may be fitted with solid stoppers, gas relief valves or unspillable vents. Gas relief valves allow the escape of gases given off during charging, but prevent the entry of air which causes deterioration of the electrolyte.

(*e*) **Electrolyte.** Potassium hydroxide is supplied in solid form in airtight containers and is dissolved in distilled water to the S.G. given on the maker's instruction label for the particular battery (commonly S.G. 1·19).

Chemical Action

32. *During discharge*, the active material in the negative plate becomes oxidized, while that in the positive plate is reduced from a higher to a lower oxide of nickel. *During charge*, the active material in the positive plate becomes more highly oxidized, while that of the negative plate is reduced in oxidation. The chemical actions are therefore, reversible. The electrolyte takes no active part in the chemical action, the S.G. remaining *constant* throughout the cycle.

Indications of Charge and Discharge

33. Since there is no variation in the S.G. of the electrolyte during the charge-discharge cycle, and since gassing on charge is continuous, the only indication of the state of charge of a cell is the terminal p.d. Fig. 8 shows how the terminal p.d. varies with time during charge and during discharge. From these curves it is seen that:—

(*a*) The terminal p.d. of a fully charged cell *on charge* should be 1·7 V approximately.

(*b*) The terminal p.d. of a fully charged cell *on normal load* should be between 1·2 V and 1·3 V approximately.

(*c*) The terminal p.d. of a *fully discharged* cell on load is 1·1 V approximately.

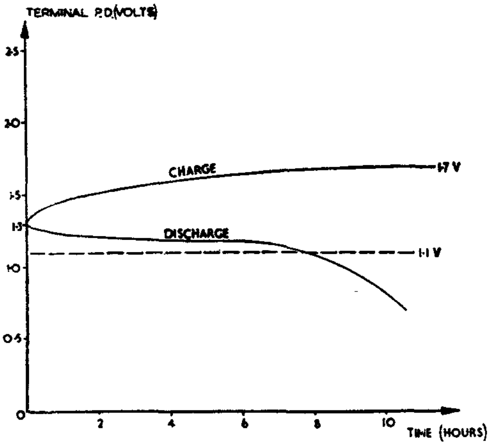


Fig. 8.—VARIATION OF TERMINAL P.D. WITH TIME FOR ALKALINE CELL

Faults in Alkaline Cells

34. Alkaline cells are relatively trouble-free owing to the robust construction of the plates and containers and to the nature of the materials used. The most common fault is loss of capacity; this is caused in most cases by undercharging, or by allowing the cell to stand idle. The active material then becomes less soluble. A full charge at two or three times the normal rate will loosen the active material and restore the capacity of the cell. There is no danger of the heavy current causing the plates to 'shed' their active material since the latter is contained in steel pockets in the plates. High charging currents, overcharging or high discharge currents will not normally cause any damage.

35. A little electrolyte is lost from all cells by gassing and the S.G. is reduced by topping up with distilled water. The minimum allowable S.G. is 0·03 below the value given in the instruction label, and when the S.G. is less than this level it is necessary to change the electrolyte completely.

Charging

36. Alkaline batteries are usually supplied fully charged and filled with electrolyte. When it is necessary to give an initial charge, the cells must be filled immediately the sealing plugs have been removed, and the initial charge carried out according to the maker's instructions. When the battery is in service, routine charging should be given when necessary. The charge is considered to be complete when the terminal p.d. per cell has remained constant on charge at approximately 1·7 V for a period of one to two hours. Moderate overcharging is often beneficial in overcoming sluggishness. Idle cells should be charged at least every six months. Maker's instructions should be followed in all these operations. If such instructions are not available the procedures detailed in R.A.F. Poster No. 21 should be followed. Poster No. 21, which provides information on the treatment of alkaline batteries, should be prominently displayed in all R.A.F. battery charging rooms.

Note:—On no account must acid-contaminated equipment be used in connection with alkaline batteries, or *vice versa*. For this reason it is preferable to have separate charging rooms for lead-acid and for alkaline batteries. Where only one charging room is available, the two types of batteries must be separated by at least three feet, and the servicing equipment kept separate.

Comparison of Alkaline with Lead-acid Cells

37. **Advantages.**

(*a*) Longer life.

(*b*) More robust.

(*c*) Will stand for long periods without deterioration.

(*d*) Able to retain charge for much longer periods.

(*e*) Higher permissible working temperature.

(*f*) Higher permissible charging rate; insensitive to overcharging.

(*g*) Over-discharge has no permanent adverse affect.

(*h*) The capacity is not affected by high rates of discharge.

(*i*) Lighter in weight for a given *capacity*.

38. **Disadvantages.**

(*a*) Lower cell voltage, requiring more cells per battery for a given voltage.

(*b*) Greater percentage range of voltage during charge and discharge cycle.

(*c*) Higher internal resistance.

(*d*) Loss of capacity at low temperature.

(*e*) Increased weight for a battery of given *voltage*.

(*f*) Containers are alive, and require effective insulation.

(*g*) Must be kept sealed to prevent deterioration of the electrolyte.

(*h*) Greater initial cost.

Use of Secondary Batteries

39. **Secondary batteries,** particularly of the lead-acid type, have a large number of uses in the Service. Two of their main uses in radio engineering are:—

(*a*) Providing a d.c. supply in aircraft. In addition to supplying electrical circuits in the aircraft, these batteries may be used to operate rotary machines which provide h.t., g.b., and l.t. supplies for certain radio equipment.

(*b*) Providing a means whereby a standby power system for low power ground communication equipment may be run in the event of failure of the mains supply.

SECTION 9

CHAPTER 2

MECHANICALLY-DERIVED POWER SUPPLIES

PART 1A SECTION 9, CHAPTER 2

MECHANICALLY-DERIVED POWER SUPPLIES

Introduction

1. Electrical power supplies that are obtained from machines by virtue of the physical movement of the latter are said to be mechanically derived. Mechanically-derived power supplies include those obtained from rotary machines and vibrator units, both of which are considered in this Chapter. Such supplies are in general use:—

(*a*) in aircraft or on other mobile systems as the main or as the auxiliary source of supply for radio equipment.

(*b*) for static equipment, as a standby in the event of failure of the mains supply.

Aircraft General Supplies

2. Virtually all the electrical supplies for an aircraft are obtained initially from generators that are driven by the aero-engines. Such generators, referred to as *engine-driven generators*, are the concern of the Electrical Fitter and will not be considered in these Notes. The supplies available from the engine-driven generators in modern aircraft are of two types:—

(*a*) *A d.c. supply*, where batteries are included in the circuit to act as a standby in the event of failure of the generators and as a 'float' to regulate the load on the generators.

(*b*) *A single-phase a.c. supply*, in which the frequency of the supply varies with the speed of the aero-engine.

3. Where a *constant frequency* a.c. supply is required, an *inverter* is used. This consists of a combined d.c. motor and a.c. generator run at a constant speed from the aircraft d.c. supply to give the required a.c. output. This may be single-phase or three-phase depending on the requirements. The a.c. supplies in an aircraft may be applied to radio equipments that are provided with 'rectifiers' (see Chap. 3) from which the necessary d.c. supplies are obtained.

Built-in Rotary Transformers

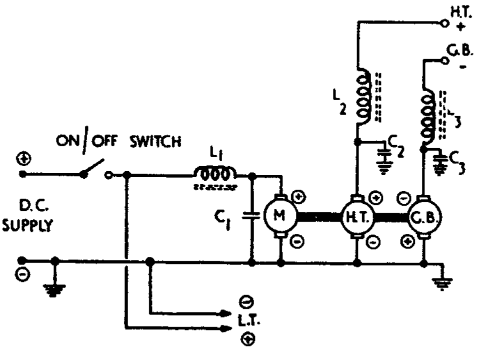


Fig. 1.—TYPICAL ROTARY TRANSFORMER CIRCUIT

4. A large number of mobile radio equipments have built-in rotary machines which operate from a low voltage d.c. supply. A typical example is shown in Fig. 1 where a rotary transformer is used to provide h.t. and g.b. supplies for the equipment, the l.t. supply being obtained direct from the d.c. source. A 'filter' circuit (L_1C_1) is inserted in the input circuit to the motor to prevent high frequency noise currents, caused by sparking at the motor commutator, from feeding back into the d.c. supply. The high reactance of L_1 and the low reactance of C_1 to such high frequency components shunts such components to earth. Similar filters are inserted in the output circuits.

Carbon Pile Regulator

5. **Purpose.** The carbon pile regulator is used to control and regulate the voltage applied to a circuit despite wide variations in the supply voltage. In a typical system, for variations in the d.c. supply voltage between 22 volts and 29 volts, the voltage applied to a circuit is stabilized at 18 volts, plus or minus 1 volt.

6. **Construction.** A simplified construction of a carbon pile regulator is shown in Fig. 2. It consists of three major portions—the electromagnet, the armature unit and the pile unit.

(*a*) *Electromagnet.* This has a cylindrical yoke (containing the coil windings), a detachable end plate and a cylindrical iron core.

(*b*) *Armature unit.* This is attached to a multi-leaved spring connected to the frame. A recessed plunger is mounted on, but insulated from, the armature and in the recess is fitted a carbon insert which makes contact with one end of the carbon pile. A terminal screw on the plunger gives one external connection to the pile.

(*c*) *Pile unit.* This is housed in a ceramic tube, the number, size and thickness of the carbon washers varying according to the type of regulator. A carbon insert, supported in an adjustable compression screw, makes contact with the outer end of the pile. The second external connection to the pile is made from a terminal screw on the compression screw bracket.

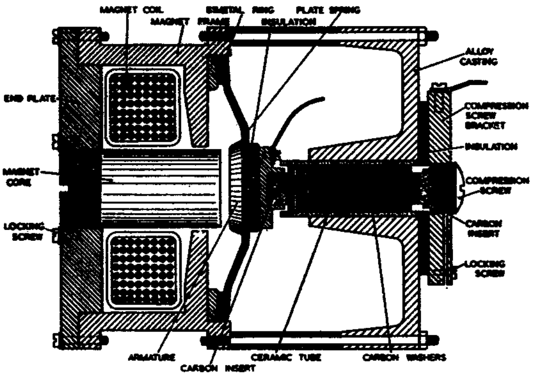


Fig. 2.—CONSTRUCTION (SIMPLIFIED) OF CARBON PILE REGULATOR

7. **Action.** Carbon has a granular surface and the resistance between two pieces of carbon depends not only upon the area of the faces in contact, but also upon the *pressure* with which the pieces are held together. For a pile of carbon washers, the resistance between the ends when the pile is tightly compressed may be as low as 5 ohms, rising to 90 ohms when the pressure on the washers is removed. Thus, if means are available to *vary* the pressure on a carbon pile, its resistance will vary and so also will the voltage drop developed across it when a current is established.

8. From Fig. 2 it is seen that the pressure acting on the carbon pile is dependent on two factors, namely:—

(*a*) the attactive force exerted by the electromagnet on the armature; this is determined by the current in the coil windings, an *increase* in current resulting in an increased attractive force on the armature and therefore a *reduction* in the pressure on the carbon pile. The resistance of the pile is thereby *increased.*

(*b*) the force exerted by the plate spring on the armature; with no current in the coil windings, the tension on the spring is such that the resultant pressure on the carbon pile gives the required *minimum* value of resistance.

9. Consider the circuit of Fig. 3. If the supply voltage *falls* below 28 volts, the current through the electromagnet winding falls. The attraction of the armature by the magnet core, is, therefore, reduced and the spring further compresses the pile, thereby *decreasing* the pile resistance. The voltage drop across the pile is then reduced. Since the carbon pile and the electromagnet winding are in series across the supply voltage and since the output is taken across the coil winding (with R_2 in series), the output voltage remains substantially constant, the decreased voltage drop across the pile compensating for the fall in supply voltage. A *rise* in supply voltage above 28 volts has the reverse effect; that is, the magnetic force initially exceeds the spring force and the movement of the armature reduces the compression of the pile to *increase* the resistance of the latter. The increased voltage drop across the pile compensates for the rise in supply voltage and keeps the output voltage constant within certain limits.

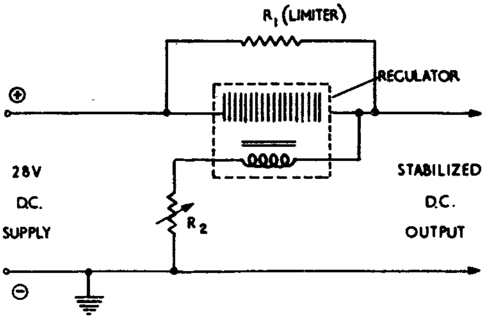


Fig. 3.

ACTION OF CARBON PILE REGULATOR

The resistance R_2 is a pre-set control which is adjusted to give the correct output voltage for an input voltage of 28 volts. The resistance R_1 acts as a shunt on the carbon pile to limit the current through the latter.

Standby Supplies for Ground Equipment

10. At static ground stations it is usual to install a standby power supply to be used in the event of failure of the mains supply. The standby equipment may consist of a gasoline or diesel engine driving a generator which provides the appropriate a.c. or d.c. supply. Such installations are also used for mobile stations.

Vibrator Units

11. The voltage of an a.c. supply can be stepped up or down as required by the use of transformers. This cannot be done with d.c., and if the source of power is a low voltage battery some other means must be found for stepping this up to the required h.t. supply. A rotary transformer may be used as previously explained. Alternatively, a vibrator unit may be used to change the d.c. to a low voltage a.c. supply which can then be stepped up by a transformer and dealt with as the normal mains supply. The way in which a vibrator does this is described below.

12. Fig. 4 shows the simplified circuit of a vibrator known as the series-drive type. Other types are available but the action of all types is similar. In the idle or rest position, the armature is connected through a contact to the vibrator coil. When the d.c. supply is connected, magnetising current is established in the coil and the armature is attracted to the top contact. Current is thus established in the top half of the transformer primary winding. This movement also breaks the circuit of the magnetising coil so that the armature is released. It therefore travels back through its rest position and, owing to its momentum, continues on to the lower contact. The armature is now in contact both with the lower contact and the coil contact. Current is thus established in the lower half of the transformer primary winding in the *reverse* direction to the previous pulse, and at the same time the magnetising coil is energised. The armature returns to the top contact and the cycle begins again. The current in the transformer primary consists of reversals of d.c. pulses which have the same effect as ordinary a.c. and produce the required high alternating voltage across the secondary.

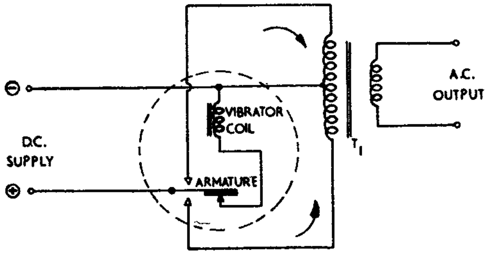


Fig. 4.—SERIES-DRIVE TYPE VIBRATOR

13. The transient voltage surges in vibrator circuits contain radio frequency noise components which cause interference in any near-by radio equipment. The methods used to suppress this interference include:—

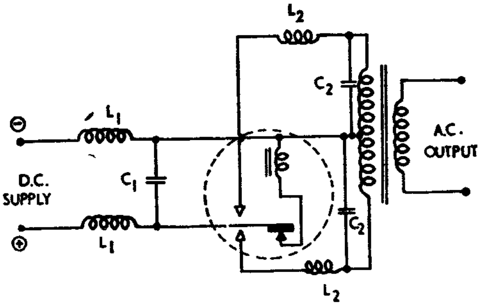


Fig. 5.—SUPPRESSION CIRCUITS

(*a*) Magnetic and electrostatic shielding.

(*b*) Good earthing.

(*c*) Radio frequency filtering in the leads to and from the vibrator.

The action of filter circuits will be considered fully in later Sections. However, in Fig. 5 the high reactance of the r.f. chokes L_1 and L_2 and the low reactance of the capacitors C_1 and C_2 to radio frequency noise components reduce such components in the output.

14. The construction of a series-drive type vibrator is shown in Fig. 6. The metal can of the vibrator provides the necessary screening and also hermetic sealing. Without the latter, the vibrator may easily stop in a humid or a rarefied atmosphere because of deterioration of the contacts. The frequency of the a.c. produced by a vibrator is generally of the order of 100 c/s. It depends principally on the mechanical inertia of the reed and the electrical characteristics of the coil. Vibrator units may be used to provide h.t. supplies in lieu of h.t. batteries, or as a standby supply for low power equipment on the ground in the event of failure of the mains supply.

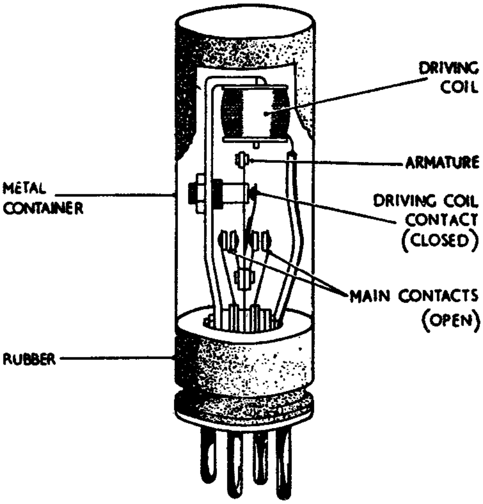


Fig. 6.—CONSTRUCTION OF SERIES-DRIVE TYPE VIBRATOR

SECTION 9

CHAPTER 3

ELECTRONIC POWER SUPPLIES

PART 1A SECTION 9, CHAPTER 3

ELECTRONIC POWER SUPPLIES

Introduction

1. Where the sources of power supply give d.c. voltage they can be used directly as h.t., l.t. and g.b. supplies to radio equipment. Where the source of power is a.c., means must be found for converting the a.c. input to the required d.c. output. This process is termed '*rectification*' and the devices used to provide rectification are known as '*rectifiers*'. The general symbols for a rectifier are shown in Fig. 1.

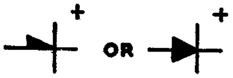


Fig. 1.—GENERAL SYMBOLS FOR A RECTIFIER

2. A rectifier is an electrical conductor with *non-linear* characteristics; that is, a device having a low resistance in one direction and a high resistance in the reverse direction so that current can flow much more easily through the device in one direction than in the other. The rectifiers considered in this Chapter are:—

(*a*) Hard-vacuum diode valves.

(*b*) Gas-filled diode valves.

(*c*) Metal rectifiers.

(*d*) Mercury-arc rectifiers.

Hard-vacuum Half-wave Rectifier

3. The basic action of a diode as a rectifier is described in Sect. 8, Chap. 1, where it is shown that the application of a.c. to a diode results in a uni-directional voltage being developed across the load. A practical half-wave rectifier circuit is shown in Fig. 2(a). A transformer is used for supplying the rectifier valve. Winding CD is a step-down low voltage winding which provides the cathode heating power. It is centre-tapped so that the alternating l.t. supply is not superimposed on the d.c. voltage developed across the load. Winding AB is the h.t. winding, the voltage of which is determined by the turns ratio according to the output voltage required. On the half-cycle when A is positive with respect to B the diode conducts and the current established develops a p.d. across the load in the polarity shown. On the next half-cycle, when A is negative with respect to B, the diode is cut off and no voltage is developed across the load. Thus, in half-way rectification, the diode conducts only on *alternate* half-cycles. The appropriate waveforms are shown in (*b*) and (*c*)

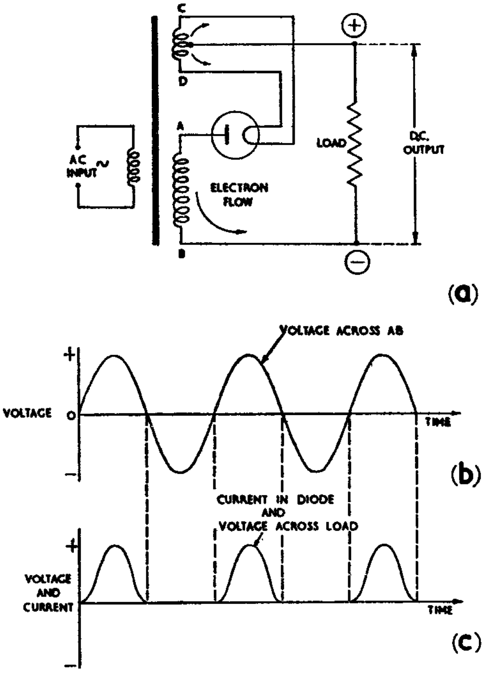


Fig. 2.—HALF-WAVE RECTIFICATION

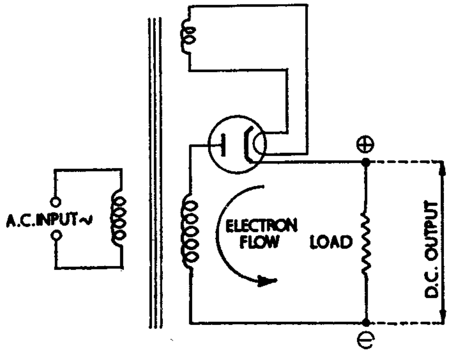


Fig. 3.—INDIRECTLY-HEATED DIODE HALF-WAVE RECTIFIER CIRCUIT

of Fig. 2. The circuit of a half-wave diode rectifier using an *indirectly-heated cathode* is shown in Fig. 3. This avoids the necessity for the centre-tap on the l.t. winding, but indirectly-heated diodes can be used only when the power requirements are low. Note that in any valve rectifier circuit, the cathode is the *positive* terminal for the output.

Hard-vacuum Full-wave Rectifier

4. This uses two diodes arranged as shown in the circuit of Fig. 4 (a). The secondary h.t. winding is centre-tapped at C, and the anodes of the two diodes are connected to opposite ends A and B. On one half-cycle of the a.c. input, when A is positive with respect to C, B is equally negative with respect to C. Thus V_1 conducts and V_2 is cut off. Electrons flow from V_1 cathode to V_1 anode, through the top half (AC) of the secondary h.t. winding, through the load and back to V_1 cathode, developing a pulse of voltage across the load in the polarity shown. On the next half-cycle of the a.c. input, the polarity of the voltage across AB is reversed, so that V_1 is cut off and V_2 conducts. Electrons flow from V_2 cathode to V_2 anode, through the bottom half (BC) of the secondary h.t. winding, through the load *in the same direction* as on the previous half-cycle and back to V_2 cathode. The pulse of voltage developed across the load is, therefore, of the *same polarity* on each half-cycle and the a.c. input has been *rectified* to give a uni-directional voltage across the load as shown in Fig. 4 (c).

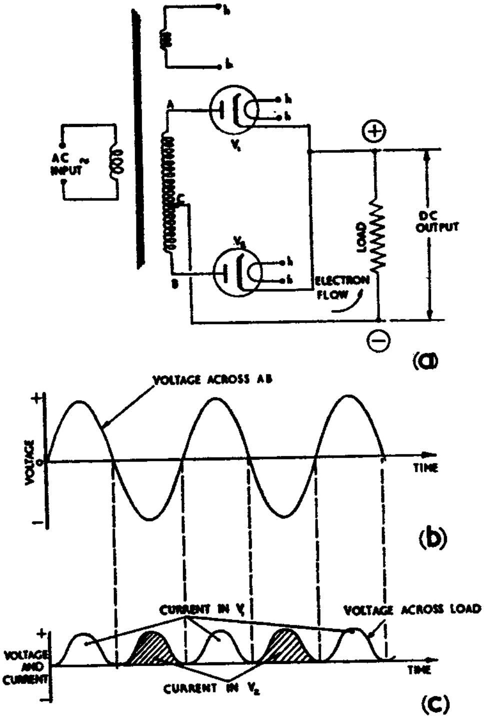


Fig. 4.—FULL-WAVE RECTIFICATION

5. In full-wave rectification, *each* half-cycle is effective. However, since the secondary h.t. winding is centre-tapped, the voltage applied to each valve has a maximum value equal to *half* the peak value of the a.c. developed across the secondary winding AB. The voltage across the load is *less* than this value because of the voltage drop across the working impedance of the conducting diode. The voltage across the secondary h.t. winding is usually given in r.m.s. values as, say, 300 — 0 — 300 volts to indicate the centre-tap. Instead of using two separate diode valves, a *double diode valve* may be used as shown in the circuit of Fig. 5. In this circuit, D_1 conducts on one half-cycle and D_2 conducts on the alternate half-cycle, to give a uni-directional voltage across the load.

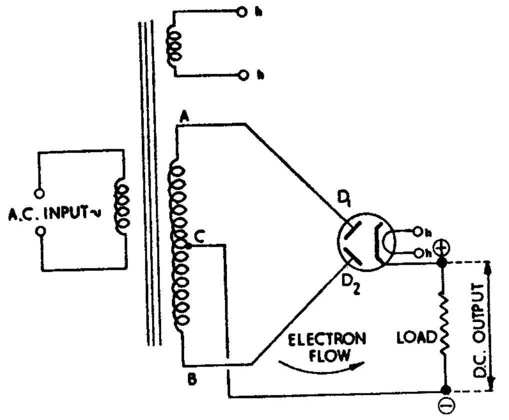


Fig. 5.—DOUBLE DIODE FULL-WAVE RECTIFIER CIRCUIT

Smoothing Circuits

6. The outputs of half-wave and full-wave rectifier circuits have the waveforms shown in Fig. 2(c) and Fig. 4(c) respectively. This pulsating d.c. is of little practical value since radio equipments require a *steady*

d.c. voltage for their operation. In order to obtain such an output, a *smoothing circuit* is inserted between the rectifier and the load. A smoothing circuit is a filter which removes any unwanted alternating component to ensure a steady d.c. voltage across the load. The two main types of smoothing circuit used are described in the following paragraphs.

7. **Capacitor input filter.** The circuit of a half-wave rectifier with a 'π-type' (capacitor input) filter is shown in Fig. 6(a). With the load disconnected, the '*reservoir*' capacitor C_1 charges to the *peak* value of the voltage appearing across the transformer secondary h.t. winding, as shown in Fig. 6(b). Current flows in the rectifier only when the secondary h.t. voltage rises above the voltage to which the capacitor is charged, so that the current gradually diminishes as shown in Fig. 6(c), finally ceasing when the capacitor is fully charged.

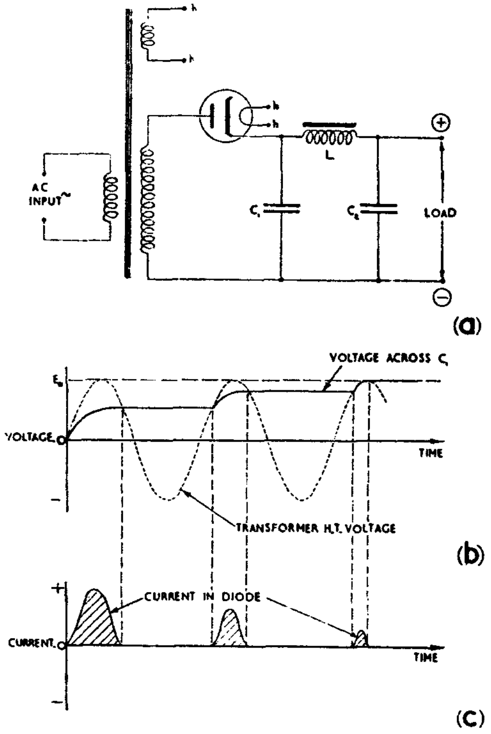


Fig. 6.—CAPACITOR INPUT FILTER, OFF LOAD

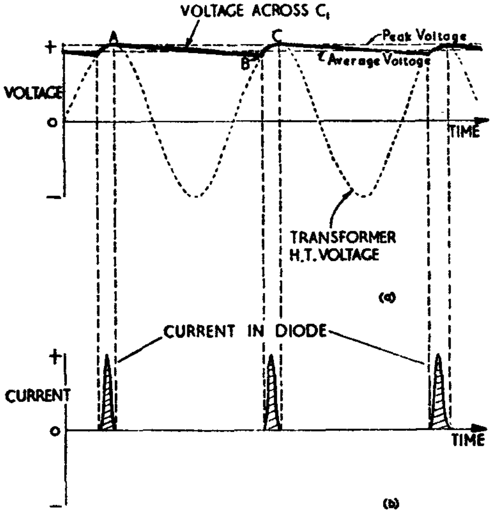


Fig. 7.—CAPACITOR INPUT FILTER, ON LOAD

8. When the rectifier is connected to a load, the current drawn by the load starts to discharge C_1 as indicated by the downward slope from A to B in Fig. 7(a). C_1 can recharge only when the transformer voltage rises above its own voltage. During this phase (B to C) C_1 has to receive enough charge to keep the load supplied continuously throughout one cycle. The value of C_1 is important:—

(*a*) If C_1 is small it will lose voltage rapidly between a.c. peaks the result being a much lower average output voltage at full load than at no load; that is, the '*regulation*' is poor (see Para. 11). In addition, there is a large ripple on the output voltage.

(*b*) If C_1 is large enough to hold the output voltage well up on load the period represented by BC in Fig. 7(a) is only a small fraction of the cycle, and since the capacitor must receive its full charge during this short period, the peak current through the rectifier is many times greater than the steady load current, and damage may result.

In practice, a value of $8 \mu F$ is typical for C_1 for power supplies of 50 c/s.

9. The voltage across the reservoir capacitor C_1 on load is a steady d.c. voltage which is almost equal to the peak value of the voltage applied to the valve, with a ripple superimposed. In half-wave rectification the frequency of the ripple component is that of

the supply. The voltage developed across C_1 is supplied to the iron-cored choke L and the smoothing capacitor C_2 connected in series to form a potentiometer across the rectifier output. At zero frequency (i.e,, for d.c.) the inductance offers a low impedance while the capacitor offers an infinite impedance; thus, practically the whole of the d.c. voltage is developed across C_2 and applied to the load. For the alternating ripple component, however, the inductance offers a high impedance, and the capacitor a low impedance; only a fraction of the ripple voltage thus appears across C_2 and across the load. Values for L and C_2 are dependent on the degree of smoothing required and are in the range 1H to 30H, and $4\mu F$ to $50\mu F$ respectively.

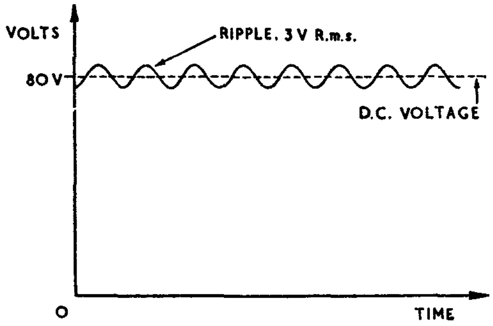


Fig. 8.—RIPPLE FACTOR

10. The '*ripple factor*' is a term used to describe the amount of ripple present in the output and is given by :—

$$\text{Ripple factor} = \frac{\textit{R.M.S. Alternating Ripple}}{\textit{Rectified D.C. Output}} \times 100 \text{ (per cent).}$$

For example, in Fig. 8, if the superimposed ripple is 3 volts r.m.s., the ripple factor is:—

$$\frac{3}{80} \times 100 = 3{\cdot}75\ \%.$$

11. Fig. 9 shows the graph of output voltage plotted against load current for a half-wave rectifier employing a capacitor input filter. This graph indicates that the *regulation* of such a system is poor since the d.c. voltage available at the output terminals falls steadily with increased load.

12. The action of a capacitor input filter in a *full-wave* rectifier circuit is similar to that for a half-wave rectifier with the exception that:—

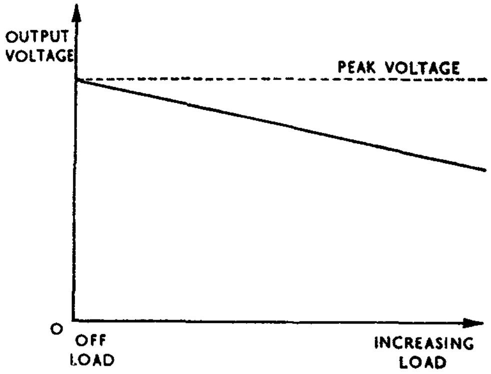


Fig. 9.—REGULATION, CAPACITOR INPUT FILTER

(*a*) The frequency of the ripple is *twice* that of the supply since, during one cycle of the a.c. supply, there are two unidirectional output pulses. Smoothing is therefore more efficient as the reactance of the smoothing choke is proportional to frequency, while that of the smoothing capacitor is inversely proportional to frequency.

(*b*) The amplitude of the ripple is *less* since the discharge of the reservoir capacitor occurs over only half the period of that of a half-wave rectifier.

(*c*) The output voltage approximates to *half* the peak voltage developed across the transformer h.t. winding because of the centre-tap.

13. **Choke input filter.** The circuit of a full-wave rectifier with an 'L-type' (choke input) filter is shown in Fig. 10(a). A choke, being a device which tends to maintain a constant current, will not permit sudden pulses of charging current such as those that occur with a capacitor input filter. The output of a full-wave rectifier has the waveform shown in Fig. 10(b). These pulses of d.c. consist of a steady d.c. component, equal to the *mean* value of the pulses, plus a number of alternating components. If this waveform is applied to the load via the choke input filter the d.c. component is virtually unaffected by the choke, so that a d.c. voltage equal to the *mean* value of the output pulses is developed across C_1 and hence across the load. The mean value of such a waveform is $\frac{2}{\pi}$ or 0·64 times the peak value. At the frequencies of the various ripple components,

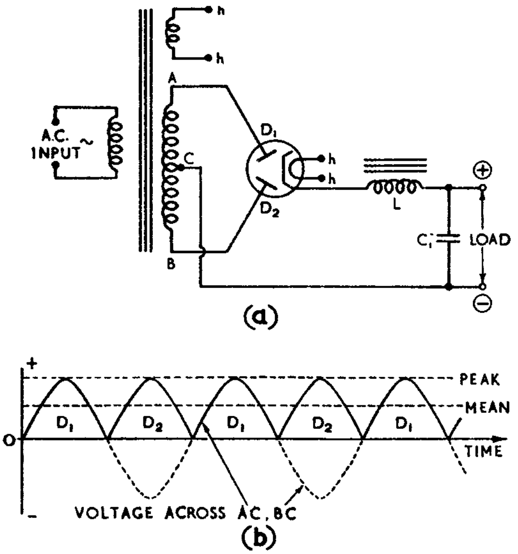


Fig. 10.—CHOKE INPUT FILTER

the inductance offers a high reactance and the capacitor offers a low reactance; only a fraction of the ripple voltage, at a frequency *twice* that of the input, appears across C_1 and thus across the load.

14. The difference in voltage between the a.c. applied to each diode and the steady d.c. output is absorbed mainly by the choke. Fig. 11 shows the form of the current through the choke, from which it is seen that the rectifier current is equal to the load current with a double-frequency ripple superimposed. If the load impedance is reduced so that the steady d.c. current increases, the rectifier current increases in similar proportion. However, since the amplitude of the *ripple* component is unaffected, the voltage drop across the choke remains tolerably constant, as does the output voltage. If, however, the load current is *reduced* to such a value that the ripple amplitude is *greater* than the steady d.c. load current, current ceases altogether at the troughs of the ripple. Under such conditions there is *no voltage* developed across the choke and the output voltage then rises towards the *peak* value of the applied a.c. The graph of output voltage plotted against load current, which indicates the regulation of the system, is shown in Fig. 12.

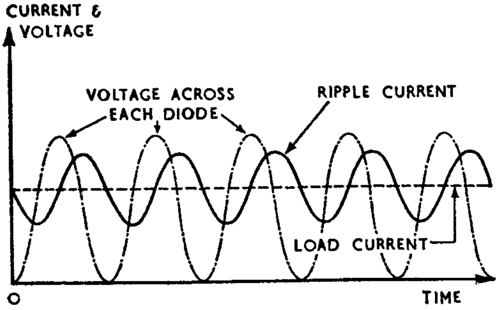


Fig. 11.—CURRENT WAVEFORM, CHOKE INPUT FILTER

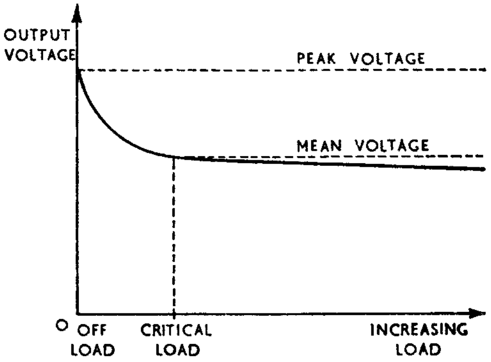


Fig. 12.—REGULATION, CHOKE INPUT FILTER

15. To prevent the sharp rise in output voltage at low values of load current:—

(*a*) Chokes, which give a rise in inductance value as the load current decreases, can be constructed. Such chokes are known as '*swinging chokes*'. They have no air gap in the iron core so that at low values of load current the inductance is high, but as the load current increases the core saturates to give a lower value of inductance (see Book 1, Sect. 7 Chap, 3

(*b*) A bleeder resistance can be connected in parallel with the smoothing capacitor to ensure an adequate minimum value of load current.

16. The choke input filter is seldom used in half-wave rectifier circuits since it is then very difficult to maintain the necessary continuous current through the choke for

satisfactory operation because of the suppressed half-cycle.

17. **Comparison of filters.** In comparison with the capacitor input filter, the choke input type:—

(*a*) has good regulation from full load down to the critical value of load (compare Fig. 9 and Fig. 12);

(*b*) prevents high amplitude charging current pulses, thereby reducing danger of damage to the rectifier;

	Capacitor Input Filter		Choke Input Filter	
	Half-wave	Full-wave	Half-wave	Full-wave
D.C. Output	650 V	325 V	416 V	208V

Table I.—OUTPUT VOLTAGES

(*c*) gives a lower value of output voltage, approximating to the *mean* value of the uni-directional pulses.

Table 1 gives approximate output voltages for the systems discussed, when the alternating voltage across the transformer secondary h.t. winding is 460 volts r.m.s. or 650 volts peak. These figures neglect voltage drops in the valves and in the smoothing circuit.

Peak Inverse Voltage

18. The peak voltage across a diode in the reverse direction during the non-conducting portion of the cycle is known as the '*peak inverse voltage*'. Fig. 13 shows the circuit of a half-wave rectifier with a capacitor input filter. On no load, the reservoir capacitor C_1 charges to the peak value Ep of the alternating voltage developed across AB. At the peak of the half-cycle during which the valve is cut off, the anode is negative with respect to its cathode by Ep across AB plus Ep across C_1. The peak inverse voltage across the rectifier is thus 2 Ep. This voltage may be very large and the circuit must be so designed that the peak inverse voltage is insufficient to damage the valve. Similar reasoning applies for choke input filters and for full-wave rectifier circuits.

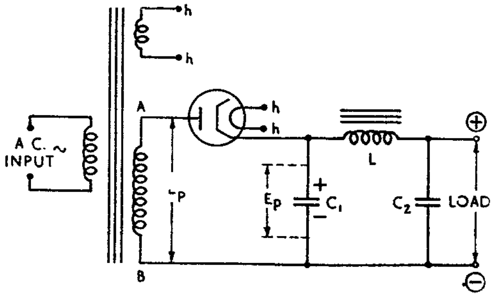


Fig.—13 PEAK INVERSE VOLTAGE

Full-wave Bridge Rectifier

19. The full-wave circuit previously described suffers from the disadvantage that only *half* the voltage across the transformer secondary is effectively applied to the load because of the centre-tap. This disadvantage is overcome by arranging four rectifiers in a bridge circuit as shown in Fig. 14. The secondary of the transformer is connected to two opposite junctions, and the load is connected as the other diagonal. When the end X of the transformer is positive, current is established, the electron flow being through diode B, the transformer secondary, diode C, the reservoir capacitor C_1 and back to diode B. During this half-cycle, diodes A and D are cut off, and the reservoir capacitor charges to the peak value E*p* of the applied voltage. On the next half-cycle, when Y is positive, diodes A and D conduct to charge C_1 in the *same direction* as previously. This is similar to normal full-wave rectifier action except that:—

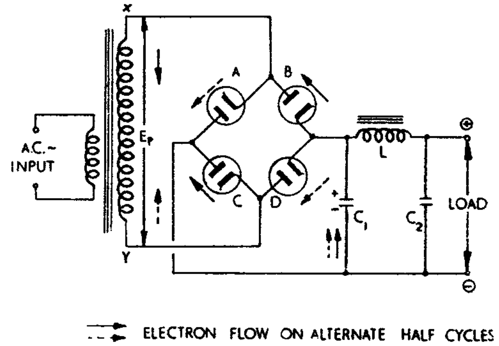


Fig. 14.—FULL-WAVE BRIDGE RECTIFIER CIRCUIT

(*a*) no centre-tap on the transformer secondary is required, and for the same output voltage the input voltage may be halved;

(*b*) the peak inverse voltage is developed across two valves in series and a higher inverse voltage is, therefore, permissible;

(*c*) four valves are now required.

Voltage Doubler

20. Fig. 15 (a) illustrates the circuit of one type of voltage doubler. The valves are so arranged that they conduct on alternate half-cycles to produce across the load a voltage which is approximately *twice* that across the secondary winding of the transformer. When A is positive with respect to B, V_2 is cut off and V_1 conducts to charge C_1 in the polarity shown. When A is negative during the next half-cycle, V_1 is cut off and V_2 conducts to charge C_2 in the polarity shown. Thus, C_1 and C_2 have each charged to approximately the *peak* value of the alternating voltage across the secondary, as shown in (b) and (c) of Fig. 15. Since C_1 and C_2 are *in series* with respect to the load, the d.c. voltage applied to the reservoir capacitor C_3 is equal to *twice* the peak value of the secondary voltage, neglecting the voltage drop across each valve. The peak inverse voltage across each valve in a voltage doubler circuit is equal to twice the peak input voltage. This, and similar systems, may be used in cascade to give four-fold to eight-fold increases in voltage. High voltages, of the order of kilovolts, can thus be obtained, but the current that this type of circuit is capable of supplying is relatively small—up to about 100 mA.

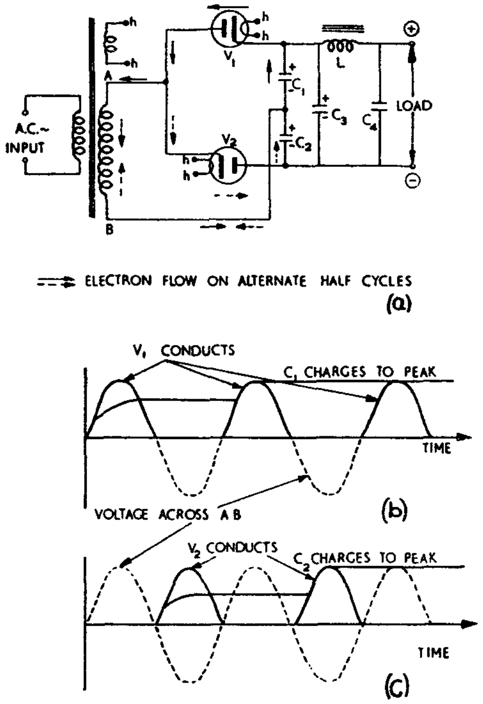


Fig. 15.—VOLTAGE DOUBLER CIRCUIT

Regulation

21. The regulation of a rectifier is defined as the fall or rise in voltage at the output terminals caused by a variation in the amplitude of the current through the internal impedance of the rectifier. Such a variation in current may result from:—

(*a*) a variation in the input supply voltage;

(*b*) a variation in the load.

Voltage regulation is normally expressed as a percentage of the no load voltage. Thus:—

$$\text{Regulation} = \frac{Ep - V_L}{Ep} \times 100 \text{ (per cent).}$$

where Ep = Output voltage off load.
V_L = Output voltage on full load.

Voltage Stabilizers

22. **Gas-filled diode.** In order to improve the regulation of a rectifier whose output voltage decreases as the load current increases, it is normal to insert a voltage stabilizer. A typical arrangement using a cold-cathode gas-filled diode is discussed in Sect. 8, Chap. 4, and the circuit is repeated in Fig. 16(a). It is so arranged that on switching on, sufficient voltage is developed across V to enable it to strike, and the voltage across the load immediately drops to the working value for the diode (Fig. 16(b)). If the load current changes due to a change in R_L, the current through V adjusts itself so that the total current and the output voltage remain substantially constant. Similarly, any change in the input voltage produces a change in current in R_s and V, but no change in the voltage across, nor in the current in, the load resistance R_L.

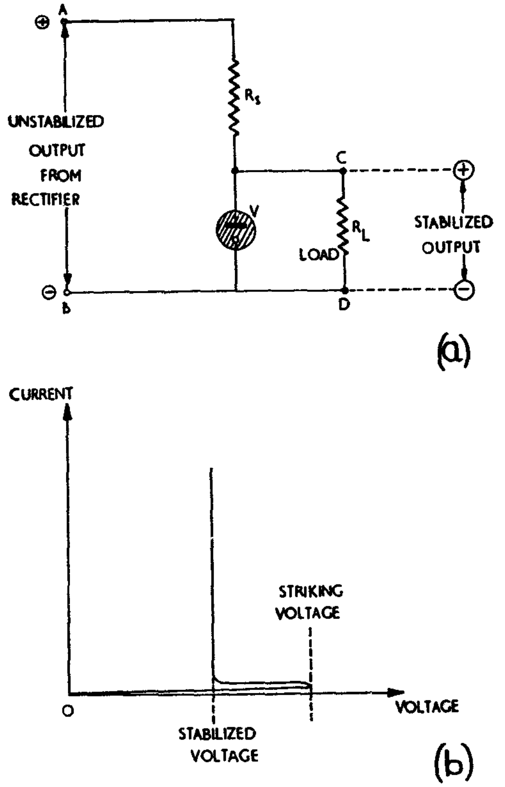


Fig. 16.—COLD-CATHODE GAS-FILLED DIODE STABILIZER

23. **Hard valve stabilizer.** A circuit using a hard valve voltage stabilizer is illustrated in Fig. 17. By using such a circuit, the output voltage will be substantially constant at some chosen value, irrespective of changes in the a.c. supply voltage or in the d.c. load current. The circuit depends for its operation on the variation of the *d.c. resistance* of a triode valve connected in series with the load. A reference voltage E_R, developed across a cold-cathode gas-filled diode, is applied to the cathode of a pentode valve V_2, while a voltage E_g, proportional to the output voltage E_o, is applied to V_2 grid. The potentiometer R_4 is so adjusted that, when the output voltage has the desired value, $E_g = E_R$. If the output voltage *rises*, E_g rises and the consequent rise in anode current in V_2 causes a reduction in V_2 anode voltage and in V_1 grid voltage. By making the grid of V_1 more negative with respect to its cathode in this way, the effective d.c. resistance of the valve is *increased*, and the increased voltage drop across V_1 causes the output voltage to fall to its original level. The action is similar, but with the polarities reversed, for an initial *fall* in output voltage. The potentiometer chain R_3, R_4 and R_5 is chosen so that R_4 can be adjusted to set the output voltage E_o for a given load and permit adequate stabilization about that setting.

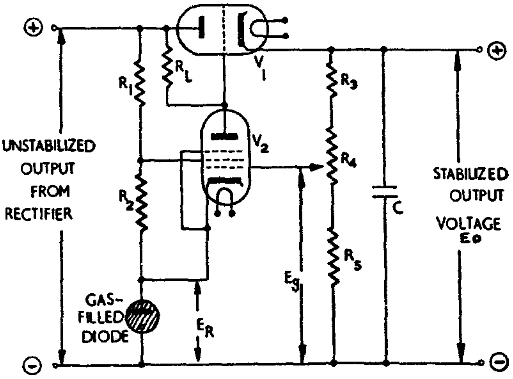


Fig. 17.—HARD VALVE VOLTAGE STABILIZER

Current Stabilizers

24. Since the resistance of a metallic conductor increases with temperature, the current actually established in a wire which has reached a steady temperature is *less* than the value of current calculated from the 'cold' value of the resistance. Current stabilizers are designed to keep the current in a circuit steady at the intended value in spite of considerable voltage fluctuations. Thus, an increase in voltage would result in an initial rise in current, but the consequent increase in temperature, and hence resistance, tends to reduce the current to its original level and stabilization is achieved. The most common form of current stabilizer is the '*barretter*', which is a lamp with a special filament of iron wire in a bulb filled with hydrogen, as shown in Fig. 18(a). The reasons for gas filling are to prevent oxidation of the filament and to permit rapid removal of heat and, therefore, speedy stabilization. The symbol for a barretter is illustrated in Fig. 18(b). Fig. 18(c) shows the voltage-current characteristic of a barretter working at 0·4 amps over a voltage range of 50 to 100 volts. The barretter, depending for its action on the generation of heat, is wasteful of power and its use is usually limited to essential circuits.

