TRANSISTOR

AUDIO AND RADIO CIRCUITS

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The bulk of the information contained in this book is based on material supplied by W. D. Benson, D. R. Hyde and R. Osborne of the Mullard Central Application Laboratory. Other contributors include K. C. Warner of the Mullard Consumer Electronics Division and engineers of the Central Application Laboratory. The material has been edited by Miss A. Peters of Mullard Central Technical Services.

FOREWORD

'Transistor Audio and Radio Circuits' is intended as a manual of established and practical circuits for use by receiver and equipment manufacturers, radio and audio service engineers, university and college students, and do-it-yourself home constructors and hi-fi enthusiasts. A wide range of circuits is presented, extending from portable radio receivers to f.m. tuners and audio amplifiers capable of the highest standards of reception and reproduction.

The circuits presented can often be used in a variety of combinations to give a wide range of equipment. For example, various pairings are possible of the h.f. and a.f. stages designed for radiograms and portable radios. And again, in the hi-fi realm, either of the f.m. tuners can be used with either of the audio amplifiers and each resulting line-up can be used with or without a stereo reception facility, and can be combined with disc and tape input sources.

Details of the practical circuits are complete, but because of the wide range of possible combinations of circuits, no attempt has been made to present precise layout and constructional information. Some general guidance is given for hi-fi amplifiers, and a chapter is devoted to some simple test equipment and its use in effectively commissioning and maintaining audio circuits. But because of the need for rather complex equipment for aligning r.f. and i.f. circuits, similar information has not been given for such stages: it is envisaged that it will be those who have access to specialised knowledge and equipment—amateur enthusiasts, beginners under expert tutelage, or professionals—who attempt to build these h.f. circuits.

'Transistor Audio and Radio Circuits' is the latest of a number of Mullard books aimed at those whose interest lies in the audio and radio fields. It reflects the advances in electronic science since the earlier books were published, both in terms of devices and circuit techniques.

CONTENTS

CHAPTER 1

SILICON AND GERMANIUM TRANSISTORS

Most transistors are made from either silicon or germanium. The early transistors were made from germanium because silicon presented many manufacturing problems, but these problems have been largely overcome by the development of the planar process. Silicon and germanium transistors are now produced concurrently, and there are applications for which each is the more suitable.

Certain characteristics of germanium semiconductor material limit the operating junction temperature of germanium transistors to below 100°C, while silicon devices can operate at temperatures in excess of 150°C. However, germanium transistors still have the advantage of a low value of forward base-emitter voltage and bottoming voltage and, since these two represent voltage losses in the circuit, germanium devices are particularly suitable for the lower-voltage type of a.f. output stage. Another advantage of germanium transistors is that they are easily matched in complementary (p-n-p/n-p-n) pairs, thus facilitating their use in complementary-pair push-pull output stages.

Because silicon has an extremely high intrinsic resistivity the leakage currents of silicon transistors are far smaller than those of germanium transistors. This advantage, together with the higher permissible junction temperature mentioned above, and the high voltage ratings available with silicon transistors, makes them ideal for a wide variety of audio applications, especially at high output powers where high voltages are necessary. Silicon planar transistors can be manufactured with high switching speeds, making these devices suitable for use at high frequencies, for example in the r.f. stages of radio receivers.

TRANSISTOR MANUFACTURE

The manufacturing processes of germanium and silicon transistors both start with the production of a crystal of the doped semiconductor material. The crystal is then cut into thin slices and it is from this point on that the manufacturing processes differ.

Germanium Transistors

Germanium transistors can be made by either of two methods—the alloy-junction method or the alloy-diffusion method. In the manufacture of a p-n-p transistor by the alloy-junction method, an emitter pellet of p-type material is fused to one side of the n-type slice and a collector pellet of p-type material is fused to the other side. In the alloy-diffusion method, a p-n-p transistor is made by depositing a layer of n-type material on one surface of the p-type slice and placing two pellets close together on the n-type layer. One of the pellets, which eventually forms the base connection of the transistor, contains n-type additives only, and the other pellet, which will form the emitter, contains p-type and n-type additives. The assembly is heated and the additives diffuse into the p-type germanium beyond the prediffusion layer. In this way a transistor is made in which the collector is formed by the original slice, and all diffusion has been effected on one side. The processes used in the manufacture of germanium transistors are now well known and will not be elaborated upon in the present book. The manufacture of a germanium alloy-junction transistor is summarised in Fig. 1.

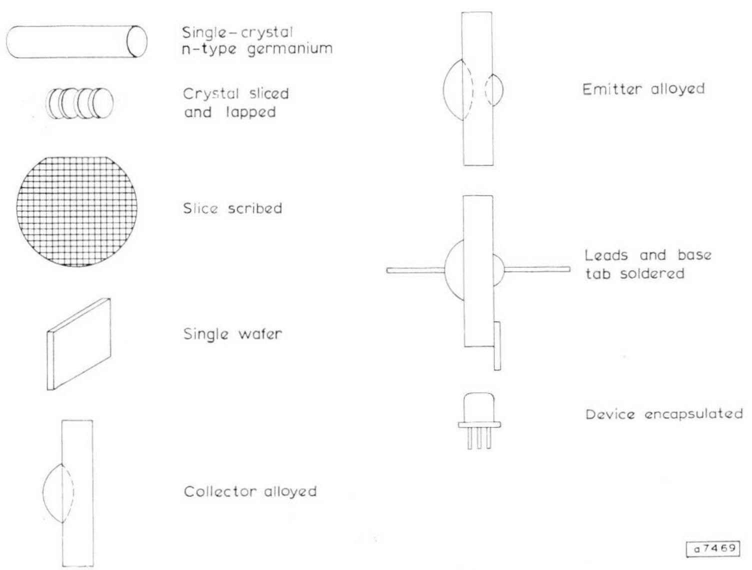


Fig. 1—Stages in the manufacture of germanium alloy-junction transistors

Silicon Transistors

A process now widely used in the manufacture of silicon transistors is the planar process illustrated in Fig. 2. A slice of n-type silicon is cleaned under very stringent conditions and is then heated in oxygen so that a layer of silicon oxide is formed all over it. The oxidised slice is attached to the middle of a high-speed spinner, and a drop of photosensitive etchant-resistant material, or photoresist, is placed on it. When the spinner revolves, the photoresist spreads out in a thin, even coating. Some transistors, such as power transistors, need a thicker layer than others; in high-frequency transistors where well-defined edges are needed, a thin layer of photoresist is used. Hence, the spinning speed for power transistors is less than that for high-frequency transistors.

Base and Emitter Diffusion

When the slice has been coated with the photoresist, it is covered with the base mask and exposed to ultraviolet light. The base mask is a photographic negative which carries a pattern formed by the proposed positions of the bases of the transistors that will be made from the slice.

The parts of the photoresist exposed to ultraviolet light change to a hard material which cannot easily be washed off, while the unexposed parts are easily removed. Therefore, after washing, there are many small windows in the photoresist, through which the silicon oxide is exposed. These windows are in the positions of the transistor bases.

The remaining photoresist is further hardened by baking the slice, which is then treated with acid to etch away the oxide exposed through the windows and reveal the silicon. The rest of the photoresist is then removed and the slice is ready for base diffusion.

When all the photoresist has been removed, the slice is again cleaned and placed in an oven with a little of the required p-type doping additive (usually boron). The additive evaporates and some of it settles in the windows. The slice is then transferred to a hotter oven filled with oxygen. In this oven, the additive is driven into the silicon to the required depth. At the same time, another oxide layer is formed all over the slice, including the windows. The base of each transistor in the slice has now been formed.

The slice is again covered with the photoresist and another pattern of windows produced as described above. This time, however, a different negative—the emitter mask—is used, and the windows are in the positions of the transistor emitters. The cleaning, diffusion and oxidation processes are repeated to form the transistor emitters. The doping additive used in this diffusion process, however, is not boron but phosphorus—an n-type substance.

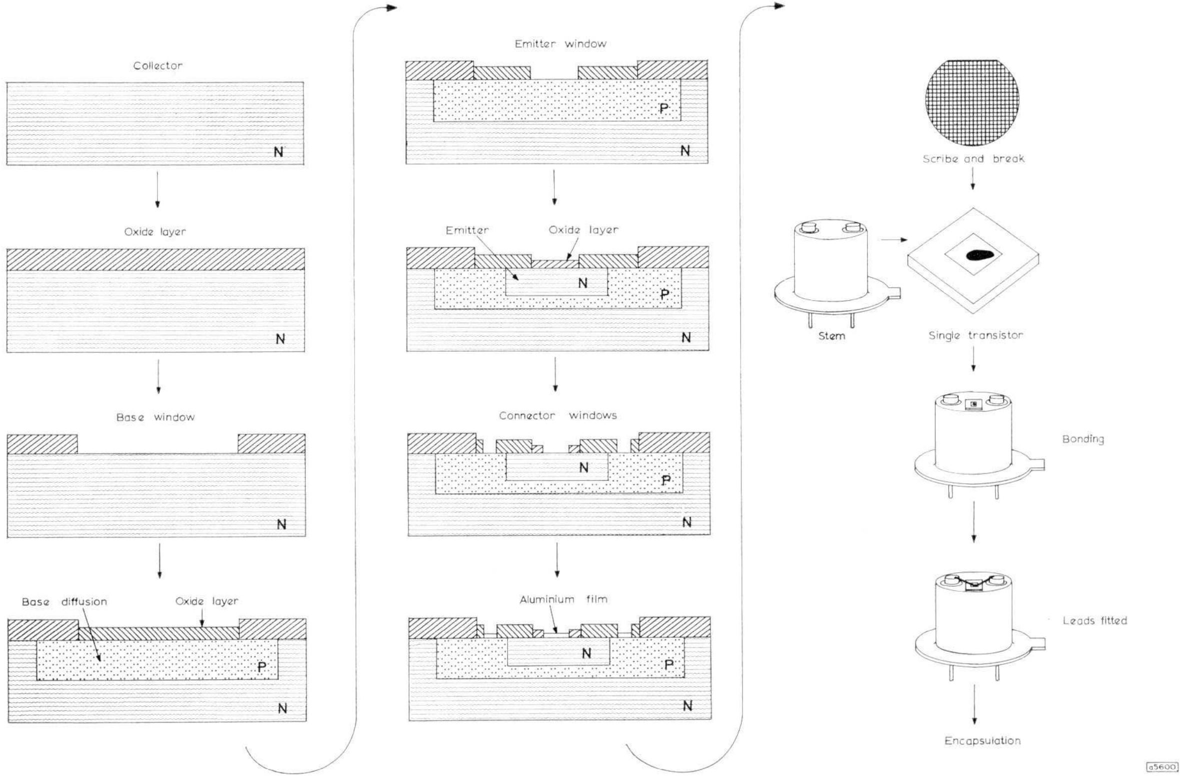


Fig. 2—Stages in the manufacture of silicon planar transistors

Lead Connections

By means of the photoresist and a third negative, windows are produced over the base and emitter areas. The slice is placed in a vacuum and a fine layer of aluminium deposited all over it. The slice is removed from the vacuum and, by means of a fourth mask and another etching process, the aluminium is removed from all areas except where it is needed to provide contacts to the bases and emitters. The remaining aluminium is lightly alloyed to the bases and emitters to form non-rectifying contacts. At a later stage, fine gold or aluminium wires will connect the aluminium layer to the transistor leads or pins.

The oxide layer is now removed from the collector side of the slice, and the thickness of the slice is reduced by etching to lower the thermal resistance and improve the knee characteristic of the transistors.

The slice now contains between 200 and 6000 transistors, depending on the type of transistor, to which no leads have been attached. Each transistor is tested for breakdown voltage and small-signal current gain by means of a special probing machine connected to a test set. A photograph of this machine is given in Fig. 3. Transistors which fail any of these tests are marked and rejected at a later stage in manufacture.

After these tests, the slice is scribed with a diamond and broken into dice, each containing one transistor. At this stage, the marked transistors are rejected.

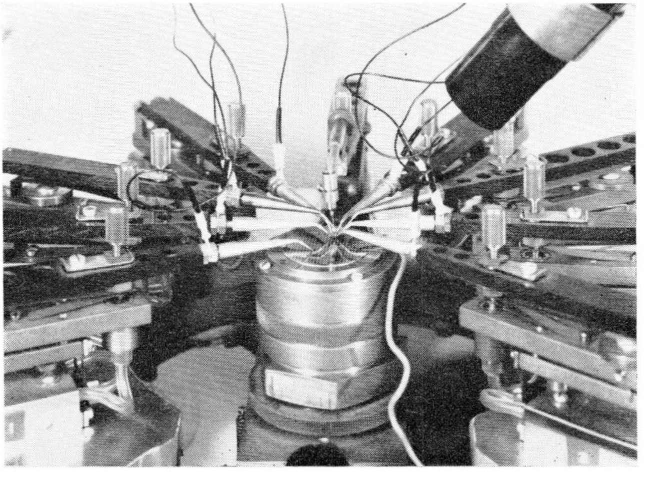


Fig. 3—Probing machine checking breakdown voltage and current gain of transistors

Each good transistor is soldered or bonded, collector side down, on the gold-plated stem or header. This operation is performed automatically on a dice bonder which fixes the transistor to the header by a silicon-gold eutectic.

The wires connecting the base and emitter to their leadout wires are usually of gold or aluminium, and are attached to the aluminium film on the transistor by means of a suitable bonder. When the leads have been attached, the transistors are encapsulated.

Lockfit Encapsulation

The familiar metal-can encapsulation requires no further description. However, a more modern encapsulation is epoxy resin, which is used on the range of Lockfit transistors. The coating of epoxy material provides the semiconductor wafer with complete environmental protection, and the junction-to-ambient thermal conductivity of the material is superior to that obtained with most metal encapsulations. The transistor leads take the form of flat pins which are designed to snap into printed-wiring board as shown in Fig. 4.

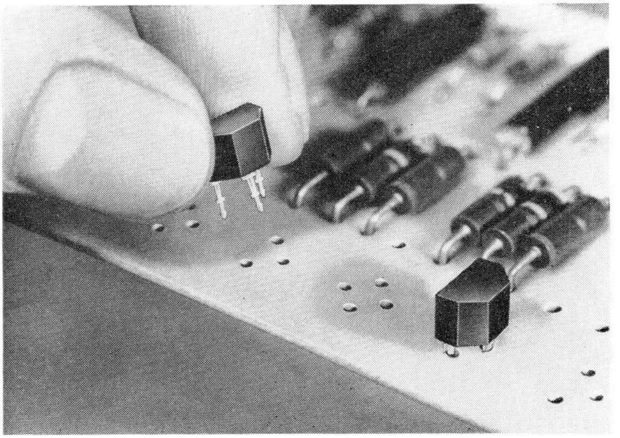


Fig. 4—Insertion of Lockfit transistors in printed-wiring board

The encapsulated transistors are subjected to a temperature cycling test, a drop test and a test to detect pinholes in the encapsulation.

Epitaxy

When special types of silicon planar transistor are required, the procedure is modified slightly. When, for example, transistors are required for r.f. and i.f. applications, the silicon substrate is more heavily doped than in other transistors, and it undergoes further treatment in the crystal-growing chamber before being cleaned and covered with the first coat of photoresist. Consequently, the slice is covered with a thin layer—5 to 20 micrometres thick—of lightly-doped silicon. This layer has the same crystal orientation as the original slice and is therefore known as the epitaxial layer. The subsequent treatment of the slice with the epitaxial layer is the same as previously described. In the epitaxial type, however, the transistor is essentially within the thin layer, and the substrate is really a low-resistance back contact which provides a convenient thickness for handling. With the non-epitaxial type, the transistor is within the substrate.

Field-effect Transistors

Despite the differences in materials and methods of manufacture, the transistors described in the earlier sections of this chapter all have essentially the same construction and principle of operation. They all consist of emitter, base and collector regions, and their properties are all largely a consequence of the movement of electrons and holes across the junctions between these regions or of the existence of the junctions themselves. The field-effect transistor, however, has a structure and characteristics differing from the conventional junction transistor.

Structure and Operation

Fig. 5 is a schematic diagram of a field-effect transistor. It will be seen that a channel of n-type semiconductor material is enclosed for part of

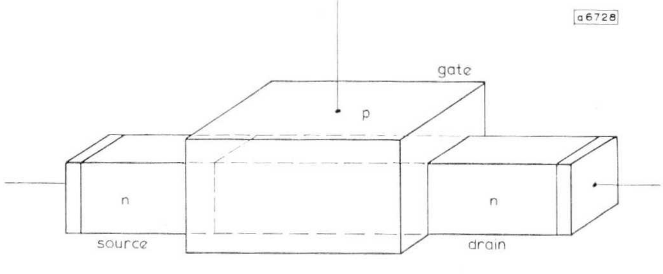


Fig. 5—Schematic diagram of a field-effect transistor

its length by a collar of p-type material. One end of the n-type channel is called the source, and the other end is called the drain. The collar of p-type material is the gate.

If a voltage is applied across the length of the n-channel so that the source is positive with respect to the drain, electrons will flow from the drain to the source; that is, current will flow conventionally from the source to the drain.

If a voltage is applied to the gate so that the gate is negative with respect to the source, the p-n junction formed between the gate and channel will be reverse-biased. No current flows across the junction, but the recombination of holes and electrons that migrate across the junction forms a region in the n-channel depleted of free electrons.

This non-conductive depletion region narrows the conductive n-channel, and thus reduces the current flowing from source to drain. Varying the gate bias voltage will thus vary the source-drain current. The action is analogous to the control of anode-cathode current in a triode valve by means of a negative bias voltage applied to the grid of the valve.

In the field-effect transistor, as in the triode valve, a comparatively small bias voltage will control a large current. Therefore amplification of a signal is possible with the field-effect transistor.

Since the gate-channel junction is reverse-biased during operation, the field-effect transistor is essentially voltage driven, and it therefore has a high input impedance. The device has a low noise figure and good high-frequency performance. For these reasons, field-effect transistors are ideal for use in such applications as the r.f. stages of high-quality f.m. tuners.

Manufacture

The structure of an n-channel field-effect transistor is shown in Fig. 6. The planar process is employed in the manufacture of these transistors.

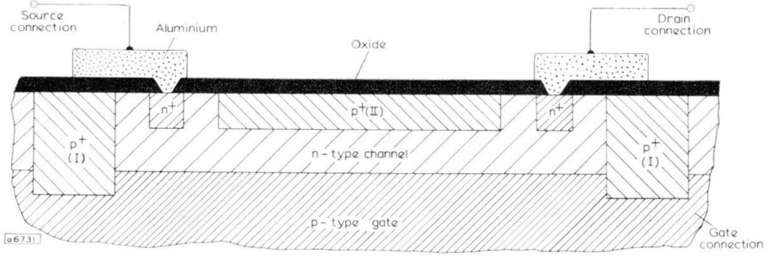


Fig. 6—Structure of a field-effect transistor

An n-type epitaxial layer which eventually will form the n-channel is grown on a substrate of high-conductivity p-type material, which will form the gate. After oxidation of the surface of the epitaxial layer, windows are etched in the oxide layer corresponding to the channel-isolating region I. Through these windows is diffused an additive rich in holes, which creates a 'super-p' or p^+ region penetrating through the epitaxial layer to the p-type substrate, and thus isolating the n-channel.

After re-oxidation, a window is etched corresponding to region II, and another p^+ region is created to form the upper part of the p-gate surrounding the n-channel.

A third diffusion is performed to create 'super-n' or n^+ regions at the ends of the n-channel. Aluminium bonds are deposited on these n^+ regions, and to these are connected the source and drain contact wires. The purpose of the n^+ region is to diminish the series resistance of the source and drain contacts. The gate contact is made to the p-type substrate, and the transistor is encapsulated in a metal envelope.

CHAPTER 2

BASIC H.F. CIRCUITS

The block schematic diagram for the high-frequency stages of a typical a.m./f.m. radio receiver is given in Fig. 7. The audio-frequency, or a.f., output is fed to one of several types of a.f. amplifier, which are discussed more fully in Chapter 3. In the present chapter, the operation of each of the circuits named on the block diagram is discussed in general terms, detailed practical circuits being given in later chapters.