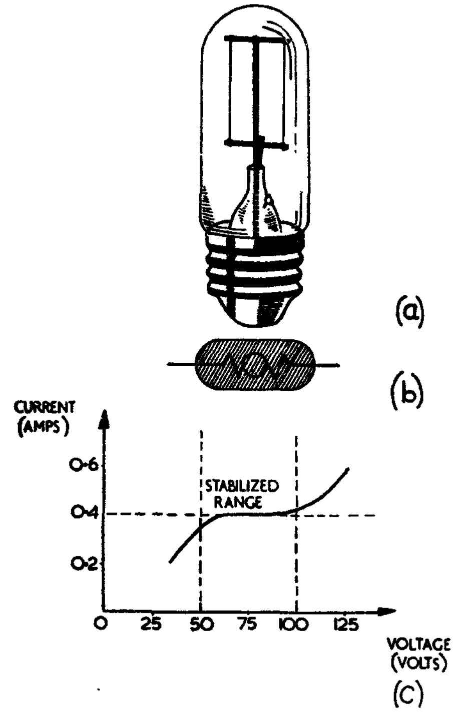


Fig. 18.—BARRETTER, CURRENT STABILIZER

Gas-filled Diode Rectifiers

25. The mercury-vapour diode, and other hot-cathode gas-filled diodes, are discussed in Sect. 8, Chap. 4 where it is shown that such valves have certain advantages and certain limitations in relation to the hard-vacuum diode. The hot-cathode mercury-vapour diode may be used in any of the rectifier circuits considered in the preceding paragraphs, but in practice it is normally used when voltages in excess of 1,000 volts at currents in excess of 0·5 amps are required. The advantages resulting from the use of a mercury-vapour diode are:—

(*a*) Lower voltage drop in the valve so that the rectifier efficiency is increased.

(*b*) Higher power outputs are possible.

26. Disintegration of the cathode emitting surface may occur if the voltage drop in a mercury-vapour diode is allowed to rise above about 22 volts. This may result from:—

(*a*) Over-running the valve; that is, allowing the load current to exceed, even momentarily, the peak rated current for

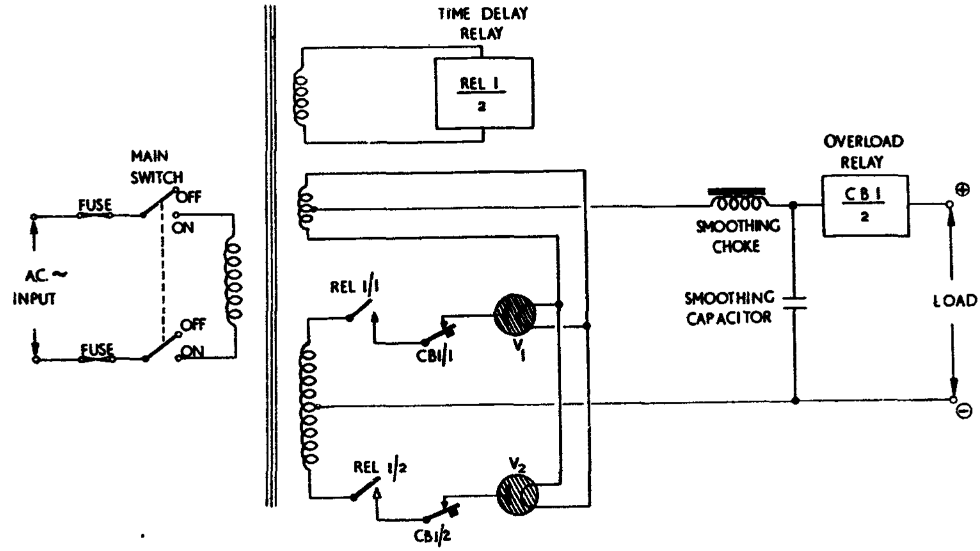


Fig. 19.—MERCURY-VAPOUR FULL-WAVE RECTIFIER CIRCUIT

the valve. *Overload relays* may be inincluded to prevent this.

(*b*) Applying the h.t. supply to the anode before the cathode has been brought to the correct operating temperature. This is normally prevented by including an automatic '*time delay*' in the h.t. supply lead.

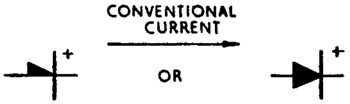
(*c*) Switching off the cathode heater supply before the h.t. is disconnected. Interlocking of anode and heater switches is widely adopted as a safety measure.

27. It is normal to use a choke-input filter with mercury-vapour rectifiers to prevent sudden pulses of charging current (see Para. 13). Such pulses, if excessive, would damage the valve as stated above. A typical circuit for a full-wave rectifier employing mercury-vapour diodes is illustrated in Fig. 19. On closing the main 'on-off' switch, the appropriate voltages are developed across the secondary windings, the l.t. supply and the supply for the time delay relay being immediately available. After a time delay, dependent on the setting of $\frac{\text{REL 1}}{2}$, contacts REL 1/1 and REL 1/2 close to apply h.t. to the anodes of V_1 and V_2. The d.c. output voltage is then available. Should the current taken by the load exceed a certain maximum safety level, the overload relay $\frac{\text{CB1}}{2}$ operates to open the circuit-breaker contacts CB 1/1 and CB 1/2 and disconnect the h.t. supply from the valves. The circuit-breaker contacts are re-set by hand after the cause of the overload has been removed.

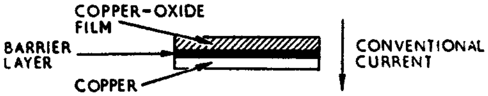
Metal Rectifiers

28. The principle of operation of metal rectifiers is considered in Sect. 8, Chap. 7. There it is shown that a semi-conductor in contact with a metal forms a rectifying element. Various semi-conductors are used in such rectifying elements, but the two types most commonly encountered are copper-oxide and selenium rectifiers.

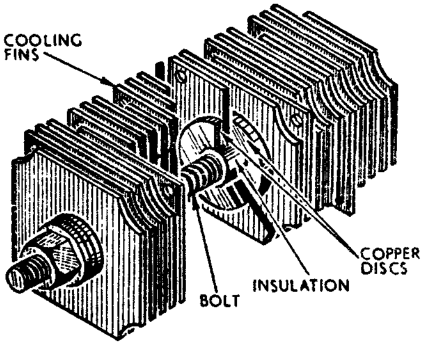
29. **Copper-oxide rectifiers.** The symbols for a metal rectifier are shown in Fig. 20(a). The direction of easy flow of conventional current is from the triangle to the plate. A copper-oxide disc (Fig. 20(b)) can withstand a reverse voltage of about 8 volts. Thus, for rectification of high voltages several discs are connected in series under adequate pressure (about 50 lbs. per sq. in.). The current needs are catered for by choosing the *area* of the elements appropriately, and cooling fins are inserted to dissipate the heat developed. The construction of a typical copper-oxide rectifier is shown in Fig. 20(c). Practical rectifiers range from assemblies of a few large discs for low-voltage high-current operation to assemblies of many small discs for high-voltage low-current operation.



(a) SYMBOLS



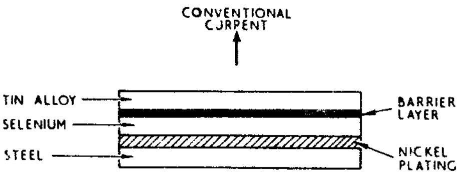
(b) RECTIFIER ELEMENT



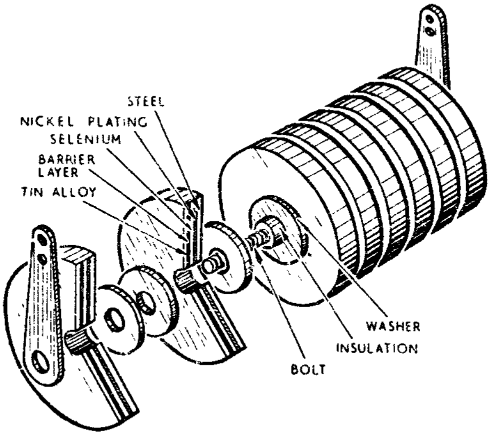
(c) CONSTRUCTION

Fig. 20.—COPPER-OXIDE RECTIFIER

30. **Selenium rectifiers.** Fig. 21 illustrates the construction of a selenium rectifier. The direction of easy flow of conventional current in this case is from the selenium to the tin alloy. Each disc (Fig. 21(a)) can withstand a reverse voltage of about 18 volts and several are connected in series for high-voltage rectification. Fewer discs are required than the number for a comparable copper-oxide rectifier. Further, the selenium rectifier does not require high pressure to ensure contact between the component layers so that the separate elements may be spread out to act as their own cooling fins (Fig. 21(b)). The area of the elements depends on the current requirements.



(a) RECTIFIER ELEMENT



(b) CONSTRUCTION

Fig. 21.—SELENIUM RECTIFIER

31. **Metal rectifier circuits.** Metal rectifiers may be used in lieu of diode valves in any of the rectifier circuits considered in this Chapter. In relation to diode valves, they have certain advantages and certain limitations:—

(*a*) No cathode-heating power is required.

(*b*) They are mechanically stronger.

(*c*) The power they are capable of supplying is limited.

The most commonly-encountered metal rectifier circuit is the full-wave bridge arrangement shown in Fig. 22.

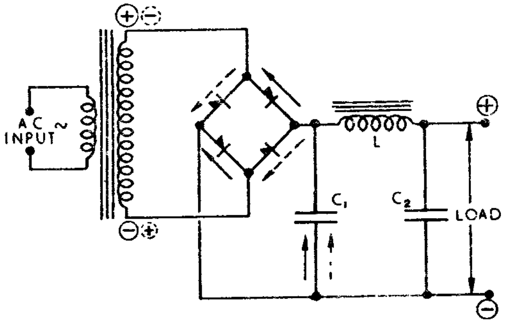


Fig. 22.—METAL RECTIFIER BRIDGE CIRCUIT

Three-phase Rectification

32. A full-wave bridge circuit used for the rectification of a three-phase supply is illustrated in Fig. 23(a). Six rectifiers are required. If the voltage V_1 of the red phase is a maximum in the polarity shown, current flows through the load via rectifiers a and b_1 As the voltage cycle progresses, rectification occurs for the yellow phase via b and c_1 and for the blue phase via c and a_1.

Similar reasoning applies for the alternate half-cycles. The individual phase output voltages are indicated by the dotted lines of Fig. 23(b), and their resultant by the full line. The ripple has, therefore, a frequency *six* times that of the fundamental so that smoothing is more efficient for a three-phase system than for a single-phase system. In addition, the power output from the three-phase system is higher.

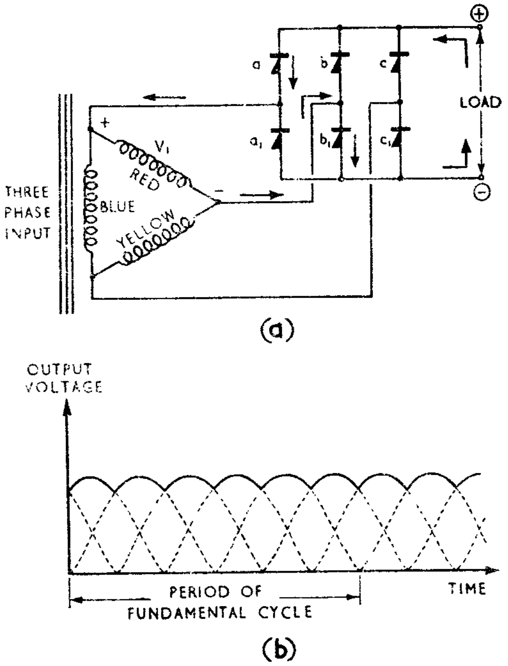


Fig. 23.—THREE-PHASE RECTIFICATION

Mercury-arc Rectifier

33. The mercury-arc rectifier consists of a glass envelope in which there are one or more carbon anodes and a pool of mercury which acts as the cathode. The tube is filled with mercury-vapour at a low pressure. Before

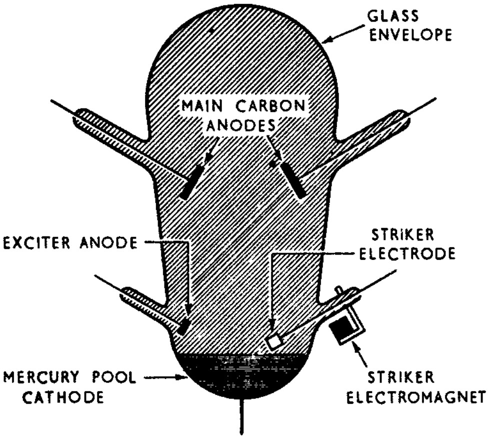


Fig. 24.—BASIC MERCURY-ARC RECTIFIER

the valve can operate, an arc must be 'struck'. This requires an additional electrode, known as the *striker electrode*, in close proximity to the pool (Fig. 24). A p.d. exists between the striker electrode and the pool, and when the striker electromagnet is energised it moves the striker into the mercury, thereby short-circuiting the electromagnet supply. The striker arm is spring biased and as it is lifted out of the pool an arc is created, the emission due to the arc ionizing the gas in the tube to establish a large current to the main anodes. The discharge is in the form of a gaseous arc, which appears as a blue glow. The positive ions produced as a result of ionization move towards the cathode, the upper surface of which is brought to a white heat as a result of the continuous bombardment. For a normal valve, cathode disintegration would rapidly occur, but for a mercury pool cathode no such action can take place and very large currents are, therefore, possible. The hot, bright cathode spot travels in an irregular manner at high speed over the surface of the pool, but if current ceases for even a short interval of time it is sufficient to cool the cathode spot and stop the emission of electrons. To prevent this, and to ensure the maintenance of the arc, an *exciter anode* working from a separate external source is provided. The

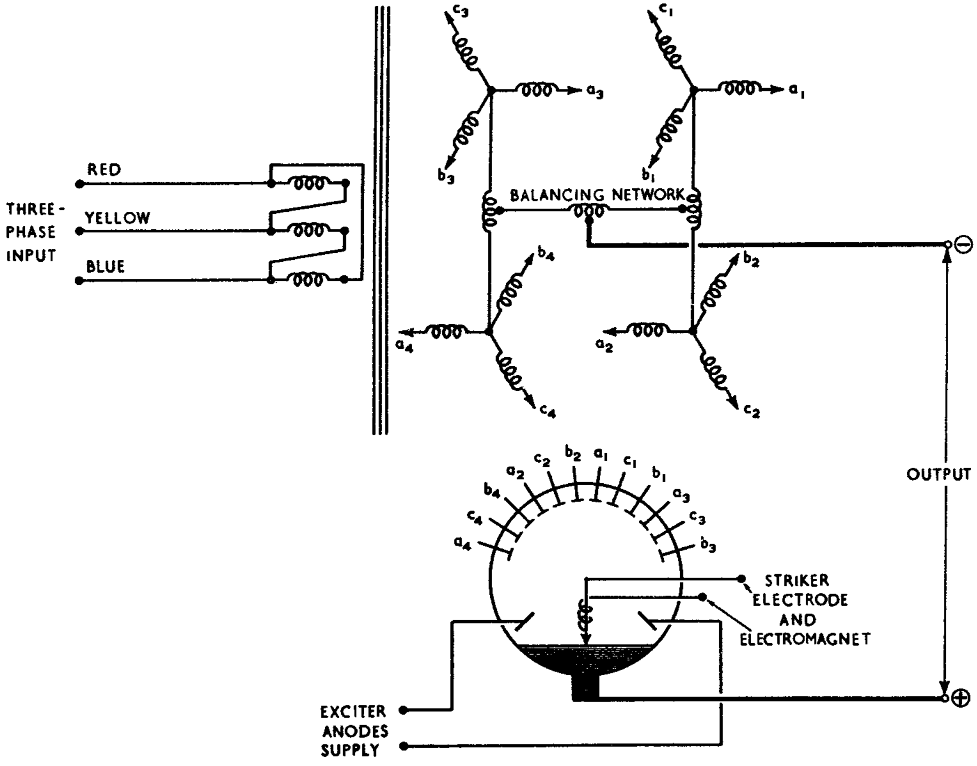


Fig. 25.—TWELVE-PHASE MERCURY-ARC RECTIFIER CIRCUIT

basic form of a mercury-arc rectifier is illustrated in Fig. 24.

34. Mercury-arc rectifiers are used exclusively for heavy power d.c. production where currents of 10 to 1,000 amperes may be required. Single-phase systems using mercury-arc rectifiers are seldom encoutered, polyphase systems being more usual at such high powers. Three-, six-, and twelve-phase systems with three, six and twelve anodes respectively are in general use. Fig. 25 illustrates the basic circuit for a twelve-phase mercury-arc rectifier used in a ground radar equipment. The main three-phase transformer, which has a delta-connected primary, feeds a twelve-phase star-connected secondary winding, between the common points of which a balancing network is inserted. The twelve-phase outputs are connected to the twelve anodes of the mercury-arc rectifier. The arc is struck by the striker electrode and maintained by the exciter anodes as explained in the preceding paragraph. On application of the main supply to the phase anodes, current is established to each anode in turn to give an output that requires very little smoothing. The output obtainable from the system shown in Fig. 25, is of the order of 500 volts at 30 amperes. A photograph of the mercury-arc rectifier used in this system is shown in Fig. 26.

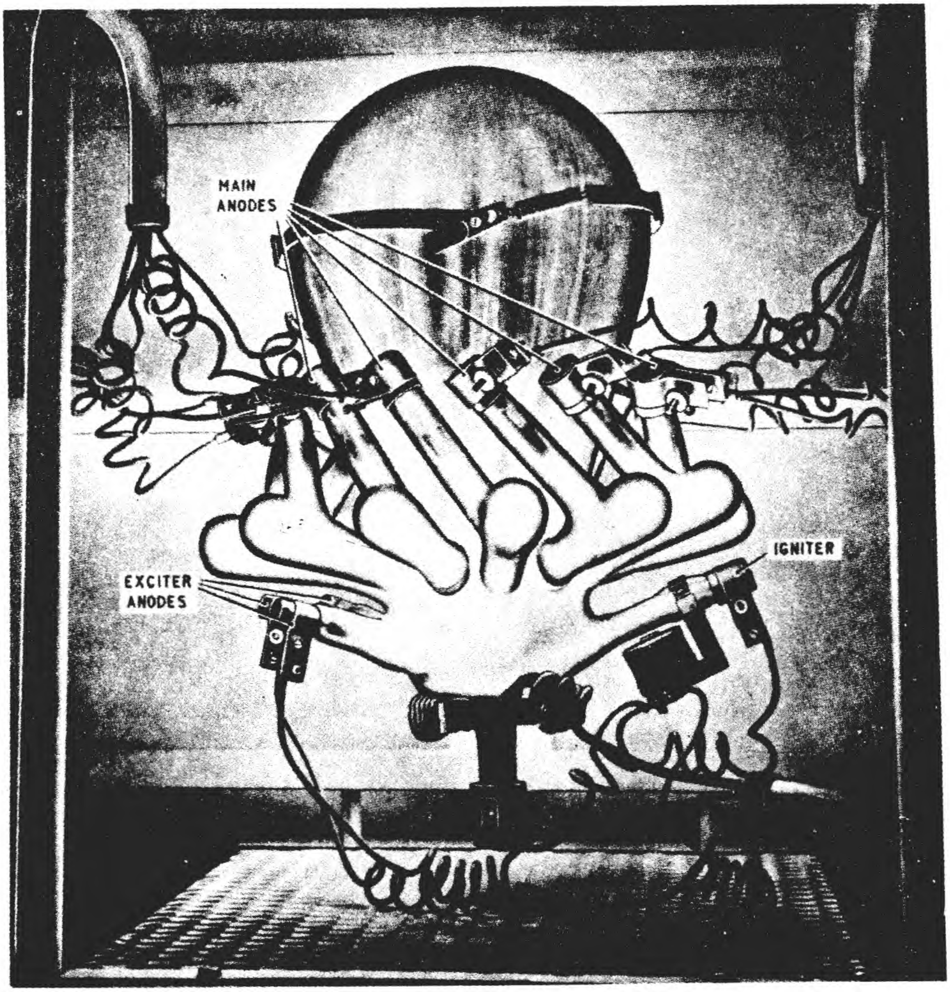


Fig. 26.—TWELVE-PHASE MERCURY-ARC RECTIFIER VALVE

SECTION 10

LOW FREQUENCY AMPLIFIERS

SECTION 10

LOW FREQUENCY AMPLIFIERS

SECTION 10

CHAPTER 1

VOLTAGE AMPLIFIERS

PART 1A, SECTION 10, CHAPTER 1

VOLTAGE AMPLIFIERS

Introduction

1. An *amplifier* is a device in which an input is used to control a local source of power so as to produce an output which is greater than, and bears a definite relationship to, the input. Triodes, tetrodes and pentodes may be used in amplifier circuits, the circuit then being termed a *valve amplifier*. Transistors may also be used. Amplifiers may be classified as :—

(*a*) **Voltage amplifiers,** which are designed to deliver a large *voltage* at negligible power to the control grid of a subsequent valve.

(*b*) **Power amplifiers,** which are designed to deliver *power* at a high efficiency to such devices as loudspeakers and aerial systems.

2. Voltage and power amplifiers may each be sub-divided into the following :—

(*a*) **Wide-band amplifiers,** which are designed to give more or less uniform amplification over a wide frequency range. In this category are :—

(i) *Audio frequency* (a.f.) amplifiers which, in the ideal case, are required to amplify equally at all frequencies in the audio band, i.e., from about 20 c/s to about 20 kc/s.

(ii) *Video frequency* amplifiers which may be required to amplify equally at all frequencies within a band of 20 c/s to about 4 Mc/s.

(*b*) **Narrow-band amplifiers,** which are designed to give amplification over a limited frequency range, the word 'narrow' meaning, in this context, that the difference between the highest and lowest frequencies amplified, is small compared with the mid-band frequency. In this category are :—

(i) *Radio frequency* (r.f.) amplifiers which may be required to amplify over a very narrow band of frequencies of, say, 10 Mc/s ± 5 kc/s.

(ii) *Intermediate frequency* (i.f.) amplifiers which may be required to amplify over a band of 460 kc/s to 470 kc/s.

3. This Section deals with wide-band amplifiers, voltage amplification being considered in this Chapter and power amplification being considered in Chapter 2. Since a.f. voltage amplifiers handle signals at frequencies within the audible range, the characteristics of sound must first be considered.

Sound

4. Sound is a form of wave motion which spreads out in all directions from the source and gives rise to external stimuli accepted through the ear and sense of hearing. The source of sound is usually a vibrating body (such as the human vocal chords) which causes near-by air particles to oscillate about their mean positions in the direction of propagation, the type of wave produced being termed a *longitudinal* wave. When a sound wave is propagated in air, the air itself is *not* passed bodily along in the direction of the wave. The oscillation of the air particles produces successive compressions and rarefactions of the air that move along in the direction of the wave, so that energy progresses in that direction at a speed of approximately 1,100 feet per second. Sound waves striking the ear cause the eardrum to vibrate at the same frequency as that of the source, and the sound emitted is heard.

5. **Characteristics of sound.** The important characteristics of sound are intensity, pitch and quality.

(*a*) **Intensity.** This depends on the *amplitude* of the sound wave. All other factors being equal, the *greater* the magnitude of the vibrations in the medium, the *louder* does the sound appear to an observer.

(*b*) **Pitch.** The pitch of any sound depends on the rate of vibration of the air particles ; that is, it depends on the *frequency* of the sound wave ; the *greater* is the frequency, the *higher* the pitch. The ear can hear sounds of frequencies from about 20 c/s to about 20,000 c/s, the exact audible limits varying from person to person. The pitch of speech is determined by the fundamental frequency of the vocal chords and the latter is about 125 c/s for a normal male voice and about 200 c/s for a normal female voice. Similarly, the pitch of a musical sound is determined by the fundamental frequency at the source, so that the *range* of pitch in an orchestra is much greater than that for speech.

(c) Quality. The tonal quality, or *timbre*, of a sound is determined by the *harmonics* of the fundamental frequency. The sound produced by a tuning fork is a 'pure' tone since it consists of a sinusoidal vibration of one frequency only. In general however, the vibrations corresponding to the tones of musical instruments and speech can be analysed into a fundamental vibration, of a frequency determining the pitch of the note, together with a number of higher frequency harmonics. It is the relative proportions of the various harmonics which distinguish between the tones of two different instruments both playing the same note, or between different people's voices. To maintain the timbre of the original sound in a communication system, a frequency response of about 100 c/s to about 9,000 c/s would be required for speech, and about 20 c/s to about 20,000 c/s for music. However, adequate quality speech and music reproduction may be obtained by eliminating some of the higher harmonics and operating over bands of 300 c/s to 3,400 c/s for speech and 30 c/s to 8,000 c/s for music.

Microphones

6. A microphone is a device for converting sound energy into electrical energy. The a.f. electrical energy so obtained may then be applied as the input to an amplifier. For a microphone to be satisfactory its electrical output should contain all the essential characteristics of the original sound input. Many different types of microphones have been designed to give this and the types used in the Service are considered in Part 2 (Communications). The symbol for a microphone is shown in Fig. 1.

Fig. 1.—SYMBOL FOR A MICROPHONE.

Telephone Receivers

7. The function of the telephone receiver is the reverse of that of a microphone; that is, the telephone receiver is a device for converting electrical energy into audible sound energy. Thus, a telephone receiver may be connected in the output of an a.f. amplifier and to be satisfactory it should reproduce faithfully into sound all the essential characteristics of the a.f. electrical output of the amplifier. Hence, if there is no distortion in the intervening circuit, the sound output from the telephone receiver should be a replica of the original sound applied to the microphone. The telephone receivers used in the Service are considered in Part 2 (Communications). The symbol for a telephone receiver is shown in Fig. 2.

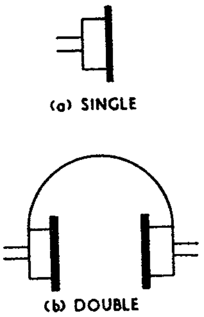


Fig. 2.—SYMBOL FOR A TELEPHONE RECEIVER.

Resistance-capacitance Coupled Amplifier

8. **Action.** The action of a basic triode amplifier is considered in Sect. 8, Chap. 2 where it is shown that a signal applied to the grid of a suitably biased valve causes an amplified version of the signal voltage to be developed across the anode load. In practice, in an a.f. voltage amplifier, the valve is biased under Class A conditons in order to

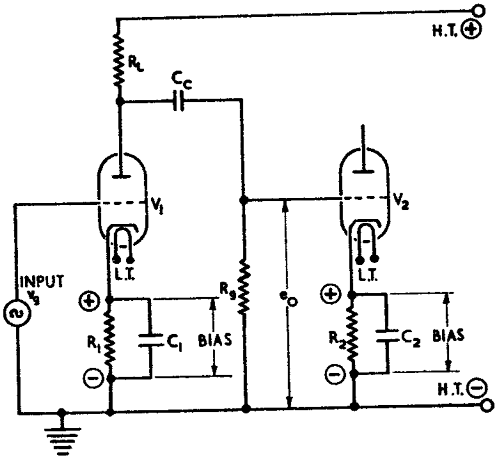


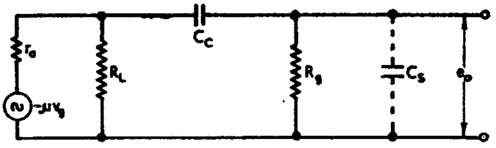
Fig. 3.—RESISTANCE-CAPACITANCE COUPLED AMPLIFIER.

obtain an undistorted output, the amplified signal voltage being applied to the input of the next stage. Only the alternating component of output voltage from the first valve is applied to the next stage and to achieve this the two stages are 'coupled' by a capacitor C_c the function of which is to 'block' the d.c. component of load voltage which is developed by the standing anode current. This Class A a.f. voltage amplifier is termed a resistance-capacitance coupled amplifier, a circuit for which is shown in Fig. 3.

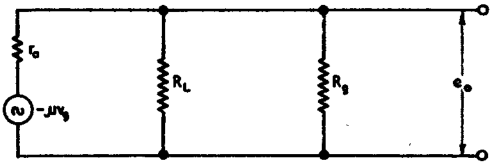
9. Under no-signal conditions ($v_g = 0$) the standing anode current of V_1 develops a d.c. voltage across R_L. Thus, V_1 anode will be at a potential of (H.T. — I_aR_L). C_c and R_g are connected in series across V_1, and C_c charges to the d.c. potential of V_1 anode. There is no p.d. across R_g. Thus, when $v_g = 0$, $e_o = 0$.

10. On the application of a signal voltage v_g, an alternating p.d. of amplitude $(v_g \frac{\mu R_L}{r_a+R_L})$ is developed across R_L, and V_1 anode potential varies by the same amount. This alternating voltage, which is superimposed on the pre-signal d.c. potential, is applied to C_c and R_g in series so that the d.c. component and part of the signal output voltage is developed across C_c, the remainder of the signal voltage being developed across R_g. It is the p.d. across R_g which gives e_o, the input to V_2, and the p.d. across C_c at the signal frequency is kept to a minimum. Thus, the reactance of C_c must be low and since in an a.f. amplifier, frequencies of the order of 20 c/s may be encountered the capacitance of C_c requires to be relatively large. A typical value is between 0·01μ F and 0·1μF (mica type capacitor). The impedance of R_g must be large in relation to that of C_c at the signal frequencies. A typical value for R_g is between 0·2 MΩ and 1 MΩ.

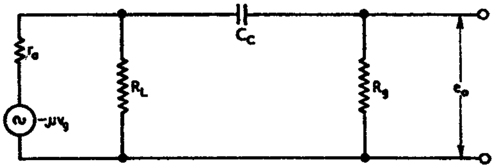
11. **Stage gain.** This is the ratio of the amplified signal voltage e_o at the grid of the second valve to the input signal voltage v_g at the grid of the first valve. This is always *less* than the v.a.f. because of the p.d. developed across C_c at the signal frequencies. In addition, the stage gain varies with frequency. The complete equivalent circuit for Fig. 3 in the constant voltage generator form is shown in Fig. 4(a). The capacitance C_s, which does not appear in the original circuit, represents the inter-electrode capacitances of the valves plus any wiring capacitance. C_s will be of the order of a few pico-farads.



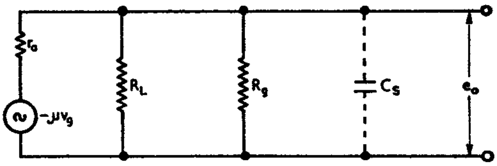
(a) COMPLETE EQUIVALENT CIRCUIT



(b) MEDIUM FREQUENCIES



(c) LOW FREQUENCIES



(d) HIGH FREQUENCIES

Fig. 4.—EQUIVALENT CIRCUITS FOR RESISTANCE-CAPACITANCE COUPLED AMPLIFIER.

(*a*) **Medium frequencies.** Over a range of frequencies in the middle of the a.f. band (say from 100 c/s to 5,000 c/s) the small capacitance C_s will have a high reactance and its effect in shunting R_g is negligible. Further, the large coupling capacitor C_c will have a low reactance which can be neglected in comparison with the resistance of R_g. The resultant equivalent circuit of Fig. 4(b) is a purely resistive circuit and the stage gain is independent of frequency over this middle range. The stage gain is less than the v.a.f. because of the presence of R_g in parallel with R_L and is given by $\frac{\mu R}{r_a + R}$, where $R = \frac{R_L R_g}{R_L + R_g}$. Provided R_g is very much greater than R_L the stage gain practically equals the v.a.f.

(*b*) **Low frequencies.** At frequencies below about 100 c/s the reactance of C_s is high and can be neglected in comparison with R_g, but the reactance of C_c may be quite large and will increase with decrease in frequency. The equivalent circuit for low frequencies is shown in Fig. 4(c). Since C_c and R_g form a potential divider across R_L, e_o and hence stage gain will decrease as the frequency decreases.

(*c*) **High frequencies.** For frequencies higher than about 5,000 c/s the reactance of C_c will be negligible compared with R_g, but the capacitance C_s will form an appreciable shunt on R_L and R_g. The equivalent circuit for high frequencies is shown in Fig. 4(d). The shunting effect of C_s across the load becomes more marked as the frequency increases and the stage gain falls accordingly.

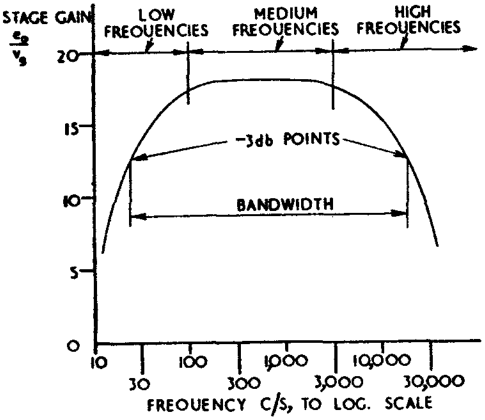


Fig. 5.—GAIN-FREQUENCY RESPONSE OF RESISTANCE-CAPACITANCE COUPLED AMPLIFIER.

12. Fig. 5 shows the way in which the stage gain varies with frequency in a typical resistance-capacitance coupled a.f. voltage amplifier stage. When the frequency of the input is such that the stage gain is 0·707 times that at the middle frequencies, the stage gain has fallen by 3db. The '*3db points*' are the factors used to determine the *bandwidth* of an a.f. amplifier.

13. To increase the stage gain at the middle frequencies in the band, large values for R_L and R_g may be selected. However, at high frequencies the shunting effect of C_s will become appreciable at much lower frequencies than before, and the stage gain will begin to fall sooner. Thus, although the stage gain at the middle frequencies has been increased, this increase is obtained at the expense of a reduction in the frequency range over which the response is flat. There must therefore be a compromise between stage gain and bandwidth.

14. **A.C. load line.** The d.c. load line for a voltage amplifier passes through the h.t. supply voltage point on the horizontal axis and has a slope of $\frac{1}{R_L}$, where R_L is the anode load. For rapid variations of anode voltage however, the coupling capacitor C_c has a low reactance and the equivalent load on the valve is R, equal to R_L and R_g in parallel. So far as a.c. variations are concerned, therefore, the load line has a slope of $\frac{R_L + R_g}{R_L R_g}$ and passes through the the operating point. The a.c. load line is not the same as the d.c. load line, as may be seen from Fig. 6, and it may be necessary to adjust the input to avoid distortion.

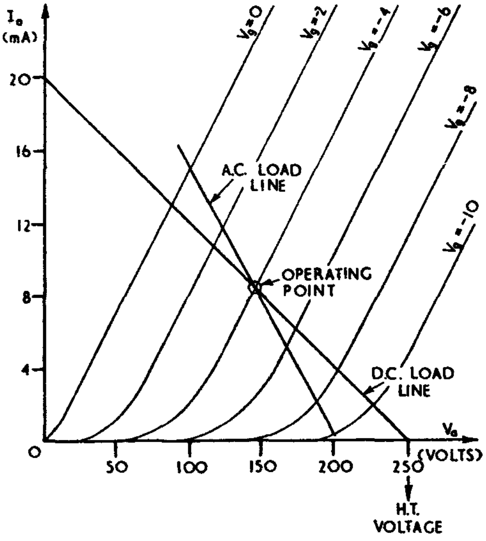


Fig. 6.—A.C. LOAD LINE.

15. **Use of pentodes.** Pentode valves are extensively used in resistance-capacitance coupled amplifiers and give a higher stage gain than is possible with triodes, because of the higher μ of a pentode. The general shape of the gain-frequency response curve is the same as that for a triode, but for a pentode the stage gain at middle frequencies in the band is given approximately by g_m R, where R is equal to r_a, R_L and R_g in parallel.

Choke Coupled Amplifier

16. Fig. 7 shows the simple circuit of a choke coupled amplifier stage. This differs from resistance-capacitance coupling only in that the anode load resistor R_L is replaced by an iron-cored inductance L. Since the impedance of a choke is ωL :—

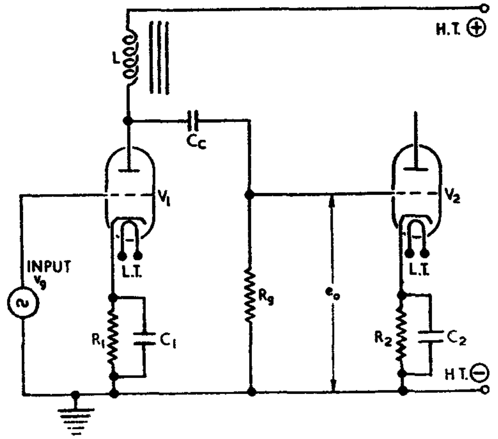


Fig. 7.—CHOKE COUPLED AMPLIFIER.

$$\text{V.A.F.} = \frac{\mu\omega L}{\sqrt{r_a^2 + \omega^2 L^2}}$$

Thus, to provide a high v.a.f. down to low frequencies, a large value for L is required. A typical value is 50 henrys. At medium frequencies within the band, ωL is large and r_a may be neglected in comparison so that at such frequencies :—

$$\text{v.a.f.} \doteqdot \mu.$$

Thus, the gain of a choke coupled amplifier is higher at the middle frequencies in the a.f. band than an amplifier with a resistive load.

17. In contrast to the resistance-capacitance coupled amplifier, the main features of this type of coupling are :—

(*a*) Although the choke offers a high impedance to the alternating components of anode current, it can be made to have a low d.c. resistance. The d.c. voltage drop across it will be small and under static conditions practically the whole available h.t. voltage is applied to the anode, thus enabling lower voltage h.t. supplies to be used. In addition, the applied signal causes the anode potential to vary *above* and *below* the h.t. supply voltage.

(*b*) The choke is an impedance varying with frequency, so that :—

(i) *At low frequencies*, the decrease in load reactance coupled with the increase in the reactance of the coupling capacitor, gives a rapid reduction in stage gain.

(ii) *At medium frequencies*, the increase in load reactance with frequency causes the stage gain to rise slightly.

(iii) *At high frequencies*, the decreased reactance of the shunting capacitance (which includes the self-capacitance of the choke) gives a rapid reduction in stage gain.

Because of these factors, the range of 'flat' response is less than the corresponding region in a resistance-capacitance coupled

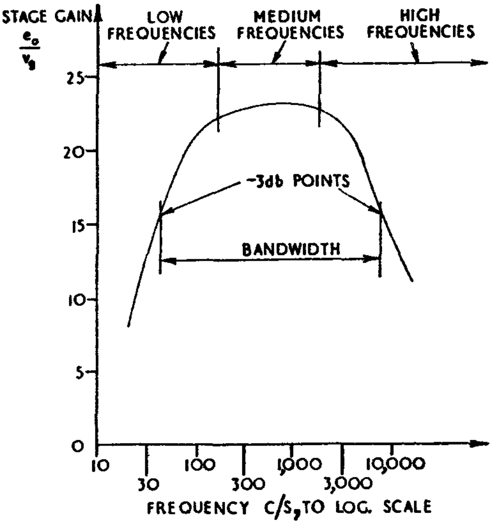


Fig. 8.—GAIN-FREQUENCY RESPONSE OF CHOKE COUPLED AMPLIFIER.

amplifier. For this reason, and that of cost, the choke coupled circuit is rarely used. A typical gain-frequency response curve is shown in Fig. 8.

Transformer Coupled Amplifier

18. The basic circuit of a typical transformer coupled a.f. voltage amplifier stage is shown in Fig. 9(a), the equivalent circuit applicable to the middle frequencies in the band being shown in Fig. 9(b). When a signal v_g is applied in series with the Class A bias voltage, the anode current of V_1 varies according to the signal to induce an amplified voltage in the secondary for application to the next stage. No coupling capacitor or grid resistor is required in this circuit since the grid of V_2 is isolated from V_1 anode by the transformer. From the equivalent circuit, if r_a is low practically the whole of the voltage μv_g is applied across the primary of the transformer, and if the latter has a step-up ratio T, a much larger voltage almost equal to $\mu T v_g$ is applied to the next stage. Thus, with transformer coupling, the stage gain may *exceed* the amplification factor of the valve.

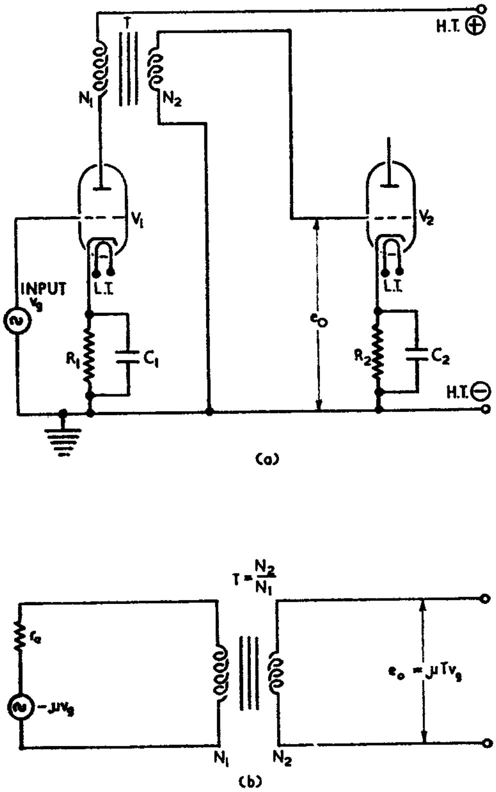


Fig. 9.—TRANSFORMER COUPLED A.F. VOLTAGE AMPLIFIER.

19. The stage gain of a transformer coupled amplifier varies with frequency in a manner similar to that of the other types of coupling considered. At *medium frequencies* in the a.f. band, the stage gain is almost equal to μT. At *low frequencies*, the stage gain decreases due to the decrease in primary inductive reactance with frequency. For a wide frequency range working down to low frequencies, the condition $\omega L > r_a$ must be maintained, and a high value of primary inductance (about 50 H) is necessary. At *high frequencies*, the stage gain tends to fall because of the shunting effect of the capacitance of the transformer windings, and the valve capacitances. At the high frequency end, resonance of the shunt capacitance with the transformer leakage inductance may occur to produce a peak in the response curve as shown in Fig. 10. The amplitude and the sharpness of this peak depends on the Q of the effective series resonant circuit, a large Q giving a large amplitude peak with a sharp frequency cut-off. This may be desirable under certain conditions. To give a reasonably flat response, Q is reduced and

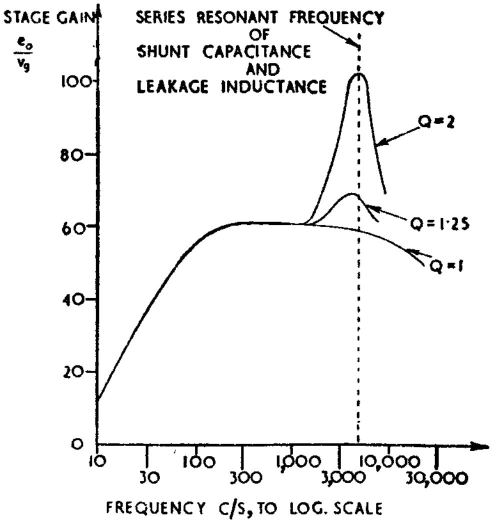


Fig. 10.—GAIN-FREQUENCY RESPONSE CURVE FOR A TRANSFORMER COUPLED AMPLIFIER.

for a given transformer there is a value of r_a which is used to give the desired Q value. A triode valve having an r_a of 10,000 ohms is suitable for most intervalve transformers. Pentode valves are not often used with this type of coupling over the a.f. range because of the high value for r_a in a pentode, but where they are used, the transformer primary is shunted by a resistance to give the desired Q value.

20. The **advantages** of transformer coupling compared with resistance-capacitance coupling are :—

(*a*) The stage gain is higher over the middle frequencies in the a.f. band, being given by μT approximately. In practice, T is a step-up ratio of up to 4 ; it is seldom more than 4 since, with a higher ratio, the effect of the impedance reflected into the primary would be such as to nullify the increased gain.

(*b*) Neither a coupling capacitor nor a grid resistor is required.

(*c*) The d.c. voltage drop across the transformer primary winding is small and, under static conditions, practically the whole available h.t. voltage is applied to the anode, so that the anode potential varies above and below the potential of the h.t. supply.

21. The **limitations** of transformer coupling are :—

(*a*) A high d.c. current may cause saturation of the iron core resulting in waveform distortion and a decrease in the value of primary inductance. To prevent this, an *air gap* is included in the magnetic circuit (see Bk 1 Sect. 7, Chap, 2, Para, 27).

(*b*) Care must be taken in the design of the transformer and its associated circuit to prevent undue peaking at the high frequency end of the range.

(*c*) The transformer should be adequately screened to prevent interference from other sources.

Parallel-fed Transformer Coupling

22. To prevent distortion of the waveform due to saturation of the core, and to ensure a high value of primary inductance and adequate step-up ratio, a parallel-fed transformer coupled circuit may be used as shown in Fig. 11. As there is no d.c. component in the primary winding, the transformer can be made more compact, with a core of very high permeability material (e.g., mumetal) and no air gap. With this arrangement, the alternating component only is applied to the primary, the d.c. component being dropped across C_c. The overall result is a flatter gain-frequency response curve, less distortion and higher stage gain, although there is still a falling-off at very low frequencies because of the coupling capacitor.

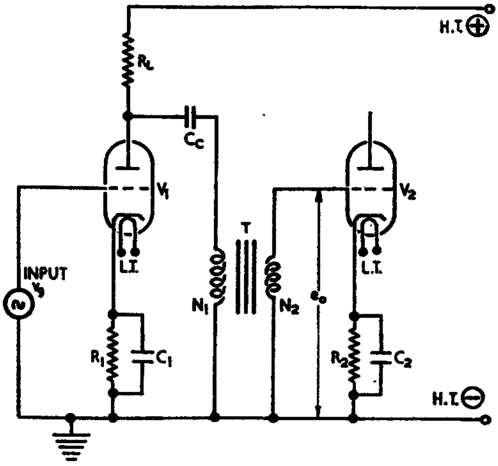


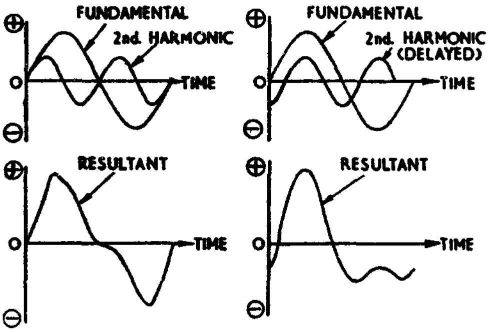
Fig. 11.—PARALLEL-FED TRANSFORMER COUPLING.

Distortion in Amplifiers

23. An amplifier is said to introduce distortion if the input and output signals are *not identical* in waveform. This change in waveform may be occasioned by a number of causes, several of which may be present simultaneously in any given amplifier :—

(*a*) **Attenuation distortion.** This applies to an amplifier where the distortion is caused by a variation of gain with frequency (sometimes called 'frequency distortion'). The types of coupling discussed in the preceding paragraphs all produce attenuation distortion to some extent, as shown by the respective gain-frequency response curves.

(*b*) **Phase distortion.** This occurs when the time taken for a signal to pass through an amplifier varies with frequency. If the phase relationships which the components in the output bear to one another are not the same as the phase relationships between corresponding components in the input, phase distortion has taken place. This is shown in Fig. 12. The phase distortion encountered in an a.f. amplifier is relatively unimportant since the ear is insensitive to small differences in phase. It is, however, important in video frequency amplifiers.



(a) INPUT (b) OUTPUT

Fig. 12.—PHASE DISTORTION

(*c*) **Non-linear distortion.** This is the general name given to a certain type of distortion that occurs when the characteristics of an amplifier are non-linear and dependent on the magnitude of the applied signal at any given instant. Non-linear distortion may be sub-divided into:—

(i) *Amplitude distortion*, where the stage gain of the amplifier depends on the amplitude of the applied signal.

(ii) *Harmonic distortion*, which is due to the production of harmonics in the output when an input of specified amplitude is applied.

(iii) *Intermodulation distortion*, which is due to the production of combination frequencies in the output when two or more voltages of specified amplitude are applied at the input.

Video Frequency Amplifiers

24. A video frequency amplifier is a wide-band voltage amplifier designed to amplify signals within a frequency range of about 20 c/s to about 4 Mc/s with the minimum distortion. In general, the amplified video signals are applied to a cathode-ray tube to produce a picture or a display as in television or radar—hence the term 'video'. It has been shown that the stage gain of a resistance-capacitance coupled amplifier falls at low frequencies due to the effect of the coupling capacitor, and falls at high frequencies due to the effect of the shunt capacitances. These effects must be counteracted if the range of uniform frequency response is to be extended to that required for video frequency amplifiers. Various methods exist for doing this and these are described in detail in Part 3 (Radar).

Direct-coupled (D.C.) Amplifiers

25. A d.c. amplifier is needed if voltages of extremely low frequency are to be amplified ; that is, in cases where the amplitude of the input voltage changes very slowly with time. Resistance-capacitance coupled amplifiers are not practicable here because of the effect of the coupling capacitor, so the coupling capacitor is omitted and direct coupling is employed. A basic circuit of a three-stage d.c. amplifier is shown in Fig. 13.