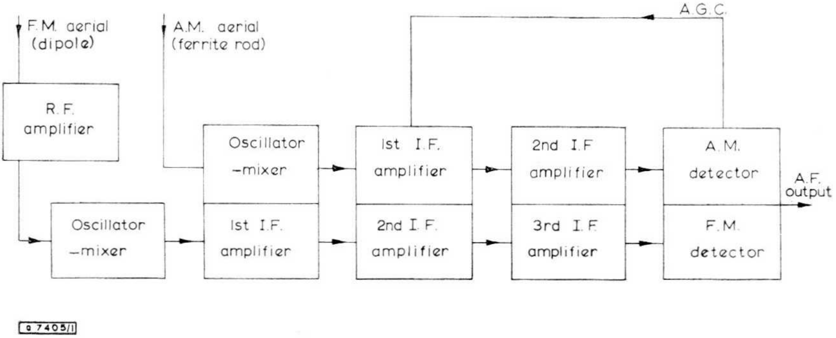


Fig. 7—H.F. stages of a typical a.m./f.m. receiver

R.F. AMPLIFIER

A typical basic r.f. amplifier circuit is shown in Fig. 8. The transistor is connected in the common-base configuration because this configuration gives the highest gain at frequencies approaching the cut-off frequency. The collector circuit is tuned and is inductively coupled to the self-oscillating mixer circuit.

MIXER

The mixer, or frequency changer, converts by heterodyne or 'beating' action the wanted frequency of the carrier signal at the aerial to a fixed frequency at the input to the i.f. amplifier. The a.m. carrier frequencies vary from 160 to 280kHz in the long waveband and from 540 to 1640kHz in the medium waveband. The f.m. carrier frequencies vary from 85 to 100MHz. The intermediate frequency is usually 470kHz for a.m. and 10·7MHz for f.m. The audio-frequency information, carried on the r.f. waveform as modulation of the amplitude or frequency, is transferred from the r.f. carrier signal to the i.f. carrier.

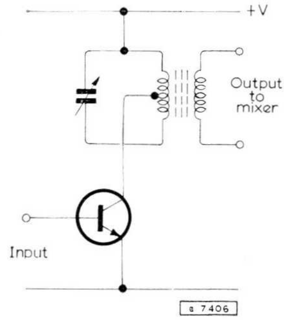


Fig. 8—Basic r.f. amplifier circuit

Heterodyne action is effected by injecting signals of different frequencies into a non-linear element such as the base-emitter diode of a transistor. In a radio receiver, one injected signal is the aerial signal, derived either directly or through an r.f. amplifier, and the other is a signal generated within the receiver by a local oscillator. The signals combine as shown in Fig. 9 to produce a complex waveform. When the waveform is applied between the base and emitter of a transistor, the main components occurring in the collector waveform are the original aerial and oscillator frequencies together with their sum and difference frequencies. The difference frequency (i.f. signal) is preferentially selected by the tuned circuit in the collector of the mixer, and further selectivity is provided by the i.f. amplifier stages. Voltage gain is obtained from the mixer stage in the same manner as the gain obtained from a normal i.f. amplifier, but somewhat reduced in magnitude.

To maintain a constant difference between the aerial and oscillator frequencies, the oscillator frequency must be changed whenever the aerial frequency is changed, that is, the oscillator tuned circuit must 'track' the aerial tuned circuit. Therefore, if the aerial is tuned to 1·5MHz, the local oscillator must be tuned to 1·97MHz, to give a difference frequency of 470kHz, and if the aerial tuning is changed to 1MHz, the oscillator must be tuned to 1·47MHz. For a.m. reception, the oscillator is usually tuned to frequencies 470kHz above the aerial frequency so that signals in the long waveband, which have frequencies below 470kHz, can be accepted. For f.m. reception the oscillator frequency may be 10·7MHz above or below the aerial frequency dependent upon other design considerations.

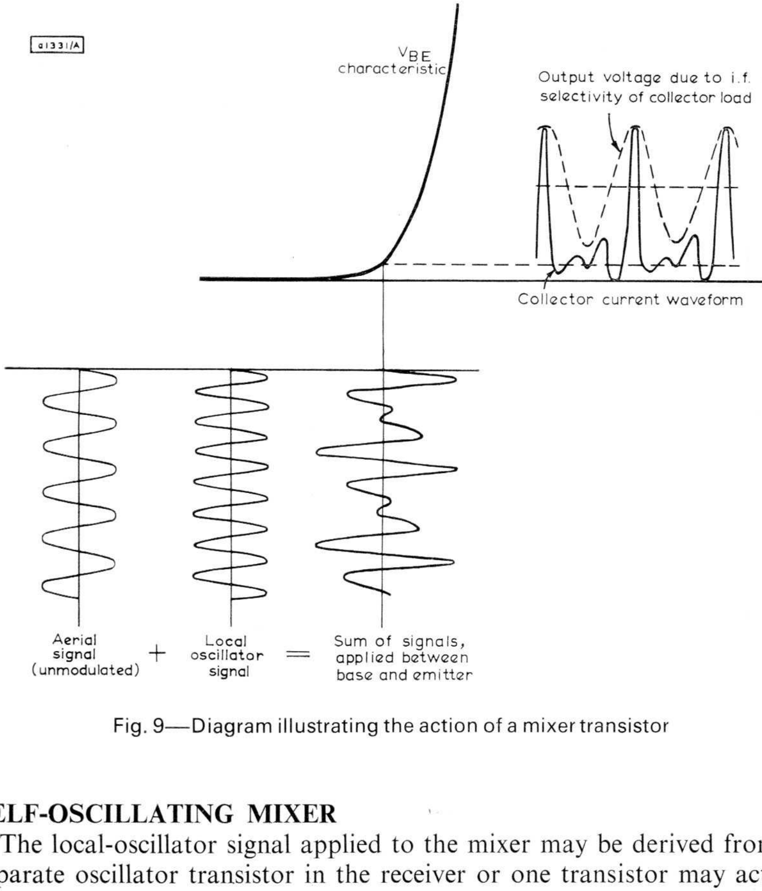


Fig. 9—Diagram illustrating the action of a mixer transistor

SELF-OSCILLATING MIXER

The local-oscillator signal applied to the mixer may be derived from a separate oscillator transistor in the receiver or one transistor may act as both oscillator and mixer. This latter arrangement is the more common and is known as a self-oscillating mixer.

The basic circuit of a self-oscillating mixer is shown in Fig. 10. At the oscillator frequency, the base of the transistor is effectively connected to earth because the aerial tuned circuit has negligible impedance. The input electrode for the oscillator action is, therefore, the emitter, which is inductively coupled to the collector of the transistor. Both the emitter winding and the collector winding are closely coupled to the oscillator tuned circuits. Positive feedback is applied from output to input through this coupling, and oscillation occurs at the resonant frequency of the oscillator tuned circuit. The aerial signal is applied from the aerial tuned circuit to the base of the transistor. Both the aerial signal and the oscillator signal are, therefore, in series with the base-emitter diode. The component of the heterodyne output signal having the required intermediate frequency is extracted by the i.f. bandpass filter or transformer in the collector circuit.

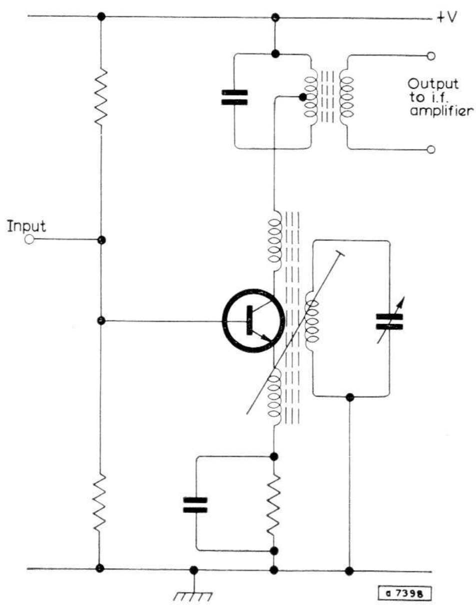


Fig. 10—Basic self-oscillating mixer circuit

A typical self-oscillating mixer for a.m. reception is shown in Fig. 11. The circuit is shown for one waveband only, the different wavebands being obtained by switching other coils into the aerial circuit and additional capacitors into the oscillator circuits. The aerial input is taken to the base of the transistor, and the oscillator feedback from the collector is taken through low-impedance coupling coils to the emitter.

To ensure that the oscillation starts easily, the transistor is biased initially for class A operation by the potential-divider bias circuit. As the oscillation increases in amplitude, rectification of the oscillating signal occurs at the base-emitter diode. This causes the loop gain after each cycle to decrease, stabilising the amplitude of the oscillation and causing the quiescent emitter current to rise slightly (from 250 to 300μA in the circuit shown in Fig. 11).

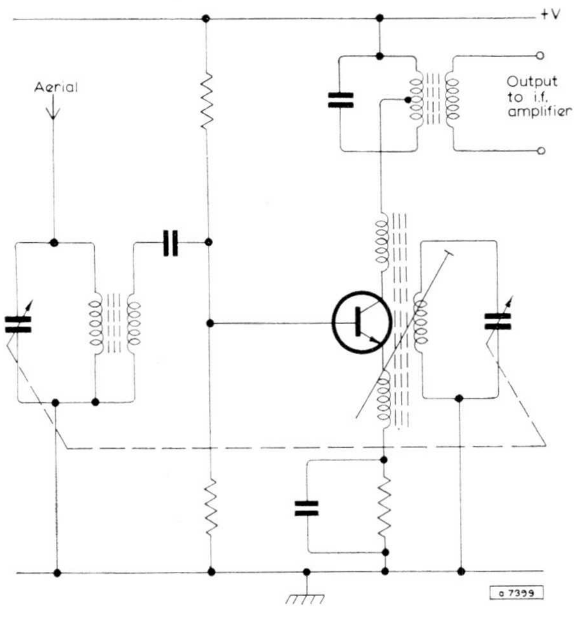


Fig. 11—Typical practical self-oscillating mixer circuit

Correct tracking of the aerial and oscillator tuned circuits to maintain the 470kHz difference in frequency can be obtained by a conventional padder circuit or by using a tuning capacitor with specially-shaped vanes for the oscillator circuit. The value of the tuning capacitance is not critical but must be sufficient to provide the required frequency coverage.

Stray capacitance can exist between the aerial and oscillator sections of the tuning capacitor and so form a path for unwanted feedback between the two sections. To prevent this feedback and the resulting possibility of spurious oscillations, a screen is usually placed between the two sections. Another source of feedback between the oscillator and aerial sections is through the stray capacitance of the leads to the wave-change switch, and these leads should, therefore, be kept as short as possible.