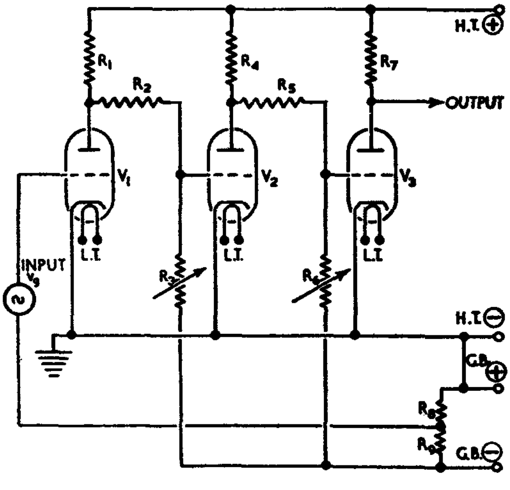
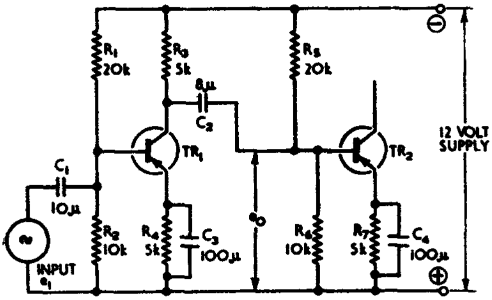


Fig. 13.—D.C. AMPLIFIER CIRCUIT.

The values of the resistors determine the bias voltages on the valves, and to avoid 'drift', high stability components are used. Small percentage changes in the potentials of the supply voltages cause serious alteration of the grid bias, and, to reduce this problem of drift, the power supplies require to be closely regulated. Since d.c. amplifiers are prone to drift, care is taken in design to counteract this.

Transistor Amplifiers

26. **Resistance-capacitance coupling.** The basic action of a transistor amplifier is considered in Sect. 8, Chap. 7. Transistors are particularly suitable for the amplification of small signals in the a.f. range because of their small size, low power requirements and absence of cathode heating power. They may be used in any of the amplifier circuits previously described, provided the supply and bias voltages are suitably adjusted. The circuit of a two-stage resistance-capacitance coupled a.f. voltage amplifier using p-n-p junction transistors is illustrated in Fig. 14. The method of obtaining bias voltages and currents is of considerable practical importance. The collector current is almost independent of collector voltage over the working range and is determined almost entirely by the emitter current. Thus, in the first stage TR_1, the emitter (and collector) current is adjusted by inserting a resistor R_4 of the correct value in the emitter lead. R_4 is by-passed by a large value capacitor C_3 (normally an electrolytic type, typically around $100\mu F$) to prevent the effects of 'negative feedback' into the input circuit (see Chap. 3). The base voltage of TR_1 is controlled by the potential divider R_1 and R_2, and the output is taken across R_3. The values of the resistances R_1 to R_4 are such that the collector is biased negatively with respect to the base (reverse bias), while the emitter has forward bias. This gives the necessary transistor amplifying conditions.



(TYPICAL VALUES SHOWN FOR A.F. WORKING)

Fig. 14.—TRANSISTOR RESISTANCE-CAPACITANCE COUPLED AMPLIFIER.

27. TR_1 is resistance-capacitance coupled to the input of TR_2. Because the input impedance between base and emitter is low, the coupling capacitor must have a large capacitance, otherwise the stage gain will fall sharply at low frequencies. Typical values for coupling capacitors are $2\mu F$ to $10\mu F$. A graph showing the variation of stage gain with frequency is shown in Fig. 15. With the circuit of Fig. 14, the stage gain is of the order or 15, the overall gain being about 200.

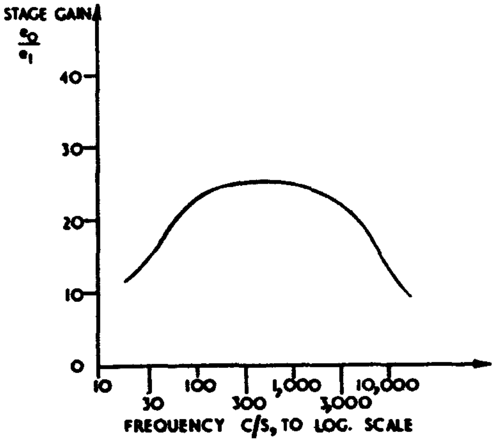


Fig. 15.—GAIN-FREQUENCY RESPONSE CURVE, TRANSISTOR RESISTANCE-CAPACITANCE COUPLED STAGE.

28. **Transformer coupling.** The circuit of a two-stage transformer coupled a.f. voltage amplifier using p-n-p junction transistors is shown in Fig. 16. The input is coupled to

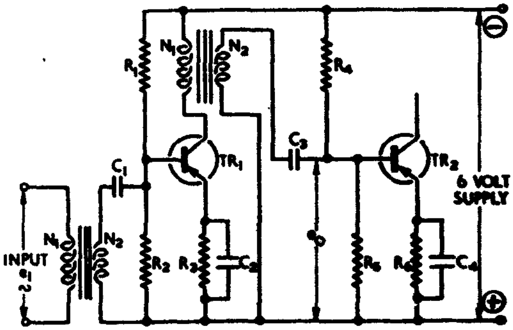


FIG. 16.—TRANSISTOR TRANSFORMER COUPLED AMPLIFIER.

the first stage by a transformer having a step-down ratio of about 30, so that correct matching to the low input impedance of the transistor is achieved. The correct bias voltages for the base and emitter are obtained in a manner similar to that for the resistance-capacitance coupled amplifier described in Para. 26. The intervalve transformer has a step-up ratio of about 3 and the circuit gives an overall gain of about 200. The response is similar to that obtainable from a comparable valve circuit.

SECTION 10

CHAPTER 2

POWER AMPLIFIERS

PART 1A SECTION 10, CHAPTER 2

POWER AMPLIFIERS

Introduction

1. In voltage amplifiers, the sole aim is the production of a large, undistorted signal *voltage* across the anode load. In power amplifiers, the important factor is the *power* developed in the anode load. No device can deliver more power than is supplied to it. Thus, in a power amplifier, the objective is to *convert* as much as possible of the d.c. supply power at the anode into signal power developed in the anode load. The ability to do this is described as the '*efficiency*' of the stage. Power amplifiers may be wide-band or narrow-band amplifiers in the same way as voltage amplifiers, wide-band a.f. power amplifiers being considered in this Chapter. Such amplifiers are used to supply the a.f. power to telephone receivers or loudspeakers in a radio receiver or public address system, and to supply the necessary a.f. power for the modulation of a radio transmitter.

2. It is usual with power amplifiers to couple the load resistance into the anode circuit of the valve by means of a transformer. The reasons for this are:—

(*a*) There is no d.c. power dissipated in the load resistance.

(*b*) The transformer primary has negligible d.c. resistance and the steady d.c. anode voltage is practically equal to the supply voltage for all operating conditions.

(*c*) The transformer turns ratio can be adjusted to give the necessary impedance matching between the load resistance and the slope resistance of the valve (see Bk 1 Sect 7 Chap. 2, Para. 15).

Class A Power Amplifier

3. An amplifier works under Class A conditions when the grid bias is such that the valve operates over the straight portion of its characteristic. The waveform of its output is then the same as that of its input, as is the case with voltage amplifiers. The gain-frequency response of a Class A power amplifier in which the load is coupled to the valve anode by a transformer is similar to that of a transformer coupled a.f. voltage amplifier. There is, therefore, a falling off in gain at both the low and the high frequency ends of the a.f. range, and a tendency to peak at the high frequency end because of self-resonance effects. The range of flat response is typically of the order of 150 c/s to 7,000 c/s, and this is adequate for most purposes.

Triode in Class A

4. The basic circuit for a triode a.f. power amplifier working under Class A conditions is shown in Fig. 1(a), while Fig. 1(b) shows the equivalent circuit applicable to the middle frequencies in the range. By applying the maximum power transfer theorem to Fig. 1(b) it is seen that maximum power is transferred to the load when $r_a = \frac{R_L}{T^2}$; that is, when the

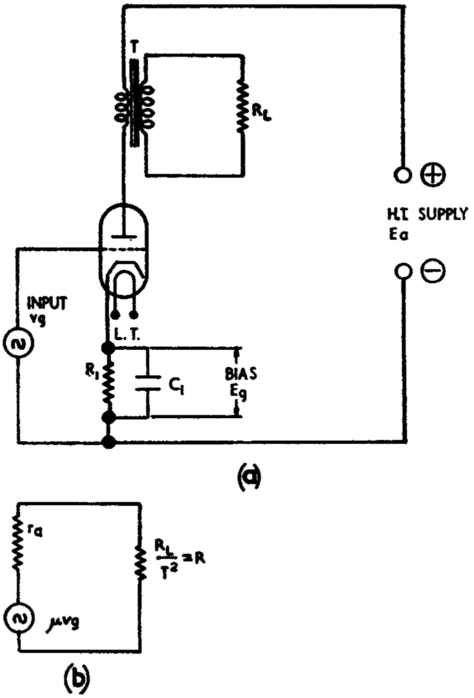


Fig. 1.—TRIODE A.F. POWER AMPLIFIER.

load R reflected across the transformer primary equals r_a. However, in an a.f. power amplifier it is required to determine the conditions that will give maximum *undistorted* power in the load R_L. This is obtained by making the signal v_g the maximum permitted by the straight portion of the dynamic characteristic for the particular value of R, and adjusting the grid bias E_g so that the operating point is at the centre of the straight portion of the characteristic. Repeating this for various values of R, the one that gives maximum output *for an acceptable degree of distortion*, may be calculated. For a triode valve, this occurs when $R = 2r_a$ for 5 per cent harmonic distortion (see Para. 9).

5. **Power output.** Fig. 2 shows the anode characteristics of a typical triode, with the anode dissipation curve superimposed. It is assumed that the characteristics are linear above a fixed minimum anode current I_{min}, the graph being shaded below this level to indicate that this portion is not used. The load line for $R = 2r_a$ is drawn with the 'usable' portion as a solid line ; it is bounded by the curved region of the characteristics at the bottom (shown shaded), and by grid current at the top of the range (i.e., $E_g > 0$); this ensures minimum distortion. The operating point Q lies on the line representing the h.t. supply voltage E_a and is adjusted so that it lies at the centre of the working part of the load line. This is done by varying the fixed bias E_g. In addition, the valve must be chosen so that the operating point lies under the dissipation curve (see Para. 6). On the application of a signal v_g, the anode current swing is from I_{min} to I_{max} about the pre-signal standing current of I_a, and the anode voltage swing is from E_{min} to E_{max} about the h.t. supply voltage E_a. The r.m.s. value of alternating current is $\frac{(I_{max} - I_{min})}{2\sqrt{2}}$, and that of the voltage is $\frac{(E_{max} - E_{min})}{2\sqrt{2}}$. Thus :—

$$\text{Power output} = \frac{(E_{max} - E_{min})(I_{max} - I_{min})}{8} \quad ..(1)$$

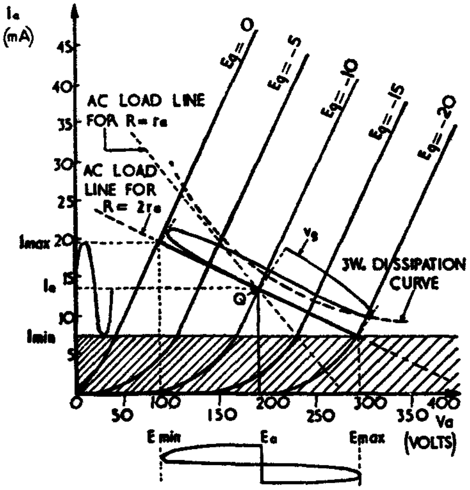


Fig. 2.—OPTIMUM WORKING CONDITIONS.

6. **Dissipation.** In a triode, the product of the d.c. anode voltage and the standing anode current represents the power dissipated in the valve itself in the form of heat. This 'anode dissipation' is not to be confused with the output power developed in the load. The amount of anode dissipation that can be tolerated is limited by the cooling arrangements for the valve, and the valve data sheets give maximum permissible values which are not to be exceeded if damage to the valve is to be avoided. Of the total d.c. power supplied to the valve, part is converted into useful power output and the remainder is dissipated in the valve. The anode dissipation is thus proportional to the output power, and, if a large power output is required, a valve having a high rated anode dissipation must be chosen.

7. **Efficiency.** The efficiency η of a power amplifier represents the efficiency with which the d.c. power supplied by the h.t. is converted into useful a.c. power output. It is given by :—

$$\eta = \frac{\text{A.C. Power Output}}{\text{D.C. Power Supplied by H.T.}} \quad .. (2)$$

$$= \frac{\text{A.C. Power Output}}{\text{A.C. Power Output} + \text{Anode Loss}}.$$

The a.c. power output is :—

$$\frac{(E_{max} - E_{min})(I_{max} - I_{min})}{8}.$$

The d.c. power drawn from the h.t. supply is $E_a I_a$, where E_a is the voltage of the supply and I_a is the standing anode current at the operating point. Thus :—

$$\eta = \frac{(E_{max} - E_{min})\,(I_{max} - I_{min})}{8\,E_a I_a} \times 100 \text{ (per cent)} \quad .. \quad (3)$$

The efficiency is always less than 100 per cent because of the anode dissipation of the valve and, in fact, in a Class A triode power amplifier, the efficiency can never exceed a maximum theoretical figure of 50 per cent ; for E_{min} and I_{min} can never be less than zero, while I_{max} and E_{max} can never be more than twice I_a and twice E_a respectively. In practice, the efficiency of a triode is rarely higher than about 30 per cent at full rated output.

8. **Matching.** The load line for $R = r_a$ is shown dotted in Fig. 2, and, although for the same grid swing the power output is higher than that for $R = 2r_a$, it is seen that the positive and negative swings of anode current are no longer equal so that distortion occurs. Thus, the '*optimum*' load is $R = 2r_a$. In the power output stage of a receiver, the loudspeaker impedance is of the order of a few ohms. The r_a of a triode output valve is much greater, so that the transformer turns ratio must be adjusted to give correct matching. For example, a triode output valve having a r_a of 3,200 ohms is supplying power to a loudspeaker of 4 ohms impedance. For optimum conditions the loudspeaker must reflect an impedance equal to $2r_a$ (6,400 ohms) into the primary. The transformer ratio is given by :—

$$T = \sqrt{\frac{Z_s}{Z_p}} \quad .. \quad (4)$$

$$= \sqrt{\frac{4}{6{,}400}}$$

$$\therefore T = 1 : 40 \textit{ step-down.}$$

9. **Distortion.** Fig. 3 shows a dynamic mutual characteristic for a triode. If the operating point Q is chosen in the middle of the 'straight' portion, this represents Class A working. If the working portion of the characteristic is not absolutely straight, the output waveform will not reproduce accurately a sinusoidal input voltage waveform and the result is harmonic distortion. In particular, for a triode output valve, the output will contain a *second* harmonic component as well as the fundamental.

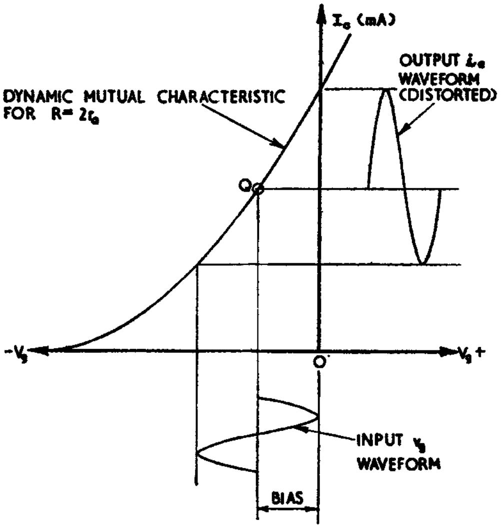


Fig. 3.—DISTORTION IN A TRIODE AMPLIFIER.

10. The distortion for triodes in Class A is usually expressed in terms of the ratio :—

$$p = \frac{\text{Amplitude of Second Harmonic}}{\text{Amplitude of Fundamental}} \quad .. \quad (5)$$

A second harmonic equal to 5 per cent of the fundamental ($p = 0{\cdot}05$) is the usual permissible value on which data published by the valve manufacturers is based, since this is the the smallest amount of harmonic distortion that can be detected by the ear. The percentage of harmonic distortion may be determined graphically. It may be shown that the ratio $\frac{\text{Pos I Peak}}{\text{Neg I Peak}}$ equals $\frac{11}{9} = 1{\cdot}22$ for 5 per cent harmonic distortion. This value must not be exceeded if the distortion is to be less than the maximum permissible value. In Fig. 4, the ratio of positive to negative peak is $\frac{PQ}{QR} = \frac{8}{7} = 1{\cdot}14$ so that the distortion is less than 5 per cent.

Class A Pentode and Beam Power Valves

11. Pentode and beam tetrode valves are fundamentally different from the triode since the anode current of such valves is substantially independent of the anode voltage over the working portion of the characteristics. In this case it is best to

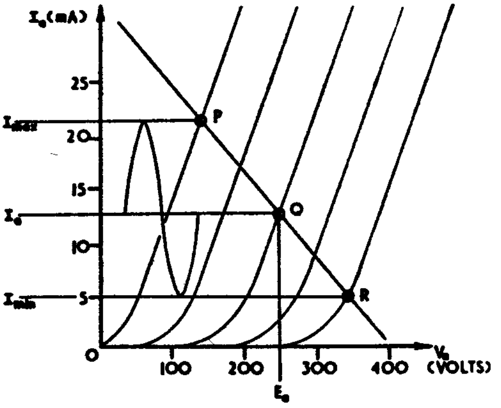


Fig. 4.—CALCULATION OF DISTORTION.

draw the anode characteristics for a given value of screen voltage and fix an operating point. The h.t. supply voltage E_a is known and the operating point is adjusted by means of the grid bias E_g so that I_a is the highest steady current permitted by the rated anode dissipation for the valve. Various load lines are drawn and the one selected is that which gives minimum distortion. This will usually be the line that is drawn from the operating point towards the 'knee' of the characteristic for $E_g = 0$, and is usually such that the impedance reflected into the primary from the load is about $\frac{1}{8}$ r_a to $\frac{1}{10}$ r_a.

12. Fig. 5(a) shows a typical pentode power output stage working under Class A conditions, the anode characteristics with three load lines superimposed being shown in Fig. 5(b). From these curves it may be seen that the load line that gives least distortion consistent with the allowable dissipation is the 10,000 ohms load line. This load therefore affords the best compromise between power output and distortion.

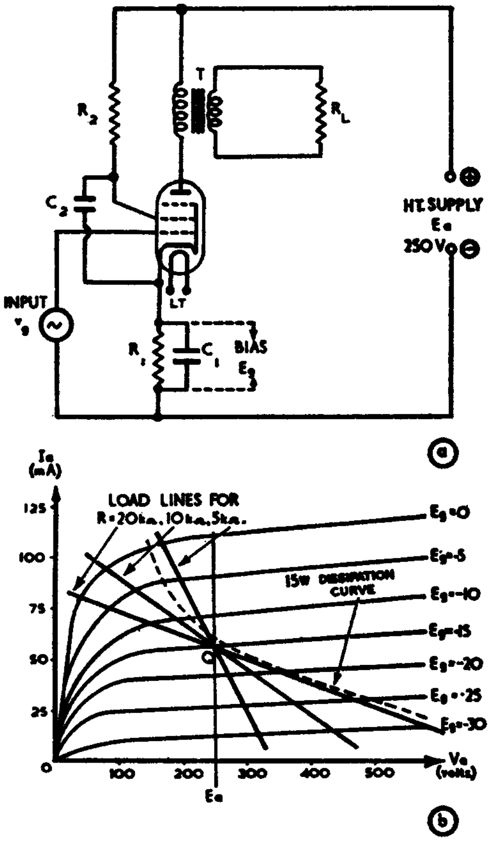


Fig. 5.—PENTODE CLASS A POWER AMPLIFIER.

13. **Efficiency.** In tetrodes and pentodes the product of screen voltage and screen current represents an additional power dissipation known as the *screen dissipation*. The total dissipation is the sum of anode and screen dissipations. However since the amplification factor of such valves is greater than that for a triode the a.c. power output given by equation (1) and also the efficiency are higher. Practical efficiencies obtainable in typical pentode and beam tetrode Class A power amplifiers without severe distortion are in the range 35 to 45 per cent.

14. **Distortion.** The dynamic mutual characteristic of a typical pentode is shown in Fig. 6. As compared with the dynamic characteristic of a triode (Fig. 3), this shows greater curvature. This is because the static mutual characteristics of a pentode are practically coincident, except at low values of anode voltage when the anode voltage then has considerable effect on the anode current (see Sect. 8, Chap. 3, Para. 18). Such a dynamic characteristic gives rise to harmonics higher than the second, and in pentode and beam tetrode valves the main distortion is *third harmonic*. The load line is adjusted to keep this distortion to a low level. The ratio of positive to negative peak in the anode current does not usually exceed 1·4.

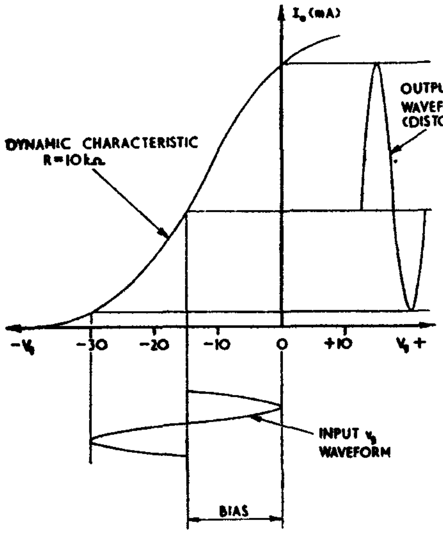


Fig. 6.—DISTORTION IN A PENTODE AMPLIFIER.

Comparison Between Triodes and Pentodes

15. The main features may be summarized as follows :—

(*a*) *In favour of pentodes and beam tetrodes.*

(i) The anode efficiency is higher than that of triodes so that for an available h.t. supply the a.c. power output will be higher.

(ii) The amplification factor is large and therefore the required driving voltage is small.

(*b*) *In favour of triodes.*

(i) The distortion is small and consists mainly of even harmonics which can be eliminated by push-pull amplification (see Para. 19).

(ii) The optimum working conditions are more clearly defined.

Valves in Push-pull

16. Where it is required to supply a greater power output than a 'single-ended' valve of a given dissipation is capable of providing, two valves may be connected in *push-pull* as shown in Fig. 7. The cathodes are at the same potential, and the grids are excited with voltages of ***equal amplitude*** and ***opposite phase*** so that when the potential of V_1 grid is rising, that of V_2 grid is falling and *vice versa*. The outputs of the two valves are combined by means of an output transformer T_2 having a centre tap. The two anode currents are in opposite directions in the two halves of the primary of T_2 and produce opposing magnetic fluxes. The resultant flux at any moment is, therefore, the *difference* between the two. Thus, the output waveform may be obtained by taking the difference between the two anode current waveforms.

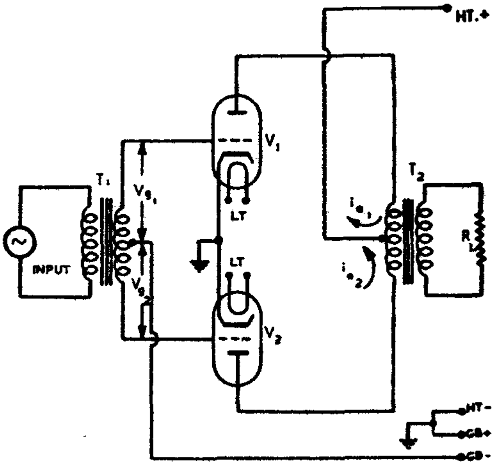


Fig. 7.—VALVES IN PUSH-PULL.

Composite Characteristics

17. With push-pull amplifiers it is usual to prepare equivalent valve characteristics for the pair of valves together. Such characteristics are termed 'composite characteristics', an example of a composite mutual characteristic being shown in Fig. 8. The mutual characteristic of V_2 is inverted and shifted along the X-axis so that its origin lies to the right of the Y-axis by the same amount as the origin of the other curve lies to the left of the Y-axis. The Y-axis represents the operating point on the mutual characteristics and this is determined by the common grid bias. The dotted line is the difference between the two currents and represents the effective current in the output transformer as a function of the input voltage.

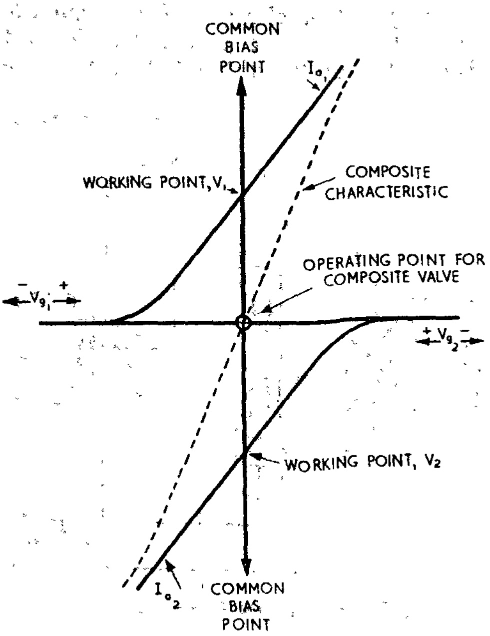


Fig. 8.—COMPOSITE CHARACTERISTICS.

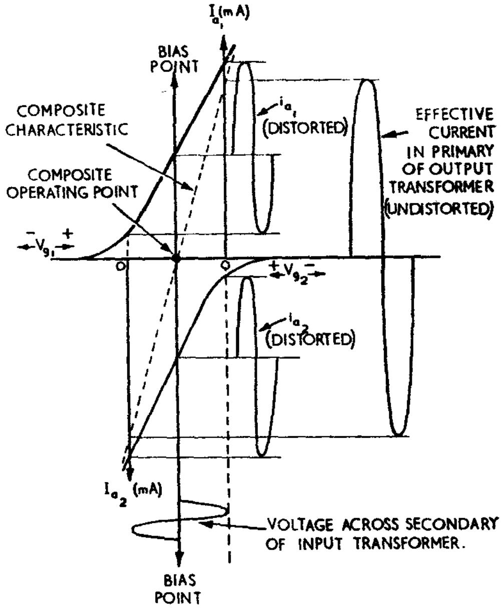


Fig. 9.—VALVES IN CLASS A PUSH-PULL.

18. For two identical valves, the features of the composite characteristic are at once apparent :—

(*a*) The effective transformer current is zero at the operating point.

(*b*) The composite characteristic is much straighter than either of the individual valve characteristics so that the composite valve may be operated down to zero current without undue distortion.

Class A Push-pull Amplifiers

19. **Action.** The common bias is such that each valve operates at the centre of the 'straight' portion of its mutual characteristic. From Fig. 9 it may be seen that the individual outputs of the two valves suffer from harmonic distortion and are 'rich' in second harmonic components. However, the combined output (obtained from the composite characteristic) is entirely free from second harmonic distortion. This is because second and other even-order harmonics cancel out in the output transformer.

20. **Power output.** The input signal that may be applied to the grids of two valves in Class A push-pull is twice that for a single valve. Thus, when each valve in a push-pull amplifier is operated in the same way as a single Class A valve the power output obtainable is *twice* that for a single valve. Owing to the cancellation in even harmonic distortion in the output, Class A valves in push-pull may be permitted to operate under the non-linear region of their characteristic so that the grid drive to each valve may be increased. In this case the maximum power output obtained for a given percentage distortion is *greater* than twice that for a single valve. The optimum anode-to-anode load is *twice* the value required for a single valve, since both valves work over the whole input cycle (360° working) and the valve impedances can be considered to be connected in series. The turns ratio of the output transformer is adjusted for correct matching.

21. **Advantages of push-pull.** The advantages of the push-pull connection, assuming identical 'matched' valves are :—

(*a*) The direct currents in the two halves of the output transformer produce opposing fluxes so that there is no resultant d.c. magnetisation and hence no d.c. saturation of the core.

(*b*) Due to the cancellation of even harmonic distortion a greater power output per valve can be obtained before the permissible distortion limit is reached.

(*c*) In the same way that the even-order harmonics are cancelled in the output transformer, so the fundamental signal frequency and all odd-order harmonics are cancelled in the centre-tapped lead to the h.t. supply. There is, therefore, no signal frequency current in the d.c. power supply and little danger of feedback to other stages via the power supply.

(*d*) If the h.t. supply is derived from the a.c. mains, there is no tendency for mains hum to be introduced in this stage since the hum currents in the two halves of the output transformer balance each other.

Class B Push-pull Amplifier

22. The efficiency of a Class B valve is greater than that of a Class A valve. Thus, where the power output with given valves and a given h.t. supply is required to be large, two valves may be operated in Class B push-pull. The Class B audio amplifier is a push-pull power amplifier in which the valves are biased to 'projected cut-off'; that is, to the point P in Fig. 10 where the straight portion of the dynamic characteristic, when extended, meets the X-axis. A single valve biased in this way would introduce severe distortion since the anode current is directly proportional to the input voltage on the positive half-cycles, while on the negative half-cycles the current is virtually zero. A single Class B valve cannot therefore be used in an a.f. amplifier stage. However, when two Class B valves are connected in push-pull one of the valves amplifies the positive half cycles of the signal voltage, while the other amplifies the negative half cycles. The output transformer then combines these half cycles in such a manner as to give an amplified reproduction of the applied signal. This is shown in Fig. 10. Class B amplification requires a high degree of equality between the two valves to avoid distortion; that is, the valves should be *'matched'*.

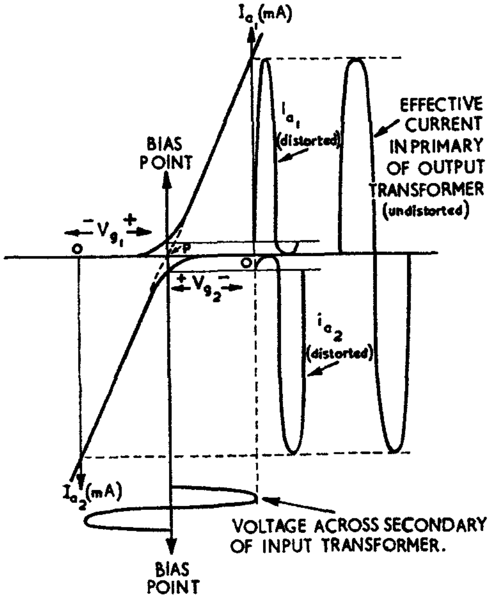


Fig. 10.—VALVES IN CLASS B PUSH-PULL.

23. **Power output.** To obtain maximum efficiency in Class B operation, the grid is allowed to become positive with respect to the cathode on the peaks of the input signal and grid current is established. Where grid current flows the arrangement is termed a Class B_2 power amplifier. This gives an increase in the maximum permissible input voltage and consequently in the output power. The maximum excitation amplitude is over twice that permissible for the same valve when working under Class A conditions, and the power output from the two valves working in Class B_2 push-pull may be from 5 to 8 times that obtainable from a single Class A valve. The optimum value of load reflected between anode-to-anode has to be *four times* the load required for a single valve, since only one valve is working at any given instant. The transformer turns ratio is adjusted accordingly. Class B push-pull amplifiers find their chief use where the amount of a.f. power to be developed is large, as is the case in public address systems and in the modulation of radio transmitters.

24. **Efficiency.** The efficiency of a Class B amplifier stage is high because of the small standing anode current in each valve. The maximum theoretical anode efficiency of such a valve can be shown to be 78·5 per cent compared with the maximum of 50 per cent under Class A conditions. In practice this figure is not realized and the actual efficiency at full rated power output is commonly of the order of 60 per cent compared with a practical Class A efficiency of 30 per

cent. Because of the low mean value of anode current and the relatively high efficiency, the average power dissipation at the anodes is much less than that for Class A operation giving the same power output. It should be noted that if a Class B valve is run at *less* than its full rated output the mean anode current is less and so also is the anode dissipation. The efficiency then rises towards its theoretical value.

25. **Regulation.** The d.c. anode current that a Class B amplifier draws from the power supply, depends upon the amplitude of the signal input. For this reason, the method of obtaining the grid bias voltage by means of cathode biasing cannot be used. The fixed value of grid bias voltage must, therefore, be obtained from a separate supply. The h.t. supply system must have good regulation to prevent variations in the d.c. component of anode current varying the supply voltage by more than a few volts.

26. **Comparison with Class A.**

(*a*) *Advantages.*

(i) Higher anode efficiency.

(ii) Negligible power dissipation when no signal voltage is applied.

(iii) Greater power output available from a given valve and given h.t. supply.

(*b*) *Disadvantages.*

(i) Distortion is greater.

(ii) Operating conditions are more critical.

(iii) The h.t. and bias supplies must have good regulation.

Class AB Push-pull Amplifier

27. Class AB amplifiers are used to obtain efficiencies greater than are obtainable from Class A amplifiers, while at the same time avoiding the critical adjustments necessary for distortion-free operation of Class B amplifiers. The valves are used in push-pull and the grid bias is adjusted to a value intermediate between that required for Class A and that required for Class B. When an alternating voltage is applied to the grid of a single valve operating under these conditions, the ouput is badly distorted. When two valves are used in push-pull however, the distortion is very small as shown in Fig. 11.

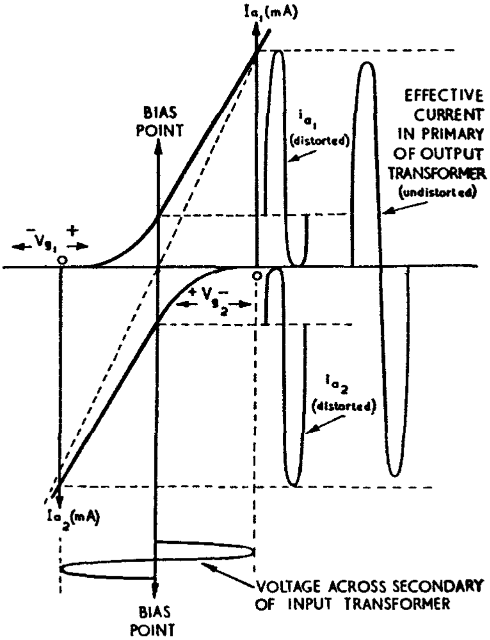


Fig. 11.—VALVES IN CLASS AB PUSH-PULL.

28. The Class AB power amplifier has operating characteristics such as anode efficiency, anode dissipation, and available power output intermediate between Class A and Class B behaviour. The efficiency of 40 to 50 per cent obtainable in practice makes the Class AB power amplifier suitable for use in public address systems where medium power output and efficiency are required and moderate distortion can be tolerated. The possibility of driving the valves into grid current (i.e., Class AB_2 operation) exists in the same way as for Class B amplification.

Triodes in Push-pull

29. Table 1 shows the power output available with triode push-pull systems relative to the power output obtainable from a single-ended Class A power amplifier. For convenience, the practical efficiency at full load and the optimum load impedance reflected across the primary of the output transformer are also shown.

Arrangement	Relative Power Output	Practical Efficiency (*Full Output*)	Optimum Load
Single Class A	1	30%	$R = 2r_a$
Class A Push-pull	2–3	30%	$R = 4r_a$ (anode-anode)
Class AB Push-pull	3–5	45%	$R = 6r_a$ (anode-anode)
Class B Push-pull	5–8	60%	$R = 8r_a$ (anode-anode)

TABLE I. TRIODE POWER AMPLIFIERS.

Pentodes in Push-pull

30. The same power triode, pentode and beam tetrode valves that are suitable for single-ended Class A power amplifiers can be used in all the push-pull arrangements discussed. The advantages obtained when pentodes or beam tetrodes are operated in push-pull are the same as those which have been explained for triode valves. The optimum load is not so easily determined as it is for triodes, but it is chosen so that with a single valve the distortion is mainly second harmonic, since the push-pull arrangement does not eliminate third harmonic components.

Push-pull Input Circuits

31. The two valves of a push-pull system are to be excited by voltages of equal magnitude but opposite phase. Thus, the input to a push-pull stage consists of two symmetrical voltages, balanced with respect to earth. This presents the problem of obtaining balanced exciting voltages from a voltage amplifier that normally develops an output voltage that is *asymmetrical* with respect to earth. One method of exciting a push-pull system is to use an inter-valve coupling transformer having a symmetrical centre-tapped secondary as shown in Fig. 12.

32. Transformer coupling has the disadvantage that transformers possess a limited frequency range. As a result, a number of push-pull exciting arrangements based on the resistance-capacitance coupled amplifier have been devised. One of the simplest of these arrangements, known as a '*phase-splitter*' is shown in Fig. 13. The resistors R_K and R_L are made large and *equal* so that on the application of the input signal voltage, the voltages v_{g2} and v_{g3} applied to the push-

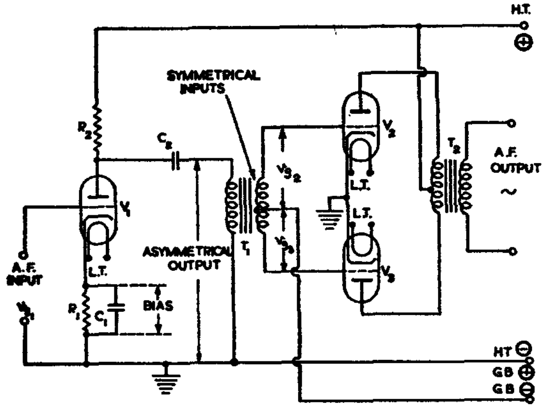


Fig. 12.—TRANSFORMER INPUT TO PUSH-PULL VALVES.

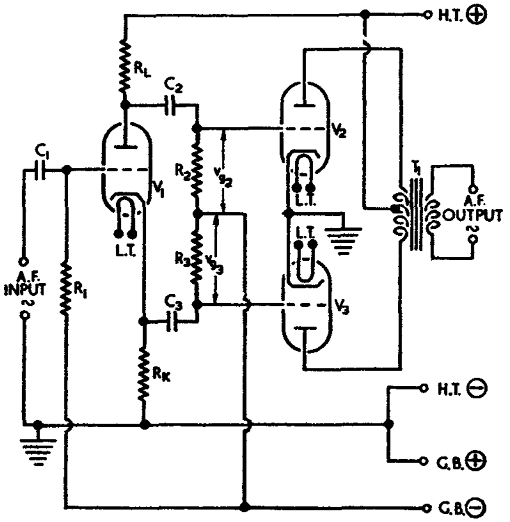


Fig. 13.—USE OF PHASE-SPLITTER TO DRIVE PUSH-PULL VALVES.

pull valves V_2 and V_3 will be equal in magnitude and opposite in phase. The detailed action of this circuit and of other 'paraphase' amplifiers is considered in Part 3 (Radar).

33. Since Class B_2 and Class AB_2 power amplifiers are driven into the region of positive grid voltage, grid current is established on the positive peaks of the driving signal to each valve. Power is therefore developed in the *grid* circuits of the push-pull stage. To supply this power the preceding stage, usually referred to as the '*driver stage*', must itself be a power amplifier, although its power output will be only a small fraction of the total power output from the push-pull stage. The driver stage is usually a single triode power valve working under Class A conditions. The grid circuits of the push-pull stage must have a *low impedance* to prevent the excessive generation of harmonics, caused by the non-linearity of the grid circuit which results from the flow of grid current. Consequently the grid signal voltage of the push-pull stage is usually supplied from the driver stage by means of a *step-down* matching transformer having a centre-tapped secondary to provide the balanced input.

Valves in Parallel

34. Another method of increasing the power output of an amplifier is to arrange two or more valves with their electrodes connected *in parallel*, as shown in Fig. 14. This combination may be replaced by an equivalent valve with the following characteristics :—

Anode slope resistance $= \frac{r_a}{n}$.

Mutual conductance $= ng_m$.

Amplification factor $= \mu$.

The *power handling* capacity is increased by n, where n is the number of valves in parallel. Because of the lowered anode slope resistance, the anode load must also be reduced if the combination is to work as efficiently as a single valve. The turns ratio of the output transformer is adjusted accordingly. When valves in parallel are used as a.f. power amplifiers they can be operated only under Class A conditions, as for a single valve. The advantages are not therefore so great as for valves connected in push-pull.

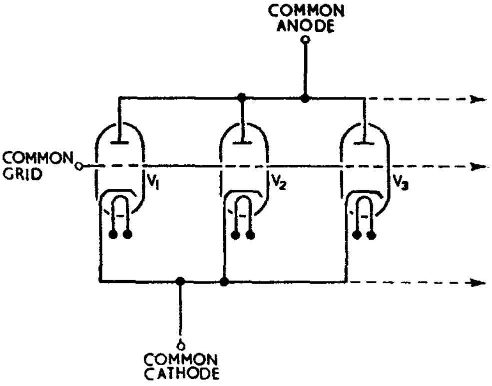


Fig. 14.—VALVES IN PARALLEL.

Transistor Power Amplifiers

35. At their present stage of development, transistors are limited in the amount of power they are capable of handling since, if the rise in temperature becomes excessive, damage to the transistor will result. The maximum output available from a single power transistor operating under Class A conditions is of the order of 3 to 4 watts. For higher power outputs, transistors may be connected in push-pull in much the same way as valves. Fig. 15 shows a transistor driver stage working under Class A conditions coupled by a transformer to a push-pull transistor stage working under Class B conditions. R_1, R_2 and R_3 provide the bias voltages for the driver stage, and R_4 and R_5 provide the Class B bias for the push-pull

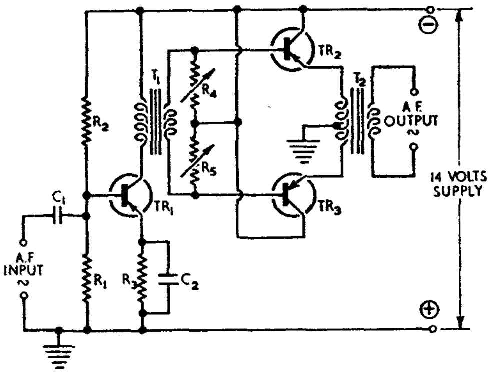


Fig. 15.—TRANSISTOR POWER AMPLIFIER.

stage, the resistance of the output transformer primary providing the emitter bias. Both transformers are step-down to provide the correct matching. With a 14 volt supply the distortion is less than 4 per cent at full rated power output of 10 watts, and the frequency response is flat from about 150 c/s to 7,000 c/s. The current consumption is only 0·75 amperes, so that the efficiency is high.

SECTION 10

CHAPTER 3

FEEDBACK

PART 1A SECTION 10, CHAPTER 3

FEEDBACK

Positive Feedback

1. 'Feedback' is the return of energy from the output to the input of an amplifier. Where the energy fed back is *in phase* with the original input signal, so that it tends to increase the gain of the amplifier, the feedback is said to be '*positive*'. Fig. 1 shows the schematic diagram of an amplifier to which positive feedback is applied. The amplifier has an initial gain A and positive feedback is applied in such a manner that a fraction β of the output voltage is fed back into the circuit in phase with the input signal. The factor β is termed the '*feedback factor*'. Since a valve is a voltage operated device no power need be fed back, and the presence of the feedback network need not affect the output voltage.

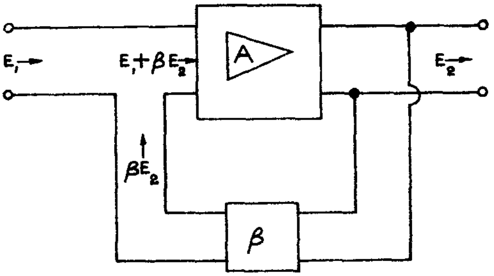


Fig. 1.—POSITIVE FEEDBACK.

2. The original signal applied to the amplifier is E_1, the output voltage is E_2 and a fraction β of E_2 is fed back to assist E_1. Hence, the input to the amplifier is $(E_1 + \beta E_2)$. This signal, amplified A times, produces the ouptut voltage E_2 so that:—

$$E_2 = A\,(E_1 + \beta E_2).$$

Thus, the gain A_o with feedback is :—

$$A_o = \frac{E_2}{E_1} = \frac{A}{1-\beta A} \quad \cdot\cdot \quad (1)$$

The application of positive feedback to the amplifier is seen to increase the gain of the amplifier from A to $\frac{A}{1-\beta A}$. However, when $\beta A = 1$, the gain becomes *infinite*. In fact, the amplifier becomes an '*oscillator*' (see Sect. 12); instability has occurred. Positive feedback is sometimes known as 'regeneration' or 'reaction' and was extensively used in the early days of radio. However, since its use led to a decrease in the stability of an amplifier, this method of increasing the amplification is seldom used nowadays.

Negative Feedback

3. Where the energy fed back from the output to the input of an amplifier is in *anti-phase* with the original input signal, the feedback is said to be '*negative*'. Fig. 2 shows the schematic diagram of an amplifier to which negative feedback is applied. The amplifier has an initial gain A and negative feedback is applied in such a way that a fraction β of the output voltage E_2 is fed back into the input circuit in opposition to the original signal E_1. Hence, the total input to the amplifier is $(E_1 - \beta E_2)$. This signal, amplified A times, produces the output voltage E_2 so that :—

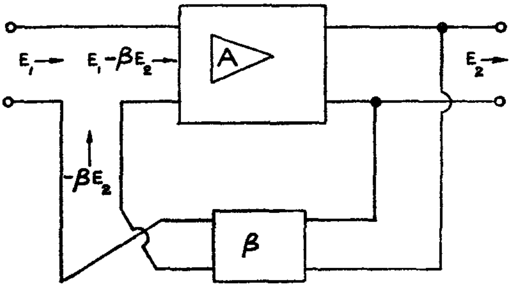


Fig. 2.—NEGATIVE FEEDBACK.

$$E_2 = A\,(E_1 - \beta E_2).$$

Thus, the gain A_o with negative feedback is :—

$$A_o = \frac{E_2}{E_1} = \frac{A}{1+\beta A} \quad \cdot\cdot \quad (2)$$

4. The application of negative feedback to the amplifier reduces the gain of the amplifier from A to $\frac{A}{1+\beta A}$. However, provided sufficient feedback is applied, a great improvement in stability and general performance of the amplifier is obtained.

In fact, the use of negative feedback converts a valve amplifier from a device whose gain depends on numerous factors, such as supply voltage and age of valves, into a precision device whose gain may be made almost independent of these external factors.

Effect of Negative Feedback on Gain

5. Since the gain A_o with negative feedback is $\frac{A}{1+\beta A}$, it is seen that if βA is large compared with 1, the denominator is approximately equal to βA, and the gain A_o becomes:—

$$A_o \simeq \frac{A}{\beta A} \simeq \frac{1}{\beta} \quad .. \quad (3)$$

(provided $\beta A \gg 1$).

Since the gain in this case is now completely independent of A, any factor that may cause a change in A will *not alter* the effective gain A_o of the negative feedback amplifier.

6. **Example.** An amplifier has an overall voltage gain A, without feedback, of 20,000. Let $\frac{1}{50}$th of the output voltage be applied as negative feedback. This reduces the gain of the amplifier to :—

$$A_o = \frac{A}{1+\beta A}$$

$$= \frac{20{,}000}{1+\frac{20{,}000}{50}}$$

$$\therefore A_o = 49{\cdot}9.$$

Suppose that for any reason the inherent gain A of the amplifier drops to 10,000. This big reduction in gain reduces the gain of the negative feedback amplifier to :—

$$A_o = \frac{10{,}000}{1+\frac{10{,}000}{50}}$$

$$\therefore A_o = 49{\cdot}75.$$

Thus, although the inherent gain of the amplifier is reduced by 50%, the overall gain of the amplifier with negative feedback drops only from 49·9 to 49·75. Of course, the gain of the negative feedback amplifier has been considerably reduced in relation to that of the amplifier without negative feedback, but the advantages of negative feedback in maintaining the *gain stability* of the amplifier and in reducing distortion (see Para. 7) more than compensate for the additional stages of amplification required.

Effect of Negative Feedback on Distortion

7. In addition to the effect it has on the stability of gain, negative feedback is used to reduce all forms of distortion produced by an amplifier.

8. **Attenuation distortion.** Distortion due to a variation in the gain of an amplifier with frequency is greatly reduced by the application of negative feedback over the range of frequencies for which βA is large, provided that β is made independent of frequency (i.e., if a simple resistive feedback network is used). This follows directly from the fact that the gain in such a case approximates to $\frac{1}{\beta}$, and if β is independent of frequency so also will be the gain. Fig. 3 shows the gain-frequency response curves of a three-stage resistance-capacitance coupled amplifier, before and after the application of negative feedback for which β was $\frac{1}{200}$. The effects of negative feedback do not become really pronounced until the inherent gain A has become large enough to make βA much greater than 1. This occurs only above 500 c/s in this case. In some circumstances *deliberate* attenuation distortion may be required. The feedback network is then made reactive so that β varies with frequency to give the desired gain-frequency response.

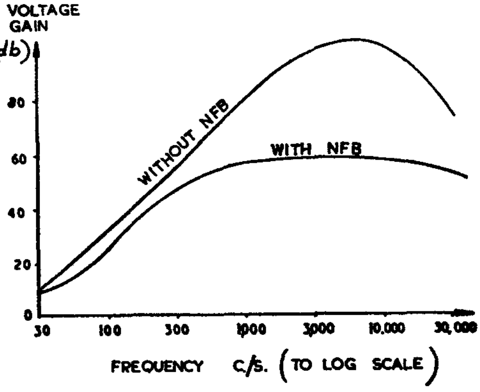


Fig. 3.—EFFECT OF NEGATIVE FEEDBACK ON ATTENUATION DISTORTION.

9. **Non-linear distortion.** As noted in Chap. 2, the non-linear characteristics of a valve may give rise to amplitude distortion, harmonic distortion and intermodulation distortion. In general, the percentage of

non-linear distortion in the output of an amplifier will be *reduced* by the application of negative feedback. For harmonic distortion, when negative feedback is applied, any harmonics in the output will be fed back into the input, will be amplified, and will appear in the output 180° out of phase with the original harmonics. Cancellation of the harmonics may then be achieved. It can be shown that the distortion is reduced according to the relation :—

$$D_o = \frac{D}{1 + \beta A} \quad .. \quad (4)$$

where D_o = Distortion with feedback,
D = Distortion without feedback.

10. **Phase distortion.** There is in general, a phase shift through an amplifier such that the components in the output do not bear the same phase relationships to each other as do the same components in the input. This phase shift is reduced by the application of negative feedback.

11. **Noise.** Noise picked up or generated *within the feedback loop* is reduced by the application of negative feedback according to the relation :—

$$N_o = \frac{N}{1+ \beta A} \quad .. \quad (5)$$

where N_o = Noise with feedback.
N = Noise without feedback.

The noise may be due to valve noise or to hum produced by a poorly smoothed h.t. supply, but provided it is generated *within* the feedback loop it is reduced by the factor shown. Noise picked up *outside* the feedback loop behaves as part of the incoming signal and negative feedback cannot improve the signal to noise ratio under those conditions.

Input and Output Impedances

12. The input and output impedances of an amplifier are also modified by the application of negative feedback. The input impedance is the impedance that appears at the input terminals when 'looking into' the circuit from the signal source, and it is important when calculating the power absorbed by the circuit when it is acting as a load on the signal source. The output impedance of an amplifier is the impedance that appears at the output terminals when 'looking back' into the circuit from the load. For a conventional valve amplifier the output impedance can be considered to be the anode slope resistance of the valve. Output impedance is important when calculating the voltage or power supplied by the amplifier when it is acting as a generator for the load. The effect of negative feedback on input and output impedances depends on the method of applying the feedback, and typical methods are discussed in the succeeding paragraphs.

Voltage Negative Feedback

13. When the voltage fed back into the input circuit is proportional to the *voltage* across the output load, and is in opposition to the input voltage, the arrangement is known as voltage negative feedback. The voltage fed back is usually obtained by placing a potential divider R_1, R_2 across the load as shown in Fig. 4. $R_1 + R_2$ is large compared with the load impedance R_L, and its shunting effect may be neglected. The output voltage E_2 is developed across R_1 and R_2 in series and the voltage fed back is that developed across R_2. In this case the feedback factor β is $\frac{R_2}{R_1 + R_2}$.

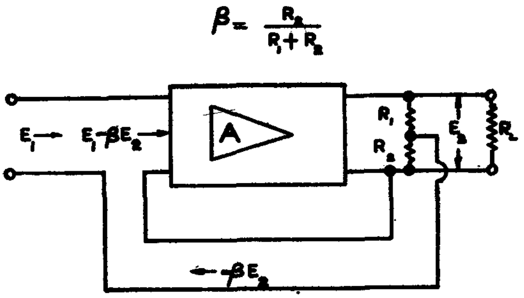


Fig. 4.—GENERAL CASE OF VOLTAGE NEGATIVE FEEDBACK.

14. In Fig. 4, if A represents the gain of a single valve amplifier with a resistance load, then without feedback :—

$$A = \frac{\mu R_L}{r_a + R_L}.$$

By substituting this for A in the expression $A_o = \frac{A}{1 + \beta A}$, the gain with negative feedback is :—

$$A_o = \frac{\frac{\mu}{(1 + \beta\mu)} R_L}{\frac{r_a}{(1 + \beta\mu)} + R_L} \quad .. \quad (6)$$

This is the formula for the v.a.f. of a stage with the amplification factor reduced to $\frac{\mu}{(1 + \beta\mu)}$ and the anode slope resistance reduced to $\frac{r_a}{(1 + \beta\mu)}$. The reduction in gain due to the application of negative feedback is already known, but in addition it is seen that the application of voltage negative feedback reduces the ouput impedance of the stage by the factor $\frac{1}{(1 + \beta\mu)}$. This reduction in impedance is of great importance in the output stage of an amplifier since it facilitates the matching of a high impedance valve, such as a pentode, to a low impedance load.

Practical Voltage Negative Feedback Circuits

15. One method of applying voltage negative feedback to a single stage amplifier is shown in Fig. 5. In Fig. 5(a), R_1 and R_2 form a potential divider across the output load, their values being such that $R_1 + R_2$ is large compared with the load impedance. C_1 is a 'blocking' capacitor to prevent the application of h.t. voltage to the grid, and its value is such that its reactance is negligible over the working range of frequencies. From the equivalent circuit of Fig. 5(b), it is seen that $\beta = \frac{R}{R + R_2,}$ where R is the combined resistance of R_1 and R_3 in parallel.

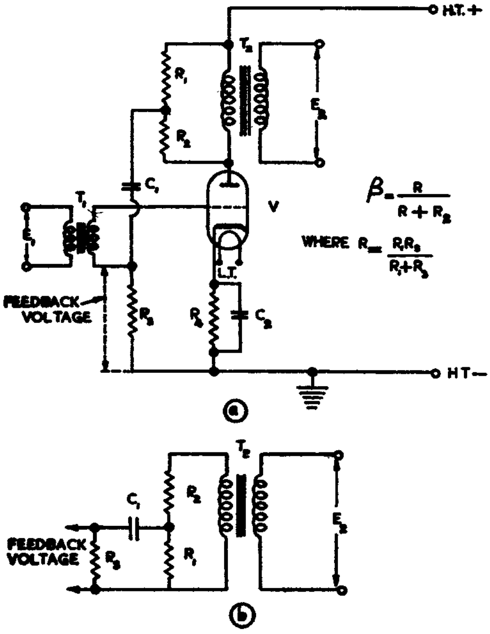


Fig. 5.—SINGLE STAGE VOLTAGE NEGATIVE FEEDBACK.

16. Another popular circuit that is used for the application of voltage negative feedback to a single stage amplifier is shown in Fig. 6.

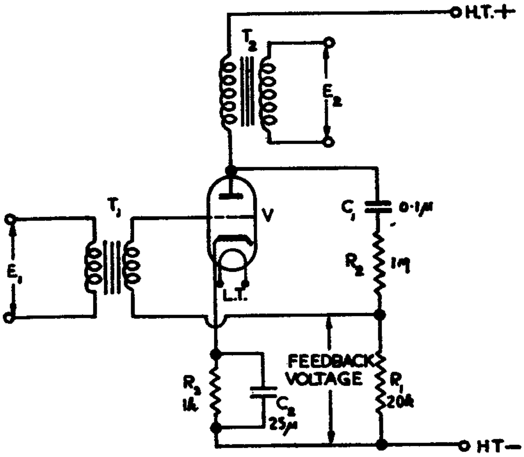


Fig. 6.—TYPICAL VALUES FOR VOLTAGE NEGATIVE FEEDBACK A.F. STAGE.

Component values suitable for an a.f. amplifier are inserted. Over the range of frequencies for which the reactance of C_1 is negligible, $\beta = \frac{R_1}{R_1 + R_2} = \frac{1}{51,}$ using the typical values shown.

17. Feedback in a multi-stage amplifier is often provided over more than one stage. Fig. 7 shows a three-stage amplifier with voltage negative feedback applied over all three stages. R_1 and R_2 form a potential divider across the primary of the output transformer, and over the range of frequencies for which the reactance of C_1 is negligible, $\beta = \frac{R}{R + R_2}$, where R is the combined resistance of R_1 and R_3 in parallel. Although triode valves have been considered in the foregoing, the arguments apply equally well to beam tetrodes and pentodes.

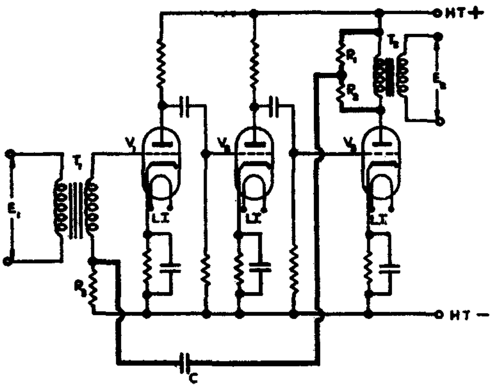


Fig. 7.—VOLTAGE NEGATIVE FEEDBACK OVER MORE THAN ONE STAGE.

Current Negative Feedback

18. Current feedback is said to have been applied to an amplifier when the voltage fed back is proportional to the *current* in the load. This form of feedback is usually obtained by placing a small resistance R *in series* with the anode load R_L, and feeding back the voltage developed across R in opposition to the incoming signal as shown in Fig. 8.

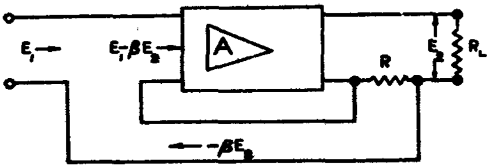


Fig. 8.—GENERAL CASE OF CURRENT NEGATIVE FEEDBACK.

19. The simplest method of applying single stage current negative feedback to a valve employing cathode bias is to disconnect the by-pass capacitor from the cathode bias resistor. The arrangement is shown in Fig. 9(a), and the equivalent circuit in Fig. 9(b). By calculating the output voltage E_2 and the grid voltage v_g in terms of the circuit constants and components it may be shown that the gain $\frac{E_2}{E_1}$ with negative feedback is :—

$$A_o = \frac{\mu R_L}{[r_a + R_K(1 + \mu) + R_L]} \quad ..\ (7)$$

This may be written as :—

$$A_o = \frac{\mu R_L}{r_a{}^1 + R_L}$$

where $r_a{}^1 = r_a + R_K(1 + \mu)$ .. (8)

Thus, the valve after current negative feedback has been applied behaves as if its anode slope resistance has been *increased* by R_K $(1 + \mu)$. Thus, *current negative feedback increases the output impedance* of an amplifier by this factor.

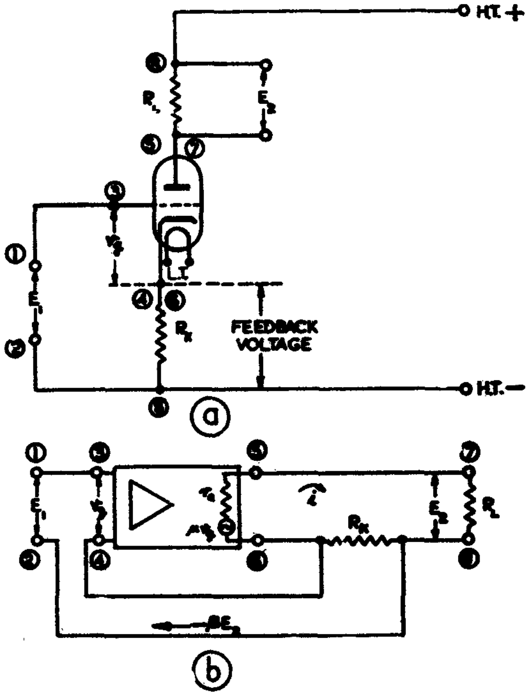


Fig. 9.—BASIC CURRENT NEGATIVE FEEDBACK CIRCUIT.

20. **Example.** A type 6V6 valve has the following constants :—

r_a = 50,000 ohms : μ = 200.

It is used as a voltage amplifier as shown in Fig. 10. It is required to determine the effect on the output impedance Z of removing the capacitor C.

Without feedback :—

$Z = r_a = 50{,}000$ ohms.

With feedback :—

$$\begin{aligned} Z &= r_a + R_K(1 + \mu) \\ &= 50{,}000 + 500\ (1 + 200) \\ &= 150{,}500 \text{ ohms.} \end{aligned}$$

Removal of C thus *increases* the output impedance by 100,500 ohms.

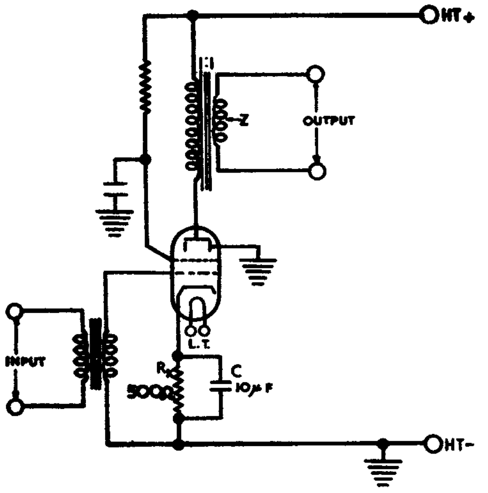


FIG. 10.—EFFECT OF CURRENT NEGATIVE FEEDBACK ON OUTPUT IMPEDANCE.

Practical Current Negative Feedback Circuits

21. Fig. 11 shows current negative feedback applied to a single amplifier stage, where the feedback resistance is :—

(*a*) *Equal* to the cathode bias resistance.

(*b*) *Smaller* than the cathode bias resistance.

(*c*) *Greater* than the cathode bias resistance.

In Fig. 11(a), R_f provides both bias and negative feedback. In Fig. 11(b), R_f provides feedback, and $R_f + R_1$ provides bias. In Fig. 11(c), $R_1 + R_2 = R_f$ provides feedback and R_1 provides bias.

22. Fig. 12 shows a single stage current negative feedback arrangement in which the feedback *decreases* as the frequency increases. The reactance of C is comparable with R over the working range. The amount of negative feedback voltage developed across C and R depends on the frequency, being greatest at the lowest frequency. Thus, the gain of the amplifier *increases* as the frequency increases.

23. Fig. 13 shows a single stage current negative feedback arrangement in which the feedback *increases* as the frequency increases because of the increase in the reactance of L. Thus, the gain of the amplifier *decreases* with increase in frequency.

24. In Fig. 14 single stage current negative feedback has been applied by not decoupling the cathode bias resistors R_{K1} and R_{K2}. Feedback over *two* stages has also been applied since all the current through the load T_2 flows through part of R_{K1} developing a voltage across R_{K1} which is in opposition to the incoming signal.

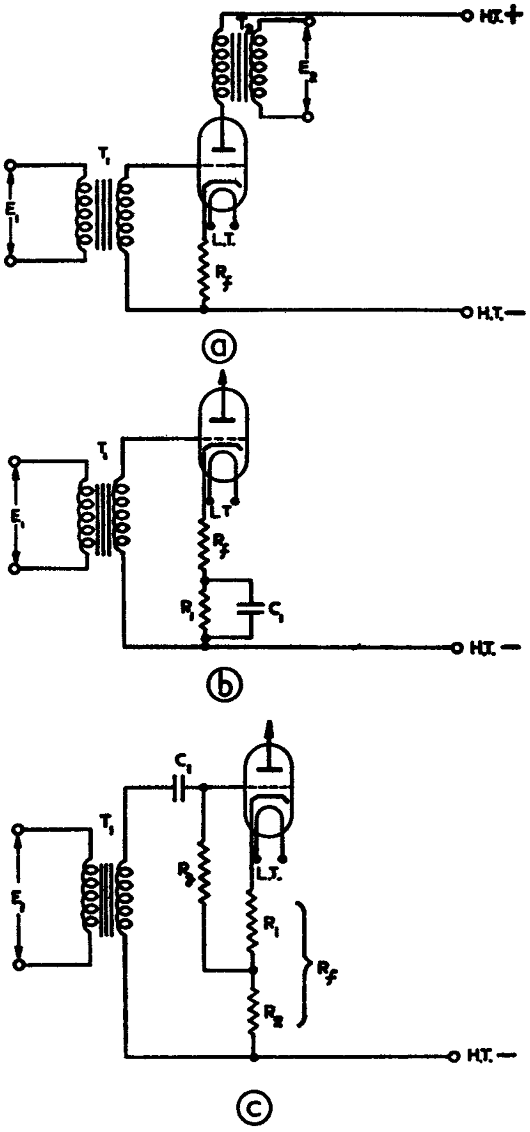


FIG. 11.—APPLICATION OF CURRENT NEGATIVE FEEDBACK TO A SINGLE STAGE.

Composite Negative Feedback

25. The application of voltage negative feedback causes the output impedance of an amplifier to be reduced. The output impedance is increased by the application of

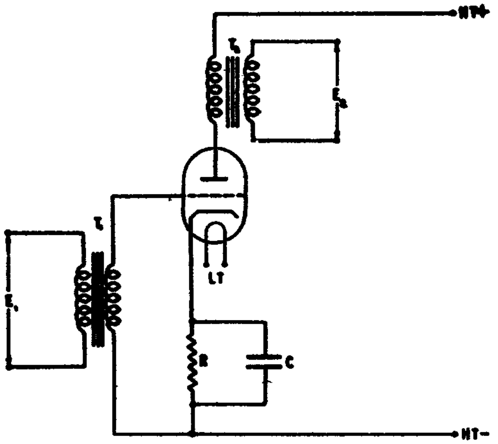


Fig. 12.—DECREASE OF NEGATIVE FEEDBACK WITH INCREASE OF FREQUENCY.

current negative feedback. Thus, by simultaneous application of both types of feedback (termed 'composite feedback') the output impedance of an amplifier can be given any desired value within reasonable limits.

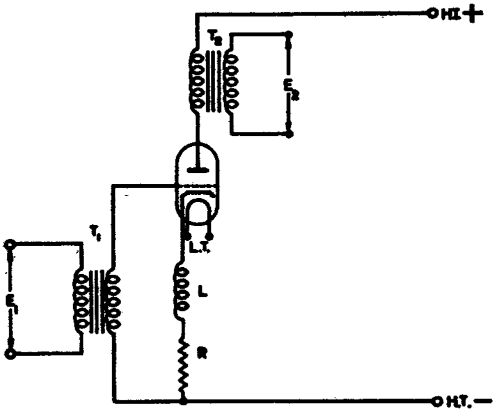


Fig. 13.—INCREASE OF NEGATIVE FEEDBACK WITH FREQUENCY.

Series and Parallel Negative Feedback

26. In all cases of feedback so far discussed, the signal fed back has been applied *in series* with the input signal. This type of feedback is known as 'series feedback'. Its distinguishing property is that the *input impedance is increased* no matter whether voltage, current or composite feedback is employed. These latter considerations affect only the

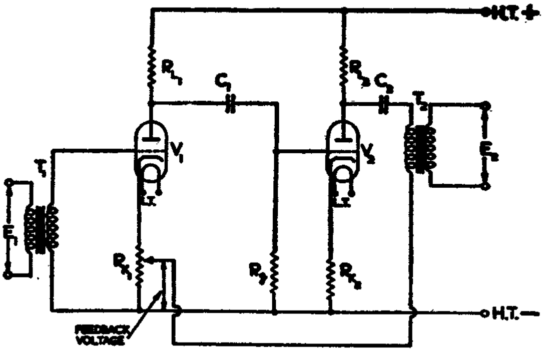


Fig. 14.—CURRENT NEGATIVE FEEDBACK OVER MORE THAN ONE STAGE.

output impedance as noted earlier. Fig. 15(a) shows an amplifier with series voltage negative feedback; the input impedance is increased, and the output impedance is reduced. Fig. 15(b) shows an amplifier with series current negative feedback; the input and the output impedances are both increased. Fig. 15(c) shows that the input impedance of the amplifier considered by itself is *in series* with the feedback resistor so that the input impedance with series feedback is *greater* than that for the amplifier without feedback.

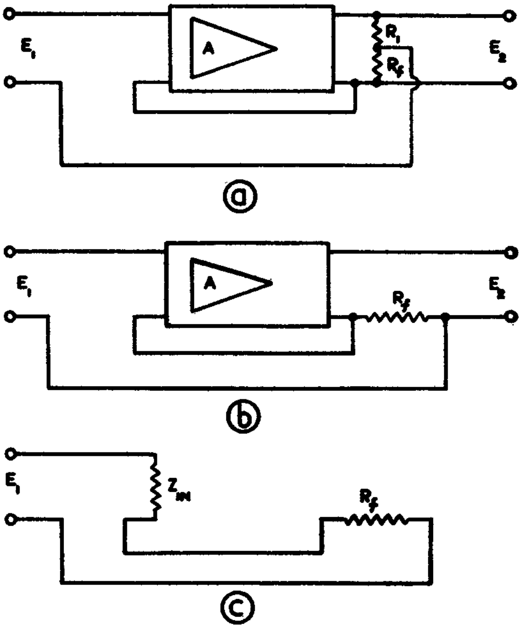


Fig. 15.—SERIES NEGATIVE FEEDBACK.

27. An alternative method of applying feedback is *in parallel* with the incoming signal. This has the effect of *decreasing the input impedance*. Voltage, current or composite feedback may be employed and the method used will determine the *output impedance*. Fig. 16(a) shows an amplifier with parallel voltage negative feedback where the input and output impedances are both reduced. Fig. 16(b) shows an amplifier with parallel current negative feedback; in this case the input impedance is reduced and the output impedance is increased. Fig. 16(c) shows that the input impedance of the amplifier considered by itself is *in parallel* with the feedback resistor so that the input impedance with parallel feedback is *less* than that for the amplifier without feedback. A low input impedance is desirable in an output stage where power is being supplied from a driver stage; distortion in the output stage is thereby reduced (see Chap. 2, Para. 33).

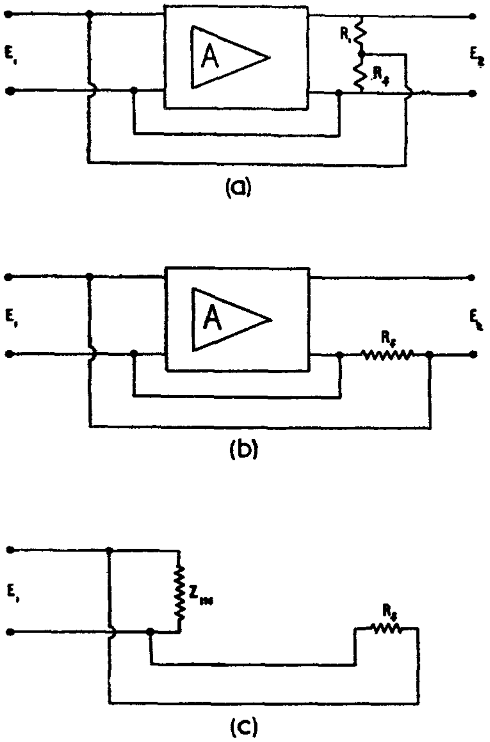


Fig. 16.—PARALLEL NEGATIVE FEEDBACK.

Summary of Negative Feedback

28. The effects of negative feedback on amplifier performance are summarized below :—

(*a*) The gain of the amplifier is reduced to $\frac{A}{1 + \beta A}$ (Para. 3).

(*b*) Any variation in the overall gain, caused by a change in the inherent gain A, is reduced (Para. 6).

(*c*) All forms of distortion are reduced (Para. 7).

(*d*) The gain-frequency response may be adjusted by choice of a suitable feedback network (Paras. 8, 22, and 23).

(*e*) Noise generated within the feedback loop is reduced (Para. 11).

(*f*) The input and output impedances of the amplifier are altered in accordance with Table 1.

	Input Impedance	**Output Impedance**
Series voltage feedback	Increased	Reduced
Series current feedback	Increased	Increased
Parallel voltage feedback	Reduced	Reduced
Parallel current feedback	Reduced	Increased

TABLE 1. INPUT AND OUTPUT IMPEDANCES.

The Cathode Follower

29. In circumstances where an amplifier is required to have a very high input impedance and a very low output impedance, a cathode follower circuit is used. This type of circuit is illustrated in Fig. 17. The circuit is really a special case of the application of *series voltage negative feedback.* The load R_K is in the *cathode* circuit and the *total voltage* developed across the load is fed back to the input in opposition to the incoming signal. The feedback factor β is therefore equal to 1, and there is always an overall *loss* instead of a gain.

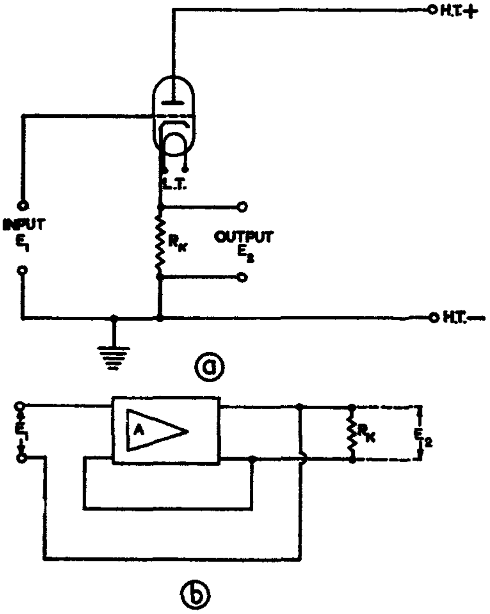


Fig. 17.—THE CATHODE FOLLOWER.

30. A rise in input voltage causes an increase in anode current which, in turn, increases the potential of the cathode with respect to earth. Thus, the cathode potential 'follows' that of the grid and the output voltage is *in phase* with the input voltage. The gain without feedback is given by :—

$$A = \frac{\mu R_K}{r_a + R_K}$$

With feedback, the gain is :—

$$A_o = \frac{A}{1 + \beta A}$$

$$\therefore A_o = \frac{A}{1 + A} \text{ (since } \beta = 1\text{)}.$$

Substituting for A :—

$$A_o = \frac{\dfrac{\mu R_K}{r_a + R_K}}{1 + \dfrac{\mu R_K}{r_a + R_K}}.$$

$$= \frac{\mu R_K}{r_a + R_K + \mu R_K}$$

$$\therefore A_o = \frac{\mu R_K}{r_a + R_K (1 + \mu)} \quad .. \quad (9)$$

Since $r_a + R_K (1 + \mu)$ must always be greater than μR_K, it follows from equation (9) that the gain with feedback is *always less than 1,* although by making μ and R_K large it may approach 1. Since the output voltage v_o is only slightly less than the input voltage v_i, the actual voltage applied to the valve, $v_g = (v_o — v_i)$ is very small and the cathode follower can handle large input voltages.

31. **Equivalent circuit.** By dividing numerator and denominator of equation (9) by $(1 + \mu)$, the gain with feedback may be written as :—

$$A_o = \frac{\dfrac{\mu}{(1 + \mu)} \quad R_K}{\dfrac{r_a}{(1 + \mu)} + R_K}$$

$$\therefore A_o = \frac{\mu^1 R_K}{r_a{}^1 + R_K} \quad .. \quad (10).$$

where $\mu^1 = \dfrac{\mu}{(1 + \mu)}$,

$r_a{}^1 = \dfrac{r_a}{(1 + \mu)}$.

Thus, the cathode follower behaves as a normal amplifier with the amplification factor reduced to $\dfrac{\mu}{(1+\mu)}$ and the anode slope resistance reduced to $\dfrac{r_a}{(1+\mu)}$. If the valve μ is very much greater than 1, as is usual, then $\dfrac{\mu}{(1+\mu)}$ approximates to 1, and $\dfrac{r_a}{(1+\mu)}$ approximates to $\dfrac{r_a}{\mu} = \dfrac{1}{g_m}$. The equivalent circuit is shown in Fig. 18.

32. **Output impedance.** Owing to the application of voltage negative feedback, the

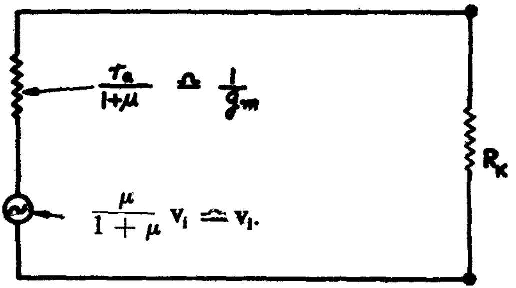


Fig. 18.—CATHODE FOLLOWER EQUIVALENT CIRCUIT.

valve behaves as if its anode slope resistance is reduced to $\frac{r_a}{(1+\mu)} \simeq \frac{1}{g_m}$. The output impedance is equal to this equivalent r_a in parallel with the load R_K as shown in Fig. 18. Thus, the output impedance is slightly *less* than $\frac{1}{g_m}$. Assuming a value for g_m in a triode of 4mA/V, the corresponding output impedance is approximately 250 ohms. The output impedance of a cathode follower is, therefore, *very low.*

33. **Input impedance.** Since *series* negative feedback is used in the cathode follower, the input impedance is *increased* (see Para. 26). This increase is due to two reasons :—

(*a*) Since there is no varying voltage on the anode, the valve inter-electrode capacitance C_{ga} has less effect, because it charges and discharges to the extent of the *grid* signal only and not, as in the conventional amplifier, to the extent of (anode variation + grid variation).

(*b*) Since the grid and cathode are in phase, the valve inter-electrode capacitance C_{gk} charges and discharges to the *difference* between the input and output voltages, whereas in the conventional amplifier it does so to the extent of the input signal.

Expressed as an equation these facts give :—

Conventional amplifier.
Input capacitance $= C_{gk} + (1 + A)\,C_{ga}$.
Cathode follower.
Input capacitance $= (1 - A)\,C_{gk} + C_{ga}$.

The input capacitance of the cathode follower is, therefore, *very low*, and since impedance is inversely proportional to capacitance (i.e., $Z = \frac{1}{\omega C}$) the valve input impedance is *high.*

Example. For a CV1091 pentode, $C_{gk} = 8{\cdot}3$ pF and $C_{ga} = 0{\cdot}007$ pF. When connected in a conventional amplifier circuit the gain A = 99 ; when used in a cathode follower circuit the gain $A_o = 0{\cdot}9$. Then :—

Conventional amplifier.
Input capacitance = 9 pF.
Cathode follower.
Input capacitance = 0·837 pF.

At a frequency of 160 kc/s the input impedances are 111 kΩ and 1·2 MΩ respectively.

34. **Grid bias.** In the basic cathode follower circuit, bias is provided by the load R_K. However, in general, the value of bias so provided is too large, and to reduce the bias it is normal to return the grid leak to a point on the cathode that is positive with respect to earth. Two circuit arrangements are shown in Fig. 19. In both cases the bias

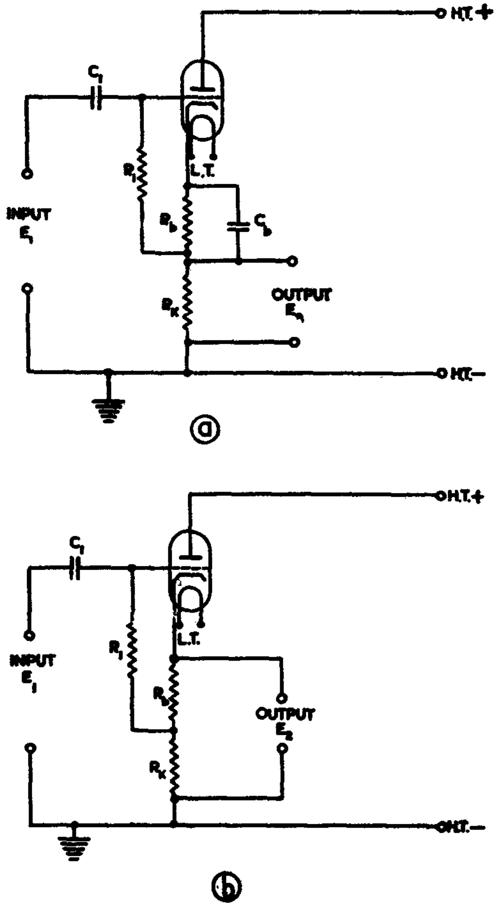


Fig. 19.—BIAS IN THE CATHODE FOLLOWER.

voltage is provided by R_b, but in (a) the output is taken across R_K only, while in (b) the output is taken across $(R_K + R_b)$.

35. **Applications of cathode follower.** This circuit is used :—

(*a*) For insertion between a circuit that has a high output impedance and a circuit that has a low input impedance to give correct matching with minimum distortion.

(*b*) As a 'buffer' stage in an amplifier. The frequency response of a normal resistance-capacitance coupled amplifier falls off at high frequencies due to the effect of the shunt capacitances. A cathode follower with its low input capacitance and low output impedance provides a suitable coupling stage.

36. **Comparison of cathode follower with conventional amplifier.** Table 2 gives a comparison of the main points discussed in the preceding paragraphs.

	Conventional amplifier	**Cathode follower**
Gain	Greater than 1	Less than 1
Phase	Input and output in anti-phase	Input and output in phase
Input impedance	Low to medium value	High value
Output impedance	High value	Low value

TABLE 2. COMPARISON OF CATHODE FOLLOWER WITH CONVENTIONAL AMPLIFIER.

Undesired Feedback

37. Feedback in an amplifier may be deliberate, as in the negative feedback circuits discussed, or it may be random and unwanted. In the latter case random variations in the gain of the amplifier and distortion will result, unless preventative steps are taken.

38. **Miller feedback.** This occurs through the valve inter-electrode capacitance C_{ga} and it will be *positive* feedback if the anode load is inductive (see Sect. 11, Chap. 1). Miller feedback normally results in the production of 'parasitic' oscillations that are above audibility, but may cause distortion. The remedy is to introduce damping by means of a '*grid stopper*' as shown in Fig 20. Parasitic oscillations occur mainly in the power stage of an amplifier.

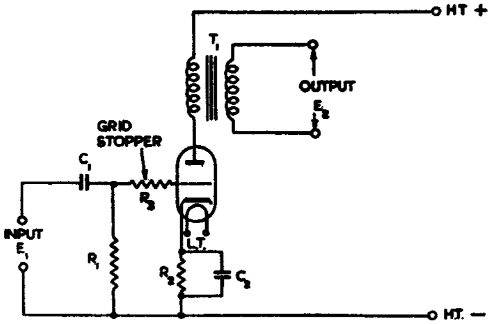


Fig. 20.—REDUCTION OF MILLER FEEDBACK.

39. **Feedback through a common impedance.** An impedance that is common to two or more circuits will always introduce interaction between those circuits, unless it is *decoupled*. Thus, in a cathode-biased valve the cathode resistor is common to both grid and anode circuits, and if it is not decoupled, current negative feedback results. The remedy is to shunt the cathode bias resistor by a large value capacitance to provide a low impedance path for the alternating component of anode current. Very little alternating voltage is then developed across the bias resistor and feedback has been eliminated.

40. Fig. 21 shows the arrangement of a two-valve amplifier with a common h.t. supply. In this case the internal resistance of the supply is an impedance common to both valves. The a.f. anode current of the

output valve flows through the primary of the output transformer and through the h.t. supply back to V_2 cathode, so that an alternating p.d. is developed across the internal impedance of the supply. This alternating p.d. is superimposed on the h.t. potential applied to the *first* valve and constitutes a feedback voltage which is developed across the grid leak of V_2, via R_1 and C_c. The feedback may be positive, in which case instability and tendency to a type of oscillation commonly referred to as '*motor-boating*' may result. To prevent such feedback it is necessary to decouple the resistance of the h.t. supply. The decoupling components are R_D and C_D in the h.t. supply lead to the anode of V_1; preceding stages, if any, are decoupled in a similar manner. R_D and C_D are connected in series across the supply terminals and their junction is taken to V_1 anode circuit. The capacitor has a value, depending on the frequency, such that it offers a low reactance to the alternating component in the h.t. supply. Thus, the alternating component is dropped across the resistor and the p.d. across the capacitor is a *steady d.c.* equal to the h.t. supply. It is this latter voltage that is applied to V_1 anode to ensure that no unwanted feedback voltage reaches V_2 grid. Typical values for R_D and C_D in an a.f. amplifier are 10 kΩ and 1μF respectively, but the degree of decoupling may be increased by increasing both R_D and C_D. There is no need to decouple the last stage in an amplifier since, if all the previous stages are decoupled, nothing can be fed back to it.

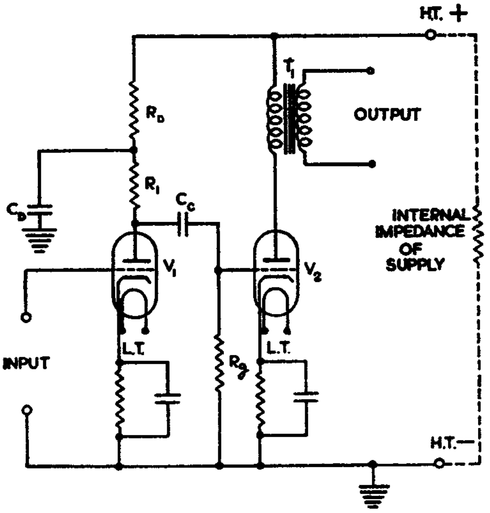


Fig. 21.—FEEDBACK THROUGH A COMMON IMPEDANCE.

41. **Feedback through stray capacitance and inductance.** Inductive coupling between transformers or between chokes, and coupling due to stray capacitances can be prevented only by careful design and adequate screening of components.

Typical High Quality Amplifier

42. The circuit of an amplifier incorporating some of the features discussed in this Section, and capable of providing an output of 10 watts at less than 1% total distortion, is illustrated in Fig. 22.

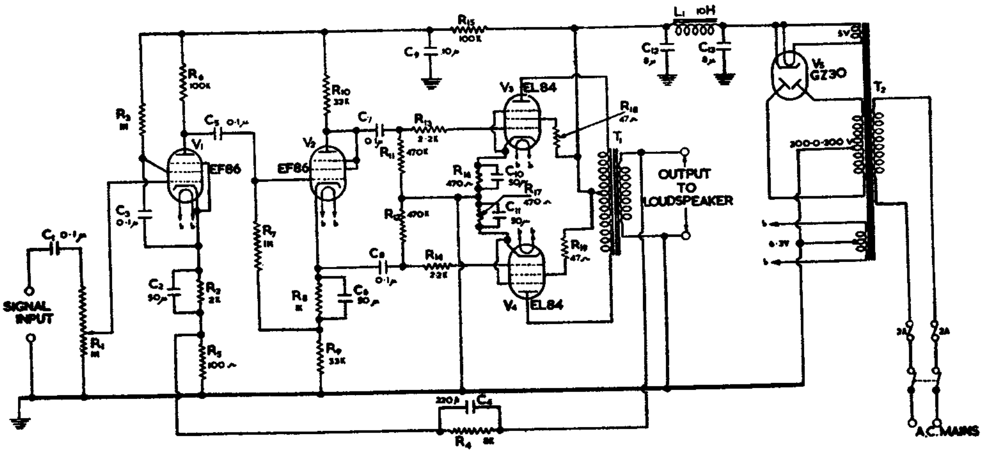


Fig. 22.—TYPICAL HIGH QUALITY A.F. AMPLIFIER.

43. (*a*) **First stage.** V_1 is a high-gain pentode voltage amplifier working under Class A conditions. The cathode bias is provided by R_2. The signal input is applied to V_1 grid via the potentiometer R_1 which operates as a '*gain*' control. Overall voltage negative feedback is applied to this stage by the potential divider R_4, R_5 connected across the secondary of the output transformer. Since R_5 is not decoupled, current negative feedback is also operative. V_1 is resistance-capacitance coupled by R_6, C_5, R_7 to the driver stage V_2, the stage gain of V_1 being approximately 120.

(*b*) **Driver stage.** V_2 is a pentode with screen and suppressor strapped to the anode. This effectively converts the valve into a triode, while still retaining the high mutual conductance of the pentode, and it enables the same type of valve to be used both for V_1 and V_2, thus reducing the 'spares' problem. V_1 operates as a phase-splitter, driving V_3, V_4 in Class A push-pull. Grid bias is provided by R_8; R_9 and R_{10} are the cathode and anode loads respectively. V_2 is coupled by C_7, C_8 and the grid leaks R_{11}, R_{12} to the grids of V_3, V_4. Anode decoupling for the stages V_1 and V_2 is provided by R_{15}, C_9.

(*c*) **Output stage.** V_3, V_4 are high-slope pentodes connected in Class A push-pull. Cathode bias is provided by R_{16} and R_{17}. The anti-phase inputs are developed across the grid leaks R_{11} and R_{12}, and applied via the anti-parasitic resistors R_{13}, R_{14} to the grids of V_3, V_4. R_{18} and R_{19} in the screen connections to V_3, V_4 are also anti-parasitic resistors which damp out any unwanted feedback. The turns ratio of T_1 is such that correct matching between the output stage and the loudspeaker is obtained.

(*d*) **Power Supplies.** V_5 is arranged in a conventional full-wave rectifier circuit to provide an h.t. supply of 320 V at 180 mA, smoothed by the capacitance input filter C_{13}, L_1, C_{12}. An additional winding on the mains transformer provides 6·3 V for the valve heaters.

Negative Feedback in Transistor Circuits

44. To reduce distortion and improve the general performance of a transistor amplifier, negative feedback may be applied in a manner similar to that for a valve amplifier. Fig. 23(a) shows a resistance-capacitance coupled transistor stage in which the decoupling capacitor across the resistance R_3 in series with the emitter has been removed. The resulting current negative feedback reduces the stage gain and percentage distortion, while increasing the input and output impedances of the amplifier. Voltage negative feedback over more than one stage is illustrated in the circuit of Fig. 23(b). The potentiometer R_3, R_4 connected across the output provides voltage negative feedback to the first stage, and reduces the output impedance of the amplifier.

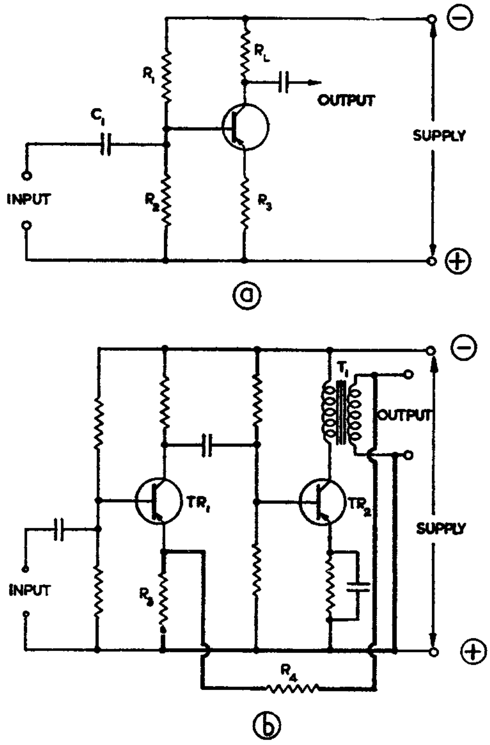


Fig. 23.—NEGATIVE FEEDBACK IN TRANSISTOR CIRCUITS.

SECTION 10

CHAPTER 4

MAGNETIC AMPLIFIERS

PART 1A SECTION 10, CHAPTER 4

MAGNETIC AMPLIFIERS

Introduction

1. In the previous Chapters of this Section, valve and transistor type amplifiers are considered. For certain purposes however, an alternative form of amplifier, known as the *magnetic amplifier*, is used. The controllable element in a magnetic amplifier is a *transductor*, the principles of which are described in Bk 1 Sect. 7. Chap 3. Because of the absence of valves, the magnetic amplifier is characterised by its robustness and insensitiveness to mechanical shock. Its operating characteristics are constant over long periods of time, unlike those of the valve. Thus, the magnetic amplifier is finding increasing application where reliability, long life and robust construction are important. Power gains of up to about 10^6 per stage may be readily obtained without instability. With the magnetic materials at present available, the supply frequencies range from 50 c/s to a few kc/s, and, because of the time constant of the transductor, this sets a limit to the signal frequency. Consequently the magnetic amplifier is well suited (when the output is rectified) for d.c. amplification. Other applications are considered later.

Basic Magnetic Amplifier

2. Fig. 1 illustrates an elementary transductor which consists of two series-connected output windings A and B, and a control winding C. The output windings are connected in *phase opposition* so that no alternating voltage is induced in the control circuit by transformer action from the output windings.

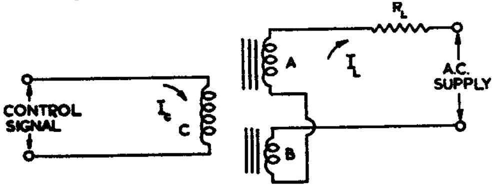


Fig. 1.—BASIC MAGNETIC AMPLIFIER.

3. When the control current I_c is zero, the cores are unsaturated and, because of the high impedance of the output windings to the applied a.c., the output current I_L in the load R_L is small. As the d.c. control current is progressively increased in either direction the cores tend towards saturation and, because of the resultant decrease in the impedance of the output windings, the output current increases (see Bk 1, Sect. 7, Chap. 3, detailed operation). A graph showing the relationship between the output current and the control current is illustrated in Fig. 2.

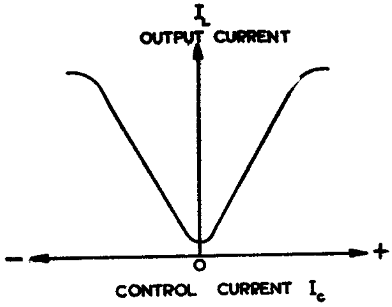


Fig. 2.—OUTPUT CHARACTERISTIC OF SIMPLE TRANSDUCTOR.

Amplification is obtained since a relatively small variation of control current gives a large variation of output current. Thus, the action of the magnetic amplifier is such that a small direct current in one winding controls a comparatively large a.c. power in the load circuit.

Need for Bias

4. A transductor having an operating characteristic similar to that of Fig. 2 cannot discriminate between control signals of different polarity, and the sign of the output remains the same for positive and negative control currents. Furthermore, the most convenient and most sensitive operating point of a transductor does not coincide with zero control current. However, a 'sensed' output and increased sensitivity can be obtained by *biasing* the transductor so that its operating point lies in the middle of one of the linear portions of the characteristic.

5. Bias is obtained by providing the transductor with an additional d.c. winding around both cores and by supplying this bias winding with current from a d.c. source.

The d.c. bias current must be of correct value and polarity to give the required sense of output. The bias is obtained either by rectifying the a.c. supply or from an external d.c. bias supply as shown in Fig. 3(a). The resultant displacement of the operating point is shown in Fig. 3(b). In this instance the transductor output current will increase with a positive control signal, and will decrease with a negative control signal.

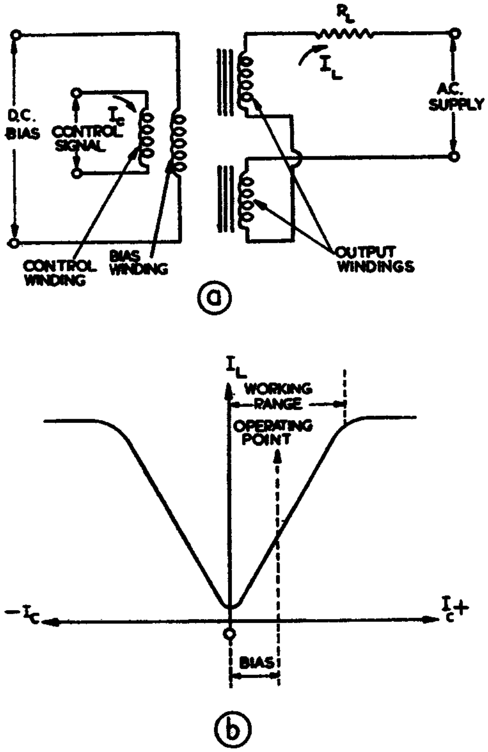


Fig. 3.—BIAS AND ITS EFFECT.

Feedback

6. The sensitivity of a magnetic amplifier may be greatly increased, and its performance improved in certain respects, by feeding a proportion of the (rectified) output signal back to the input circuit in such a direction as to assist the control signal. This *positive feedback* may be applied to a transductor by means of an additional 'feedback winding'. A typical circuit arrangement is shown in Fig. 4(a). The current in the output windings is rectified by a full-wave bridge rectifier so that it flows through the feedback winding in the same direction at all times and thus establishes a magnetic flux which assists that due to the d.c. control signal. The magnitude of the feedback is proportional to that of the output current and, hence, that of the input signal.

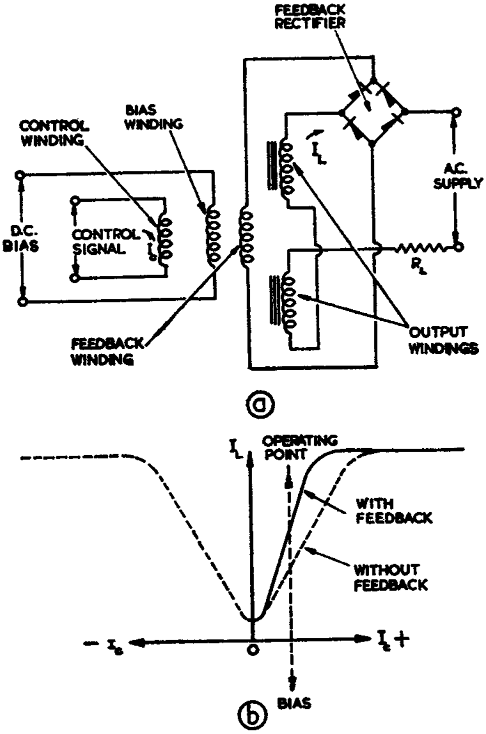


Fig. 4.—POSITIVE FEEDBACK AND ITS EFFECT.