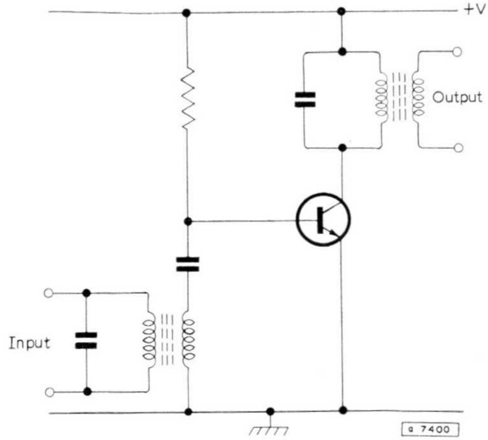


Fig. 12—Basic i.f. amplifier circuit

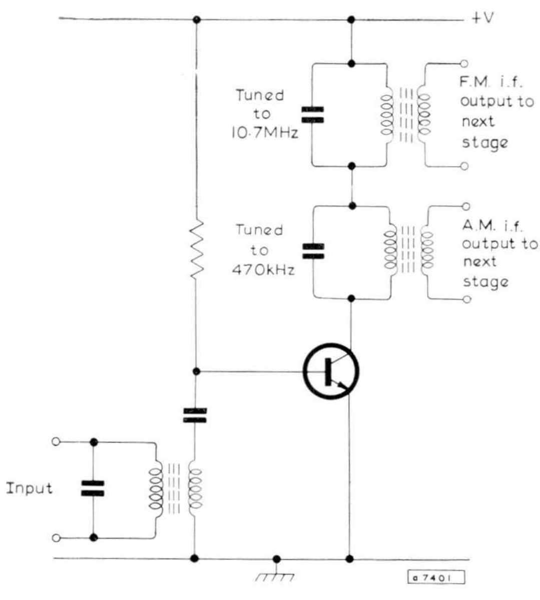


Fig. 13—Basic i.f. stage for an a.m./f.m. receiver

I.F. AMPLIFIER

The basic circuit of an i.f. amplifier stage is shown in Fig. 12. The input and output transformers are tuned to the intermediate frequency, the selectivity and bandwidth being improved if both windings of the transformers are tuned. In an a.m./f.m. receiver it is necessary for two i.f. transformers to be provided between stages—one tuned to 470kHz and the other to 10·7MHz. A method of connecting these i.f. transformers is illustrated in Fig. 13. The remaining i.f. stages are similar to the first except that, in an a.m./f.m. receiver, the a.m. and f.m. outputs must be separated so that each can be fed to its own detector circuit.

DETECTOR

The a.m. detector of a transistor receiver is usually a semiconductor diode feeding a potentiometer load. The potentiometer forms the volume control, which feeds the demodulated signal to the audio-frequency stages of the receiver. The potentiometer is usually shunted by a 10nF capacitor, as shown in Fig. 14, to filter out the i.f.

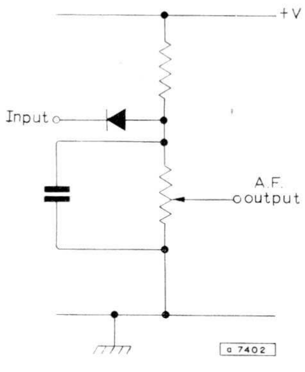


Fig. 14—Basic a.m. detector circuit

The detector stage for f.m. signals is usually of the type known as a ratio detector, the basic circuit of which is shown in Fig. 15.

The ratio detector depends for its operation on the fact that as the frequency of the signal varies, the phase relationships in a circuit tuned to the nominal frequency change. The transformer has three windings, the primary and secondary tuned to the intermediate frequency and the tertiary, close-coupled to the primary, injecting a voltage into the secondary

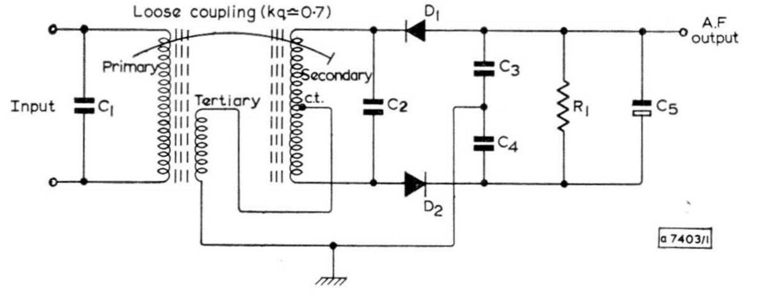


Fig. 15—Basic ratio detector circuit

in series with the voltage across each half of the secondary. Figs. 16 and 17 show the phase relationships. In Fig. 16, the voltage V_S induced in each half of the secondary is 90° out of phase with the injected primary voltage V_P at the intermediate frequency. Voltage V_{D1} and V_{D2}, applied to the detector diodes, are equal in magnitude, so there is no audio output. Fig. 17 shows the phase conditions when the frequency of the signal differs from the nominal value. The effect of a shift in frequency is to change the phase of the secondary voltage with respect to the injected primary voltage. As the frequency of the signal sweeps above and below the nominal value the variation of the relative phase of the injected primary and secondary voltages produces a variation in the amplitudes of V_{D1} and V_{D2} at audio frequency. The voltages are rectified by the two diodes and the a.f. signal taken from across either C_3 or C_4. The long-time-constant circuit R_1C_5 suppresses any unwanted a.m. signals.

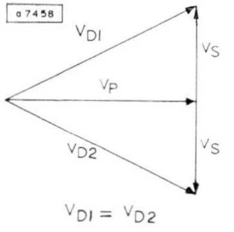


Fig. 16—Vector diagram of voltages in a ratio detector circuit with signal at nominal frequency

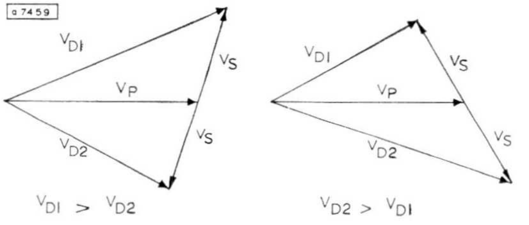


Fig. 17—Vector diagram of voltages in a ratio detector circuit with signal varying from nominal frequency

AUTOMATIC GAIN CONTROL

It is necessary for the sensitivity of r.f. and i.f. stages in a receiver to be high, so that weak signals are amplified to give a useful output. However, if no automatic control of gain were incorporated in an a.m. receiver, clipping would occur when the receiver was tuned to a strong signal. The automatic gain control (a.g.c.) signal, proportional to the i.f. output, is derived from the detector and applied in series with the base-emitter voltage of the first i.f. transistor (and the r.f. transistor in car radios). The a.g.c. signal therefore controls the bias, and hence the gain, of the transistor. The simplest method of modifying the basic detector and i.f. amplifier is shown in Fig. 18.

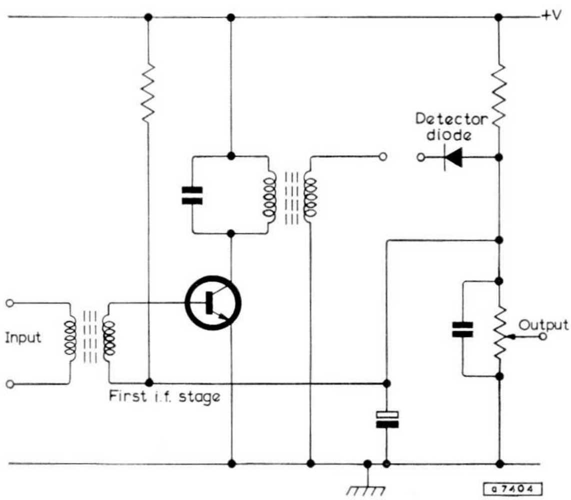


Fig. 18—Simple a.g.c. circuit

To eliminate the audio- and radio-frequency signals from the a.g.c. voltage, the decoupling capacitor of the potential divider feeding the base of the first i.f. transistor is large—8μF is a typical value. The a.g.c. characteristic may be improved by adding a damping diode across the first i.f. transformer. This diode ensures that the a.g.c. action is effective over a larger range of r.f. input voltages than in a normal circuit.

In f.m. receivers the waveform is already limited for a.m. suppression and a.g.c. is generally thought to be unnecessary, although it may be included to improve the performance of the i.f. stages.

BIAS AND STABILISATION AGAINST TEMPERATURE VARIATIONS

In the foregoing discussion of r.f. and i.f. circuits, no mention is made of biasing requirements. The operating current of a transistor is determined by the biasing voltages or currents applied to it, and these are usually set by means of resistors in the base, emitter and collector circuits. Because of the superior high-frequency performance of silicon transistors, germanium devices are not used in any of the r.f. and i.f. stages described in this book. Therefore only biasing arrangements for silicon transistors in these stages are described in detail.

Ideally, biasing circuits should allow a particular value of operating current to be obtained irrespective of changes in supply voltage and temperature; and spreads in the current amplification factor, h_{FE}, and the base-emitter voltage, V_{BE}, of the transistor. Unless an external stabilising element is used, it is impossible to eliminate the effects of variations in supply voltage; but in a good biasing arrangement, the resulting variations in current will not be much greater than the percentage variations in the supply voltage.

Different biasing arrangements satisfy these requirements to different degrees. The biasing arrangement shown in Fig. 19 has been almost universally adopted for low-level class A amplifiers using germanium transistors. In this arrangement the current is only slightly influenced by spreads in h_{FE}, but a value of emitter voltage greatly in excess of V_{BE} is required if the current variations resulting from supply voltage variations are to be small. On the other hand, a single biasing resistor in the base circuit, as shown in Fig. 20, gives negligible changes in current with V_{BE} variations, but the current in this circuit is fully dependent on spreads in h_{FE}.