7. The effect of positive feedback in increasing the sensitivity of the amplifier is shown by the increased slope of the transductor operating characteristic (Fig. 4(b)). Using positive feedback and choosing suitable core materials and winding turns-ratios, it is possible to construct transductors with a power amplification of the order of 10^6. It should be noted that *negative feedback* may be applied in a similar manner in instances where the amplification is already adequate and greater stability is required.

Transductors in Cascade

8. As the amplification of a transductor stage is increased, so its response time is increased in direct proportion (see Bk 1, Sect. 7, Chap. 3, Para. 24). Thus, a high stage gain is accompanied by a sluggish response to an applied signal, and a rapid response can be obtained only by making the gain quite low. To obtain high amplification without in-

curring an unduly large response time, it is necessary to limit the amplification per stage and to connect several stages in cascade. The a.c. output of the first stage is rectified and applied to the control circuit of the second stage, this process being repeated according to the number of stages. The overall gain is the *product* of the individual stage gains, while the total response time is approximately equal to the *sum* of the individual response times.

Transductors in Push-pull

9. The output current of a single transductor is never zero, and even with zero control current there will be a standing current in the output windings (see Fig. 4(b)). To obtain an output current which is zero for zero input signal, two transductors are balanced against each other in a push-pull circuit. A push-pull circuit gives the added advantage that a truly 'sensed' output is obtained ; that is, the *direction* of the output changes according to the direction of the control signal. A typical push-pull circuit is shown in Fig. 5. The signal input is applied in series with the control windings of both transductors, and the signal output is applied to the load R_L via a centre-tapped transformer T.

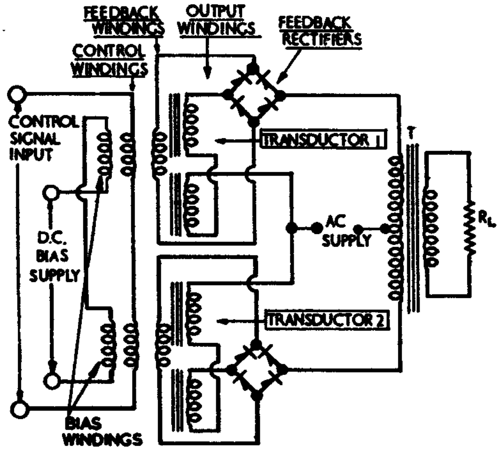


Fig. 5.—TRANSDUCTORS CONNECTED IN PUSH-PULL.

10. With zero input signal, equal standing currents are passed by both transductors. However, the standing currents are in opposition through the primary of the transformer T and effectively cancel one another. No output current is induced in the secondary, and no current flows in the load resistor R_L.

11. The transductors are biased in opposite senses (as shown by the connections to the bias windings in Fig. 5), and the bias is adjusted so that the working points of the transductors coincide with zero control signal, as shown in Fig. 6. Thus, although the general form of each individual characteristic is unchanged, the composite characteristic will be as shown. With this form of characteristic, it is seen that a positive input signal will produce an increase in the output current of transductor 1 and a corresponding decrease in the output current of transductor 2 ; *vice versa* for a negative input signal. The differential output is shown dotted in Fig. 6 and it is seen that a positive-going input produces a positive-going output, and *vice versa*. To avoid distortion the control signal is kept within the limits of the working range shown.

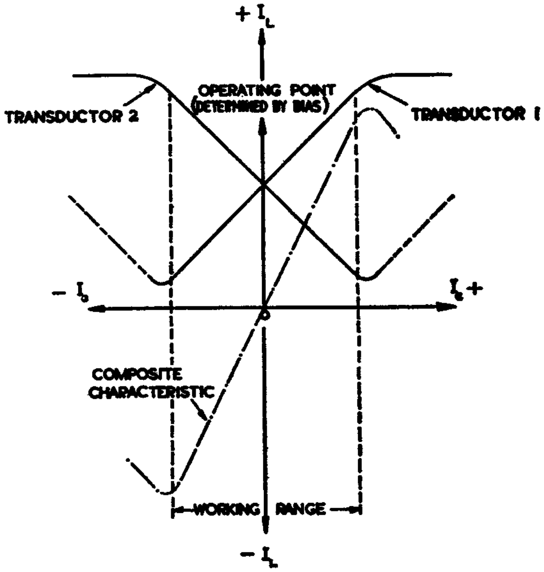


Fig. 6.—PUSH-PULL OPERATION OF TRANSDUCTORS.

Comparison with Valve Amplifiers

12. In relation to amplifiers employing valves, the magnetic amplifier has certain advantages and certain limitations. These are summarized below :—

(*a*) *Advantages.*

(i) Its robust construction makes the transductor reliable, and ensures long life with little maintenance.

(ii) The transductor is immediately available from the moment of switching on.

(iii) No special power pack is necessary.

(iv) The high electrical efficiency (about 75%) enables a given amplification to be obtained with less energy wasted in the form of heat.

(*b*) *Limitations.*

(i) A source of a.c. supply is required.

(ii) It gives a delayed response to an applied signal because of its relatively long time constant.

(iii) The signal frequency is limited to a fraction of the supply frequency, which in turn is limited with present techniques to a few kc/s.

Practical Applications

13. The magnetic amplifier may be preferred to a valve amplifier in applications where the signal frequency is comparatively low and where high reliability and efficiency are required. The biggest difficulty in the design and operation of valve d.c. amplifiers is the inherent tendency to drift. A valve amplifier set up and adjusted so that there is zero output for zero input will in time drift away from this alignment, mainly due to the varying characteristics of the valves themselves. Thus valve d.c. amplifiers are constantly needing adjustment. In contrast to this, the magnetic amplifier has remarkably good long-term stability and is, therefore, suitable for use as a d.c. amplifier for measurement or relay work. An example of the latter consists of a small magnetic amplifier, taking power at 12 volts 50 c/s, feeding a standard Post Office relay type 600. This relay is a robust mechanism but requires a power of the order of one watt to operate it. Greater sensitivity can only be obtained by a much more delicate construction involving loss of robustness. A simple magnetic amplifier provides the power required to operate the relay with an input of a few microwatts. Because of the stability of the magnetic amplifier the on-off conditions remain constant.

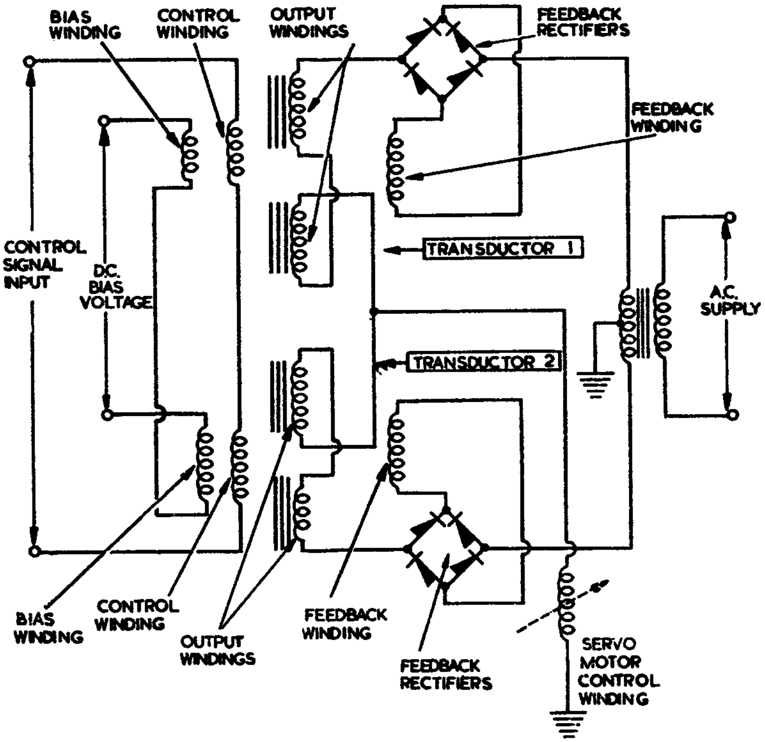


Fig. 7.—MAGNETIC AMPLIFIER USED TO CONTROL SERVO-MOTOR.

14. Economy of space, weight and power consumption, and high reliability and efficiency make the magnetic amplifier eminently suitable for use in aircraft to control the power for driving servo-motors. Fig. 7 shows the simplified circuit of a magnetic amplifier used in an airborne radar equipment. The output of the amplifier is used to control the speed and direction of an induction servo-motor, the latter driving mechanical devices for measurement purposes. Since a push-pull arrangement is used for the magnetic amplifier a truly sensed output is obtained. Thus, a positive d.c. input to the control windings gives an output to drive the motor in one direction, while a negative input to the control windings causes the motor to be driven in the reverse direction. The magnitude of the input determines the magnitude of the output and, hence, the speed of the motor.

15. Other applications of the magnetic amplifier include :—

(*a*) D.C. to a.c. convertor for feeding valve amplifiers.

(*b*) Control of the frequency and voltage of rotary inverters.

(*c*) Use of several control windings, whereby the operation may be made to depend on the sum or the difference of several input signals.

16. If the a.c. supply is at a high frequency, the time constant of the transductor may be reduced to such an extent that it would be possible to control and amplify audio frequency currents. A typical application is the modulation of a radio transmitter, which dispenses with high power modulation valves. Suitable magnetic materials to work at these frequencies are being developed.

SECTION 11

RADIO FREQUENCY AMPLIFIERS

SECTION 11

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SECTION 11

CHAPTER 1

TUNED VOLTAGE AMPLIFIERS

PART 1A, SECTION 11, CHAPTER 1

TUNED VOLTAGE AMPLIFIERS

Introduction

1. *Tuned voltage amplifiers* are voltage amplifiers that use a tuned circuit as the anode load. Such amplifiers may be used to amplify aerial signal-frequency voltages at the input to a radio receiver, in which case they are termed '*r.f. voltage amplifiers*'. The anode load may be tuned to a fixed frequency or it may employ variable tuning. In either case, the use of a tuned circuit as the anode load makes the tuned amplifier *selective* with respect to frequency. Thus, r.f. voltage amplifiers differ from the a.f. voltage amplifiers discussed in Sect. 10, Chap. 1 in that the former are required to amplify voltages over a *selected narrow band* of frequencies in the r.f. spectrum, whereas the latter are required to amplify voltages over a *wide range* of frequencies in the a.f. band.

Band	Frequency	Wavelength	Typical Uses
Very low frequency (VLF)	3–30 kc/s	Very long wave (VLW) 100,000–10,000 metres	Long distance point-to-point communication
Low frequency (LF)	30–300 kc/s	Long wave (LW) 10,000–1,000 metres	Navigational aids, broadcasting, long distance communication
Medium frequency (MF)	300–3,000 kc/s (3 Mc/s)	Medium wave (MW) 1,000–100 metres	Navigational aids, broadcasting
High frequency (HF)	3–30 Mc/s	Short wave (SW) 100–10 metres	Long distance communication
Very high frequency (VHF)	30–300 Mc/s	Metre wave 10–1 metres	Short distance communication, television, radar
Ultra high frequency (UHF)	300–3,000 Mc/s (3 Gc/s) (G=Giga=1,000 Mc/s)	Decimetre wave 10–1 decimetres (100–10 centimetres)	Long-distance 'scatter' and short-distance communication, television, radar
Super high frequency (SHF)	3–30 Gc/s	Centimetre wave 10–1 centimetres	Radar
Extremely high frequency (EHF)	30–300 Gc/s	Millimetre wave 10–1 millimetres	Experimental

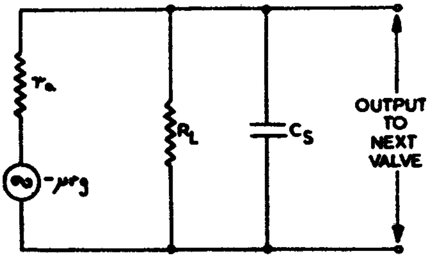
TABLE 1.—RADIO FREQUENCY BANDS

Radio Frequency Bands

2. Radio frequencies are used for a variety of purposes and cover a very wide frequency range. This range is sub-divided into bands which are classified as shown in Table 1.

Need for Tuned Circuit

3. In a resistance-capacitance coupled amplifier the effect of the shunt capacitance, formed by the inter-electrode capacitances of the valves and stray wiring capacitances, is to decrease the stage gain as the frequency increases. The equivalent circuit of such an amplifier is shown in Fig. 1. At a low frequency, the reactance of C_S is so high that its shunting effect can be neglected. The stage gain is then given by $A = \frac{\mu R_L}{r_a + R_L}$, and



$\mu = 30 : r_a = 20 k\Omega \quad R_L = 40 k\Omega : C_S = 30 pF$

Fig. 1.—EQUIVALENT CIRCUIT FOR RESISTANCE-CAPACITANCE COUPLED AMPLIFIER, HIGH FREQUENCIES

for the values shown in Fig. 1, $A = \frac{30.40}{60} = 20$. At a high frequency of 10 Mc/s however, the reactance of C_s is $\frac{1}{\omega C_s} \simeq 500$ ohms. This reactance, in parallel with R_L, gives an effective load of approximately 500 ohms and a stage gain of approximately

$$\frac{30 \cdot 0{\cdot}5}{20 + 0{\cdot}5} = 0{\cdot}73.$$

The effect of the shunt capacitance at this frequency is to introduce such a loss as to render the amplifier useless.

4. By using a parallel tuned circuit as the anode load, as shown in Fig. 2(a), this limitation can be overcome. From the equivalent circuit of Fig. 2(b), the shunt capacitance C_s is now in parallel with the tuned circuit capacitance and alters merely the minimum frequency of operation of the tuned circuit. Thus, a circuit of this type overcomes the effects of the shunt capacitance by absorbing the latter into the tuned circuit capacitance. The v.a.f. is given by $A = \frac{\mu R_D}{r_a + R_D}$, where R_D is the dynamic resistance $\frac{L}{CR}$ of the resonant circuit. An additional advantage in using a circuit of this type is that only signals in the narrow band of frequencies to which the circuit is tuned are amplified.

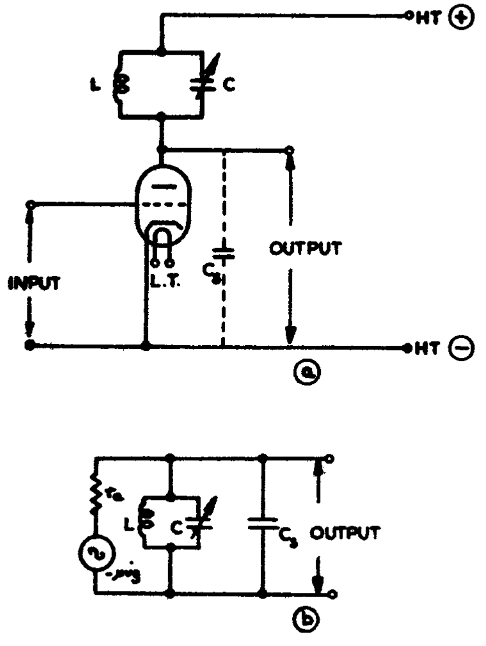


Fig. 2.—USE OF TUNED CIRCUIT AS ANODE LOAD

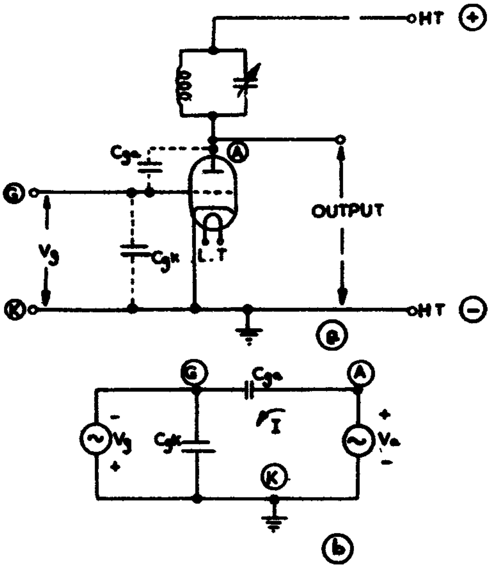


Fig. 3.—MILLER FEEDBACK

Miller Effect

5. It is shown in Sect. 8, Chap. 2, Para. 57 that 'Miller effect' sets an upper limit to the frequency at which a normal triode valve may be used for amplification. Fig. 3(a) shows the simplified circuit of a triode tuned voltage amplifier, and Fig. 3(b) shows the equivalent circuit for Miller feedback through C_{ga}. The change in anode voltage v_a, caused by a variation in grid voltage v_g, is given by $v_a = Av_g$, where $A = \frac{\mu R_D}{r_a + R_D}$. Since at resonance the tuned circuit impedance is a pure resistance, v_a and v_g are 180° out of phase, and the polarity of the voltages at one given instant may be as indicated in Fig. 3(b). The voltage across C_{ga} then is:—

$$\begin{aligned} Vc_{ga} &= v_g + v_a \\ &= v_g + Av_g \\ \therefore\ Vc_{ga} &= v_g (1 + A). \end{aligned}$$

The current I in $C_{ga} = \frac{\text{Voltage}}{\text{Reactance}}$.

$$\therefore\ I = v_g (1 + A)\, \omega C_{ga}.$$

The input impedance Z_{IN} of a valve is the input voltage v_g divided by the current in the input circuit. Thus, the input impedance *due to Miller feedback* is:—

$$Z_{IN} = \frac{v_g}{I}$$

$$= \frac{v_g}{v_g(1 + A)\ \omega C_{ga}}$$

$$\therefore\ Z_{IN} = \frac{1}{\omega C_{ga}\ (1 + A)}$$

This indicates that the input impedance due to feedback is *capacitive*, the effective capacitance having a value of $C_{ga}\ (1 + A)$. The *total* input capacitance includes the inter-electrode capacitance C_{gk} and, for a pure resistive load, is given by:—

$$C_{IN} = C_{gk} + C_{ga}\,(1 + A) \quad .. \quad (1).$$

6. The effect of Miller feedback on the valve input impedance depends to a large extent on the anode load, and the effect on the input circuit for three types of anode load is:—

(*a*) **Pure resistive load.** The input capacitance is much increased as shown in Para. 5 and by equation (1).

(*b*) **Pure capacitive load.** Because of the phase shift between current and voltage in the anode load, the input capacitance is *less* than that for a pure resistive load. In addition there is a *resistive* component in the input impedance which results in damping and power loss in the input circuit.

(*c*) **Pure inductive load.** Again, because of the phase shift between current and voltage in the anode load, the input capacitance is *less* than that for a pure resistive load. In addition, there is a '*negative resistance*' element which cancels some of the positive damping losses in the input circuit and gives rise to general instability and a tendency to oscillation. The term 'negative resistance' applies to an element or a circuit in which the current *decreases* as the voltage applied to it *increases*. It therefore tends to counteract the effects of 'positive' resistance in a circuit.

In a tuned voltage amplifier any slight change in the conditions which puts the anode circuit off tune, makes the anode load either inductive or capacitive and causes positive feedback (with instability) or negative feedback (with damping) respectively.

Reduction of Miller Effect

7. The main factors determining the magnitude of Miller effect are ω, C_{ga} and A. High values for all these factors will affect considerably the operation of the valve. To obtain a high gain (A) at high frequencies($\frac{\omega}{2\pi}$), the factor C_{ga} must either be reduced to a very small value or its effect be neutralized. For a triode, C_{ga} has a relatively large value and unless steps are taken to cancel the effects of Miller feedback, a conventional triode amplifier circuit becomes useless at high frequencies. In some triode r.f. amplifiers, an additional external path from anode to grid is inserted to provide feedback of the exact opposite phase to that produced through C_{ga}. This process is termed '*neutralization*'. It is commonly employed in r.f. *power* amplifiers and is considered in Chap. 2. In tuned voltage amplifiers it is more usual to use pentode or tetrode valves rather than triodes, in order to take advantage of the much lower values for C_{ga} in such valves. Miller feedback is consequently reduced, and with a pentode or a tetrode, high values of A may be obtained at high values of ω (up to frequencies of the order of 200 Mc/s) without the instability that would result from the use of a triode amplifier.

Input to R.F. Voltage Amplifier

8. One purpose of a r.f. voltage amplifier is to select a signal at a given frequency from among many signals induced in the aerial, to amplify the required signal and to apply it to a succeeding stage. The first essential is to provide some means of transferring the signal voltage from the aerial system to the input of the first r.f. amplifier. The usual

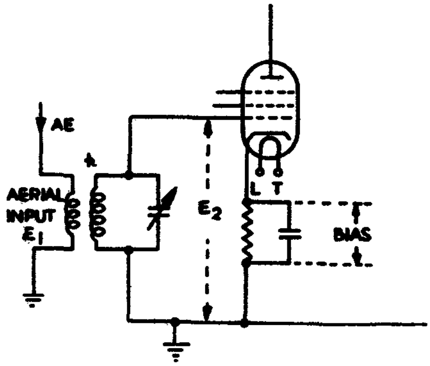


Fig. 4.—AERIAL TRANSFORMER COUPLING

method is to use some form of aerial coupling transformer that will give the desired signal voltage a larger amplitude than that of any other r.f. voltages which may also be present at the receiver input terminals. There are many possible forms of aerial coupling arrangements, but the most common type is the inductively coupled transformer, consisting of a primary winding and a secondary winding, the latter being tuned by a variable capacitor as shown in Fig. 4. The voltage E_2 developed across the tuning capacitor is applied as the input to the valve, and the gain $\frac{E_2}{E_1}$ of the coupling circuit is proportional to the Q of the tuned secondary (typically, $Q = 100$) and to the coefficient of coupling (typically, $k = 0{\cdot}2$). The *selectivity* of the system is also proportional to the Q of the tuned secondary, the half-power (3db) bandwidth being given by $\frac{f_o}{Q}$ (see Book 1, Sect. 5, Chap. 3, Para. 18).

9. Another common type of aerial coupling arrangement is shown in Fig. 5, where an auto-transformer replaces the inductively coupled transformer of Fig. 4. The tapping point on the auto-transformer determines the degree of coupling between primary and secondary, the gain and selectivity of this arrangement being similar to that given in Para. 8.

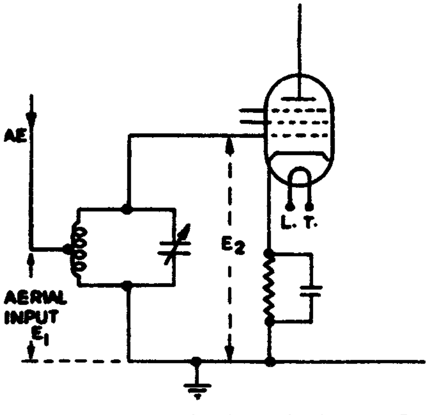


Fig. 5.—AERIAL AUTO-TRANSFORMER COUPLING

Intervalve Coupling for R.F. Voltage Amplifiers

10. **Capacitance coupled stage.** An anode load consisting of a parallel tuned circuit that is capacitively coupled to the grid leak of the next stage is sometimes used in r.f. voltage amplifiers. A basic circuit is shown in Fig. 6. The grid circuit is tuned to the required signal frequency, the selected signal voltage being applied between grid and cathode of V_1. Voltage amplification occurs, and the amplified signal voltage developed across the anode load is coupled via C_6 to the grid leak R_4. The ratio of E_2 to E_1 represents the stage gain of V_1.

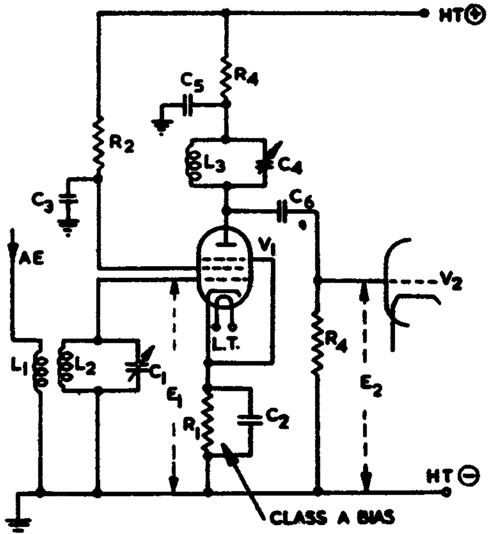


Fig. 6.—TUNED CIRCUIT CAPACITIVELY COUPLED TO THE FOLLOWING STAGE

11. It is shown in Sect. 8, Chap. 3, Para. 20 that the voltage amplification factor for a pentode amplifier is:—

$$\text{V.A.F.} \simeq g_m Z_L,$$

Where g_m is the valve mutual conductance, Z_L is the load impedance.

When the anode tuned circuit is resonant to the input signal, its impedance Z_L is the dynamic impedance R_D, where $R_D = \frac{L_3}{C_4 R}$.

Thus:—

$$\text{V.A.F.} \simeq g_m R_D \quad \ldots \quad \ldots \quad (2).$$

Now:—

$$R_D = \frac{L_3}{C_4 R} = \frac{\omega L_3}{\omega C_4 R} = Q\omega L_3,$$

$$\text{where } Q = \frac{1}{\omega C_4 R}.$$

Thus, the amplification at resonance may also be written as:—

$$\text{V.A.F.} \simeq g_m Q \omega L_3 \quad \ldots \quad \ldots \quad (3).$$

12. The amplification varies with frequency about resonance in accordance with the response curve for the anode load, as shown in Fig. 7. Selectivity is determined by the effective Q of the anode tuned curcuit, the half-power bandwidth being given by $\frac{f_0}{Q}$. Note, from Book 1, Sect. 5 Chap. 3, Para. 22, that the selectivity will be high and the bandwidth narrow only if the internal impedance of the supply (i.e. the r_a of the valve) is high in relation to the load impedance. With a pentode this condition is realized. The bandwidth required depends on the "type" of signal being received. In communication and broadcast receivers it is required to pass only a narrow band of frequencies (± 5 kc/s about the signal frequency on the medium-wave and long-wave bands). In television and radar receivers the bandwidth is required to be very much larger.

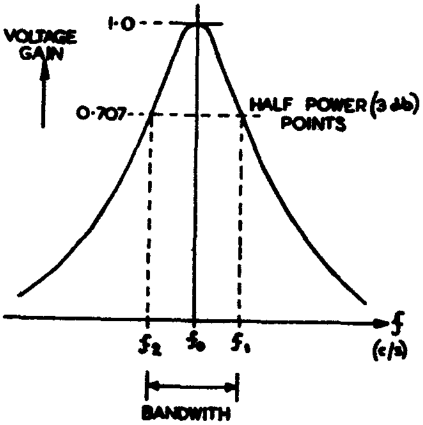


Fig. 7.—GAIN-FREQUENCY RESPONSE CURVE FOR SINGLE TUNED CIRCUIT VOLTAGE AMPLIFIER

13. **Tuned transformer coupled stage.** The most common arrangement for coupling a r.f. voltage amplifier to the following stage is by means of a transformer having a tuned secondary, as shown in Fig. 8. The *primary* winding has a value such that it resonates (with the valve inter-electrode capacitances and stray capacitances) either above or below the tuning range of the amplifier, in order to avoid large changes in gain and selectivity. The necessity for using a grid leak and coupling capacitor is now eliminated. The anode slope resistance r_a of a pentode is high. Thus, assuming that r_a is very much greater than the impedance reflected into the primary from the tuned secondary, the amplification at resonance is:—

$$\text{V.A.F.} \simeq g_m Q \omega M \quad .. \quad .. \quad (4).$$

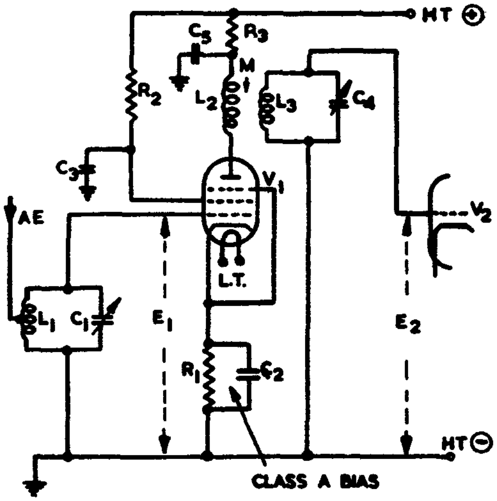


Fig. 8.—TUNED TRANSFORMER COUPLING

Comparison of equations (3) and (4) shows that they differ only in that M in (4) has replaced L in (3). Thus, when the same tuned circuit is involved, the only difference in behaviour between transformer coupling and tuned circuit capacitance coupling is that the amplification is modified in accordance with the ratio $\frac{M}{L}$. The selectivity and the 3db bandwidth remain the same. Similar results may be obtained if the *primary* is tuned and the secondary untuned, but in this case the tuning capacitor is no longer isolated from the h.t. supply.

Multi-stage R.F. Voltage Amplifiers

14. When there is more than one stage of r.f. voltage amplification the overall characteristic is the *product* of the amplification curves of the individual stages. As a result, the half-power bandwidth of the multi-stage system is *less* than the bandwidth of a single stage, as illustrated in Fig. 9. Thus, in order to maintain a given bandwidth, it is necessary that the effective Q of the individual stages be reduced as the number of stages is increased. Because of the tendency to instability and the difficulties

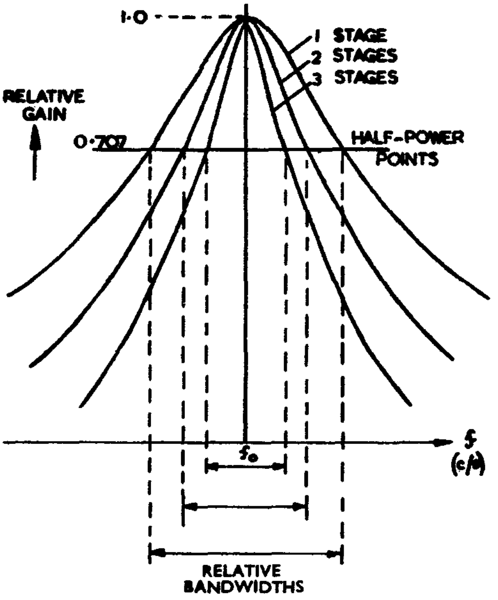


Fig. 9.—BANDWIDTH IN MULTI-STAGE AMPLIFIERS

involved in tuning multi-stage amplifiers at high frequencies it is not normal to employ more than one or two stages of r.f. voltage amplification.

Band-pass Coupling

15. Intermediate frequency (i.f.) voltage amplifiers are used in superheterodyne receivers in order to improve gain and selectivity (see Bk 3. Sect. 14)·The bandwidth of such amplifiers (i.e. the band of frequencies to be passed in relation to the mid-band frequency) has to be larger than that for r.f. voltage amplifiers, since the intermediate frequency is lower than the signal frequency. One method of obtaining the required increase in bandwidth is the use of coupled circuits consisting of tuned primary and tuned secondary windings, as discussed in Book 1. Sect. 7, Chap. 1 Para. 10).

16. The circuit of an i.f. voltage amplifier coupled to the following stage by means of a band-pass coupled circuit is shown in Fig. 10. The stage works under Class A cathode bias conditions. For critically coupled circuits, with identical primary and secondary circuits, the v.a.f. at resonance is:—

$$\text{V.A.F.} \simeq \frac{k_c Q \omega L}{2} \quad \ldots \quad \ldots \quad (5).$$

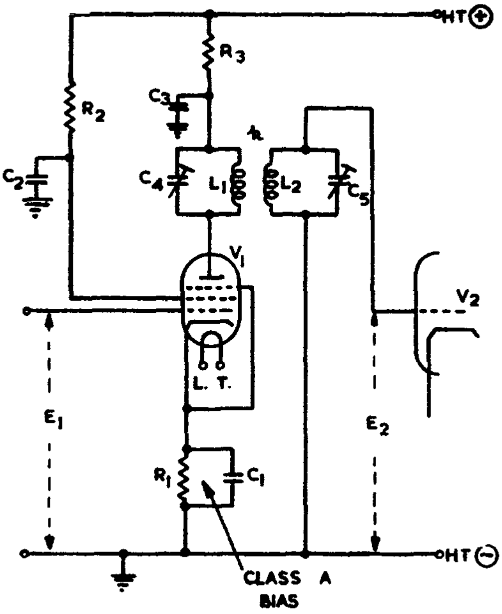


Fig. 10.—BAND-PASS COUPLING IN I.F. VOLTAGE AMPLIFIER

Thus, the amplification is *half* that for a single tuned circuit (see equation (3)). The shape of the response curve depends on the coefficient of coupling. In Fig. 11 curve 'a' shows the result of tight coupling, curve 'b' illustrates critical coupling and curve 'c' loose coupling.

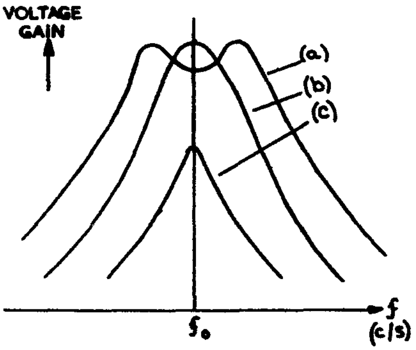


Fig. 11.—EFFECT OF VARYING k IN BAND-PASS COUPLED CIRCUITS

17. The bandwidth in the case of a band-pass coupled amplifier, is defined in the same manner as for a system employing single tuned circuits. When the Q values of the two coupled circuits are the same and the coefficient of coupling equals the critical

value, the half-power bandwidth of the band-pass stage is $\sqrt{2}$ times that of a single tuned stage using the same tuned circuit. Compared with a single tuned stage having the same half-power bandwidth, the gain of the band-pass stage is *higher* and it possesses a response that is flatter in the vicinity of resonance and has steeper sides. When band-pass coupled circuits are used in i.f. voltage amplifier stages, it is usual to make the coefficient of coupling slightly *less* than the critical value. In this way by causing the response curve to have a slight peak, it is easier to adjust the individual circuits to resonance at a common frequency. When a number of identical band-pass coupled stages are connected in cascade, the overall bandwidth of the system is reduced in the same manner as for single tuned stages connected in cascade (see Para. 14).

Methods for Increasing Bandwidth

18. In some applications (e.g. radar and television receivers) the band of frequencies that is required to be amplified is greater than the bandwidth of the conventional band-pass coupled stage described in Para. 17. Two methods are commonly used in order to further increase the bandwidth.

19. **Damped tuned circuits.** Damping resistors shunting the resonant circuits are often inserted as shown in Fig. 12. This lowers the Q values of the circuits and since the half-power bandwidth equals $\frac{f_0}{Q}$, the required bandwidth may then be obtained. However, a reduction of Q gives a proportionate reduction in the v.a.f. of the stage, and further stages of amplification may be required.

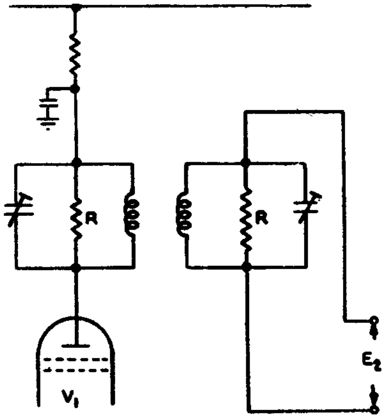


Fig. 12.—REDUCTION IN Q BY USE OF DAMPING RESISTORS

20. **Staggered tuning.** Two or more stages with single tuned circuits in the anode may be used, the resonant frequency of each individual circuit being off-set about the mid-frequency of the band to be amplified. For two staggered tuned stages, one circuit is tuned below the centre frequency, and the other tuned above it by the same amount,as shown in Fig. 13(a), so that the centre frequency of the band is at the common half-power points of the individual circuits. The overall response of the two-stage staggered tuned pair is shown in Fig. 13(b). From this it is seen that the bandwidth has

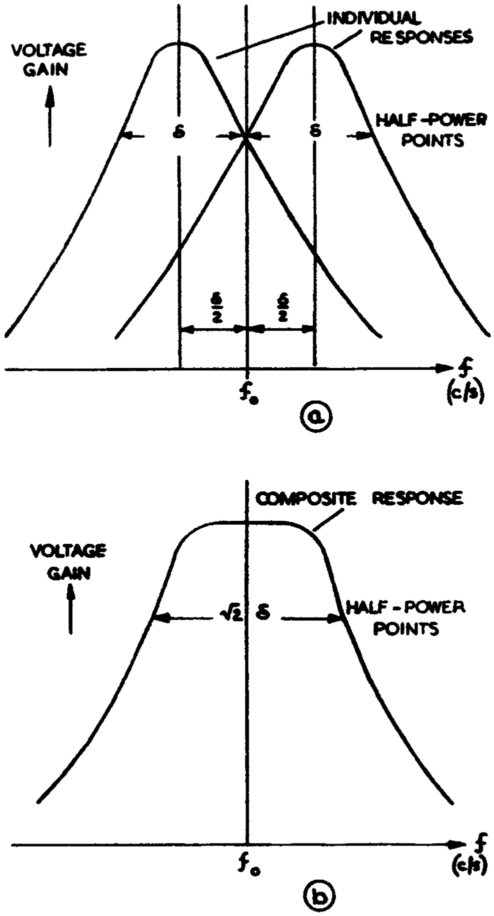


Fig. 13.—STAGGERED TUNING

been increased to $\sqrt{2}$ times that of the individual circuits. By adding more staggered tuned stages, a very large bandwidth with a reasonable gain may be obtained (see Book 1, Sect. 7, Chap. 1, Para. 17).

Grounded-grid Triode

21. There are three basic ways of connecting a triode valve as an amplifier. These ways may be expressed in terms of the electrode that is earthed or 'grounded' so far as the signal voltage is concerned; the other two electrodes are used for input and output respectively. The most common arrangement is the grounded-cathode, in which the input is applied to the grid and the output taken from the anode (Fig. 14(a)). A second arrangement is the grounded-anode, better known as the "cathode follower". The input is applied to the grid and the output taken from the cathode (Fig. 14(b)). The last arrangement is the *grounded-grid*, in which the input is applied to the cathode and the output taken from the anode (Fig. 14(c)).

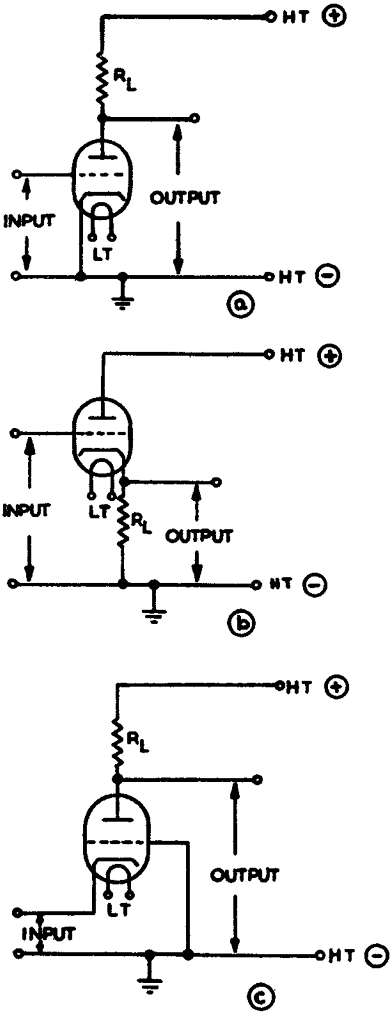


Fig. 14.—METHODS OF CONNECTING A TRIODE

22. In a r.f. voltage amplifier the most usual choice of valve is a pentode, in which Miller feedback is practically eliminated and which has a high gain. However, a pentode is considerably noisier than a triode (see Sect. 8, Chap. 3, Para. 50) so that if a triode can be made to amplify satisfactorily at high frequencies it may be preferred to a pentode in cases where a high signal-to-noise ratio is required. A grounded-grid triode fulfils these conditions, and is used as an amplifier at radio frequencies.

23. Fig. 15 shows the simplified circuit of a r.f. voltage amplifier using a grounded-grid triode. Here the control grid acts as an earthed shield between the output and input circuits, the output circuit being effectively between anode and earth, and the input circuit between cathode and earth. As a result, energy transfer between input and output circuits through the inter-electrode capacitance C_{ak} is very small because of the grounded grid between anode and cathode. C_{ga} now appears across the anode tuned

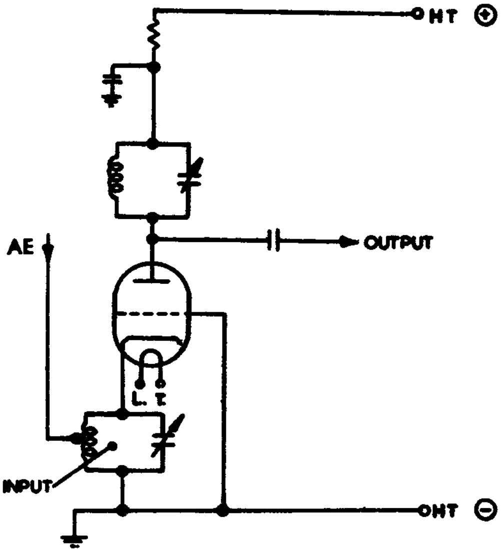


Fig. 15.—GROUNDED-GRID TRIODE R.F. VOLTAGE AMPLIFIER

circuit, and C_{gk} appears across the cathode tuned circuit, these two capacitances being absorbed into their respective tuned circuits to alter merely the minimum operating frequency. A *rise* in cathode potential relative to earth has the same effect on the anode current as a fall in grid potential relative to the cathode, so that the anode current falls and the anode voltage *rises*. The input and output signals are, therefore *in phase*.

24. From the equivalent circuit of Fig. 16, the gain of the grounded-grid amplifier may be found:—

$$\text{Total e.m.f.} = v_g + \mu v_g = v_g(1+\mu)$$

$$\text{Current } i_a = \frac{v_g(1+\mu)}{r_a + R_D}$$

$$\text{Output voltage } v_o = i_a R_D = \frac{v_g(1+\mu)R_D}{r_a + R_D}$$

$$\text{V.A.F.} = \frac{v_o}{v_g} = \frac{(1+\mu)R_D}{r_a + R_D} \quad \ldots \quad (6).$$

Compared with the usual expression for v.a.f. $= \frac{\mu R_D}{r_a + R_D}$, the v.a.f. of a grounded-grid amplifier is slightly *greater* than that of a conventional grounded-cathode amplifier.

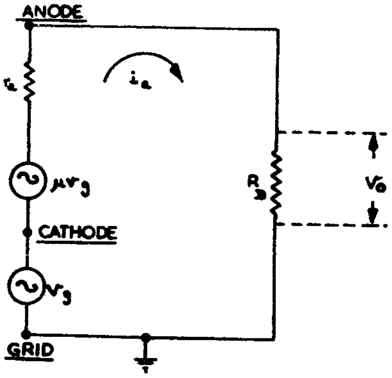


Fig. 16.—EQUIVALENT CIRCUIT FOR GROUNDED-GRID TRIODE

25. The input impedance of a grounded-grid amplifier is *low*.

$$\text{Anode current } i_a = \frac{v_g(1+\mu)}{r_a + R_D}$$

$$\text{Input impedance } Z_{IN} = \frac{v_g}{i_a} = \frac{v_g}{v_g(1+\mu)/r_a + R_D} = \frac{r_a + R_D}{1+\mu}$$

If $r_a > R_D$, and $\mu > 1$,

$$Z_{IN} \simeq \frac{r_a}{\mu} = \frac{1}{g_m} \quad \ldots \quad \ldots \quad (7).$$

Thus, the input impedance of a grounded-grid amplifier is approximately the same as the *output* impedance of a cathode follower circuit. This low input impedance damps the input tuned circuit and may be an advantage where a wide bandwidth is required. However, correct matching of the aerial input to the low input impedance of the valve is not achieved and a large signal power is required to develop the signal voltage across the valve input terminals. The main use of a grounded-grid triode is for wideband working at v.h.f. where a high signal-to-noise ratio is required. The noise introduced by a pentode r.f. amplifier may be sufficient at v.h.f. to reduce the signal-to-noise ratio below the acceptable level.

The Cascode Amplifier

26. The name '*cascode*' may be regarded as an abbreviated form of '*casc* (-aded triode amplifier having characteristics similar to, but less noisy than, a pent-) *ode*'. The cascode amplifier consists of two triodes in cascade; the first stage is normally a conventional grounded-cathode circuit and the second stage uses a grounded-grid arrangement. A simplified cascode circuit is shown

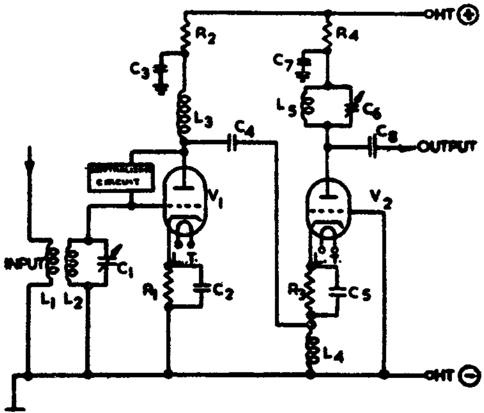


Fig. 17.—CASCODE AMPLIFIER

in Fig. 17, where the output of V_1 is capacitively coupled to the cathode input circuit of V_2. Although V_1 is a grounded-cathode stage, Miller feedback is reduced to negligible proportions by neutralizing action (see Chap. 2) and because the gain A of this stage is kept low. The cascode combines the low noise of a triode, the high stability of the grounded-grid arrangement and the high gain of a pentode. In addition, it overcomes the disadvantage of the grounded-grid circuit, namely low input impedance. The input impedance of the grounded-cathode stage V_1 is high, so that damping of the aerial input circuit L_2C_1 is small. Correct matching of L_2C_1 to the input impedance of V_1 may now be achieved, so that the signal power required to develop a voltage across L_2C_1 is correspondingly less than that for a grounded-grid amplifier. Because of the reduction in required signal power the cascode amplifier can be used to give 'fringe' reception at v.h.f. while still retaining the high signal-to-noise ratio obtainable with the grounded-grid triode. The bandwidth is catered for by the circuit components rather than by the low input impedance of the grounded-grid circuit.

SECTION 11

CHAPTER 2

R.F. POWER AMPLIFIERS

R.F. POWER AMPLIFIERS

Introduction

1. Radio frequency power amplifiers are used in radio transmitters (see Bk 3. Sect. 13) to *convert* the d.c. power supplied to the stage into r.f. power. The valve controls this conversion of power in such a way that a high value of r.f. power output, at the required frequency and of the required waveform is available at the aerial for transmission.

2. Amplifiers may be classified as:—

(*a*) **A.F. voltage amplifiers.** These are used to amplify voltages without regard to power over a wide frequency range in the audio band. Ideally, they should amplify equally well at all frequencies within the range 20 c/s to 20 kc/s, but a more limited range gives adequate quality for communication purposes. A.F. voltage amplifiers are used in receivers, public address systems and transmitters to provide the drive to succeeding a.f. power amplifier stages (see Sect. 10, Chap. 1).

(*b*) **A.F. power amplifiers.** These operate over a frequency range similar to that for a.f. voltage amplifiers and are used to convert d.c. power to the required a.f. power. They are used in receivers and in public address systems to provide the a.f. power to operate telephone receivers and loudspeakers (from a few milliwatts up to the order of 30 watts). They are also used to provide the a.f. power for modulation of a transmitter, the power requirements in this case varying from a few watts to many kilowatts (see Sect. 10, Chap. 2).

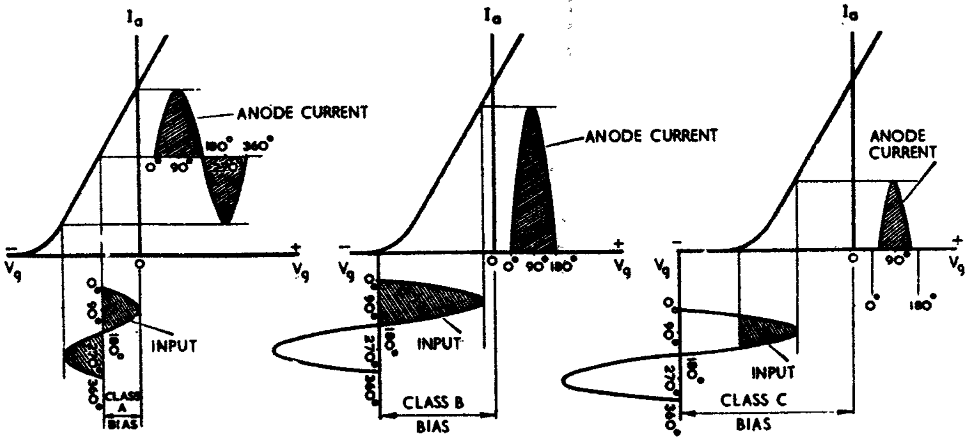
(*c*) **R.F. voltage amplifiers.** These are tuned amplifiers used to amplify voltages over a selected narrow band of frequencies in the r.f. spectrum. In receivers, r.f. amplifiers are designed to amplify small signal voltages of the order of millivolts or microvolts. In transmitters, r.f. voltage amplifiers provide the drive to succeeding r.f. power amplifier stages (see Sect. 11, Chap. 1).

(*d*) **R.F. power amplifiers.** These are tuned in the same way as r.f. voltage amplifiers and operate over a similar frequency range, converting d.c. power to the required r.f. power. They are used in transmitters to supply r.f. power to the aerial for transmission, the power requirements varying from a few watts to many kilowatts. Such amplifiers are considered in this Chapter.

Bias Conditions in R.F. Power Amplifiers

3. R.F. power amplifiers are operated under Class A, Class B or Class C bias conditions, depending on the requirements:—

(*a*) **Class A.** The valve is biased above cut-off so that current flows during the whole of every cycle of input; that is, the '*angle of anode current flow*' is 360°, as shown in Fig. 1(a). Such amplifiers are



ⓐ 360° WORKING CLASS A

ⓑ 180° WORKING CLASS B

ⓒ LESS THAN 180° WORKING CLASS C

Fig. 1.—BIAS CONDITIONS IN R.F. POWER AMPLIFIERS

used in low power circuits, where preservation of the input waveform is important and the efficiency of power conversion relatively unimportant.

(*b*) **Class B.** The valve is biased approximately to cut-off so that current flows during *alternate* half-cycles only; the angle of current flow is 180° (Fig. 1(b)). Such amplifiers are used in high power circuits where preservation of the input waveform and power considerations are both important.

(*c*) **Class C.** The valve is biased beyond cut-off so that current flows for only a *part* of every alternate half-cycle; the angle of current flow is *less* than 180° (Fig. 1(c)). Class C power amplifiers are used in high power circuits where efficiency is of the greatest importance.

4. For a parallel resonant circuit, the voltage across it and the current circulating in it remain very nearly sinusoidal even if the current supplied to the circuit departs very much from a sinusoidal waveform. A tuned circuit is similar to a flywheel in its action in that a pulse of current applied at the correct instant of time in relation to the resonant period of the tuned circuit will produce a nearly sinusoidal voltage across the tuned circuit. Thus, a single valve operating under Class B or Class C conditions may be used in conjunction with a tuned anode load to provide r.f. power amplification. However, where the r.f. input to a power amplifier is '*amplitude-modulated*' (see Bk 3, Sect.13) must be operated under either Class A or Class B bias in order to preserve the modulated waveform. Class C amplifiers are confined to amplification of unmodulated or frequency-modulated inputs: that is, under conditions where the *amplitude* of the r.f. input is *constant*.

Circuit Arrangements

5. Fig. 2 illustrates simplified circuits of typical arrangements used in r.f. power amplifiers. A basic series-fed circuit is shown at (a). Fig. 2(b) shows a shunt-fed circuit, the anode choke L_3 having a reactance large compared to that of the tuned circuit inductance L_4 or to the reactance of C_3, so that L_3 carries only a small part of the r.f. current circulating in the anode circuit. L_4 carries two 'taps'; the *aerial tap* varies the turns ratio between the load and the tuned

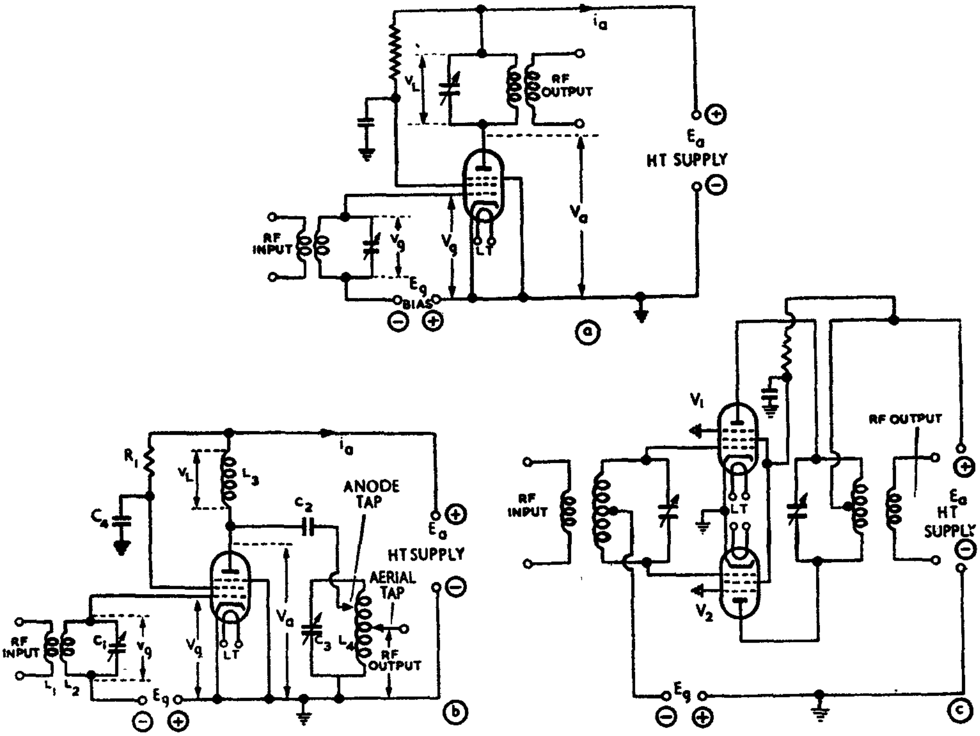


Fig. 2.—RADIO FREQUENCY POWER AMPLIFIER ARRANGEMENTS

circuit to enable matching between them to be achieved; the *anode tap* varies the turns ratio between the tuned circuit and the valve, for matching between them. Push-pull arrangements are also common in r.f. power amplifiers, and a typical circuit is shown at (c). The value of bias in the three circuits determines whether the amplifier works under Class A, Class B or Class C conditions.

Neutralization

6. When possible, beam valves or pentodes are used in preference to triodes for r.f. power amplification, because Miller feedback through the comparatively large C_{ga} of a triode gives rise to instability (see Chap. 1, Para. 5). However, tetrode and pentode valves are not normally available for power outputs much greater than a few kilowatts because of the difficulty in dissipating the heat developed in the inner electrodes. Thus, at higher outputs, triodes are used. It is then essential to provide some means of 'neutralizing' the effect of C_{ga} because power amplifiers must be prevented from self-oscillating and taking control of the frequency. A grounded-grid circuit may be used to prevent Miller feedback, but in r.f. power amplifiers it is more usual to arrange a path by which a second current flows into the input circuit, of opposite phase to the current through C_{ga}.

7. **Split-anode circuit.** A typical neutralizing circuit is shown in Fig. 3(a). The output circuit is centre-tapped at B and effectively connected to earth through the h.t. supply. One end A, of the tuned circuit is connected to the anode, and thence through C_{ga} to the grid D. The opposite end C, of the tuned circuit (which is 180° out of phase with A) is also connected to the grid D, via a neutralizing capacitor C_N. By correct adjustment of C_N, the energy transfer that would normally occur between input and output tuned circuits through C_{ga} can be exactly balanced by the anti-phase feedback through C_N. The equivalent circuit of Fig. 3(b) shows that the neutralizing circuit is effectively an a.c. bridge. When this bridge is correctly balanced in both magnitude and phase by adjustment of C_N:—

$$\frac{\dfrac{1}{\omega C_{ga}}}{\omega L_b} = \frac{\dfrac{1}{\omega C_N}}{\omega L_a}$$

$$\frac{C_N}{C_{ga}} = \frac{L_b}{L_a}$$

$$\therefore \quad C_N = C_{ga} \cdot \frac{L_b}{L_a}.$$

The neutralizing capacitor C_N is adjusted to this value to obtain a balance. At balance, there is no transfer of energy between input (BD) and output (AC) terminals, and the amplifier operates in a stable state.

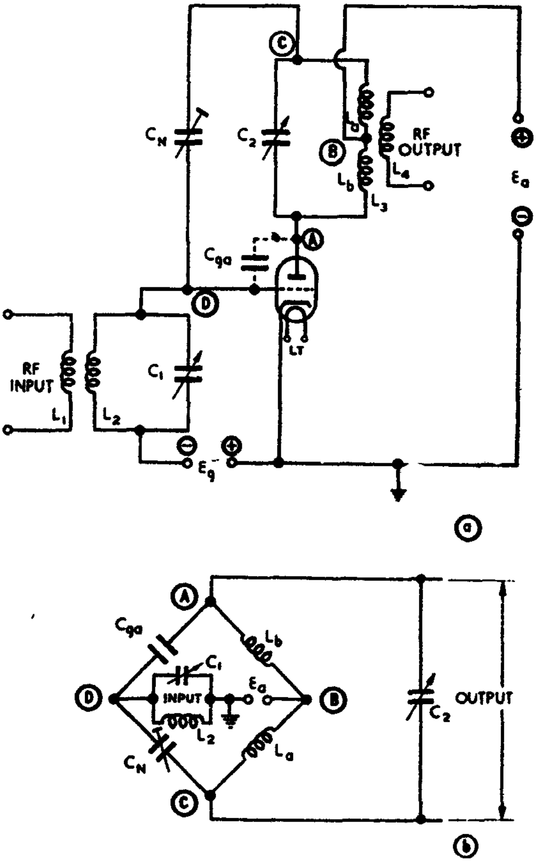


Fig. 3.—SPLIT-ANODE NEUTRALIZING CIRCUIT

8. **Split-grid circuit.** The input tuned circuit shown in Fig. 4(a) is centre-tapped to earth at C, one end D being connected to the grid and the other end B connected to the anode through C_N. By adjustment of C_N the current through the neutralizing capacitor from the grid circuit to the anode circuit, and *vice versa*, can be made equal and opposite to the corresponding current through C_{ga}. Note, however, in this case that since the input circuit is centre-tapped, only *half* the available input is applied to the valve, and for the same power output from the stage a

greater drive is necessary. The bridge equivalent circuit of Fig. 4(b) shows that for balance:—

$$C_N = C_{ga} \cdot \frac{L_a}{L_b}.$$

9. **Push-pull circuit.** The push-pull neutralizing circuit is shown in Fig. 5(a). The principle of neutralizing in this case consists of connecting each grid through a capacitor C_N to a point at a potential 180° out of phase with its anode circuit. Such a point is the anode of the *other* valve. Thus, V_1 grid is connected to V_2 anode through C_{N1}, and V_2 grid is connected to V_1 anode through C_{N2}. The equivalent circuit of Fig. 5(b) shows that for balance:—

$$\frac{C_{N2}}{C_{ga1}} = \frac{C_{ga2}}{C_{N1}}.$$

This circuit is symmetrical and gives perfect stability and complete freedom from feedback up to very high frequencies.

Class A Power Amplifier

10. In order to obtain the required high power at the aerial of a transmitter, the r.f. power output valve is normally allowed to run into grid current. Power is therefore developed in the grid circuit of the output valve and this presents variable 'loading' on the previous stage. A variable load of this kind in the initial stages of a transmitter is undesirable and it may be necessary to insert a 'driver stage' before the output valve. The driver stage must itself be a power amplifier in order to supply the grid power of the output valve, but the power output from the driver valve will be only a small fraction of the total power output of the transmitter. The driver stage is usually a single power valve operating under Class A conditions; grid current is therefore not established in this valve so that the load on the preceding stage is low and constant.

11. When the anode tuned circuit is resonant to the applied frequency, it presents a purely resistive load $R_D = \frac{L}{CR}$ to the valve. The phase relationships between current and voltage in the circuit, when the bias is such that the angle of anode current flow is 360°, are then as shown in Fig. 6. The d.c. voltage drop in the resonant circuit

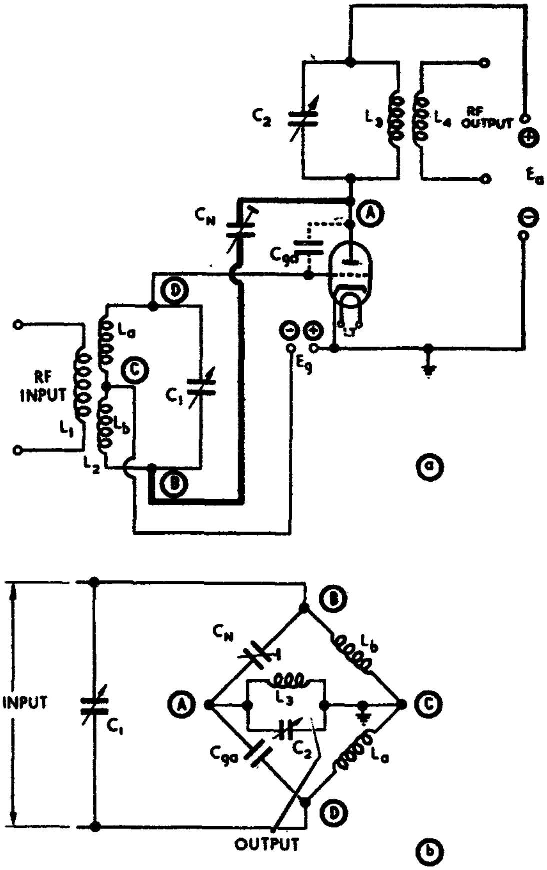


Fig. 4.—SPLIT-GRID NEUTRALIZING CIRCUIT

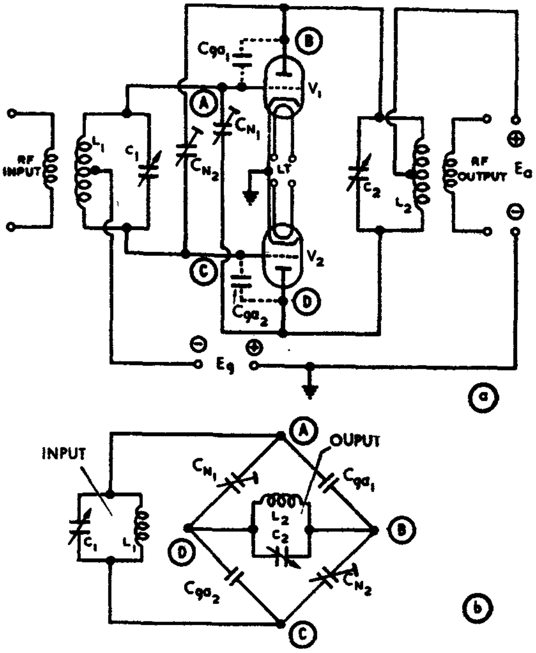


Fig. 5.—PUSH-PULL NEUTRALIZING CIRCUIT

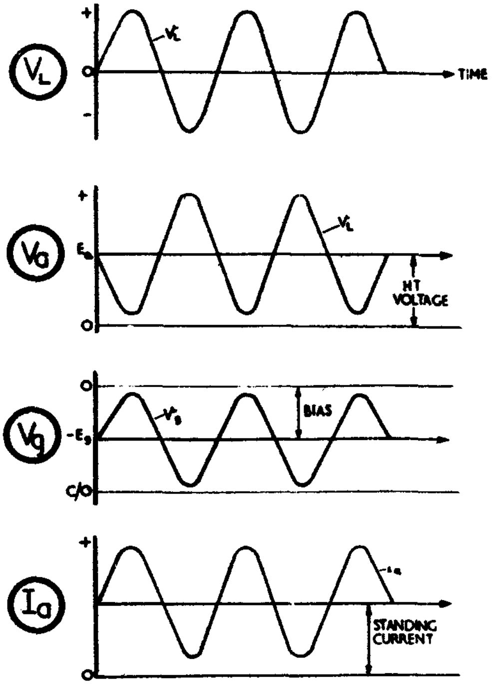


Fig. 6.—PHASE RELATIONSHIPS IN A CLASS A OPERATED AMPLIFIER

is negligible, and the voltage v_L across the load is purely alternating. Thus the voltage V_a actually applied to the anode is the h.t. voltage E_a minus the alternating voltage drop v_L across the load; that is, $V_a = E_a - v_L$. Hence the voltage across the valve rises above the h.t. value during parts of the cycle as shown in Fig. 6. The voltage V_g applied to the grid is the sum of the grid bias voltage $-E_g$ and the signal voltage v_g; that is, $V_g = -E_g + v_g$. The two voltages V_a and V_g are in phase opposition, while v_g, v_L and i_a are in phase with each other.

Class B Power Amplifier

12. **Use.** A Class B r.f. power amplifier (Fig. 7), is used as a final power amplifier where a high power output at a reasonable level of efficiency is required and where the input to the stage is amplitude-modulated. It is usual to find Class B power output valves employed in systems using '*low level modulation*', i.e. where modulation is effected at an early stage working at low power. With a valve biased to projected cut-off the angle of current flow is 180°, and the output is proportional to the grid input voltage; distortion of the modulated envelope is thereby prevented.

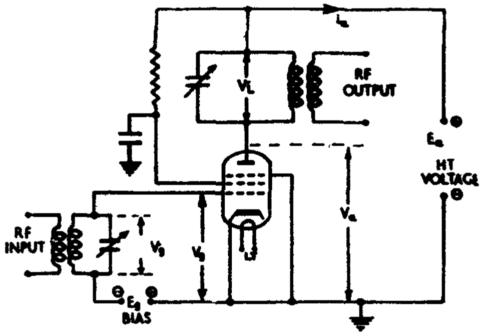


Fig. 7.—BASIC CIRCUIT FOR A CLASS B POWER AMPLIFIER

13. **Bias.** The r.f. input to a Class B valve is normally varying in amplitude at the modulation frequency, and since the bias is required to be constant, automatic self-bias from a grid leak and capacitor is not permissible. The bias voltage must therefore be obtained from an external bias-supply, or cathode bias may be used, or a combination of both methods. Cathode bias is permissible since the *mean* d.c. valve current remains constant.

14. **Grid drive.** With an amplitude-modulated input to a Class B valve, the load placed on the preceding stage varies considerably during the modulation cycle. This is illustrated in Fig. 8. At the troughs of modulation the grid is negative and the required driving power is negligible. At the peaks of modulation however, considerable grid current flows and the driving power required is large. It is therefore necessary to supply the Class B valve from a driver stage (see Para. 10).

15. **Efficiency and power output.** A graph showing the phase relationships between current and voltage in a Class B valve at the peak of the modulation cycle is shown in Fig. 9. The relationship between v_L, V_a and V_g are the same as those given in Para. 11. However, since the valve is biased to cut-off, the anode current i_a waveform is a half sine wave (i.e., the angle of flow θ_a is 180°) and contains d.c., fundamental and harmonic components. Grid current i_g is established during the angle θ_g (i.e., V_g above zero volts). If the tuned

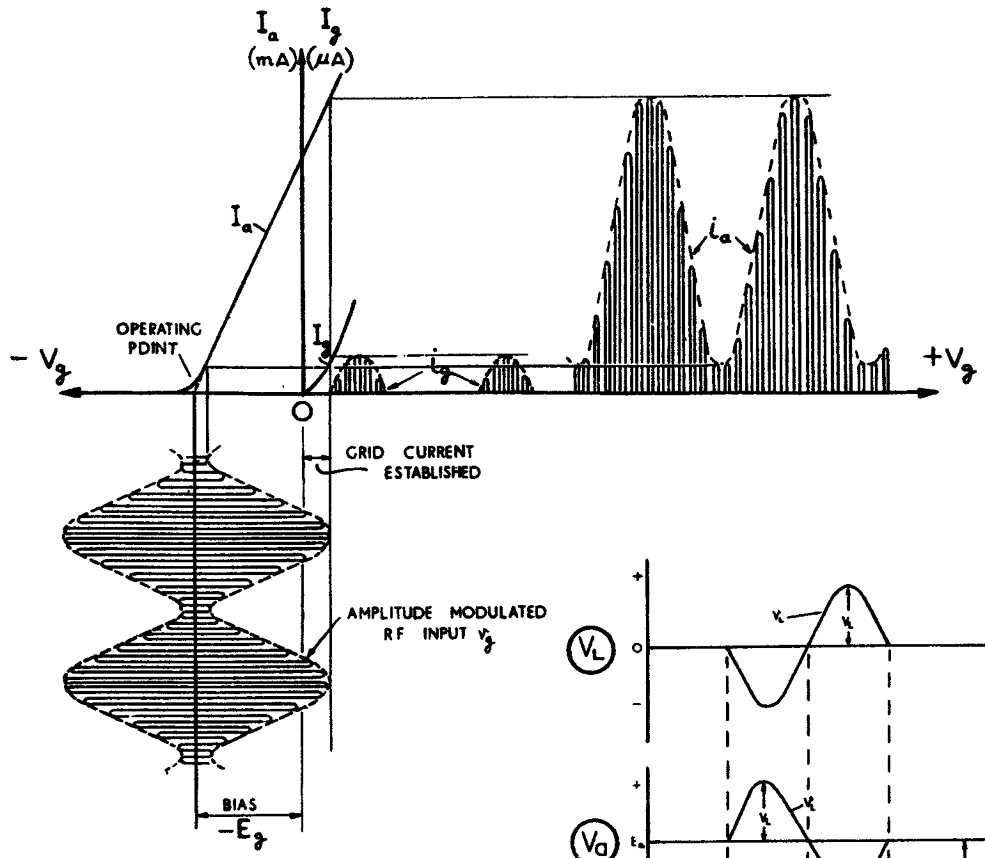


Fig. 8.—VARIABLE GRID LOADING FOR A CLASS B POWER AMPLIFIER WITH AMPLITUDE-MODULATED INPUT

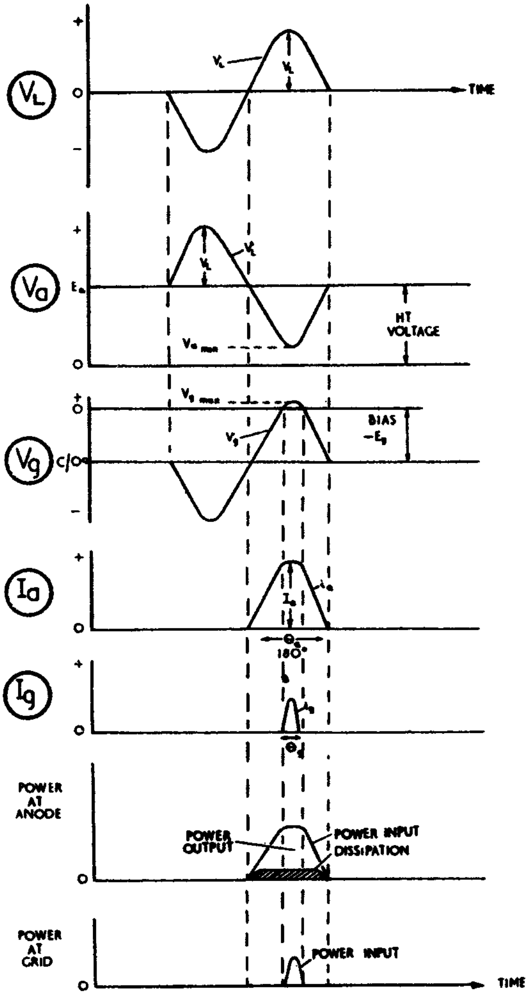


Fig. 9.—PHASE RELATIONSHIPS IN A CLASS B POWER AMPLIFIER

anode load is correctly adjusted, the harmonic components of anode current produce no output voltage, the output being at the fundamental frequency only. The fundamental component of current in a Class B valve has a r.m.s. value of $\frac{I_a}{2\sqrt{2}}$; the r.f. voltage at the fundamental frequency across the load has a r.m.s. value of $\frac{V_L}{\sqrt{2}}$; the r.m.s. power output is given by the product of these two components:—

$$\text{Power output} = \frac{I_a\, V_L}{4} \quad .. \quad (1).$$

16. From Fig. 9 it is seen that peak current is established at the point in the cycle when the anode voltage is at a minimum. The dissipation in the valve is, therefore, relatively small and the efficiency relatively high. The d.c. power supplied to the valve is given by the product of the h.t. voltage E_a and the the d.c. component of anode current; for a half sine wave the latter is $\frac{I_a}{\pi}$. Thus:—

$$\text{Power input} = \frac{I_a\, E_a}{\pi} \quad .. \quad .. \quad (2).$$

The *difference* between the input power and the output power gives the power wasted or dissipated as heat in the circuit, mainly in the valve. The *ratio* of power output to power input gives the efficiency η of the amplifier. Thus:—

$$\eta = \frac{\text{Power Output}}{\text{Power Input}} \times 100 \text{ (per cent).}$$

$$= \frac{\frac{I_a\, V_L}{4}}{\frac{I_a\, E_a}{\pi}} \times 100 \text{ (per cent).}$$

$$\therefore \eta = \frac{\pi}{4} \cdot \frac{V_L}{E_a} \times 100 \text{ (per cent.)} \quad .. \quad (3).$$

The maximum theoretical efficiency is, therefore, $\frac{\pi}{4} \times 100$ (per cent), or 78·5%.

Practical efficiences are of the order of 55% to 65% at the modulation peaks. However, since the amplitude of the anode current pulses is proportional to the grid driving voltage, the power output and efficiency are less at other parts of the modulation cycle. The *average* efficiency is, therefore, only about 40%.

Class C Power Amplifier

17. **Use.** Class C r.f. power amplifiers are used when a high power output at a high level of efficiency is required and when the input to the stage is a radio frequency of constant amplitude, i.e., unmodulated or frequency-modulated. A Class C stage cannot be used to amplify an amplitude-modulated input since, as shown in Fig. 10, anode current ceases altogether at the troughs of the input modulation envelope, and distortion of the output waveform results. When amplitude-modulation is required at this stage, it is obtained by adjustment and operation of the Class C amplifier itself (termed '*high level modulation*').

18. **Circuit.** The basic circuit of a Class C r.f. power amplifier is similar to that given in Fig. 7. The valve is biased beyond cut-off, the bias voltage being obtained from a fixed external source of supply, from the use of a grid leak and capacitor or from a combination of these methods.

19. **Waveforms.** Fig. 11 shows a graph illustrating the phase relationships between current and voltage in a Class C stage. The voltage V_a applied to the anode of the valve is the h.t. voltage E_a minus the alternating voltage drop v_L across the anode load; that is, $V_a = E_a - v_L$. The voltage V_g applied to the grid is the grid bias voltage $- E_g$ plus the signal voltage v_g; that is, $V_g = - E_g + v_g$ The two voltages V_a and V_g are sinusoidal, since both are developed across tuned circuits, and they are in phase opposition. Because of the bias, the anode current i_a flows for *less* than a half-cycle; that is, the angle of current flow θ_a is less than 180°. Since the grid is driven positive at the peak of the grid swing, grid current i_g is established during the angle θ_g (less than θ_a).

20. **Efficiency and power output.** The anode current waveform in a Class C amplifier contains d.c., fundamental frequency and harmonic components. Provided the tuned anode load is correctly adjusted, only the fundamental frequency component is effective in producing the required r.f. power output. The fundamental frequency component of anode current I_f has a r.m.s. value of $\frac{I_f}{\sqrt{2}}$;

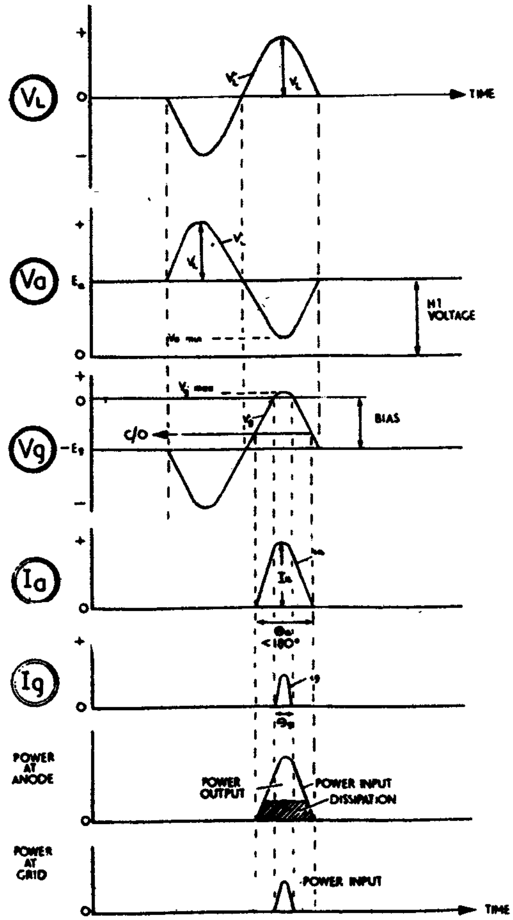


Fig. 11.—PHASE RELATIONSHIPS IN A CLASS C POWER AMPLIFIER

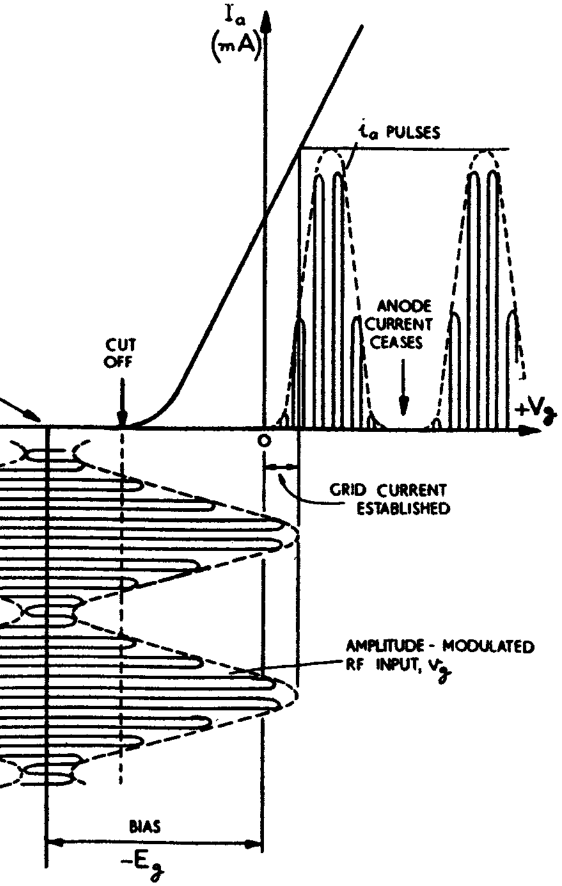


Fig. 10.—DISTORTION OF AMPLITUDE-MODULATED WAVE IN A CLASS C POWER AMPLIFIER

the r.f. voltage at the fundamental frequency across the load has a r.m.s. value of $\frac{V_L}{\sqrt{2}}$; the r.m.s. power output is given by the product of these two components:—

Power output

$$= \tfrac{1}{2}\, I_f V_L = \tfrac{1}{2}\, I_f^2 R_D = \frac{V_L^2}{2R_D} \quad (4).$$

where R_D = dynamic impedance of load. The power supplied to the valve is the product of the h.t. voltage E_a and the d.c. component of anode current I_{dc}. Thus:—

$$\text{Power input} = E_a I_{dc} \quad .. \quad (5).$$

The *difference* between the d.c. power input P_i and the a.c. power output P_o is dissipated as heat, mainly at the valve anode, and gives the anode loss P_a. The conversion efficiency η of a power amplifier is defined as the *ratio* of the a.c. power output to the d.c. power input to the anode. Thus:—

$$\eta = \frac{P_o}{P_i}$$

$$\eta = \frac{P_o}{P_o + P_a}$$

$$\therefore P_a = P_o\left(\frac{1}{\eta} - 1\right) \quad .. \quad .. \quad (6).$$

As valves are rated by their anode dissipation, a high conversion efficiency increases the output power available for a given type of valve and a given d.c. power supply.

Example. A Class A power amplifier delivering 1 kW of a.c. power at 25 per cent efficiency will dissipate 3 kW at the anode, and will require an input of 4 kW.

A Class C power amplifier at 75 per cent efficiency will deliver 3 kW of a.c. power with the same input and a dissipation of 1 kW, thus permitting smaller valves to be used.

Tank Circuit

21. The anode tuned circuit is sometimes termed a 'tank' circuit because one of its functions is to store energy during parts of each cycle so that a sinusoidal voltage is maintained across it even if the valve current is in the form of pulses. The energy delivered to the tank circuit is not always continuous because in a Class C amplifier the anode current is established only for the angle θ_a. However, the energy delivered by the tank circuit to the load is continuous. Thus, the energy delivered to the tank circuit from the valve during the time I_a is flowing must be considerably in excess of the power delivered to the load at any instant. This excess energy is in effect, 'stored' by the tank circuit to supply the output during the periods when the anode current is zero.

Tuning Adjustment in a Class C Amplifier

22. For maximum power output the tank circuit must be tuned to resonance with the frequency of the grid excitation voltage. At resonance, the tuned circuit impedance is a maximum and purely resistive $\left(R_D = \frac{L}{CR}\right)$. Thus at resonance, V_L is a maximum and the minimum value of V_a is at its lowest (see Fig. 11). Since peak anode current coincides with minimum anode voltage, the peak current at resonance drops, so that the d.c. component of anode current is at a minimum. Thus, the tank circuit is tuned for *minimum* reading in a d.c. milliammeter placed in the anode lead. In most transmitters it is necessary to reduce the d.c. anode current for the period the tank circuit is off resonance, otherwise an excessive current will flow because V_L is small off resonance and the minimum value of V_a correspondingly high. Therefore, until the circuit is correctly tuned, with the d.c. anode current at a minimum, the h.t. voltage is reduced. The external load on the tank circuit will affect the tank circuit impedance and therefore V_L. Thus, after tuning for a 'dip' in the meter and adjusting the coupling correctly between the the tank circuit and the load, it is necessary to retune.

Power Handling Capacity

23. The triode, beam and pentode valves used for Class A and Class B amplification are equally suitable for Class C operation. Beam and pentode valves normally have a higher gain than triodes, so that for the same r.f. power output a smaller grid excitation power is required. However, beam and pentode valves produce heat at the screen in addition to the heat produced at the anode, and because it is often difficult to dissipate this heat at the inner electrodes, triode valves are preferred for the higher power outputs.

The power that any valve can handle is determined by:—

(*a*) The voltage that may safely be applied to the anode and other electrodes.

(*b*) The electron emission from the cathode.

(*c*) The amount of power that can be dissipated within the valve without overheating.

24. Glass-envelope valves are commonly used to develop output powers up to about 1 kW. For good radiation of heat to the glass walls, the anode structure is normally carbonized to take advantage of the high radiating efficiency of a black body. For output powers in excess of 1 kW up to about 20 kW, the copper anode of the valve constitutes part of the containing envelope of the valve and is fitted with cooling fins; forced-air cooling of the anode by a blower motor is used. For still higher output powers the anode, which again constitutes part of the valve envelope, is fitted with a jacket through which water circulates. All cooling supplies need to be started before the application of any voltage. An example of a water-cooled triode capable of supplying a power output of 70 kW is illustrated in Fig. 12. Anode voltages as high as 10,000 to 20,000 volts are used in the higher powered valves. The cathodes are usually directly-heated tungsten types capable of giving adequate emission, although oxide-coated indirectly-heated cathodes find use in valves that operate at anode potentials of the order of 1,000 volts and less. For new high-power transmitting valves and valves which have been out of service for a considerable time, it is necessary to condition or '*season*' the valve by a gradual application of the h.t. voltage. A low h.t. voltage is first applied and left on for some time; after this period the voltage is gradually increased up to the full value in easy stages. Typical ratings for valves used as Class C power amplifiers are given in Table 1.

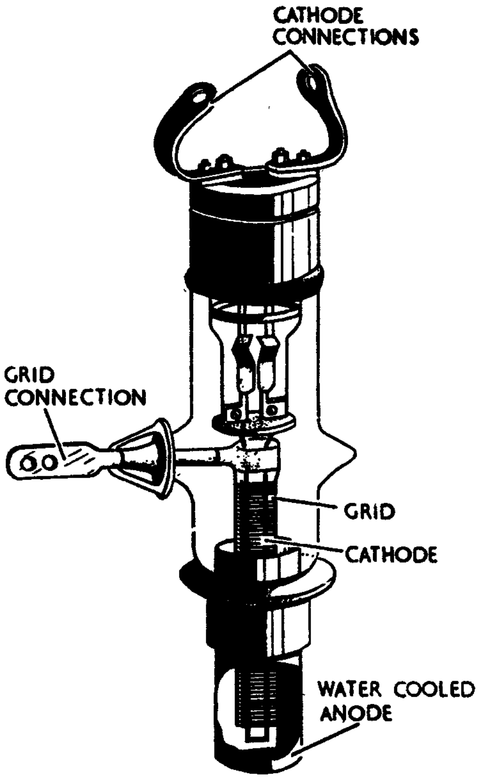


Fig. 12.—TYPICAL HIGH-POWER WATER-COOLED TRIODE

Type	Cathode		Permissible Dissipation	μ	H.T., E_a	Bias E_g	Grid Swing	D.C. I_a (mA)	D.C. I_g (mA)	Grid Power	Output Power	Efficiency
	Volts	Amps										
CV 3523 Tetrode	6	1·2	20 W	4	600 V	–85 V	100 V	113	3	0·3 W	52 W	74%
CV 2657 Triode	7·5	3	35 W	15	1,250 V	–175 V	300 V	70	15	4 W	65 W	74%
CV 2131 Tetrode	5	14·5	250 W	5	4,000 V	–225 V	300 V	312	9	2·5 W	1,000 W	80%
5762 (USA) Tetrode	12	29	2,500 W	29	5,000 V	–750 V	1,075 V	1,100	250	240 W	4,100 W	75%
5671 (USA) Triode	11	285	25,000 W	39	15,000 V	–1,500 V	2,270 V	6,000	1,000	2,000 W	70,000 W	77%

TABLE 1.—TYPICAL RATINGS FOR CLASS C AMPLIFIERS

Valves in Push-pull and in Parallel

25. To obtain an increased r.f. power output, tuned Class B and Class C power amplifiers may be connected in push-pull or in parallel. The push-pull and parallel connections of valves are discussed in Sect. 10, Chap. 2, where the advantages and limitations are listed. The power output available from a valve is determined by the amount of power that can be safely dissipated within the valve. However, by connecting two valves in push-pull or in parallel, the total permissible dissipation for the stage is doubled, and by making adjustments to the input power and grid excitation as appropriate, the power output available, relative to that of a single valve of similar rating, may be doubled. For very high power outputs of the order of 100 kW, two or more parallel banks of valves in push-pull may be used. One such arrangement using two parallel banks is shown in Fig. 13. V_1 and V_3 are connected in parallel, as are V_2 and V_4, while V_1 — V_3 and V_2 — V_4 are in push-pull. A conventional push-pull input circuit is used, the output circuit is shunt-fed, and a push-pull neutralizing arrangement is included. The two points marked 'h.t.+' are common. Balancing networks are inserted across the valve heaters so that the negative end of h.t., which is connected to earth, supplies electrons equally to both ends of the valve heater, in order to give a uniform emission of electrons.

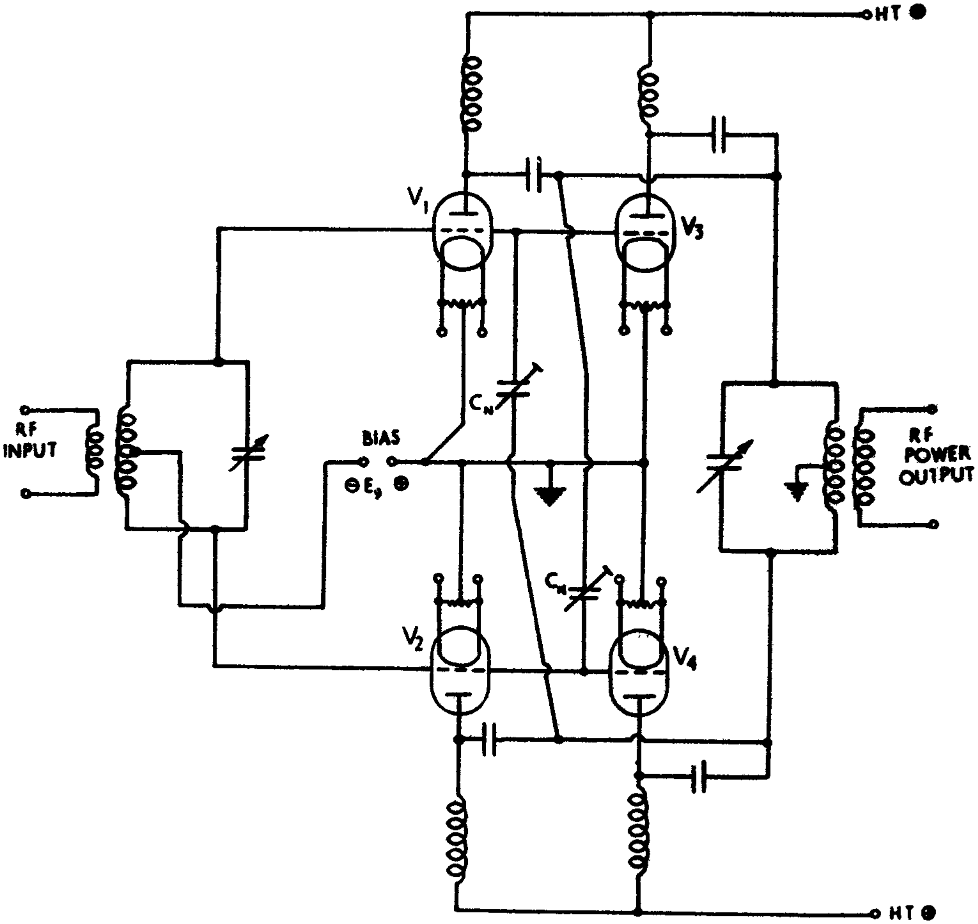


Fig. 13.—HIGH-POWER PARALLEL PUSH-PULL POWER AMPLIFIER

Frequency-multipliers

26. Frequency-multiplier circuits (sometimes termed '*harmonic generators*') are used both in radio transmitters and radio receivers, particularly at v.h.f., to obtain high frequency outputs from low frequency sources. It is often much easier to obtain certain effects at a low frequency and multiply the resultant, rather than obtain those same effects directly at the high frequency.

27. The simplified circuit of a frequency-multiplier is shown in Fig. 14, and is similar to that of a conventional Class C amplifier.

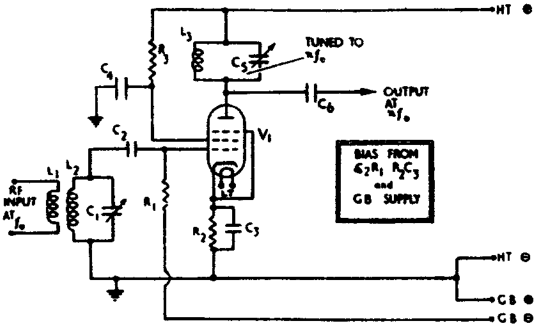


Fig. 14.—BASIC FREQUENCY-MULTIPLIER CIRCUIT

The anode current pulses of a Class C amplifier operated from a constant frequency source, are rich in harmonics. For normal Class C working, the tank circuit is tuned to the fundamental frequency of the input, and the harmonic components of anode current have then so little effect that they can be neglected. If however, the tank circuit is tuned to one of the harmonics, an output power at the harmonic frequency results. The amplitudes of the harmonic components contained in the anode current pulse depend on the waveform of the pulse, and this is controlled by the angle of flow θ_a. Thus, by altering the grid bias, θ_a is adjusted to give a large amplitude to the required harmonic. By increasing the grid bias (θ_a smaller) the amplitudes of the higher order harmonics are increased, and in a triode for a second harmonic output θ_a is about 110°, reducing progressively to a value of about 65° for a fifth harmonic output. In relation to the power output obtainable from the same valve used as a normal Class C amplifier, the harmonic power output decreases with the order of the harmonic, being approximately inversely proportional to the order of the harmonic. With triode valves, it is not usual to obtain more than five times the frequency of the fundamental from any one stage; if biased sufficiently negative, pentodes can be employed to produce harmonics up to the ninth.

28. For frequency-doubling, a single valve circuit is not the most efficient arrangement because the tank circuit receives a pulse of current only on *alternate* cycles of its swing. If a two valve circuit is arranged with a push-pull input and a parallel output (a 'push-push' system), and the valves biased beyond cut-off, a more efficient doubling arrangement results. The circuit is shown in Fig. 15. As each valve in this circuit conducts on alternate half-cycles of the input, the tank circuit now receives a pulse of current *every cycle* of its swing.

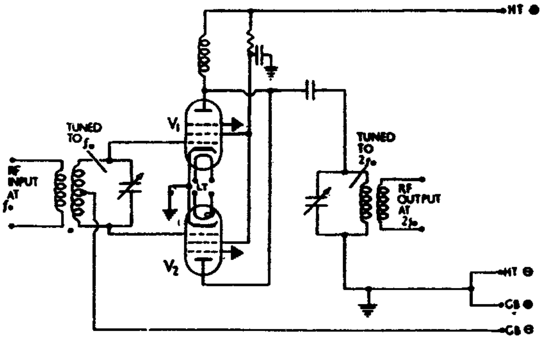


Fig. 15.—'PUSH-PUSH' DOUBLING CIRCUIT

29. A *conventional* push-pull amplifier cancels all even-order harmonics in its output circuit. Thus, if a push-pull arrangement is used with its output circuit tuned to the third harmonic of the input, an efficient frequency-trebler is obtained.

SECTION 12

VALVE OSCILLATORS

SECTION 12

VALVE OSCILLATORS

SECTION 12

CHAPTER 1

TUNED CIRCUIT (LC) OSCILLATORS

PART 1A, SECTION 12, CHAPTER 1

TUNED CIRCUIT (LC) OSCILLATORS

Introduction

1. Basically, an *oscillator* does nothing more than generate an alternating voltage at a desired frequency. A valve circuit, or a transistor circuit, arranged to act as an a.c. generator constitutes an oscillator. The frequency at the output is determined by the circuit constants and components, and may be anything from a few cycles per second to many mega-cycles per second; the frequency may be fixed, adjustable in steps or continuously variable.

2. Oscillators are used extensively in communication, radar and test equipments to produce:—

(*a*) radio frequency alternating voltages, which may be subsequently amplified and modulated in a *radio transmitter* for radiation by the transmitter aerial over long distances;

(*b*) Audio frequency alternating voltages for *modulation* in certain radio transmitters;

(*c*) Audio frequency or radio frequency alternating voltages in *test equipment* (signal generators and frequency meters) to enable checks to be carried out on transmitters, receivers and other main equipment;

(*d*) radio frequency alternating voltages in *receivers* to provide 'frequency changing';

(*e*) radio frequency alternating voltages in *radar* systems for timing and switching.

It is seen then that oscillators form an essential part of many systems.

3. For most purposes, the output waveform of the oscillator is required to be *sinusoidal*, and the types of oscillator used to produce a sine wave output include:—

(*a*) tuned circuit (LC) oscillators, considered in this Chapter;

(*b*) crystal - controlled oscillators, considered in Chapter 2;

(*c*) phase-shift (RC) oscillators, considered in Chapter 3.

In other instances, the oscillator is required to produce a *square-wave* output waveform. Such oscillators are called 'relaxation' oscillators; they are introduced in Chapter 3.

Free Oscillations

4. Anything which swings backwards and forwards in a uniform way is said to be 'oscillating'; a child's swing moving back and forth 'oscillates'; a pendulum swinging on a clock 'oscillates'. An electronic oscillator is a simple circuit consisting of a capacitor and an inductor connected in parallel. A current established in this circuit will flow first in one direction round the circuit and then in the other direction; the current moving back and forth in this way represents an '*oscillatory*' current. If a capacitor C, charged to V volts, is connected to an inductor L of negligible resistance as shown in Fig. 1(a), the capacitor will discharge through the inductor and oscillations of current and voltage will occur at a frequency determined by the values of L and C. Initially the energy $\frac{1}{2}CV^2$ is stored in the capacitor, where V is the maximum p.d. between the plates. When V has its maximum value, the current in the circuit is zero (C fully charged).

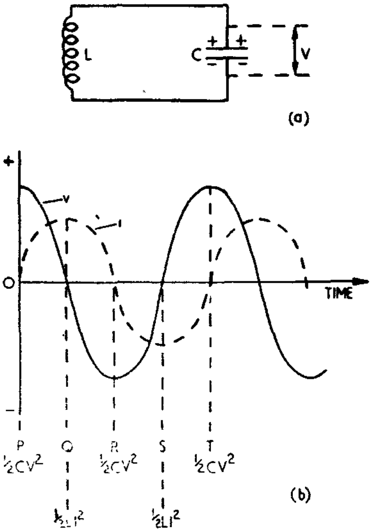


Fig. 1. FREE OSCILLATIONS

As C discharges through L, V falls and the current in the circuit rises until, when C is fully discharged, the current I is a maximum and the energy $\frac{1}{2}CV^2$ originally stored in the electric field of the capacitor is transferred as energy $\frac{1}{2}LI^2$ to the magnetic field of the inductor. Since there is no p.d. to sustain the current, the magnetic field begins to collapse and induces a back e.m.f. in L, with the result that current continues to flow *in the same direction* through L to charge C in the opposite sense from its original charge. When C is fully charged, current ceases and the energy $\frac{1}{2}CV^2$ is again stored in the electric field of the capacitor. This sequence of events is shown by the points P, Q and R in Fig. 1(b). The whole process is repeated in the opposite direction, as shown by the points R, S and T, until the capacitor is again fully charged in the initial sense.