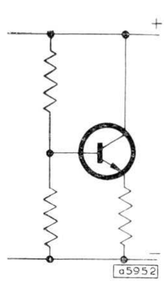


Fig. 19—Conventional voltage - biased transistor

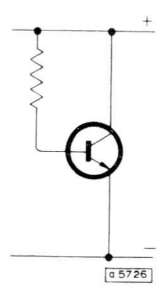


Fig. 20—Conventional current-biased transistor

The two major differences between the characteristics of silicon and germanium transistors that affect the biasing arrangements occur in the base-emitter voltage V_{BE} and collector leakage current I_{CBO}. The base-emitter voltage of silicon transistors at low currents is 0·6 to 0·7V, and of germanium transistors, 0·1 to 0·2V; the typical mutual characteristics of a silicon and a germanium transistor are compared in Fig. 21. The difference in V_{BE} is equivalent to a difference in 'starting' voltage, but the rate of change of current with base-emitter voltage above this 'starting' voltage is identical for silicon and germanium transistors. The only consequence of this difference in V_{BE} is that, for conventional voltage biasing shown in Fig. 19, a larger drop across the emitter resistor will be required with silicon transistors than with germanium transistors for comparable changes in operating current with supply voltage, unless a stabilised source of bias voltage is available.

The collector leakage current in silicon transistors is negligible up to the highest temperatures likely to be encountered in radio equipment. Consequently, it would seem to be possible to use a high value of d.c. base source resistance, and to dispense with the normal stabilising circuit, which applies negative d.c. feedback to stabilise the operating conditions. However, other effects need consideration and these may necessitate retaining the stabilising circuit and limiting the magnitude of the source resistance. These effects are variations in supply voltage, spreads in V_{BE} and h_{FE}, and variations in V_{BE} and, to a lesser extent, h_{FE} with temperature.

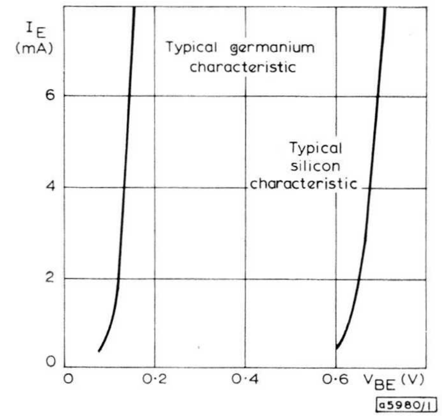


Fig. 21—Typical mutual characteristics of a germanium and a silicon transistor

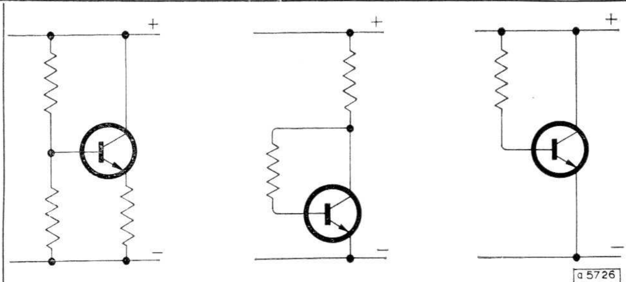
The three most common biasing arrangements are voltage biasing with emitter feedback, current biasing with collector feedback and simple current biasing. The choice of biasing arrangement depends very much on the supply voltage, the proportion of this voltage that can be dropped across a feedback resistor, and the type of transistor used. A comparison of the spread in emitter operating current resulting from spreads in h_{FE} and V_{BE}, for each of the three biasing arrangements, is given in Table 1. The changes in operating current when the supply voltage falls to half its nominal value are also compared. An analysis of the biasing arrangements, and worked examples showing their relative effectiveness, are given in Appendix 1.

TABLE 1

Comparison of Biasing Arrangements

Nominal values of parameters:

$V_S = 7V$; $V_{BE} = 0{\cdot}7V$; $h_{FE} = 100$
$I_E = 1mA$ (approx.)

Biasing arrangement			
Parameter variation		I_E variation	
h_{FE} ±50%	+1, -1·6%	+23, -35%	±49%
V_{BE} ±7%	±1·7%	±0·95%	±0·86%
V_S −50%	−62%	−55%	−56%

CHAPTER 3

BASIC A.F. CIRCUITS

Audio-frequency amplifiers range from the simple amplifier used in small portable radio receivers, where size and cost are the important factors, to the large multi-stage circuit where cost is of little importance but reproduction of the highest quality is required. In the present chapter, only the basic amplifying stages are considered, the refinements necessary for high-fidelity reproduction being considered in Chapter 7.

The only class of one-transistor amplifier stage in common use is class A. Push-pull stages, either using two similar transistors or a matched complementary pair, may operate in class AB or class B. A special type of class AB push-pull amplifier sometimes encountered is known as 'π-mode'.

CLASS A OPERATION

Class A operation is that in which the values of bias and signal voltage applied to the transistor ensure that collector current always flows. Class A output stages are used mostly in car radios, where current drain is relatively unimportant, although class A circuits are invariably used as the earlier stages of multi-stage amplifiers where the current level is comparatively low. Class A, one-transistor output amplifiers are inefficient and have the additional disadvantage of suffering from even as well as odd harmonic distortion, although they are simple and enable moderate powers to be obtained from circuits consisting of very few components.

A basic class A output stage is shown in Fig. 22. The bias is provided by resistors R_1 and R_2, and resistor R_3 provides thermal stabilisation.

PUSH-PULL OPERATION

Push-pull output stages are used almost universally in modern amplifier circuits, and push-pull circuits are often used in the driver stages of multi-stage amplifiers. The output power obtained from a pair of transistors connected in a push-pull configuration is considerably more than double the power obtained from one transistor. Push-pull operation also gives rise to less of the even harmonics sometimes associated with single-transistor, class A amplifiers. The use of transistor push-pull driver and

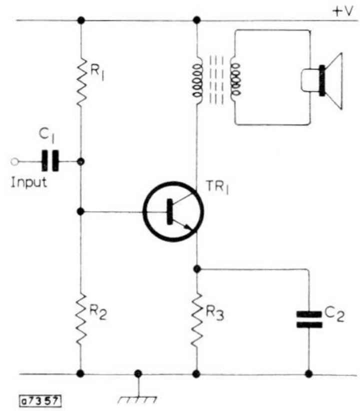


Fig. 22—Basic class A output stage

output circuits has eliminated the need for driver and output transformers, which were an essential part of valve push-pull circuits.

In a class B push-pull output stage, two transistors are biased nearly to cut-off, so that only a small quiescent current flows under zero-signal conditions. The signal is applied in antiphase to each transistor by means of some kind of phase-splitting arrangement, so that one transistor is cut off while the other conducts. With push-pull operation, the efficiency is 60 to 65% and the average current drain is low compared with class A operation. This low current drain is, of course, a great advantage in portable battery receivers.

If both transistors are biased exactly to cut-off, a form of distortion occurs which is extremely unpleasant to the listener. This is called 'cross-over' distortion and is overcome by applying a slight forward bias to each transistor.

Output transistors dissipate considerable amounts of power and are, therefore, usually fitted with some kind of heatsink.

Because push-pull circuits using complementary pairs have virtually superseded those using two similar transistors in all but the output stages of high-power amplifiers, they are considered first.

Complementary Push-Pull Circuits

A basic complementary push-pull output stage is shown in Fig. 23. The signal from the driver transistor TR_2 is applied simultaneously to the bases of the two output transistors, TR_3 and TR_4, which conduct during alternate halves of the waveform. The amplified signal, occurring as variations in the potential at point A, is applied via capacitor C_4 to the loudspeaker. The collector load of the driver transistor is returned to the live end of the loudspeaker so that the signal is applied between base and emitter of the output transistors. The output devices therefore operate in common-emitter, giving maximum amplification.

The quiescent bias voltage for the output transistors is developed across a resistor R_7 placed in the collector circuit of the driver transistor. Since the value of R_7 is much lower than the value of R_8 its inclusion does not introduce any significant imbalance at the input of the output transistors. Resistors R_9 and R_{10} are included in the emitter circuits of the output transistors to improve thermal stability.

Transistor TR_1 acts as a pre-amplifier for a.c. signals and as a d.c. difference-amplifier comparing the voltage V_1, derived from the potential divider R_1, R_2, R_3, with voltage V_2. The high loop gain of the circuit keeps the small difference between V_1 and V_2 constant, so that V_2 is defined with respect to V_1 regardless of spreads in the characteristics of transistors TR_1 and TR_2 and the tolerance variations in values of R_5, R_6, R_7 and R_8.

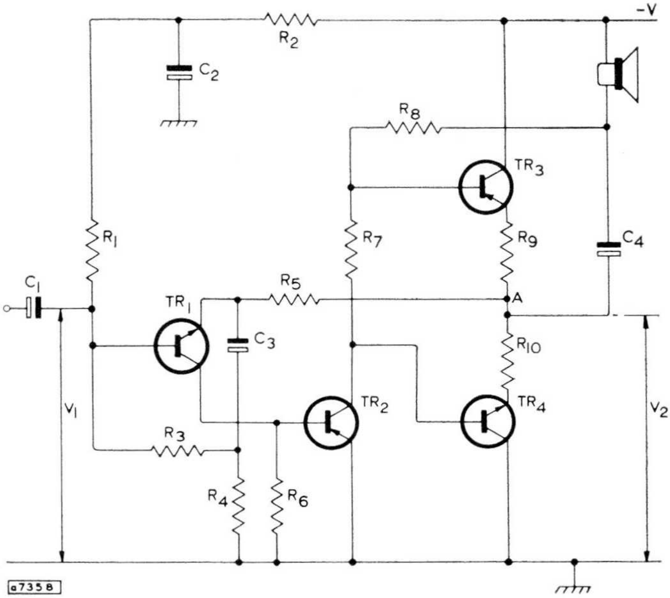


Fig. 23—Basic complementary push-pull circuit

Negative feedback is taken via resistor R_5 from the output to the emitter of TR_1. Decoupling is provided by capacitor C_3 to remove a.c. feedback between these two points, and a known small amount of feedback is re-introduced by means of R_4 to reduce gain spreads and distortion.

Crossover distortion can be eliminated by biasing the output transistors to give a linear combined transfer characteristic. For a given output stage, an optimum bias condition (and quiescent current) exists for which the crossover distortion is a minimum. Changes in supply voltage or ambient temperature, however, will cause departures from this optimum and therefore increase the crossover distortion.

Considerable spreads about the nominal current may result from tolerance spreads in resistor values and transistor characteristics, and from variations in ambient temperature and supply voltage. The effects of changes in ambient temperature can be minimised by the use of a thermistor in conjunction with the biasing resistor, R_7, which can be preset to compensate for component spreads. The effects of supply voltage changes can be minimised by means of a diode or other stabilising device from which the bias can be derived.