5. If there is no energy loss in the circuit the above process repeats itself indefinitely, the natural frequency f_n of the oscillations occurring in the circuit being given by:—

$$f_n = \frac{1}{2\pi\sqrt{LC}} \quad \dots \quad (1).$$

Since however, there is always some resistance R present with practical circuit elements, the amplitude of each successive oscillation decreases until eventually all the energy is dissipated, mainly in the form of heat in R, and oscillations cease. The oscillations are said to be '*damped*' as illustrated in Fig. 2, the natural frequency being modified by the presence of R to give:—

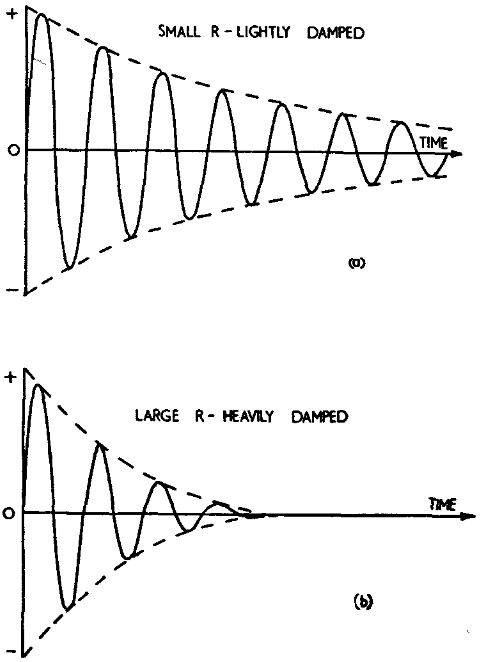


Fig. 2. DAMPED OSCILLATIONS

$$f_n = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{4L^2}} \quad \dots \quad (2)$$

The rate at which the oscillations die away depends on the factor $\sqrt{\frac{R^2}{4L^2}} = \frac{R}{2L}$; this is called the '*damping coefficient*'. A large value for R causes the oscillations to die away rapidly.

Note. The natural frequency of a circuit is not quite the same as the resonant frequency f_0 (see Bk 1, Sect. 5, Chap. 2, Para The former applies to the frequency of oscillations that occur when the circuit is allowed to oscillate freely; the latter applies to a 'driven' circuit and is the frequency at which the impedance of the circuit is a pure resistance.

Maintenance of Oscillations

6. In order to maintain oscillations in a tuned circuit, one of two systems is employed:—

(*a*) *Feedback oscillators*, in which extra energy from some external source is fed into the tuned circuit. Provided the added energy equals that being lost by damping, oscillations are maintained.

(*b*) *Negative resistance oscillators*, in which some device having a 'negative' resistance equal in value to the positive damping resistance of the circuit is placed in parallel with the tuned circuit. The two resistances neutralize each other and oscillations are maintained.

FEEDBACK OSCILLATORS

Principle of Operation

7. In feedback oscillators, the d.c. power supply is the required external source of energy, and the valve controls the energy that is added to the tuned circuit in order to counteract the damping losses. To maintain oscillations in the tuned circuit, the added energy must be of the correct *phase* and *amplitude*. The addition of energy requires the injection into the tuned circuit of a

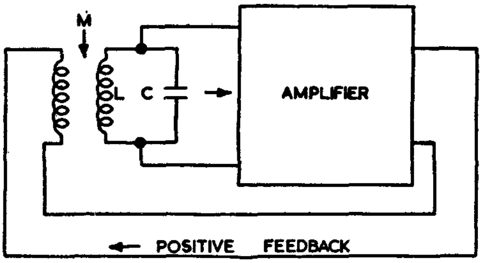


Fig. 3. FEEDBACK OSCILLATOR PRINCIPLE

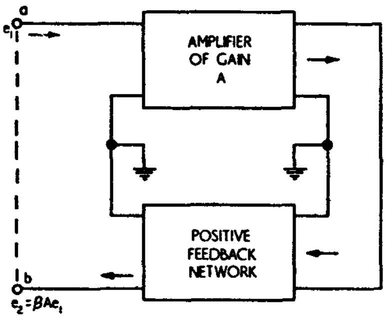


Fig. 4. AMPLIFIER SUPPLYING ITS OWN INPUT

sinusoidal e.m.f. *in phase* with the current circulating in the oscillatory circuit. The normal method of doing this is to connect the tuned circuit to the input of an amplifier and couple the output of the amplifier back to the oscillatory circuit; that is, the basic arrangement (Fig. 3) constitutes a *positive feedback* amplifier.

8. Provided that the voltage fed back to the oscillatory circuit is of sufficient amplitude and of the correct phase, the circuit will oscillate at a frequency determined by the circuit constants. If the amplitude of the feedback voltage is insufficient to neutralize altogether the effect of the damping losses, oscillations will decay, although at a rate slower than that without the application of feedback. If the amplitude of the feedback voltage is greater than that required for the replacement of the lost energy, oscillations increase in amplitude. This continues until the grid swing is such that the valve anode current is limited by the bottom bend and by grid current; the amplifier gain and subsequent feedback then fall to such values that the circuit damping is exactly neutralized and the oscillator settles down to deliver an output of steady amplitude.

9. A valve acting as an a.c. generator requires a circuit arrangement whereby the valve provides its own input of such magnitude and phase that the energy lost by the circuit is replaced, and oscillations are maintained. In Fig. 4, e_1 is the instantaneous voltage at a fixed frequency applied to an amplifier of gain A. The output of the amplifier is applied to a feedback network that develops an output voltage $e_2 = \beta Ae_1$. If this circuit is adjusted so that $e_2 = e_1$ in both magnitude and phase, then on joining points *a* and *b* the amplifier continues to supply its own input at the original frequency and the circuit oscillates. Thus, *any amplifier circuit will oscillate at the frequency at which the feedback voltage is equal in magnitude and phase to the original input voltage.* The condition for the maintenance of oscillations is:-

$$e_2 = e_1 \text{ (in magnitude and phase)}$$

$$\therefore \beta Ae_1 = e_1$$

$$\therefore \beta A = 1\angle 0^\circ \quad \ldots \quad (3).$$

10. Equation (3) shows that for the maintenance of oscillations, the gain of the amplifier and feedback network together (from the input, through the amplifier and the network, back to the input) should be unity; that is, the loss introduced by the oscillatory circuit and feedback network must be exactly neutralized by the gain of the amplifier. If βA is greater than unity initially, the gain drops as explained in Para. 8, and the circuit settles down to produce oscillations of such magnitude that $\beta A=1$. The angle $\angle 0^\circ$ is inserted to show that the *phase* as well as the magnitude of the feedback voltage must be correct. In a single stage amplifier, a phase shift of 180° exists between grid and anode voltages. Thus, to provide a phase shift of 0° in a single stage oscillator, a further 180° phase shift must be introduced into the feedback path from anode to grid.

11. A basic circuit arrangement that satisfies the required conditions is shown in Fig. 5, where the oscillatory circuit in the anode is coupled by mutual inductance to the grid of the valve. On applying h.t., the anode current rises and induces a back e.m.f. in L_1; C_1 charges and the oscillatory action commences. L_1 is coupled by mutual inductance M to L_2 and the oscillatory current in L_1

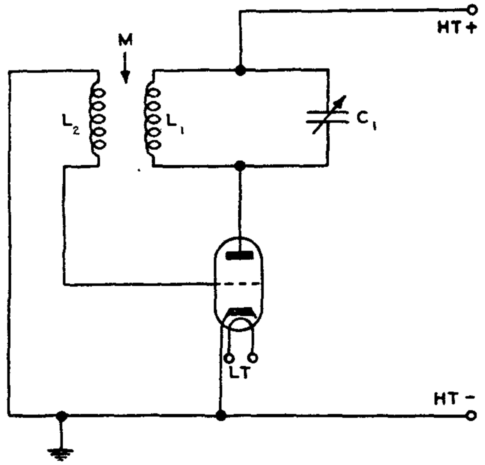


Fig. 5. BASIC FEEDBACK OSCILLATOR CIRCUIT

induces an oscillatory e.m.f. in L_2 which is applied between grid and cathode of the valve. In this way, the necessary feedback voltage is obtained. The two conditions necessary for the maintenance of oscillations are:—

(*a*) the *magnitude* of the feedback voltage must be sufficient to overcome all the losses in the circuit; that is, βA must be unity:

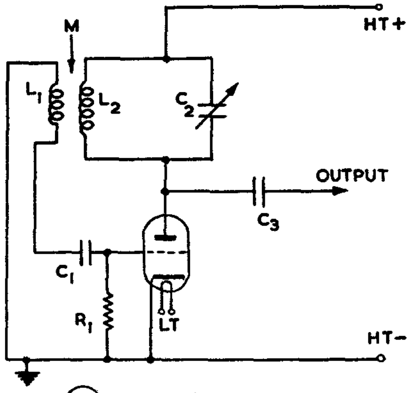
(*b*) the *phase* of the feedback voltage must be zero; that is $\theta = 0^\circ$.

The first condition is determined by the *magnitude* of the mutual inductance M and this may be adjusted by altering the coupling between L_1 and L_2 until $\beta A = 1$. Since a phase shift of 180° occurs between grid and anode in the valve, the second condition is satisfied by introducing a *further* 180° phase shift into the feedback path between anode and grid; this is accomplished by transformer action between L_1 and L_2 when L_2 is wound in the *correct direction* relative to L_1. In this way, oscillations are maintained. There are several circuit arrangements that satisfy the conditions necessary for the maintenance of oscillations; some of these are considered in the succeeding paragraphs.

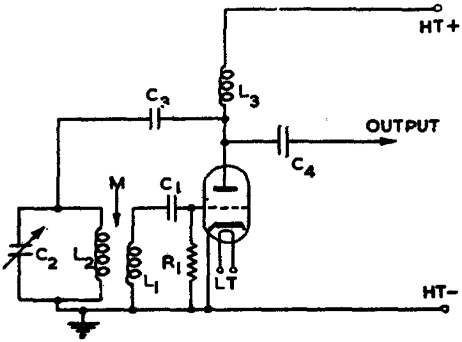
Tuned Anode Oscillator

12. Fig. 6 shows two arrangements for an oscillator known as the tuned anode oscillator. In (a), the h.t. is applied to the valve through the tuned circuit and represents a '*series-fed*' circuit. With this arrangement, the tuned circuit is at a high d.c. potential with respect to earth, and this introduces practical disadvantages. The components require to be well insulated, and since the rotor plates of the tuning capacitor are not earthed, there is the danger of high voltage to the operator; in addition, a capacitance exists between the rotor plates and the chassis (earth). These disadvantages are overcome in the '*shunt-fed*' circuit shown in Fig. 6(b). In this case, the tuned circuit is in parallel with the valve, the blocking capacitor C_3 being inserted to prevent short-circuit of the supply through the tuned circuit; C_3 has a low reactance at the operating frequency. The r.f. choke L_3, which has a high reactance at the operating frequency, acts as the anode load across which the oscillator output is developed; without this choke the oscillator would be effectively short-circuited by the low impedance of the h.t. supply. Both circuits use an automatic grid biasing arrangement, R_1C_1.

13. The circuits of Fig. 6 are 'shock-excited' into oscillation by applying h.t. to



(a) SERIES-FED CIRCUIT



(b) SHUNT-FED CIRCUIT

Fig. 6. TUNED ANODE OSCILLATOR

the emitting valve, and oscillations are maintained in the manner described in Para. 11; that is, the *magnitude* of M determines the amount of energy fed back, and the *sign* of M determines the phase of the feedback voltage. The resonant frequency of the tuned anode oscillator is affected by the resistance R associated with the circuit, and by the anode slope resistance r_a of the valve; it is given by:—

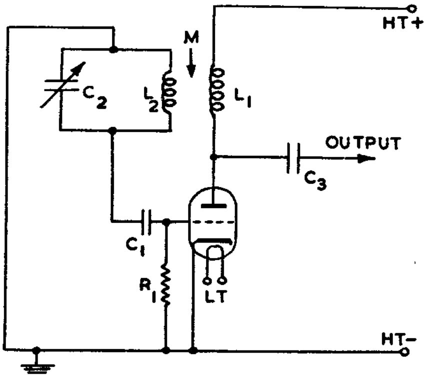
$$f_o = \frac{1}{2\pi\sqrt{L_2C_2}}\sqrt{1+\frac{R}{r_a}} \quad \ldots (4)$$

In general, R/r_a is very small and for most purposes may be neglected, when:—

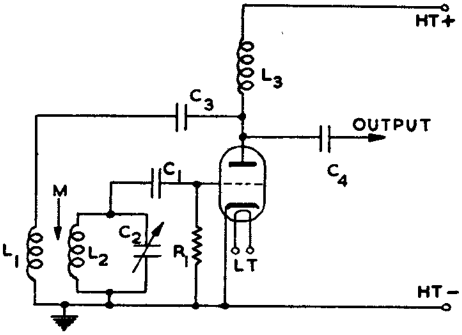
$$f_o = \frac{1}{2\pi\sqrt{L_2C_2}} \quad \ldots (5).$$

Tuned Grid Oscillator

14. Fig. 7 shows typical series-fed and shunt-fed tuned grid oscillators. These are similar to the tuned anode oscillator, the only point of difference being that in the tuned grid circuit the grid winding of the coupling transformer is tuned and the anode winding is untuned. The *magnitude* and *phase* of the feedback voltage from anode to grid is again determined by the *value* and *sign* respectively of the mutual inductance M, and the approximate frequency of oscillation is $f_o \simeq \frac{1}{2\pi\sqrt{L_2C_2}}$.



(a) SERIES-FED CIRCUIT



(b) SHUNT-FED CIRCUIT

Fig. 7. TUNED GRID OSCILLATOR

Meissner Oscillator

15. Fig. 8 shows the circuit of a Meissner oscillator. It differs from the tuned anode and tuned grid oscillators (sometimes loosely termed 'Meissner' oscillators) in that there is no *direct* connection between the oscillatory circuit and the valve, and the frequency of oscillation is more independent of circuit resistance. The feedback voltage is determined by the magnitude and sign of the combined mutual inductance in the circuit.

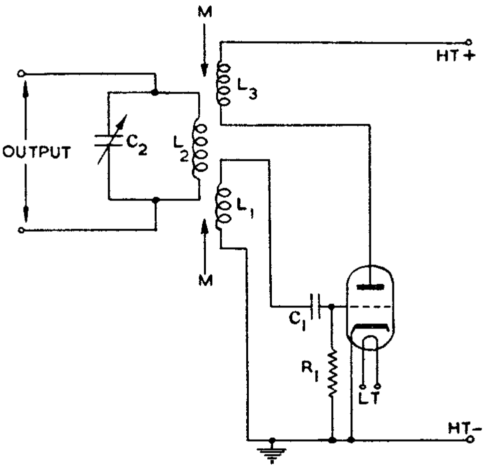
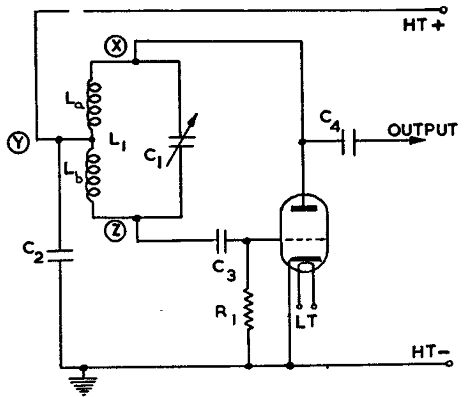


Fig. 8. MEISSNER OSCILLATOR

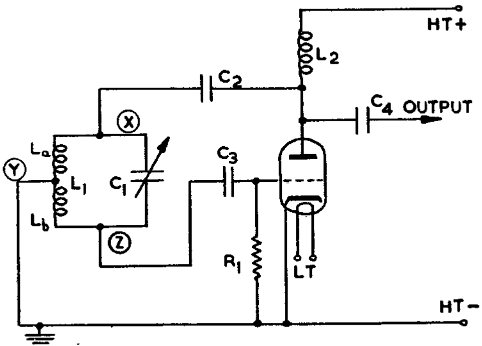
Hartley Oscillator

16. Fig. 9 shows series-fed and shunt-fed Hartley oscillator circuits. In this type of oscillator, the tuned circuit is connected between grid and anode, and the cathode is connected to a tapping on the tuned circuit inductance L_1. The capacitor C_2 is a d.c. blocking capacitor that prevents short-circuit of the h.t. supply but provides a relative short-circuit for r.f. current. Its value is such that its reactance is negligible at the oscillatory frequency so that point Y

is at earth potential to r.f. in both circuits. Thus when the circuit is oscillating, the potential at X with respect to Y is 180° out of phase with the potential at Z with respect to Y. In this way the 180° phase shift in the external feedback path between anode and grid is obtained, and the condition $\beta A = 0$ is satisfied. The other condition for the maintenance of oscillations is the *magnitude* of the feedback voltage and this is dependent on the ratio $\frac{L_b}{L_a}$. The nearer Y is to the anode end of L_1, the greater is the proportion of oscillatory voltage used as feedback, but at the same time the effective anode load of the valve is decreased and so also is the gain. The approximate frequency of oscillation for this circuit (neglecting resistance) is $f_o \simeq \frac{1}{2\pi\sqrt{L_1C_1}}$.



(a) SERIES-FED CIRCUIT



(b) SHUNT-FED CIRCUIT

Fig. 9. HARTLEY OSCILLATOR

17. In the 'modified' or 'inverted' Hartley oscillator circuit, shown in Fig. 10, the *anode end* of the tuned circuit is connected directly to earth. Since the anode is at earth potential to r.f. (reactance of C_3 is low) the required Hartley connections are satisfied. This circuit has the advantage over the conventional Hartley that the tuning capacitor is directly earthed. No blocking capacitors in series with the tuned circuit are necessary.

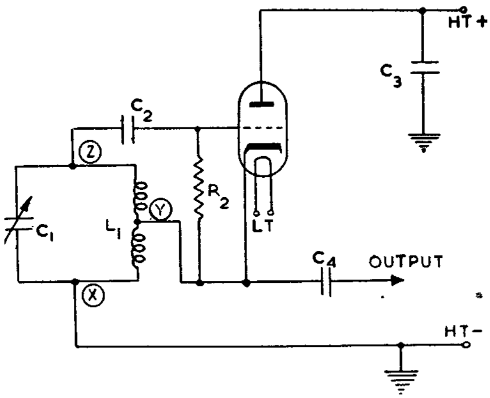


Fig. 10. INVERTED HARTLEY OSCILLATOR

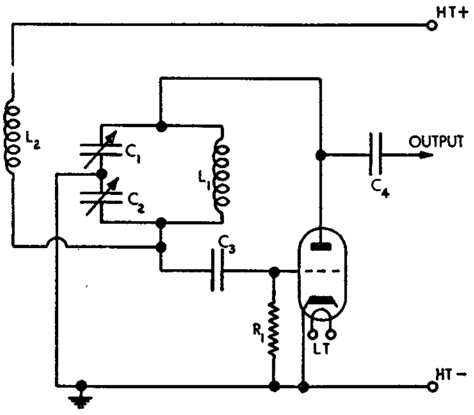
18. A Hartley oscillator offers advantages over the tuned anode and tuned grid oscillators, the greatest advantage being that the feedback winding is part of the tuned circuit and does not offer the same difficulties as the other circuits in which the feedback winding may resonate with stray capacitances. Because of this the useful upper frequency limit for tuned anode and tuned grid oscillators is about 50 Mc/s, whereas the Hartley oscillator can give satisfactory operation to frequencies as high as 150 Mc/s.

Colpitts Oscillator

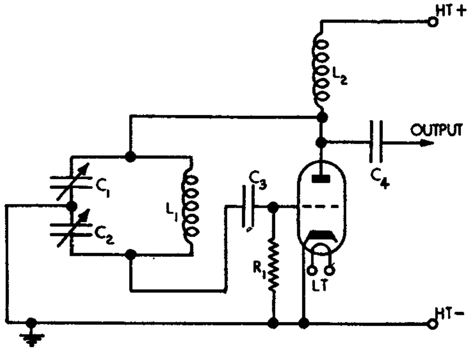
19. In the Colpitts oscillator, one end of the tuned circuit is connected to the anode and the other end to the grid; the cathode is taken to the junction of two capacitors in the tuned circuit. The required 180° phase shift between anode and grid in the external feedback path is therefore obtained in a manner similar to that of the Hartley oscillator. Fig. 11(a) shows a series-fed circuit, the choke L_2 preventing short-circuit through the h.t. supply to earth of the r.f. input to the grid. The more usual shunt-fed circuit is shown in Fig. 11(b). The magnitude of the feedback voltage must be sufficient to maintain oscillations, and this is deter-

mined by the ratio $\frac{C_1}{C_2}$. The approximate frequency of oscillations is $f_o \simeq \frac{1}{2\pi\sqrt{L_1C}}$, where C is the capacitance of C_1 and C_2 in series.

The Colpitts oscillator is used extensively on all frequency ranges up to frequencies of the order of 200 Mc/s. It is not so liable to parasitic oscillations as the Hartley type because the valve inter-electrode capacitances merely augment the tuned circuit capacitances C_1 and C_2, and at the higher frequencies may be used instead of actual capacitors. An 'inverted' Colpitts oscillator, similar in construction to the inverted Hartley circuit of Fig. 10, is in common use. A typical arrangement is shown in Fig. 12. The choke L_2 is inserted to ensure a d.c. return to earth for the cathode, while at the same time preventing a short circuit of the oscillatory voltage developed across C_1.



(a) SERIES-FED CIRCUIT



(b) SHUNT-FED CIRCUIT

Fig. 11. COLPITTS OSCILLATOR

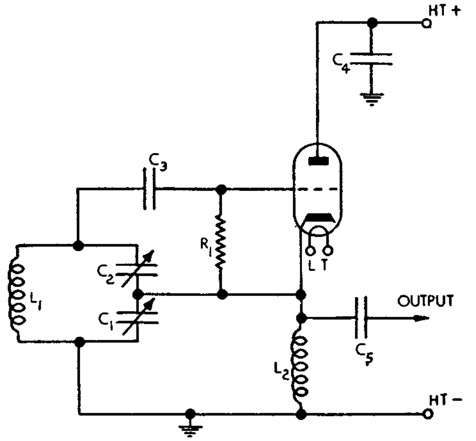


Fig. 12. INVERTED COLPITTS OSCILLATOR

Push-pull Oscillator

20. Fig. 13 shows a push-pull arrangement that may be used to provide the conditions necessary for the maintenance of oscillations. If the voltage across the tuned circuit LC makes the grid of V_1 positive with respect to earth and that of V_2 negative, the anode current of V_1 increases and that of V_2 decreases. The anode potential of V_1 thus *decreases* and that of V_2 *increases*, these changes being applied via the feedback networks C_1R_1 and C_2R_2 to the grid of V_2 and the grid of V_1 respectively. Since V_2 anode is in phase with V_1 grid, and V_1 anode is in phase with V_2 grid, the feedback is in the *correct phase* for the maintenance of oscillations. The capacitors C_1 and C_2 in the feedback networks are d.c. blocking capacitors that are inserted to prevent short-circuit of the h.t. supply to earth through the input tuned circuit; their values are such that they have a low reactance at the operating frequency. The values of the equal resistors R_1 and R_2 in the feedback networks determine the *magnitude* of the feedback voltage, and their values are such that, together with the gain of the stage, $\beta A = 1$. With the two conditions necessary for the maintenance of oscillations satisfied, the resultant current in the primary of the output transformer produces the oscillatory voltage across the output terminals. The

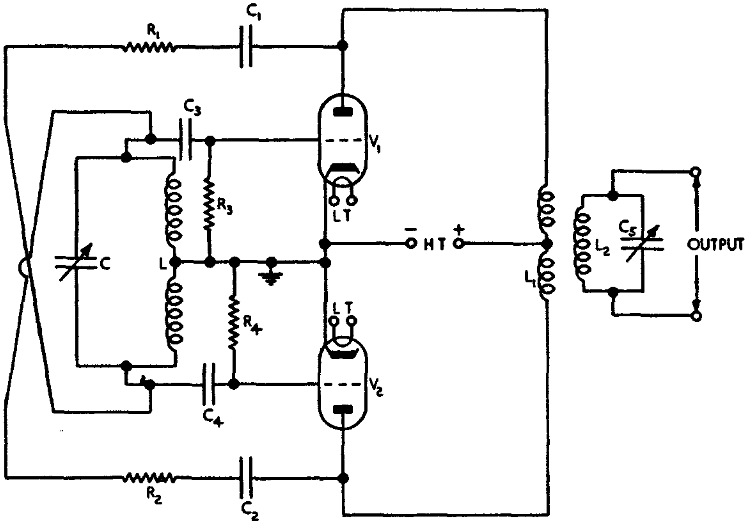


Fig. 13. PUSH-PULL OSCILLATOR

approximate frequency of oscillation is $f_0 \simeq \frac{1}{2\pi\sqrt{LC}}$. Each valve is biased separately, V_1 by C_3R_3 and V_2 by C_4R_4. The self-bias produced is then proportional to the drive to the valve, so that the output from each valve is equal. The push-pull oscillator has the same advantages as the push-pull amplifier (see Sect. 10, Chap. 2, Para. 21); the main advantage is the cancellation of even-order harmonics which means that a greater power output per valve can be obtained with less distortion.

Operating Conditions for Oscillators

21. Oscillators are operated under Class A, Class B or Class C bias conditions, depending on the requirement. Class A operation is used when special requirements are to be met; for example, in high quality laboratory instruments where purity of waveform is essential. Something approaching Class B operation is common for local r.f. oscillators in superheterodyne receivers where the bias is obtained from a grid leak and capacitor. Class C operation is generally employed for transmitters, as it gives high anode efficiency and maximum power output for a given valve.

22. **Use of grid leak bias.** Automatic grid leak bias (see Sect. 8, Chap. 2, Para. 23(c)) is the method normally used in Class B and Class C oscillators. It has the advantage that before oscillations begin the bias is zero and the valve is operating at the steepest part of its characteristic, giving the best starting conditions. As the oscillations build up, so does the bias, and when the oscillations reach a steady amplitude the bias is just sufficient to limit grid current and give the correct operating conditions for

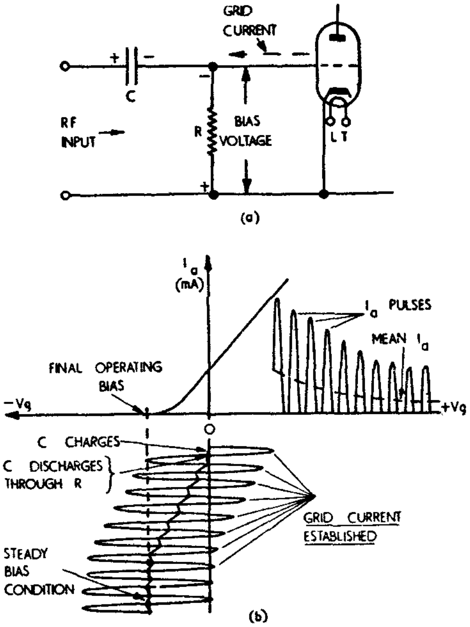


Fig. 14. AUTOMATIC GRID LEAK BIAS

stability of oscillations (Fig. 14). An oscillator employing this method of biasing is self-starting, but if for any reason the oscillator does not oscillate, the grid has zero bias and the anode current may rise to a sufficiently high value to cause damage to the valve; to avoid this, an additional 'safety' bias is sometimes employed, usually in the form of cathode bias. In an oscillator employing grid leak bias, the anode current drops when oscillations commence; this may be used as an indication that the oscillator is working.

23. **Squegging.** The term squegging implies a state of intermittent oscillations. For an oscillator operated correctly under grid leak bias, a variation in the magnitude of the oscillations results in a variation in the magnitude of the feedback voltage, and the bias voltage automatically adjusts itself to maintain stable oscillations. If however, the *time constant* of the grid bias circuit is large enough, a decrease in oscillation amplitude is not *immediately* followed by a decrease in bias voltage. Thus, the gain of the circuit falls and the condition necessary for the maintenance of oscillations ($\beta A = 1$) no longer exists and the oscillation dies down. Oscillations recommence when the grid capacitor has discharged to the point at which anode current flows once more. This sequence of events may be repetitive, the output of the oscillator having the waveform shown in Fig. 15. Under certain circumstances this output waveform is required, in which case squegging of the oscillator is introduced deliberately.

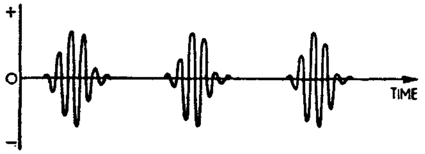


Fig. 15. SQUEGGING OSCILLATIONS

Frequency Stability

24. When oscillation is maintained in a circuit by means of a valve supplied from an external power unit, it is found that the oscillator frequency varies with time from the moment of switching on. The magnitude of this frequency drifts depends on a variety of separate factors (see Para. 25) and efforts are made to reduce frequency instability to a minimum. This is particularly important in communication systems for two main reasons:—

(*a*) if the frequency of the oscillator in a transmitter departs from its assigned value, interference with other transmissions on adjacent channels will result;

(*b*) if either the frequency of the transmitter or the frequency of the local oscillator in the receiver is subject to frequency instability, the operator at the receiving end of the communication link is required to continually re-adjust his receiver.

The importance of frequency stability in a transmitter may be illustrated by an example. On the medium-wave band, broadcast stations on adjacent channels have a separation of 9 kc/s between their mid-band frequencies; each transmitter may have a bandwidth of 8 kc/s so that the separation between adjacent 'sidebands' is only 1 kc/s as illustrated in Fig. 16(a). Assume the frequency assigned to one transmitter is $f_1 = 1{,}500$ kc/s and that assigned to the other is $f_2 = 1{,}509$ kc/s.

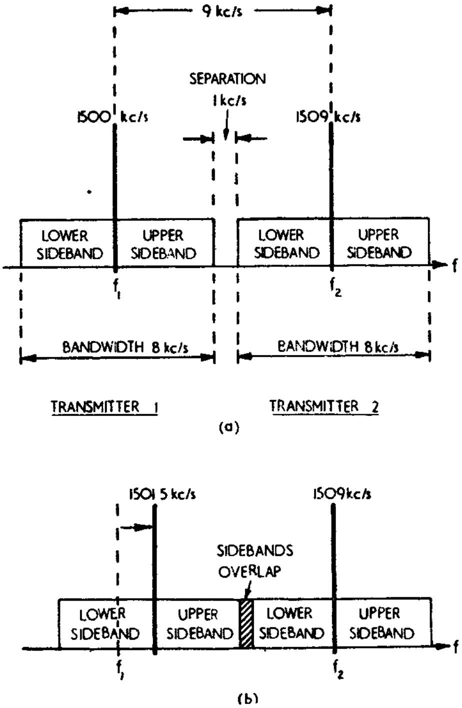


Fig. 16. INTERFERENCE CAUSED BY FREQUENCY INSTABILITY

If the first transmitter has a frequency instability of 1 part in 1,000 (0.1%), its frequency variation is (±) 1.5 kc/s. Assuming that the frequency drifts to 1,501.5 kc/s, Fig. 16(b) shows that the upper sideband of the first transmitter overlaps the lower sideband of the second transmitter, and interference results. This is one reason for the poor reception on the medium-wave broadcast band. To reduce such interference the frequency stability of oscillators must be kept as high as possible. For LC oscillators a frequency stability of the order of 1 part in 20,000 (0.005%) or better is required for most purposes.

25. Frequency instability in LC oscillators is caused by variations in:—

(*a*) *the oscillatory system*, i.e., by changes of L and C;

(*b*) *the maintaining system*, i.e., by changes in the power supply and in the valve;

(*c*) *the utilization system*, i.e., by changes in the load connected to the oscillator output.

26. **The oscillatory system.** The values of L and C are the main factors determining the frequency stability of an oscillator, since the frequency of oscillation is given approximately by $f_o = \frac{1}{2\pi\sqrt{LC}}$; any change of either inductance or capacitance will produce a corresponding variation of frequency. A change in the values of inductance and capacitance may be caused by mechanical vibration, but the most important single cause is variation in *temperature* resulting in expansion and contraction of the components. Suitable steps are taken during design to ensure that these components:—

(*a*) are mechanically stable;

(*b*) have a low temperature coefficient of expansion;

(*c*) are shielded from high temperature sources, e.g., valves. Some oscillator circuits use a capacitor having a *negative* temperature coefficient to counteract the effect of the positive temperature coefficient of the inductor. In other oscillator circuits, the tuned circuit is placed in a temperature-controlled oven to keep the temperature constant.

27. **The maintaining system.** Whilst the values of inductance and capacitance in the oscillatory system are the main factors controlling the frequency, the maintaining system exerts some influence:—

(*a*) *Valve changes.* Frequency changes occur because of changes in the valve inter-electrode capacitances due to temperature changes in the valve itself, mainly during the initial warming-up period. By arranging that the valve capacitances represent only a small part of the tuned circuit capacitance, frequency stability is improved.

(*b*) *Changes in supply voltages.* Variations in the supply voltages to the valve cause a change in the valve constants. As a result, a change occurs in the gain of the stage, and by virtue of Miller effect the input capacitance is altered and hence the frequency. *Stabilization* of the power supplies is essential for good frequency stability.

28. **The utilization system.** When the oscillator is connected directly to a load, any variation in the load affects the frequency. It is shown in Para. 13 that provided the ratio of load R to valve slope resistance r_a is small, its effect on frequency can be neglected. To ensure a small ratio of R to r_a:—

(*a*) A valve with a high value of r_a is used.

(*b*) The load R on the oscillator is kept as small and as constant as possible. To do this, one of three methods is employed:—

(i) The oscillator output is *loosely* coupled to the load.

(ii) A 'buffer' amplifier is inserted between the oscillator and the load (see Para. 29).

(iii) An electron-coupled oscillator is used (see Para. 30).

Buffer Amplifier

29. The most satisfactory way of reducing frequency instability due to a variation in load conditions is to insert another valve operating as a Class A amplifier stage between the oscillator and the load. Such an amplifier is known as a 'buffer' amplifier. Since the oscillator works straight into the grid circuit of the buffer amplifier, where no grid current is established, the load on the oscillator is negligible and constant, and the frequency stability is high. A typical circuit is shown in Fig. 17. In some circuits the input to the buffer amplifier is from the

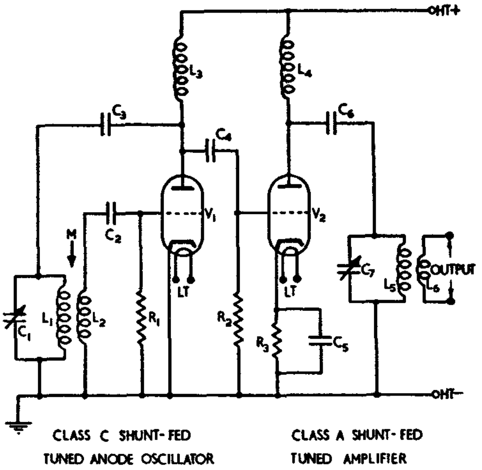


Fig. 17. USE OF BUFFER AMPLIFIER

grid circuit of the oscillator because the waveform here is more free from harmonics than is the anode circuit.

Electron-coupled Oscillator

30. An electron-coupled oscillator, an example of which is shown in Fig. 18, although employing only one valve, is equivalent to an oscillator followed by a buffer amplifier, and has the same high frequency stability. Coupling between the oscillator and the load is by the electron stream within the valve. The valve used is a pentode in which the cathode, control grid and screen grid are operated as a triode oscillator with the screen grid acting as the effective 'anode'. In Fig. 18, the circuit arrangement consists of a series-fed Hartley oscillator, the tuned circuit being connected between the control grid and the screen grid, while the output is taken from the anode. In a pentode, the anode voltage has little effect on the anode current, the latter being varied mainly by changes of control grid and screen grid potentials. Thus, any change of anode voltage caused by a variation of either supply voltage or load, has little effect on the oscillator circuit, and the frequency stability is high. On the other hand, as the screen grid and control grid potentials are varied by the oscillatory action, so the anode current varies to produce the required output voltage across the load.

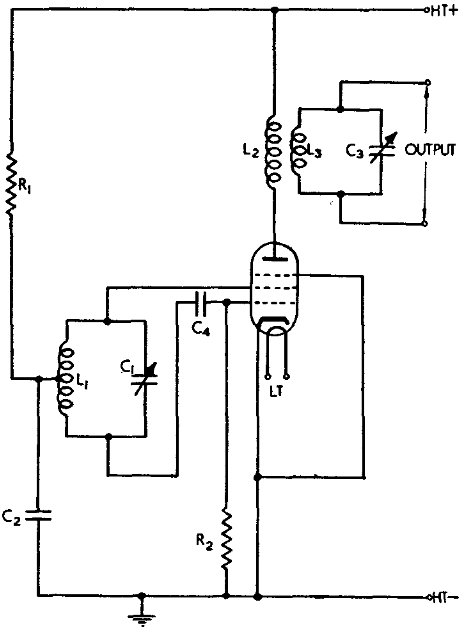


Fig. 18. ELECTRON-COUPLED OSCILLATOR

Franklin Oscillator

31. The 'Franklin' oscillator is essentially a two-stage amplifier providing 180° + 180° phase shift. The circuit incorporates many features designed to give a high frequency stability. A typical circuit is shown in Fig. 19. The tuned circuit, which is temperature-controlled, is connected between grid and cathode of V_1. V_1 is resistance-capacitance coupled to V_2 grid, the feedback necessary for the maintenance of oscillations

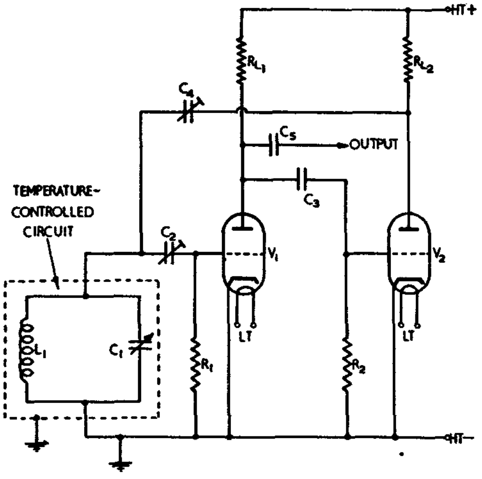


Fig. 19. FRANKLIN OSCILLATOR

being from V_2 anode to V_1 grid via the capacitor C_4. The *phase* of the feedback voltage is therefore correct since there is a phase shift of 180° across each valve. The increased gain resulting from the use of two valves enables the tuned circuit to be very *loosely* coupled to the valves, so that a high degree of frequency stability is achieved. To obtain loose coupling, the values of C_2 and C_4 are of the order of a few pico-farads. The output is usually taken from the anode of V_1 rather than from the anode of V_2, since the latter is connected to the tuned circuit through C_4 and frequency instability may result from loading the oscillator at V_2 anode. With additional precautions (e.g., stabilization of supply voltages) the Franklin oscillator can give a frequency stability approaching 1 part in 1 million (0.0001 %).

Variable Frequency Oscillators

32. Oscillators that are continuously variable in frequency over a given range, so that any required frequency within that range may be selected, are termed 'variable frequency oscillators' (v.f.o.). The frequency range over which they are capable of being tuned depends on the ratio of maximum capacitance to minimum capacitance of the variable capacitor in the tuned circuit. A typical variable air dielectric capacitor used in communication receivers has a variation in capacitance from 50 pf to 500 pf, i.e., the ratio $\frac{C_{max}}{C_{min}} = 10$. Since the frequency of oscillation is approximately equal to $\frac{1}{2\pi\sqrt{LC}}$, the frequency is proportional to $\frac{1}{\sqrt{C}}$ and for the variable capacitor quoted, the ratio of maximum frequency to minimum frequency is about 3. A typical frequency range would be 500 kc/s to 1,500 kc/s. If the oscillator stage is followed by frequency multipliers, this range will be extended. For instance, if an oscillator capable of being tuned over the range 500 kc/s to 1,500 kc/s is followed by a frequency doubler, the output covers the range 1,000 kc/s to 3,000 kc/s.

Oscillators for V.H.F.

33. In v.h.f systems, most master oscillators operate at a much lower frequency than that finally required, the latter being obtained by the use of frequency-multiplier stages as necessary. One such system is illustrated in Fig. 20 where V_1 acts as an oscillator-trebler, its output being applied to V_2 which acts as a frequency trebler; the overall multiplication is therefore, *nine* times the oscillator fundamental frequency f_o, so that for an output of 90 Mc/s, the oscillator need only operate at 10 Mc/s. V_1 cathode, grid and screen are arranged in an electron-coupled Colpitts oscillator circuit, with the screen at earth potential to r.f. because of C_5. Since the cathode is at a high r.f. potential, L_2 is inserted to ensure a d.c. return to the cathode while preventing a low impedance path to r.f. Automatic grid leak bias is provided by $C_g - R_g$, and R_1 is inserted to act as a cathode safety bias component. The anode load of V_1 is tuned to the third harmonic of the oscillator fundamental frequency f_o, giving an output at $3f_o$ to V_2. This latter stage operates as a conventional frequency trebler to give an output at $9f_o$.

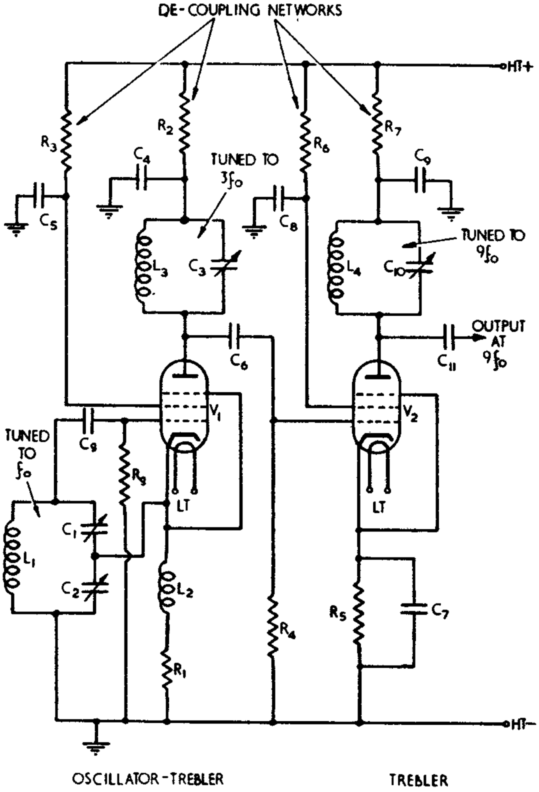
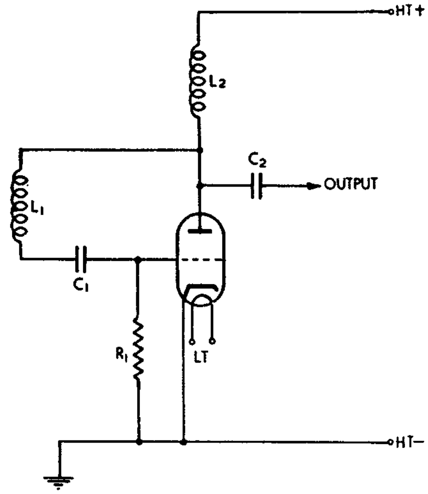
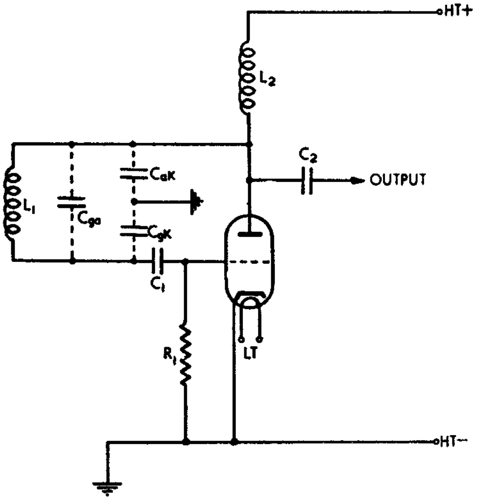


Fig. 20. OSCILLATOR-MULTIPLIERS FOR V.H.F.

34. In some circumstances, the oscillator itself is required to operate at very high frequencies. The operation at such fre-



(a) ACTUAL CIRCUIT



(b) EQUIVALENT CIRCUIT

Fig. 21. V.H.F. OSCILLATOR

quencies is complicated by transit-time effects, by valve lead inductances and by valve inter-electrode capacitances (see Sect. 8, Chap. 3, Para. 35). As a result, oscillators for the higher frequencies employ valves specially designed for high frequency operation and make use of special circuit arrangements. A typical oscillator circuit for use at v.h.f. is shown in Fig. 21(a). When the inter-electrode capacitances are taken into account, the equivalent circuit will have the form shown in Fig. 21(b). This is a Colpitts circuit in which the magnitude of the feedback voltage is determined by the ratio of C_{ak} to C_{gk}.

35. The normal inductor and capacitor resonant circuit becomes very small in size and difficult to design when it is to tune frequencies above about 200 Mc/s. Under such circumstances '*lecher bars*' may be used instead. Lecher bars (see Bk 3, Sect. 15) are specified lengths of a transmission line that exhibit properties similar to those of a tuned circuit, with certain added advantages. Radar oscillators employing lecher bars are considered in Part 3, and lecher bar systems for wireless transmitters are dealt with in Part 2.

Transistor Oscillator Circuits

36. Provided suitable adjustment is made to the power supply and bias arrangements,

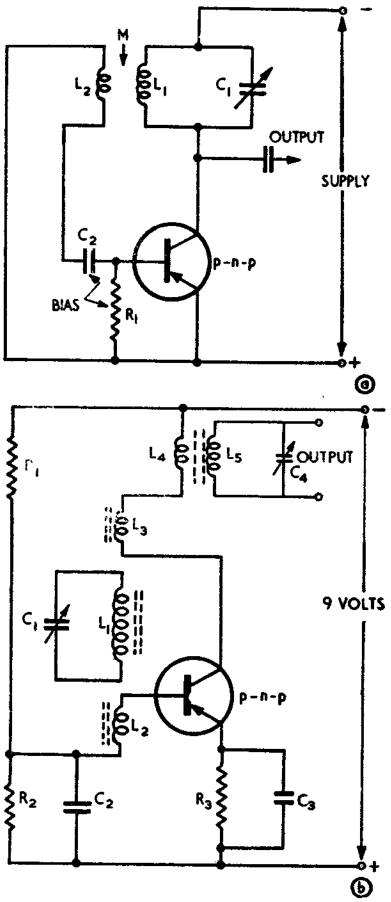


Fig. 22. TRANSISTOR OSCILLATOR CIRCUITS

transistors can be used in place of valves in any of the feedback oscillators considered in the previous paragraphs. Because of the limitations on the power handling capacity and in the frequency response of transistors, their use as oscillators is confined to low power, low frequency circuits. Fig. 22(a) shows the basic circuit of a simple type of transistor feedback oscillator. The circuit behaves in a similar way to a tuned anode valve oscillator, in which cathode, grid and anode are analogous to emitter, base and collector respectively. A more practical circuit is shown in Fig. 22(b), which represents a transistor Meissner type oscillator. Energy is fed back through the transformer from the collector (L_3) to the base (L_2) in such a way as to maintain oscillation at a frequency determined by L_1C_1. The oscillatory current component in L_4 produces the required output voltage across L_5C_4. To ensure that the oscillator starts easily, the transistor is biased initially for Class A operation with the normal stabilization circuit R_1R_2; C_2 is a r.f. by-pass capacitor. As the amplitude of oscillation increases, the d.c. voltage developed at the emitter by R_3C_3 biases the transistor into Class B conditions. With a power supply of 9 volts, a circuit of this type operating at 60 kc/s gives a power output of 20 mW.

NEGATIVE RESISTANCE OSCILLATORS

Principle

37. When free oscillations occur in a tuned circuit, the oscillations are damped because of the resistance losses associated with the circuit (see Para. 5). The resistance causing the damping is a real, 'positive' resistance, and power is dissipated in the circuit. To maintain oscillations in the negative resistance type of oscillator, the solution is to introduce a circuit element that has a '*negative*' resistance; that is, a device in which the current is decreased by increasing the voltage applied to it. The negative resistance component has such a value that it completely counteracts the effect of the positive damping resistance and oscillation is maintained. Fig. 23 shows a negative resistance $-r_a$ connected in parallel with a parallel resonant circuit; at resonance, the impedance of the tuned circuit is the dynamic resistance $R_d = \frac{L}{CR}$. For oscillations to be maintained, the total resistance across AB must be zero or have a *negative* value, since no power is to be dissipated. At resonance, the total resistance across AB is $\frac{-r_a\ R_d}{R_d - r_a}$ and the necessary conditions for the maintenance of oscillations is that r_a must be *equal to*, or *less than* R_d.