Zobel Network

The purpose and operation of the Zobel network can be explained by considering the simple complementary-pair output stage shown in Fig. 24. The operating limits of this circuit are defined by the continuous line in Fig. 25. While the load line remains within the enclosed area, the transistors will not be damaged by overloading. With a purely resistive load, this is a fairly easy matter to arrange. However, if the load is inductive, the limits could be exceeded during transients when the amplifier receives a large signal.

In extreme cases, too large a signal to the driver approximates to a large square wave. When such a signal first bottoms the driver transistor, current flows from TR_2 into the load. On the next sharp edge of the input signal, TR_1 is rapidly cut off. With a purely resistive load, TR_2 is cut off, TR_3 conducts, and the current reverses immediately. With an inductive load, however, the inductance opposes the change of current and the potential at point A rapidly decreases.

If the line voltage is less than the maximum permissible voltage of TR_2, the potential at A will become slightly negative until TR_3 acts as a reverse transistor. The energy stored in the inductance is then safely dissipated before breakdown occurs.

However, if the line voltage exceeds the maximum permissible voltage of TR_2, the voltage across TR_2 will reach the maximum permissible voltage before A becomes negative. Consequently, the energy stored in the inductance will be dissipated in the breakdown region of TR_2 and may be sufficient to cause failure of the transistor. Failure of TR_2 will be followed by failure of TR_3 as this will then be subjected to the full line voltage.

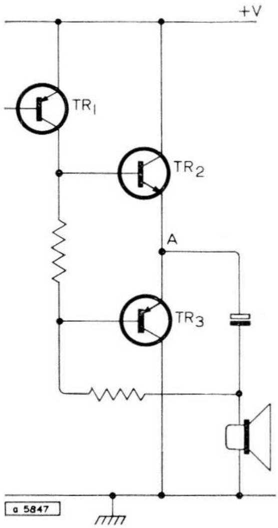


Fig. 24—Basic complementary-pair output stage

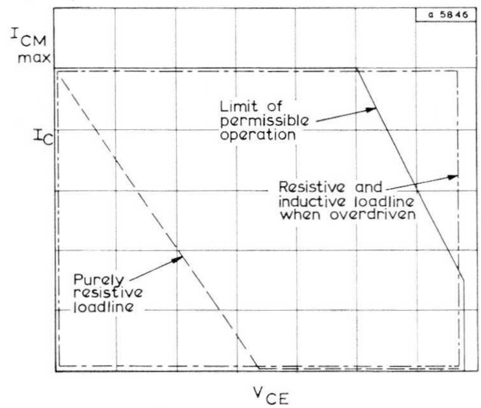


Fig. 25—Operating limits and load lines for a transistor in a complementary pair

As nearly all loudspeakers contain inductance, output circuits must be designed to allow for it. This can be done either by using transistors with high breakdown voltages or by taking precautions to prevent the conditions for transistor breakdown being satisfied. If the transistors were presented with a purely resistive load, the breakdown of TR_2 would not occur. This can be achieved by connecting a Zobel network in parallel with the loudspeaker as shown in Fig. 26. The Zobel network is a capacitance, C_Z, in series with a resistance, R_Z, the values of these components having a certain relationship to the inductance and resistance of the loudspeaker. If the loudspeaker is equivalent to a resistance, R_S, in series with an inductance, L_S, the components of the Zobel network must have values given by

$$R_Z = R_S$$

and

$$C_Z = \frac{L_S}{R_S^2}.$$

When these conditions are satisfied, the parallel combination of the loudspeaker and the Zobel network will present a purely resistive load to the output transistors. In practice, of course, the impedance of the loudspeaker is only approximately a resistance in series with an inductance. Consequently, the use of a Zobel network will not result in a perfect resistive load, but the similarity is very close.

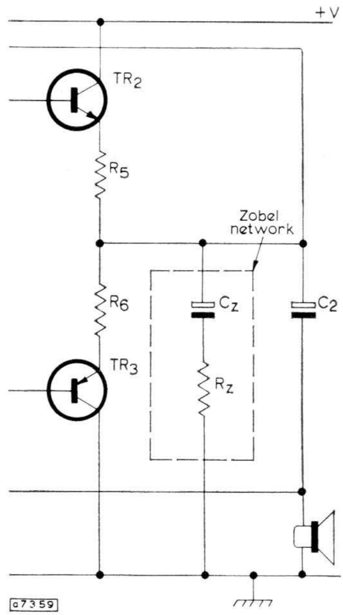


Fig. 26—Basic complementary push-pull circuit with Zobel network

Push-pull Circuits using Two Similar Transistors

The complementary push-pull circuit is normally used for amplifiers with low output powers, but the push-pull circuit using two similar transistors is still often used as the output stage of high-power amplifiers. This is because n-p-n silicon transistors capable of handling high powers are more readily available than p-n-p transistors. The most usual arrangement for high-power amplifiers is for an output stage consisting of two n-p-n transistors connected in a push-pull configuration to be driven by a complementary push-pull circuit as shown in Fig. 27 (quasi-complementary configuration).

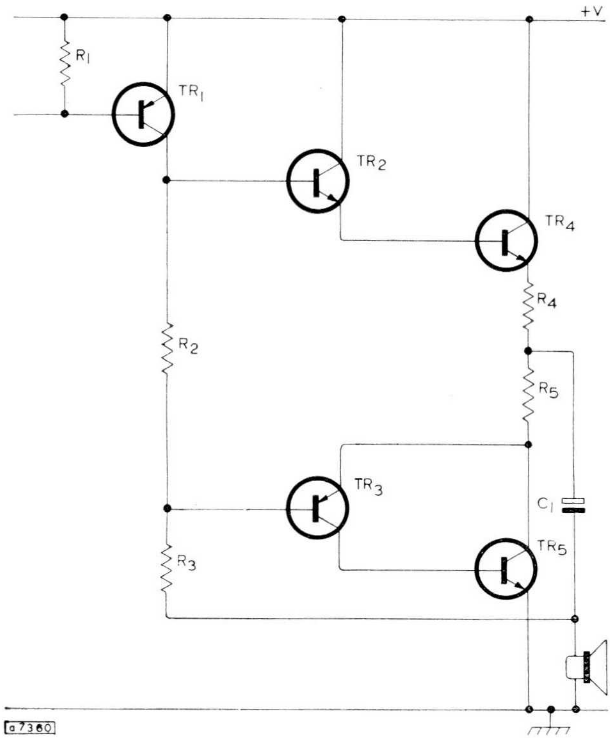


Fig. 27—Push-pull output stage using two n-p-n transistors driven by a complementary push-pull circuit

Pi-Mode Operation

Class AB operation is that in which the values of bias and signal voltage applied to the transistor cause collector current to flow for appreciably more than a half, but less than a whole, cycle of the signal voltage.

A class AB amplifier which operates in the π-mode differs from a normal class AB amplifier in that the total current remains constant with drive. Increasing the level of drive causes the operating point of the transistors to change first from class A to AB and then to class B.

With π-mode operation, the problem of crossover distortion associated with straight-forward class B operation does not arise. Regulation of the power supply is not important since, with π-mode operation, the current drain from the supply is constant and adequate smoothing of the power supply can be obtained by simple RC filtering. Distortion at normal listening levels is very low since the transistors operate in class A at low power levels and, with the use of emitter-loaded operation, the distortion of the amplifier over the full power range can be controlled to specific limits by a suitable choice of driver-stage source impedance.

The chief disadvantage of π-mode operation is that it requires a relatively large number of transistors to give the required power output.

CLASS D OPERATION

Class D operation differs from the modes of operation already considered in that the output transistors are used as switches in a single-ended push-pull arrangement. The transistors are fed with a square wave which carries the audio information as pulse-width modulation. The transistors switch the load between positive and negative voltages, the duration for which it is switched to either supply depending on the mark/space ratio of the input waveform. The transistor output is coupled to the load through a low-pass filter which removes the high-frequency content and feeds an audio-frequency waveform to the load.

This arrangement makes use of the fast switching times and low saturation voltages of the modern switching transistor. It offers potential benefits over other modes of operation: only the amount of power supplied to the load at any instant is drawn from the power supply and no excess power is converted to heat by the output transistors; the linearity of the circuit is independent of the transistors; large component tolerances are acceptable because the transistors alternate between fully 'off' and fully 'on', and no biasing currents are required; efficiencies of 95% can be realised and a large range of load impedances can be connected without substantially affecting efficiency. However, the high-speed circuitry and radiation due to fast switching present problems, and this concept is still in the development stage. No class D circuits are given in this book.

LOUDSPEAKER DIVIDING NETWORK

In the practical circuits which follow this chapter, the only loudspeaker specification quoted is impedance. However, it should be pointed out that the choice of loudspeaker is important in any audio system and that a dual system is invariably preferable to a single unit.

In a dual system, two speakers—one for high frequencies (the tweeter) and one for low frequencies (the woofer)—are used, together with a suitable crossover network to divide the audio output into two ranges. Loudspeaker manufacturers often recommend a suitable network and it is advisable to follow their recommendations. However, an example of a two-way network is shown in Fig. 28. The values of inductor L and capacitor C are given in terms of loudspeaker impedance R and crossover frequency f by the following equations:

$$L = \frac{R\sqrt{2}}{2\pi f},$$

$$C = \frac{\sqrt{2}}{4\pi f R}.$$

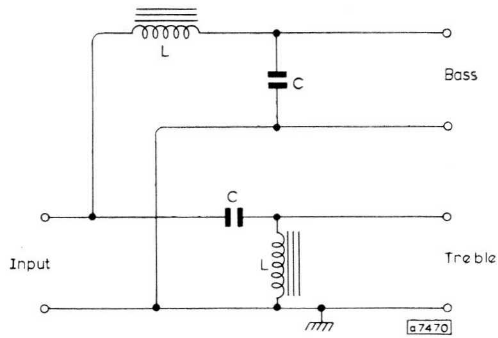


Fig. 28—Loudspeaker dividing network

OUTPUT POWER RATINGS

In practical circuits, a number of different ratings may be used to define the output power from an audio amplifier; they are:

1. Speech and music: the equivalent sine-wave output that will be obtained when peaks occur in low-level passages of speech or music. (Since this energy is taken from the reservoir capacitor, the impedance of the supply does not influence this figure.)
2. Sustained music: the equivalent sine-wave output that will be obtained during sustained high-level passages of music. (This may be lower than the first rating because the supply voltage may fall due to regulation.)
3. Sine-wave: output power attained when the amplifier is sine-wave driven.

These ratings may be quoted:

(a) at a specified distortion; the output power obtained at the specified level of distortion, often 10%;

(b) at clipping; the output power at which clipping commences, that is when the working point of the transistors reaches the knee of the output characteristic resulting in a distortion of the output waveform.

The first two ratings constitute a useful measure of the output power capability of the amplifier. The third indicates the power to be obtained when the amplifier is tested with a sine-wave drive and is a lower figure than the first two.

In the I.H.F.M. (Institute of High Fidelity Manufacturers) system, the following two ratings are used:

1. Continuous power: the power an amplifier is capable of delivering for at least 30 seconds with sine-wave drive.
2. Music power: the power obtainable after the application of a signal, measured during a time interval so short that the supply voltages in the amplifier have not changed from their zero-signal values.

HEATSINKS

In determining the size of heatsink required for output, and sometimes driver, transistors in audio amplifiers three factors are taken into consideration:

1. the maximum permissible junction temperature of the device;
2. the thermal conductivity of the semiconductor material and encapsulation of the device, and of the heatsink material;
3. maximum ambient temperature.

The circuits given in this book are designed for operation in ambient temperatures up to 45°C and the minimum size of heatsink, to be mounted vertically in free air, has been specified for each circuit. It should be remembered that if a different heatsink material is used the size of heatsink must be adjusted accordingly.

CHAPTER 4

RADIOGRAMS, RECORD PLAYERS AND PORTABLE RADIOS

The chapter is divided into two sections comprising:

1. A.F. amplifiers with output powers of 1W, 3W and 10W; a pre-amplifier for use with the 3W and 10W amplifiers; a 3W record-player amplifier.
2. R.F. and i.f. stages for an a.m. receiver and an a.m./f.m. receiver, both using a 9V supply.

Component lists for the circuits are given at the end of the chapter.

Circuits for high-quality stereo amplifiers are given in Chapter 7.

1W AUDIO AMPLIFIER

The circuit diagram for a 1W transformerless audio amplifier is shown in Fig. 29. The circuit includes a complementary output pair comprising one AC127 and one AC128 transistor, an AC128 driver transistor and an AC127 pre-amplifier. The use of high-gain output transistors allows a simple amplifier, with capacitively-coupled load, to be designed with a relatively low quiescent current. The n-p-n pre-amplifier transistor contributes to very good d.c. stability.

Circuit Description

The features of this design are the economy effected in the number of components required, and the extremely good stability of the d.c. potential V_M. The latter implies less variation of output with temperature changes compared with other forms of output stage using a capacitively-coupled load.

The load is capacitively-coupled to the emitters of TR_3 and TR_4, the bases being driven directly from the collector of TR_2. The collector load R_7 of TR_2 is returned to the live end of the loudspeaker, so that the output transistors operate in common-emitter configuration, giving maximum amplification.

A small quiescent bias voltage is required for transistors operating in class B push-pull. This is developed by the collector current of TR_2 flowing through R_9. Thermistor R_8 is used to compensate for the temperature dependence of the quiescent current.

The transistor TR_1 serves a dual purpose. It acts as a pre-amplifier for a.c. signals and also as a d.c. difference-amplifier comparing the voltage V_1 derived from the potential divider R_1, R_2, R_3, with the voltage V_M. The high loop gain of the circuit keeps the small difference between V_M and V_1 constant, so that V_M is defined with respect to V_1 , irrespective of spreads in the characteristics of the transistors TR_1 and TR_2 and the tolerance variations in the values of R_5, R_6, R_7, R_8 and R_{10}.

Negative feedback is taken via the resistor R_5 from the output stage to the emitter of TR_1. Decoupling is provided by the large-valued capacitor C_2 to remove a.c. feedback between these points, and a small known amount of a.c. feedback is re-introduced by means of R_4 to reduce gain spreads and distortion.

Performance

Output power at 10% distortion	1·2W r.m.s.
Output power at 4% distortion	0·8W r.m.s.
Sensitivity (source impedance $R_S = 2k\Omega$):	
output 50mW, 6dB feedback	2·2mV
output 1·2W, 6dB feedback	10mV
Frequency response (−3dB points)	100Hz to 20kHz
Driver current	7·7mA
Peak collector current	470mA
Total quiescent current	13mA

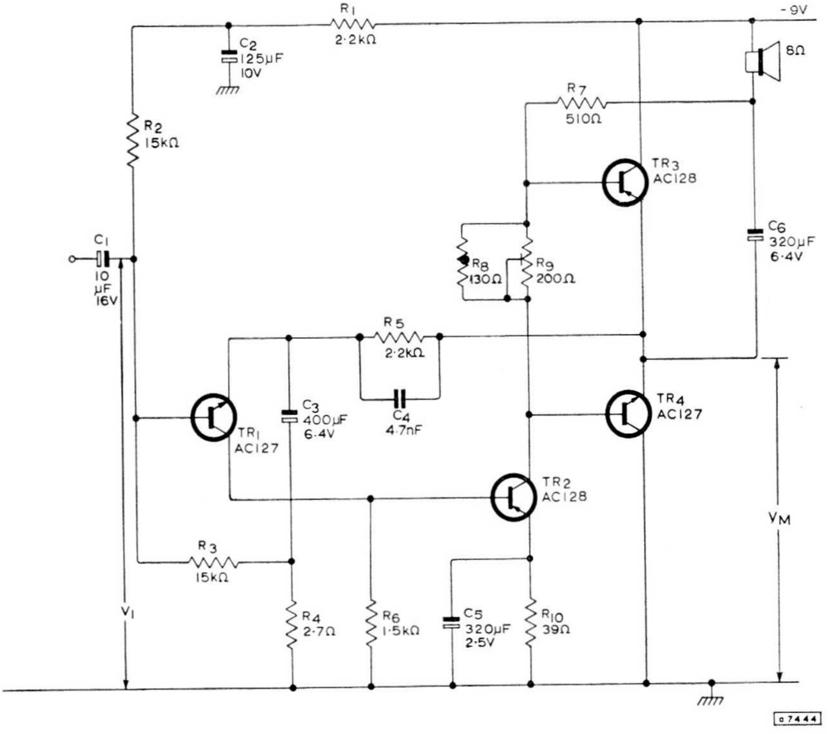


Fig. 29—1W audio amplifier circuit

Tone Control Circuit

As feedback is applied over three stages in the amplifier of Fig. 29, negative-feedback tone controls can cause instability and passive controls at the input should be used. Fig. 30 shows a simple arrangement for obtaining variable bass and treble cut, without significant loss of gain at mid-frequencies.

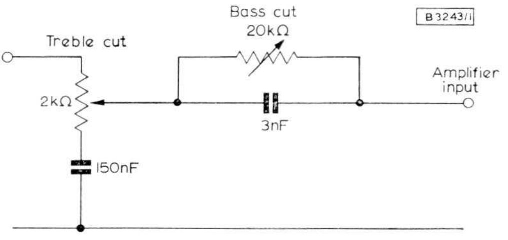


Fig. 30—Tone control circuit

3W AUDIO AMPLIFIER

The circuit for a directly-coupled 3W audio amplifier is shown in Fig. 31. The complementary matched output pair comprises one AC176 and one AC128 transistor, the driver is an AC128 and the pre-amplifier is a BC148 or BC108 transistor.

Circuit Description

A conventional four-transistor configuration delivers 3W into a 12Ω load. The use of a fixed resistor R_7 for bias is acceptable because of the relatively high values of emitter resistors and the large amount of feedback. It is recommended that the resistors which determine the working point—that is resistors R_1, R_2, R_3, R_7, R_8—should be 5% tolerance components.

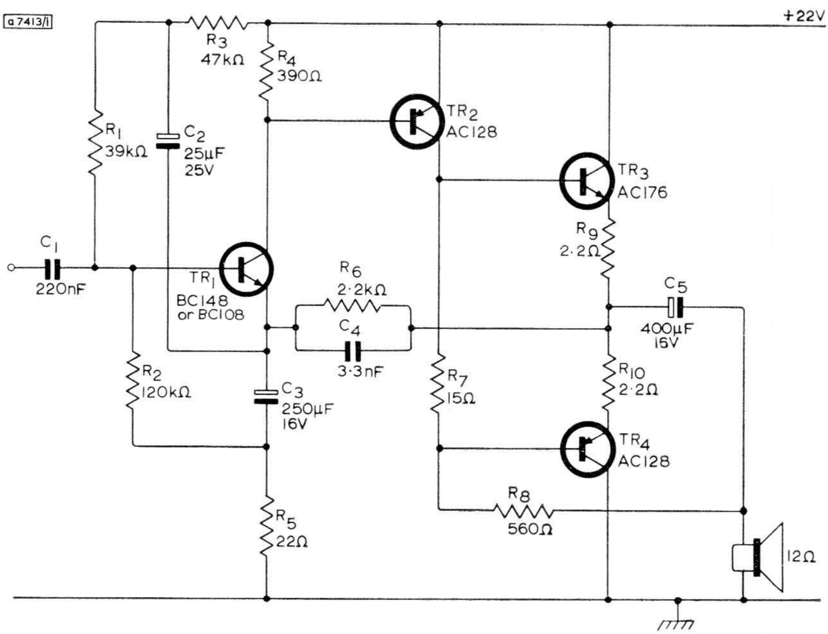


Fig. 31—3W audio amplifier circuit

Power Supply

A suitable power supply is shown in Fig. 32; the requirements are:

Quiescent consumption	25mA
Full power consumption	200mA
Effective output resistance	$\geqslant 20\Omega$

Performance

Output power	3W r.m.s.
Sensitivity for 3W output	70mV
Frequency response (—3dB points)	40Hz to 20kHz
Input resistance	80kΩ
Feedback	20dB
Distortion at 3W output (see Fig. 33)	1·7%