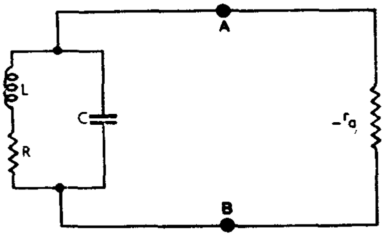


Fig. 23. PRINCIPLE OF NEGATIVE RESISTANCE OSCILLATORS

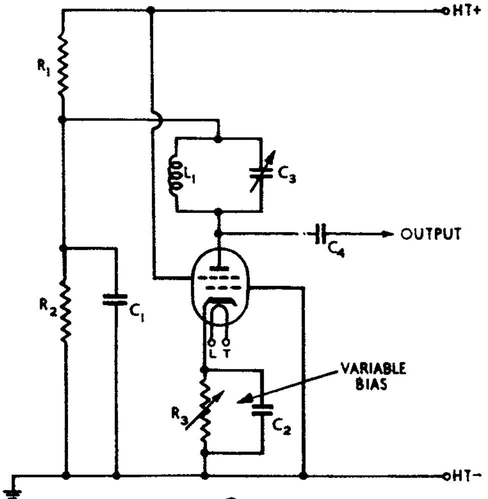
Dynatron Oscillator

38. A screen-grid tetrode valve connected as shown in Fig. 24(a), constitutes one type of negative resistance oscillator known as a '*dynatron*'. The anode characteristic of the tetrode is illustrated in Fig. 24(*b*), and over the portion of the curve AB an increase in anode voltage produces a decrease in anode current. Thus the anode slope resistance r_a is *negative* in this region.

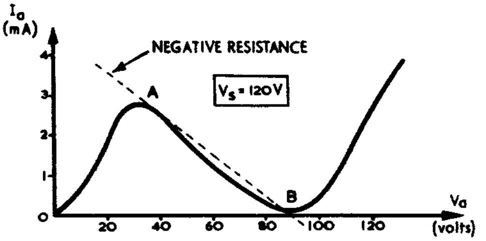
The 'kink' in the anode characteristic is due to the effects of secondary emission and occurs when the anode is at a potential lower than that of the screen (see Sect. 8, Chap. 3, Para. 8). Consequently, when a tetrode valve is operated in this manner in conjunction with a tuned circuit, oscillations are maintained when r_a is equal to or less than the dynamic resistance $\frac{L}{CR}$, the value of negative resistance being adjusted by varying the bias on the control grid. The frequency of oscillation is determined by the tuned circuit and is given approximately by $f_o = \frac{1}{2\pi\sqrt{LC}}$. A practical upper limit of frequency for a dynatron oscillator is 20 Mc/s.

Transitron Oscillator

39. A pentode valve may be used as a negative resistance device. The suppressor grid of a pentode has virtually no control on the total space current in the valve, but it



ⓐ DYNATRON CIRCUIT



ⓑ TETRODE CHARACTERISTIC

Fig. 24. DYNATRON OSCILLATOR

does affect the *division* of current between the anode and the screen. Thus, if the suppressor is driven positive the space current remains constant, but the anode current increases and the screen current drops. This fact may be used to give a negative resistance in the circuit shown in Fig. 25. If an alternating voltage is applied to the screen, the screen potential will rise on the positive half cycle; so also will that of the suppressor (since the suppressor is connected to the screen by a coupling capacitor C_2) and this causes the screen current to fall, because the suppressor potential has a greater effect than the screen potential on the screen current. Thus, a *rise* in screen potential corresponds to a *fall* in screen current; and a fall in screen potential produces a rise in screen current. There is, therefore, a *negative* a.c. resistance between screen grid and earth, and if a parallel tuned circuit is connected in the screen lead, oscillation will be maintained at the frequency of the resonant circuit, provided that the screen negative resistance r_s is equal to or less than the dynamic resistance $\frac{L}{CR}$. This type of oscillator is known as a '*transitron*'.

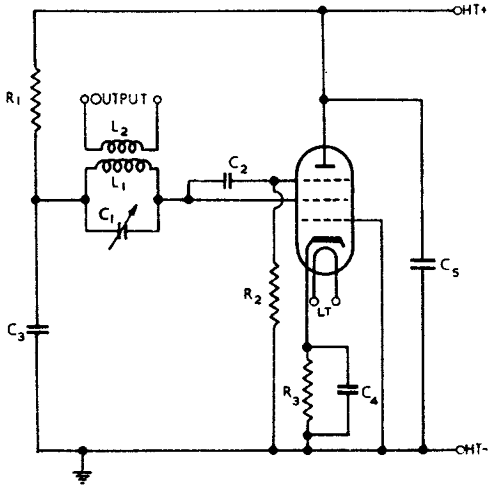


Fig. 25. TRANSITRON OSCILLATOR

Tuned Anode—Tuned Grid Oscillator

40. If a parallel tuned circuit is connected between grid and cathode of a triode valve, oscillations will occur at the frequency of the resonant circuit, provided that the resistance component of the input impedance of the valve is *negative* and has a value equal to or less than the dynamic resistance $\frac{L}{CR}$ of the tuned circuit. It is shown in Sect. 11, Chap. 1, Para. 6 that this condition will be obtained

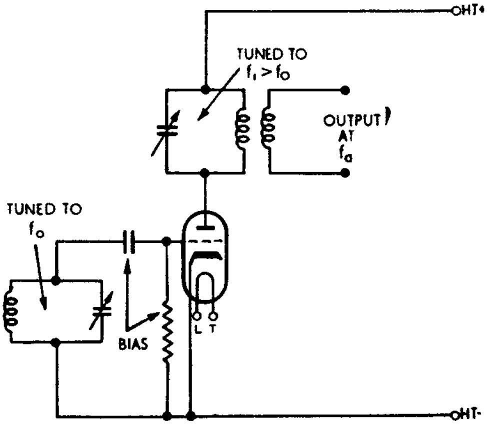


Fig. 26. TUNED ANODE-TUNED GRID OSCILLATOR

in a triode valve, by reason of Miller effect, if the anode load is *inductive*. This fact can be utilized to give a negative resistance oscillator known as the tuned anode-tuned grid (t.a. — t.g.) type. In the t.a. — t.g. oscillator (Fig. 26), the frequency of oscillation f_o is mainly determined by the *grid* tuned circuit. The *anode* circuit is tuned to a frequency f_1 slightly *higher* than f_o, so that at the oscillation frequency it presents an inductive load. Thus, the tuned grid circuit has a negative resistance across it, of a value determined by the 'mistuning' of the anode load, and oscillation at the frequency f_o takes place. Note that there is no mutual coupling between the tuned circuits.

SECTION 12

CHAPTER 2

CRYSTAL-CONTROLLED OSCILLATORS

CRYSTAL-CONTROLLED OSCILLATORS

Introduction

1. Certain crystalline substances, of which quartz is the best material for most purposes connected with radio, exhibit a phenomenon known as the '*piezo-electric*' (pressure-electric) effect; that is, if a section cut from a quartz crystal is subjected to mechanical tension or compression, an electric charge is produced across the crystal. Conversely, if a voltage is applied across the faces of a crystal section, a mechanical stress is produced in the crystal. This effect can be used to produce and maintain oscillations in a valve circuit, since the application of an exciting voltage causes the crystal to vibrate, the vibrating crystal producing its own voltage which, in turn, increases the intensity of the vibrations. In this way, oscillations at the mechanical resonant frequency of the crystal are built up, very little energy being required from the valve circuit to maintain the oscillations.

2. The ability of a quartz crystal to oscillate in this manner means that it can be used in place of a tuned circuit in an oscillator. In relation to a tuned circuit, a quartz crystal:—

(*a*) has a much higher Q value and therefore, a greater degree of frequency stability (see Para. 10):

(*b*) can oscillate satisfactorily at only *one* frequency in any given circuit; the frequency depends on the physical dimensions of the crystal, a thin plate oscillating at a frequency higher than that of a thick plate.

Because of these factors, quartz crystal oscillators are used in radio transmitters and receivers where an accurate, pre-determined 'spot' frequency is required, e.g., in ground transmitters for assigned point-to-point communication, and in aircraft transmitter-receiver systems for air-to ground communication. They cannot be used in circuits where selection of any frequency within a given range is required, a variable frequency oscillator (v.f.o.) of the LC type being normal in this case.

Crystal Construction

3. **Axes.** To obtain a quartz plate with piezo-electric properties, special 'cuts' of the crystal are made. In its natural state, a quartz crystal has a hexagonal cross-section with pointed ends as shown in Fig. 1(a).

The properties of such a crystal can be expressed in terms of three sets of axes known as the Z, X and Y axes, which are all at right angles to each other:—

(*a*) The Z (or optical) axis joins the points at the ends of the crystal (Fig. 1(a)).

(*b*) The three axes X_1, X_2, X_3 (the electrical axes) are at right angles to the Z axis and pass through the corners of the hexagon (Fig. 1(*b*)).

(*c*) The three axes Y_1, Y_2, Y_3 (the mechanical axes) are at right angles to the Z axis and are perpendicular to the main faces of the crystal (Fig. 1(c)).

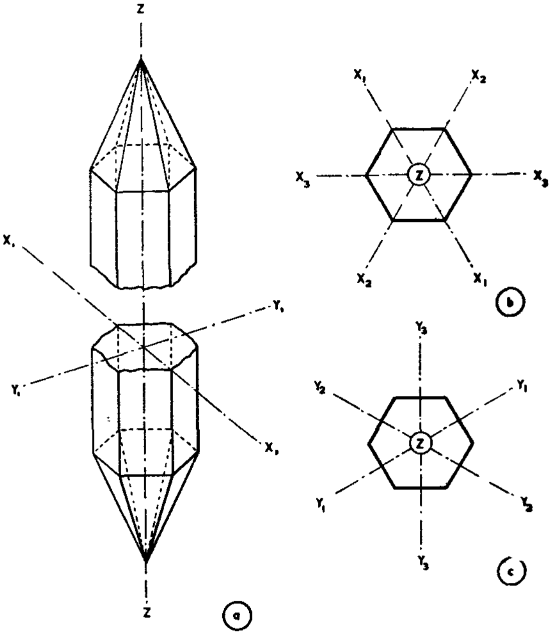
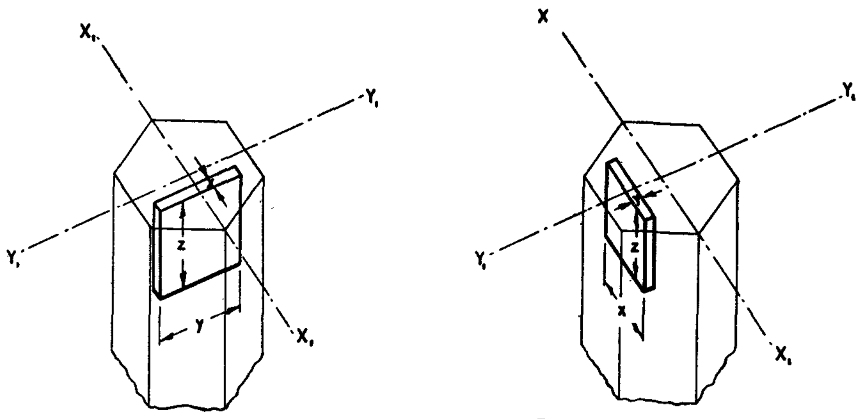


Fig. 1. QUARTZ CRYSTAL AXES

4. **X- and Y-cut crystals.** Plates cut from a crystal are normally referred to the axes. An X-cut crystal is a plate that has been cut with its flat sides perpendicular to an X axis (Fig. 2(a)), and a plate so produced exhibits strong piezo-electric effects across its faces.



(a) X-CUT CRYSTAL

(b) Y-CUT CRYSTAL

Fig. 2. X- and Y-CUT QUARTZ CRYSTALS

A Y-cut crystal is cut with its flat sides perpendicular to a Y axis (Fig. 2(b)), and a plate so produced exhibits strong piezoelectric effects. Since however, the resonant frequency of X- and Y-cut crystals varies considerably with temperature, their use is limited to circuits where *extreme degrees* of frequency stability are not required.

5. **Inclined-angle cut crystals.** To obtain crystals whose temperature coefficients are small, other cuts at certain specified angles to the Z, X and Y axes are introduced. Typical of these is the *AT-cut* crystal whose frequency is hardly affected at all by temperature changes over a wide range. An AT-cut crystal is cut in the same plane as a Y-cut crystal but is rotated away from the Z axis by an angle of 35°, as shown in Fig. 3. AT-cut crystals, and other inclined-angle cut types, are used extensively in high-stability, crystal-controlled oscillators over a frequency range of 1 kc/s to 20 Mc/s. By using harmonic vibrations of the crystal ('overtones') this range can be extended to 100 Mc/s. The factors determining the resonant frequency of a quartz plate are:—

(*a*) Type of cut;

(*b*) dimensions of plate;

(*c*) temperature at which the crystal operates;

(*d*) method of mounting the crystal.

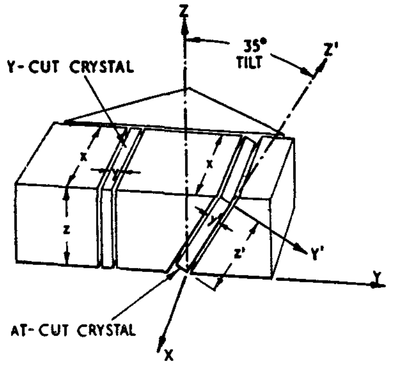


Fig. 3. AT-CUT QUARTZ CRYSTAL

Crystal Holders

6. One form of crystal holder in common use at the lower frequencies is shown in Fig. 4(a). The quartz plate is clamped lightly between two electrodes by means of contact springs, the electrodes being so shaped that the crystal can vibrate in a small air gap. The whole is enclosed in a container. At the higher frequencies, the dimensions of the quartz plate are much reduced and in this case the crystal electrodes are commonly thin metal films formed directly on the surface of the crystal by spraying. The crystal is supported by flexible wires soldered to the film electrodes as shown in Fig. 4(b) and the wires are taken through a glass pinch to two pins in a valve type base. The mounted crystal is enclosed in a sealed glass envelope to protect the crystal from the effects of moisture, and in some cases the envelope is evacuated to eliminate the damping effect of the air. The crystal frequency is normally stencilled on the container. In circuits where improved frequency stability is required it is

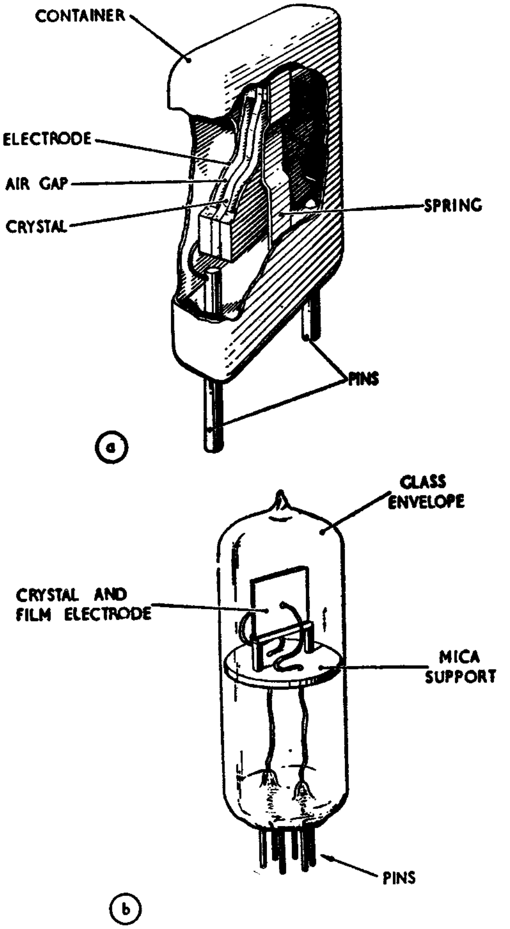


Fig. 4. QUARTZ CRYSTAL HOLDERS

normal to insert the crystal holder in a temperature-controlled oven for the purpose of maintaining the crystal at a specified temperature.

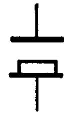
Crystal Equivalent Circuit

7. When a quartz crystal, the symbol for which is shown in Fig. 5(a), is inserted in a circuit, it is found to behave like the equivalent electrical circuit of Fig. 5(b). In this circuit, L, C_x and R are the electrical equivalents of corresponding mechanical properties of the crystal, and their values are determined by the dimensions and cut of the crystal. C_H is the capacitance produced by the electrodes and holder used in mounting the crystal. Measurements of the effective values of L, C_x, C_H and R for quartz crystals have been made, and for a specimen AT-cut crystal of 3 Mc/s fundamental frequency:—

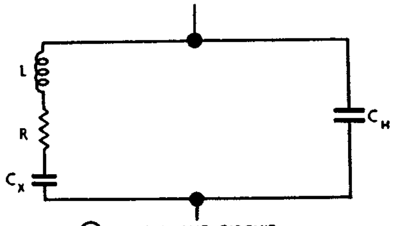
$$C_x=0.02\text{pF} \ : \ C_H=8\text{pF} \ ; \ R=30\Omega; \ L=0.127 \text{ H}.$$

$$\therefore Q=\frac{\omega L}{R} = 80{,}000.$$

The small value for the equivalent resistance R means that the vibration of the crystal, once started, can be maintained with very little external energy; the high $\frac{L}{C}$ ratio gives a frequency response curve that has a sharp peak; and the high Q value gives a good frequency stability. In comparison with the normal form of electrical resonant circuit, the Q value is extremely high and may lie between 50,000 and 500,000 compared with values of the order of 100 to 500 for high grade inductor and capacitor circuits.



(a) SYMBOL



(b) EQUIVALENT CIRCUIT

Fig. 5. ELECTRICAL PROPERTIES OF CRYSTAL

Crystal Resonant Frequencies

8. **Series resonance.** L, C_x and R form a series circuit of high Q, resonating at a frequency fixed by their respective values. The equivalent circuit is shown in Fig. 6(a) and the impedance-frequency response curve is shown in Fig. 6(b). At resonance, the impedance is equal to R and the high Q gives a very narrow bandwith. Crystals operating in the series mode are used extensively in certain types of filter circuits (see Bk 3 Sect. 15)

9. **Parallel resonance.** L and R, shunted by the *series* combination of C_H and C_x (Fig. 7(a)), form a parallel circuit resonating at a frequency slightly *higher* than that for the series mode. As C_x is only a fraction of a pF and C_H is some hundred times larger, quite large changes in C_H can be made without affecting appreciably the resultant value of

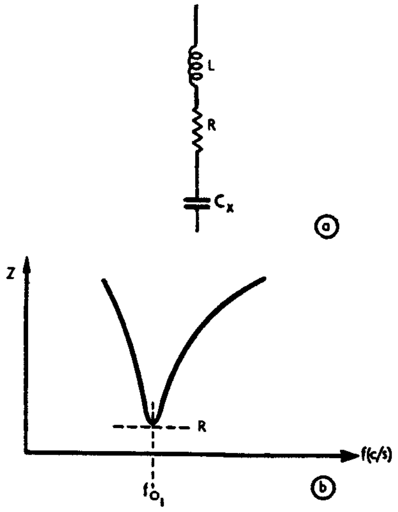


Fig. 6. SERIES MODE FOR CRYSTAL

C_H and C_x in series. Consequently, variations in C_H and in any shunt capacitance connected across the crystal have only a small effect on the resonant frequency of the crystal. Crystals operating in the parallel mode are used extensively in oscillator circuits, where the crystal behaves as a parallel tuned circuit. The impedance-frequency response curve for the parallel mode is shown in Fig. 7(b) and the effect of the high Q value in producing a narrow bandwidth can be seen. The separation between the series resonant frequency fo_1 and the parallel resonant frequency fo_2 is very small, being of the order of 0.5 per cent of fo_1. This is shown in Fig. 8 which illustrates the impedance-frequency response curve for a crystal over a range of frequencies.

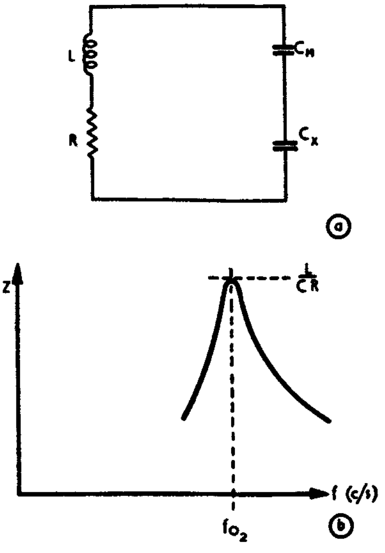


Fig. 7. PARALLEL MODE FOR CRYSTAL

Crystal-controlled Oscillators

10. Crystals are used in place of parallel LC circuits to control the frequency of oscillators. It is because of their exceptionally high Q values that quartz crystals exert such a stabilizing effect on the frequency of oscillating circuits. Not only do they hold the frequency constant despite variations in the associated circuit values, but as the resonant frequency is governed by the physical dimensions of the crystal, they can be ground for a particular frequency and will hold this frequency almost indefinitely. This is an advantage where a 'spot' frequency is required. Quartz crystals are affected to some extent by changes in temperature, but for modern crystals of the AT-cut or similar cuts, the variation of frequency with temperature is very small; this can be reduced still further if the crystal is mounted in a temperature-controlled oven. When all possible precautions have been taken in crystal-controlled oscillators, the frequency stability approaches 1 part in 10^8. One B.B.C. crystal-controlled transmitter has the very high stability of ± 1 part in 10^9. This compares with a stability of about 1 part in 10^6 for the best LC oscillators.

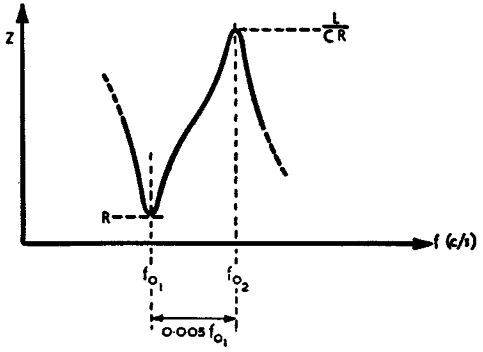


Fig. 8. SEPARATION BETWEEN SERIES AND PARALLEL RESONANT FREQUENCIES

Tuned Anode—Crystal Grid Oscillator

11. The basic circuit for a tuned anode-crystal grid (t.a-x.g.) oscillator is shown in Fig. 9. The crystal acts as a parallel tuned circuit, so that the circuit of Fig. 9 is similar in operation to the t.a.-t.g. oscillator discussed in Chap. 1, Para. 38. Oscillations occur

when the anode resonant circuit is tuned to a frequency slightly *higher* than the parallel resonant frequency of the crystal. The anode load is then *inductive* and, by virtue of Miller effect, the required *negative resistance* is introduced into the grid circuit to counteract the crystal damping losses. Automatic grid leak bias is used, the bias voltage being developed by grid current through the grid leak R_1; cathode safety bias from R_2 is also included to prevent excessive anode current should the crystal drive cease. In some cases, a variable capacitor is connected across the crystal to give a fine adjustment of frequency. The range of variation obtainable in this way is very small, being about 1 part in 10^4. The meter in the cathode circuit is for tuning purposes.

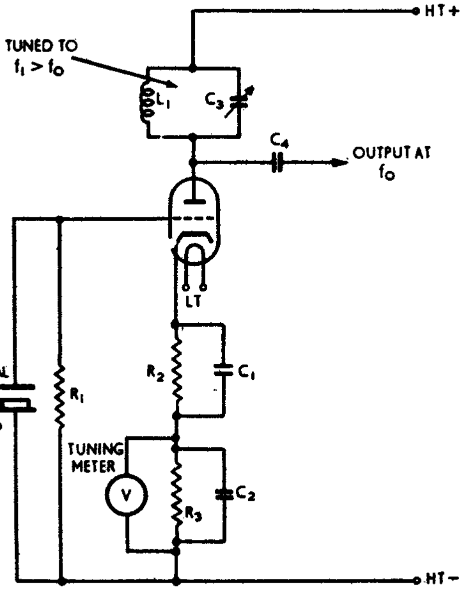


Fig. 9. TUNED ANODE—CRYSTAL GRID OSCILLATOR

12. As the anode circuit resonant frequency is reduced from a high value oscillations build up. The bias, therefore increases and the d.c. valve current falls. The d.c. voltage developed across R_3C_2 by this current also falls and a 'dip' occurs in the voltmeter reading. Further reduction in the frequency of the anode circuit produces a sudden fall in r.f. output and a sudden rise in the value of the d.c. anode current until oscillations cease just before the anode circuit comes into tune with the crystal frequency. It is necessary, therefore, that the resonant frequency of the anode circuit be *higher* than that of the crystal. The setting of the anode circuit variable capacitor should be such that the resonant frequency of the anode load corresponds to point A on the graph of Fig. 10, i.e., the circuit is tuned to the 'slow' side of the dip obtained in the tuning meter.

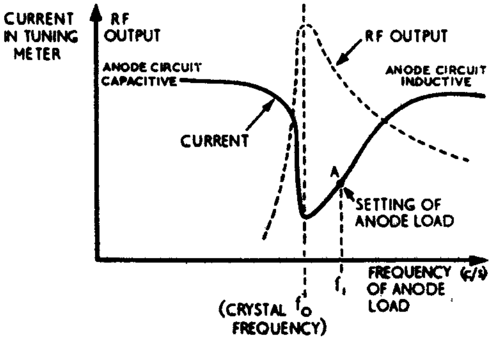
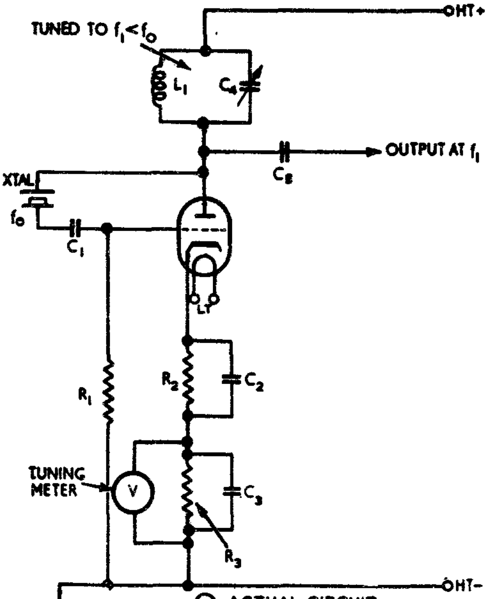


Fig. 10. TUNING OF TUNED ANODE—CRYSTAL GRID OSCILLATOR

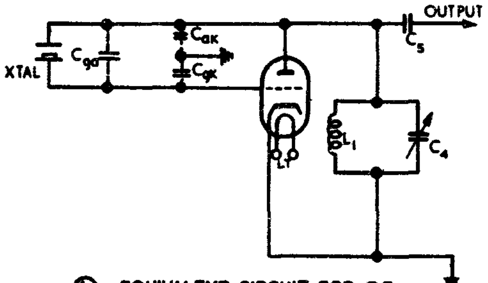
Pierce Oscillator

13. Fig. 11(a) illustrates the basic circuit of a Pierce oscillator. If the valve inter-electrode capacitances are considered, the circuit becomes that shown in Fig. 11(b), which is a form of Colpitts oscillator where the phase and the magnitude of the feedback voltage necessary to maintain oscillations is determined by C_{ak} and C_{gk}. The crystal operates at a frequency just sufficiently *below* its parallel resonant frequency to give an inductive reactance that resonates with the valve capacitances C_{ak} and C_{gk}. The circuit operates under automatic grid leak bias developed by C_1R_1, and cathode safety bias components R_2C_2 are inserted. The meter in the cathode circuit is for tuning purposes.

14. As the anode circuit resonant frequency is increased from a low value, oscillations build up and the d.c. anode current falls (because of the increase in bias) to produce a dip in the tuning meter. Further increase in the frequency of the anode circuit produces a sudden fall in r.f. output and a sudden rise in the value of the d.c. anode current until oscillations cease just before the anode circuit comes into tune with the crystal frequency. It is necessary therefore, that the resonant frequency of the anode circuit should be below that of the crystal, as illustrated in the graph of Fig. 12. This is the reverse of the t.a.-x.g. oscillator.



ⓐ ACTUAL CIRCUIT



ⓑ EQUIVALENT CIRCUIT FOR R.F

Fig. 11. PIERCE OSCILLATOR

Crystal-controlled Oscillator and Trebler

15. Fig. 13 shows the circuit of the crystal-controlled, electron-coupled Colpitts oscillator and frequency trebler used in an aircraft v.h.f. communication transmitter-receiver. The cathode, control grid and screen are the oscillator electrodes; the screen is at r.f. earth potential through C_4 and the oscillator circuit is said to be 'inverted'. L_1 is inserted to ensure a d.c. return to the cathode while maintaining the cathode at a high r.f. potential. C_1R_1 provide self bias from the crystal drive and R_2 gives cathode safety bias. Oscillations at the parallel resonant frequency of the crystal f_0 are maintained, the correct phase and magnitude of the feedback voltage being ensured by the Colpitts arrangement of C_2 C_3. The anode load L_2 C_5 is tuned to the third harmonic of the crystal frequency so that an output at $3f_0$ is obtained. The anode load is tuned for maximum drive to a following stage, this condition being obtained with the aid of a tuning meter in the next stage. In this equipment, any crystal within the frequency range 5.5 Mc/s to 8.7 Mc/s may be used and since the oscillator-trebler stage is followed by further frequency multiplier stages to give an output at 18 times the crystal frequency, the frequency at the output is in the range 100 Mc/s to 156 Mc/s. The order of frequency stability with this arrangement is 1 part in 10^5.

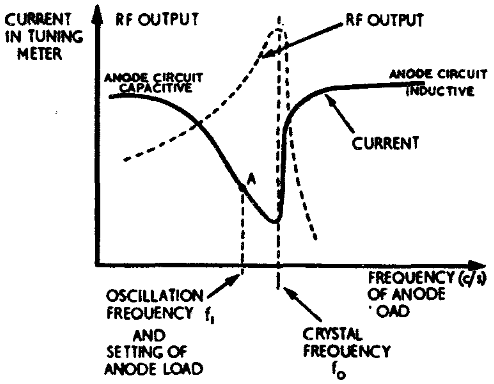


Fig. 12. TUNING OF PIERCE OSCILLATOR

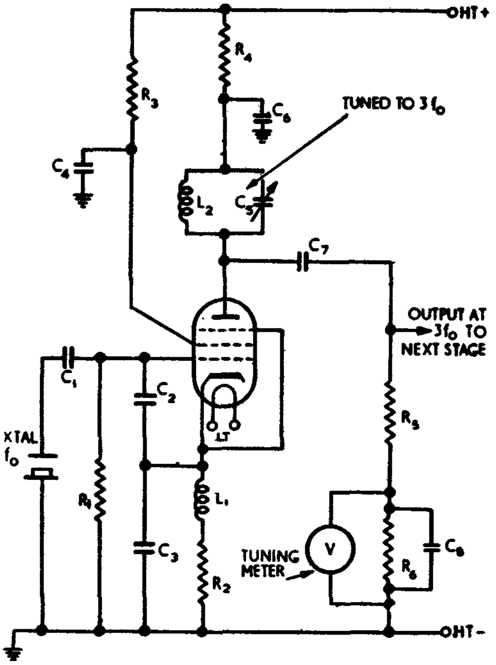


Fig. 13. CRYSTAL-CONTROLLED OSCILLATOR-TREBLER

SECTION 12

CHAPTER 3

PHASE-SHIFT (RC) AND RELAXATION OSCILLATORS

PHASE-SHIFT (RC) AND RELAXATION OSCILLATORS

Introduction

1. In the oscillators so far discussed, tuned circuits have been used as a means of controlling the frequency. However, oscillations can be maintained without the use of a tuned circuit. For oscillations to be maintained in a single valve circuit it is only necessary that the feedback network be capable of shifting the phase of the valve's output by 180° without attenuating it more than the valve amplifies. Almost any combination of reactances and resistances can be used for the feedback network, but because of the amount of inductance required at low frequencies, it is more usual to find resistance-capacitance (RC) combinations. The *phase-shift* oscillator therefore differs radically from those previously described in that it employs an RC, rather than an LC network, for both positive feedback and frequency control. It is useful for producing an output of good sinusoidal waveform at frequencies less than about 80 kc/s.

Single Stage RC Oscillator

2. A simple combination of one resistance and one capacitance can be used to cause a phase-shift of anything between (but not including) 0° and 90°, depending on the values of R and C and on the frequency of the applied voltage. An alternating voltage V applied to the circuit of Fig. 1(a) establishes an alternating current I which produces voltages V_C and V_R across C and across R respectively. If R and X_C (the reactance of C) are equal, then V_R is 45° leading on V, and V_C is 45° lagging on V; both V_R and V_C are 0.707 as large as V (Fig. 1(b)). If R is made smaller relative to X_C, by reducing R or C or the frequency, the *phase lead* of V_R is *increased*, but the *magnitude* of V_R is *reduced* (Fig. 1(c)). The lead can be increased to 90° only by making V_R zero. Thus, any increase in phase advance is made only at the expense of a reduction in amplitude. Similar remarks apply to V_C. Note that if the output is taken across R, the output voltage leads the applied voltage and the network is *phase-advancing*; if the output is taken across C, the network is *phase-retarding*.

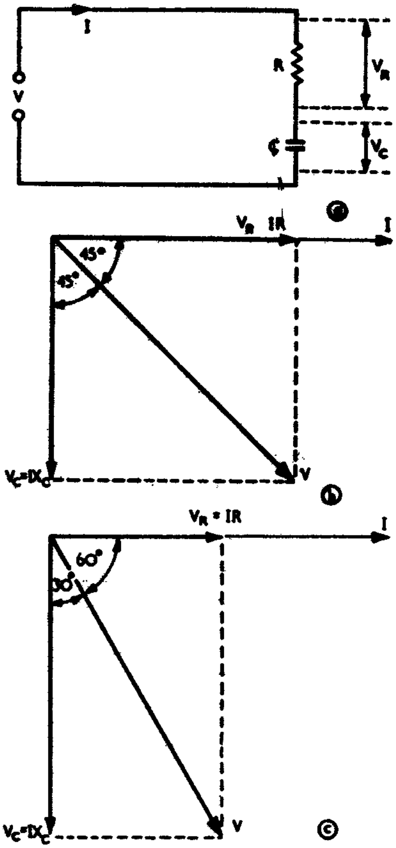


Fig. 1. PHASE SHIFT, C AND R IN SERIES

3. If two RC networks are connected as shown in Fig. 2(a), the output across R_2 will lead V_{R1}, and hence lead the applied voltage V by a greater angle at a further sacrifice in voltage (Fig. 2(b)). Obviously even this cannot give a total phase shift of 180°, but the necessary phase shift can be obtained by using *three* RC meshes (Fig. 2(c)) or *four* RC meshes (Fig. 2(d)). Each mesh in a network does not produce *exactly* the same phase shift even with identical components,

because the impedance into which each mesh works is different. However, the phase shift across each mesh in a three-mesh network approximates to 60°, and that in a four-mesh network to 45°, the overall phase shift for each network being 180°.

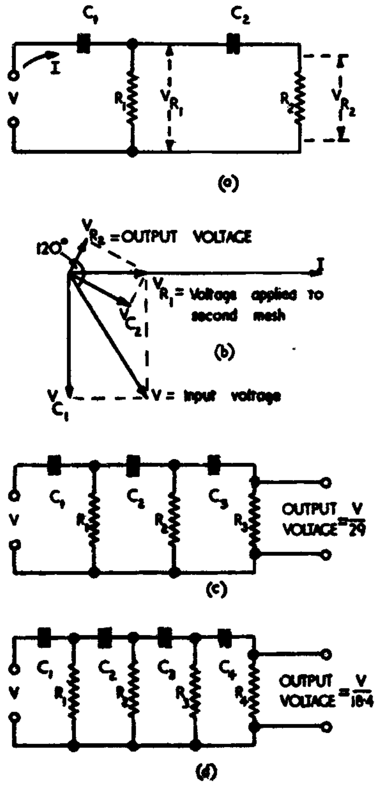


Fig. 2. RESISTANCE-CAPACITANCE PHASE-SHIFTING NETWORKS

4. The voltage available across the last resistor (or capacitor) in a RC network is smaller than the voltage applied into the network; that is, *attenuation* takes place. The amount of attenuation depends on:—

(*a*) the phase shift across each individual RC mesh;

(*b*) the total number of meshes.

In a three-mesh RC network, where each mesh gives approximately 60° phase shift, the overall attenuation is 29; that is, the output voltage is $\frac{1}{29}$th of the applied voltage. Thus, to maintain oscillations in a single valve circuit, the voltage amplification of the valve must be at least 29 to counteract the attenuation of the network. In a four-mesh RC network, where each mesh gives approximately 45° phase shift, the figure for the network attenuation and the corresponding valve amplification is 18.4.

5. A typical three-mesh phase-shift oscillator is shown in Fig. 3. With given values of C and R, there is *one frequency only* at which the circuit oscillates; that is, the frequency automatically adjusts itself to that which causes the reactances to be such that a 180° phase-shift through the network is produced. By altering the value of R or C in any of the meshes, the frequency of oscillation is changed. With the circuit shown (assuming equal values for R and equal values for C) the frequency of oscillation is:—

$$f_0 = \frac{1}{2\pi CR\sqrt{6}}.$$

Thus, unlike the LC oscillator, the frequency is inversely proportional to R or C instead of to the *square root* of L or C. Hence, with RC oscillators a greater tuning range is possible. The RC oscillator is very stable and gives a good sinusoidal output waveform. It is used mainly in test equipment for low frequency operation in the range 100 c/s to 80 kc/s.

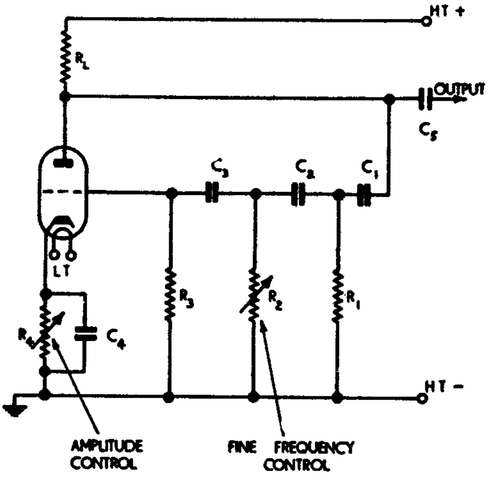


Fig. 3. BASIC THREE-MESH PHASE-SHIFT OSCILLATOR

Wien Bridge Oscillator

6. This is a two-valve RC oscillator. With the normal type of resistance-capacitance coupling between the two valves as shown in Fig. 4, the output from V_2 is in phase with the input to V_1. The feedback network is therefore relieved from the duty of creating

a phase reversal. All it has to do, to satisfy the conditions for the maintenance of oscillation at the desired frequency, is to transmit the necessary *amplitude* of feedback voltage with *no phase shift*. There are various ways of arranging resistors and capacitors to achieve this, but the simplest way is to use a feedback network of the type shown in Fig. 5. Assume R_1 equals R_2, and C_1 equals C_2. Then, at the frequency where the reactance of C_1 (or C_2) is equal to the resistance R_1 (or R_2), the voltage V_{IN} is *in phase* with V_0 and equal to *one-third* of V_0. This occurs at *one frequency only* for fixed values of C and R. Thus, if V_0 is the output of the two-stage amplifier of Fig. 4, V_{IN} will supply the necessary input voltage for oscillation, provided that the voltage amplification of the valves is *at least* 3 to counteract the attenuation of the network. A simple circuit of a Wien Bridge oscillator is shown in Fig. 6. The oscillator may be used to provide a variable frequency output, within the range 100 c/s to 80 kc/s, by using a two-gang capacitor for C_1 and C_2, with R_1 and R_2 usually equal and fixed. Assuming equal values for R and equal values for C, the frequency of oscillation is:—

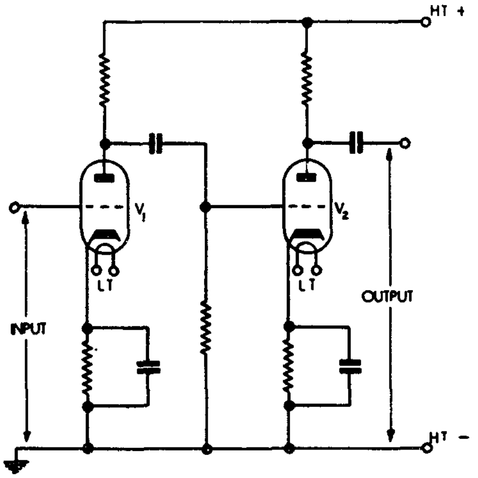


Fig. 4. BASIC CIRCUIT OF TWO-STAGE RESISTANCE-CAPACITANCE COUPLED AMPLIFIER

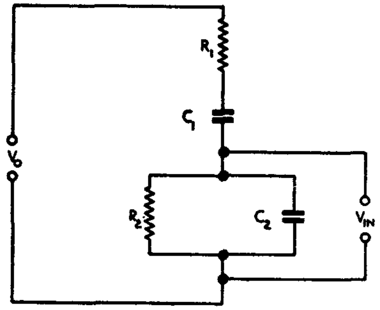


Fig. 5. NETWORK WITH ZERO PHASE-SHIFT AT ONE FREQUENCY

$$f_0 = \frac{1}{2\pi CR}.$$

The tuning range obtainable with the usual type of variable capacitor is thus of the ratio of 9 to 1 against that of 3 to 1 for LC oscillators. For a good sinusoidal output waveform over the tuning range, the voltage amplification of the two valves in cascade is kept down to about 3 by the application of *negative* feedback to V_1 (not shown in Fig. 6).

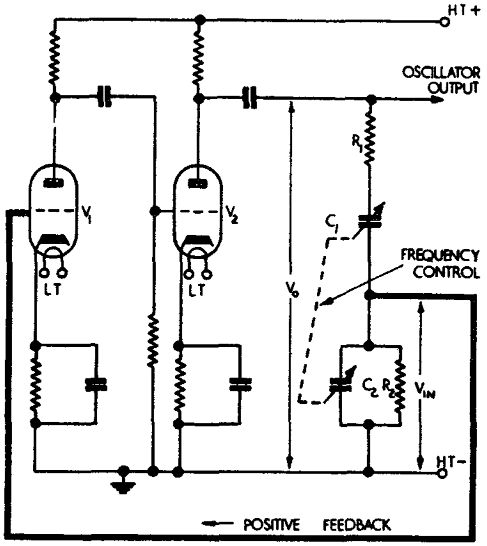


Fig. 6. SIMPLIFIED CIRCUIT OF WIEN BRIDGE OSCILLATOR

Relaxation Oscillators

7. A relaxation oscillator is one in which each cycle of oscillation consists of a period during which energy is stored in a reactive element (usually a capacitor) followed by a period of transition or relaxation during which the reactance discharges. One particular kind of relaxation oscillator is called

the '*multivibrator*' and is described below. Unlike the oscillators so far considered which give sinusoidal outputs, the multivibrator produces *rectangular* or *square pulses.*

8. A simplified circuit of a multivibrator is shown in Fig. 7. It consists of a two-stage resistance-capacitance coupled amplifier with the output from the anode of V_2 fed back through C_1R_1, to the input at the grid of V_1. As the total phase shift through such an amplifier is 360°, the circuit will oscillate at a frequency determined by the component values.

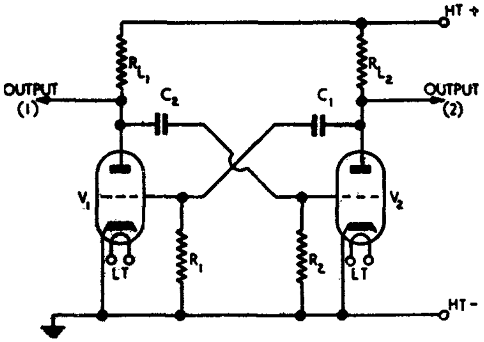
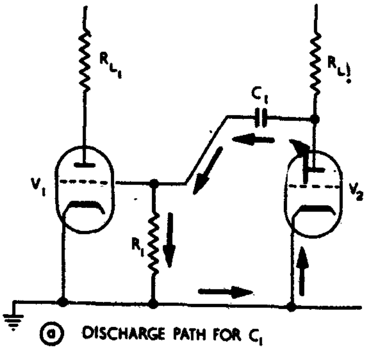


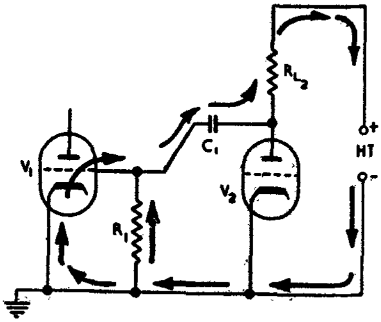
Fig. 7. BASIC MULTIVIBRATOR CIRCUIT

9. When h.t. is first applied to such a circuit, the valves do *not* begin to conduct at the same instant or to the same extent. Assume for example, that V_2 conducts first; then I_{a2} rises and V_{a2} falls, this fall in potential being applied to V_1 grid via C_1 to cause V_{g1} to fall. I_{a1} then falls rapidly, V_{a1} rises and this rise in potential is transferred to V_2 grid via C_2 to cause V_{g2} to rise. I_{a2} therefore rises still further and V_{a2} falls to a low value. This action is cumulative and results in V_1 being rapidly cut off while V_2 is conducting heavily. The condition then existing is that V_{a1} is at h.t. potential, V_{a2} is at a low value, V_{g2} is limited to zero volts by grid current and V_{g1} is below cut-off but rising exponentially as C_1 discharges through R_1 as shown in Fig. 8(*a*). With reference to Fig. 9, the action is continued thus:—

(*a*) *A to B.* V_{g1} reaches cut-on and I_{a1} flows. V_{a1} falls and since C_2 cannot change its charge instantly, V_{g2} falls by the same amount. I_{a2} then falls and V_{a2} rises causing V_{g1} to rise via C_1. I_{a1} rises further and the action is cumulative. The fall in V_{g2} quickly cuts off V_2 and V_{a2} rises to h.t. potential as C_1 charges through R_{L2} as shown in Fig. 8(b).



ⓐ DISCHARGE PATH FOR C_1
(C_2 SIMILAR)



ⓑ CHARGE PATH FOR C_1
(C_2 SIMILAR)

Fig. 8. CHARGE AND DISCHARGE PATHS FOR COUPLING CAPACITORS

(*b*) *B to C.* The circuit is stable with V_{g1} equal to zero volts, V_{a1} at a low value, V_{g2} below cut-off and V_{a2} at h.t. potential. During this period C_2 discharges through R_2 and V_{g2} rises exponentially towards cut-on.

(*c*) *C to D.* V_{g2} reaches cut-on and I_{a2} flows. V_{a2} falls causing V_{g1} to fall. I_{a1} falls and V_{a1} rises, causing V_{g2} to rise. I_{a2} rises further and the action is cumulative. V_1 is cut off and V_{a1} rises to h.t. potential as C_2 charges through R_{L1}. V_2 is conducting heavily with V_{a2} at a low value.

(*d*) *D to E.* The circuit is stable with V_{g2} equal to zero volts, and V_{g1} below cut-off but rising exponentially as C_1 discharges through R_1.

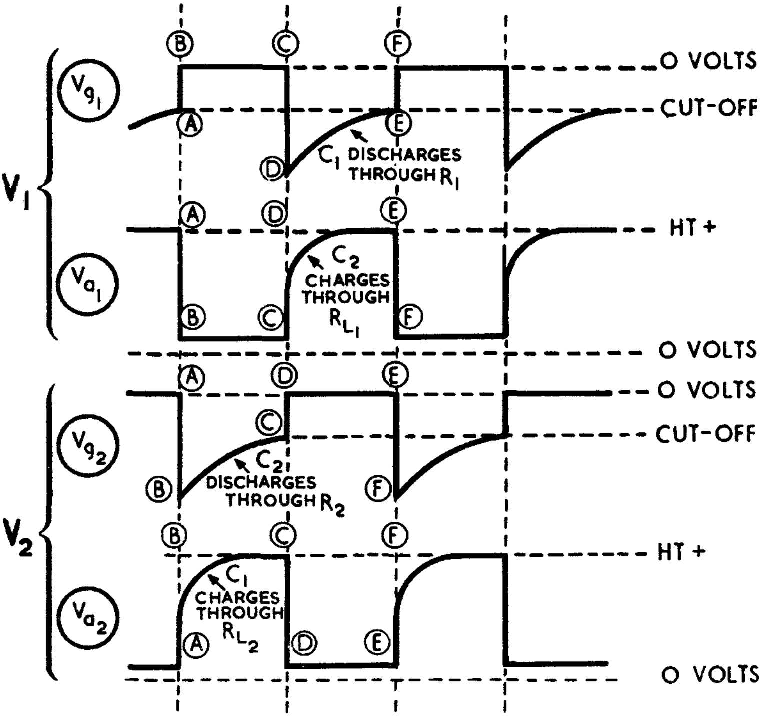


Fig. 9. BASIC MULTIVIBRATOR WAVEFORMS

(*e*) *E to F.* V_{g1} reaches cut-on, I_{a1} flows and the cycle repeats as from (*a*).

10. The square-wave output can be taken from the anode of V_1, or the anode of V_2 or both. The oscillation frequency is primarily determined by the time constants of the coupling components, C_1R_1 and C_2R_2 and may be altered by using variable components.

Multivibrators are used to generate frequencies from 1 c/s to approximately 1 Mc/s; they are especially useful as timing and controlling devices in oscilloscopes, television sets and radar equipment. Further details on this and other types of relaxation oscillators are to be found in Part 3 (Radar)·

INDEX

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