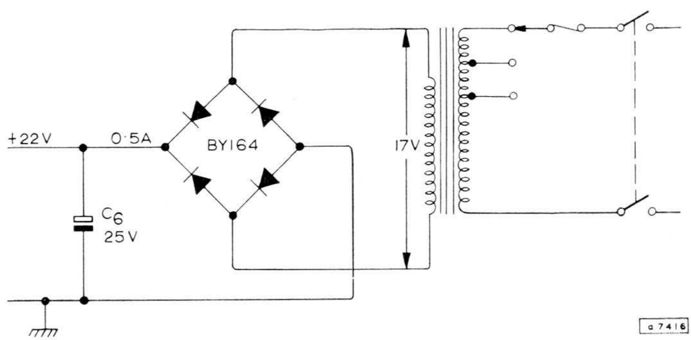


Fig. 32—Power supply circuit for 3W amplifier

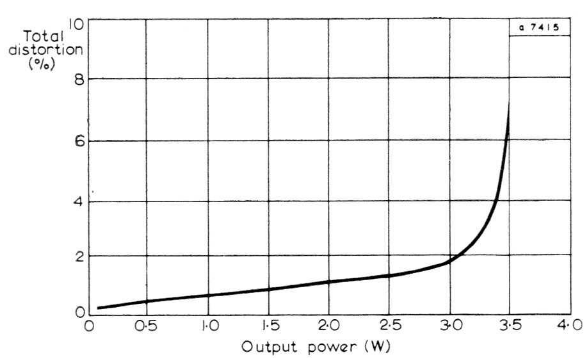


Fig. 33— Variation of total harmonic distortion with output power

10W AUDIO AMPLIFIER

The circuit of a 10W amplifier is shown in Fig. 34. In this circuit, a complementary matched pair push-pull output stage is used, and a Zobel network is incorporated.

Circuit Description

The emitter of transistor TR_1 is connected to the output stage via R_5 to provide a path for negative feedback. The quiescent current of TR_3 and TR_4 is set to approximately 15mA by means of R_9. The thermistor R_8 compensates for temperature changes and maintains a constant quiescent current. The amplifier can deliver an output of 10W into a 4Ω loudspeaker with a distortion of less than 3% (see Fig. 36).

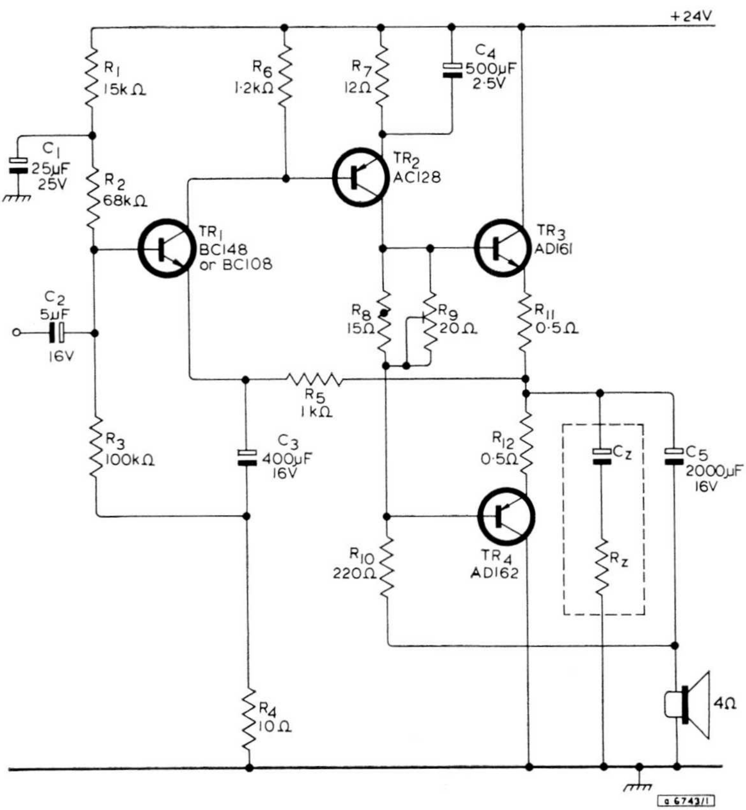


Fig. 34—10W audio amplifier circuit incorporating a Zobel network

Zobel Network

The purpose and operation of the Zobel network is explained in Chapter 3. From the formulae given on page 27, and for a 4Ω loudspeaker with an inductance of 200μH

$$R_Z = 4\Omega,$$

$$C_Z = \frac{200 \times 10^{-6}}{16} = 12{\cdot}5\mu F.$$

(A 25V component is required.)

Power Supply

The nominal line voltage required by the amplifier is +24V and the absolute maximum voltage is +27V. The maximum value of direct current per channel is 0·8A.

The circuit of a simple power supply suitable for use with a stereo system is shown in Fig. 35. The mains transformer has a secondary output of 18V which is rectified by a diode bridge and smoothed by capacitor C_6.

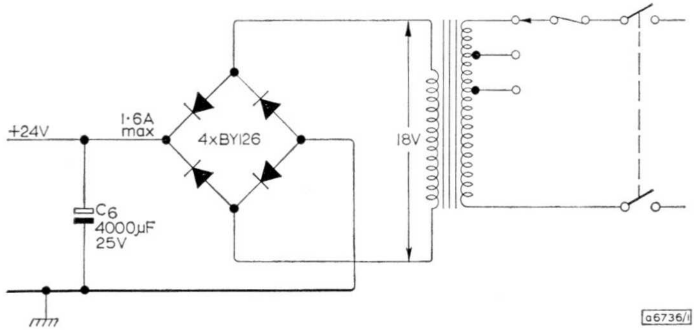


Fig. 35—Power supply circuit for 10W amplifier

Performance

Output power	10W r.m.s.
Sensitivity (voltage at base of TR_1 for rated output)	80mV
Frequency response (—3dB points)	40Hz to 20kHz
Input impedance	30kΩ
Negative feedback	15dB
Distortion at 10W output (see Fig. 36)	2·9%

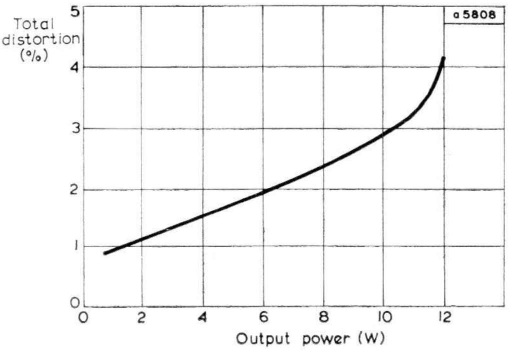


Fig. 36—Variation of total harmonic distortion with output power

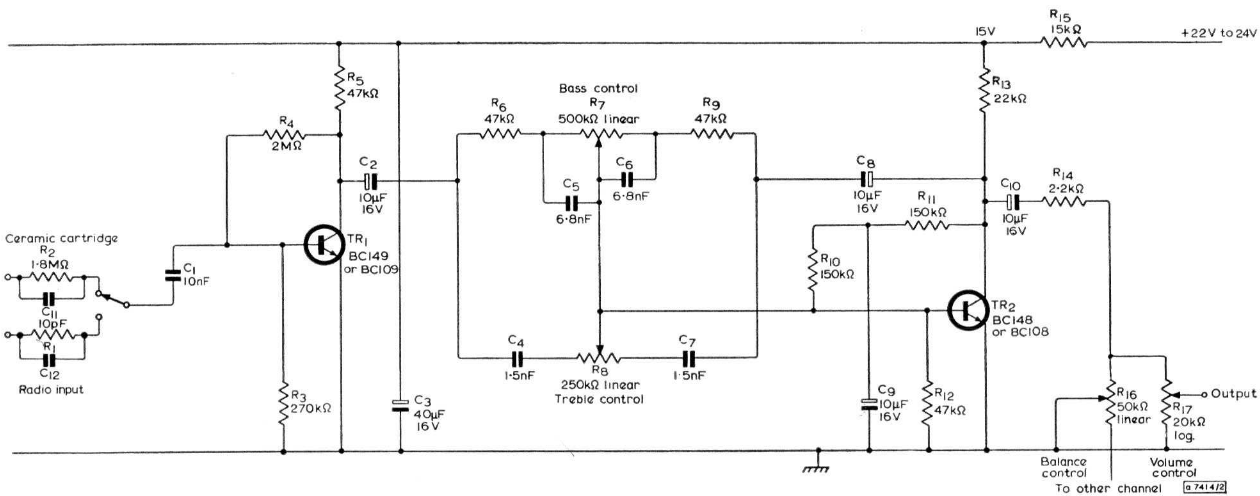


Fig. 37—Pre-amplifier circuit for use with 3W and 10W amplifiers

PRE-AMPLIFIER FOR USE WITH THE 3W AND 10W AMPLIFIERS

The circuit for a two-stage pre-amplifier which may be used with the 3W and 10W amplifiers already described is shown in Fig. 37. The first stage uses a low-noise BC149 or BC109 transistor and the second stage a BC148 or BC108 transistor.

Circuit Description

The input stage provides matching for a ceramic cartridge and for radio, the resistor R_1 being chosen to give the required sensitivity in the latter case. A capacitor C_{12} is connected in parallel with R_1, the value of capacitance being chosen to give, with R_1, a time-constant of 18μs. The second stage provides comprehensive tone control, with 10dB boost and cut at 100Hz and 10kHz. This stage is followed by a simple balance control for stereo use and a volume control.

Performance

Sensitivity:	
ceramic input, for 80mV output	180mV
radio input, with $R_1 = 470k\Omega$	45mV
Signal handling, output at clipping	1·5V

3W RECORD-PLAYER AMPLIFIER

The amplifier shown in Fig. 38 is suitable for use with cartridges giving an output of 300mV at 1cm/s, and may be adapted to include comprehensive tone control (see Fig. 41). The circuit uses a complementary output pair of one AC176 and one AC128 transistor, a BC107 driver transistor and a BC108 transistor.

Circuit Description

A crystal pick-up requires a high load, 1 to 2MΩ, if the bass response is to be adequate. A simple way to provide this load, and at the same time minimise gain spreads, is to apply feedback in series with the input. This technique is used in this amplifier.

Transistor TR_1 is an emitter follower supplying the driver transistor TR_2. Because of the large amount of feedback used in the circuit, the output transistors, TR_3 and TR_4, can have simple resistive bias. The output voltage is divided by resistors R_{11} and R_{12} and fed back to the input. The network C_4 and R_{10} forms a simple treble cut control.

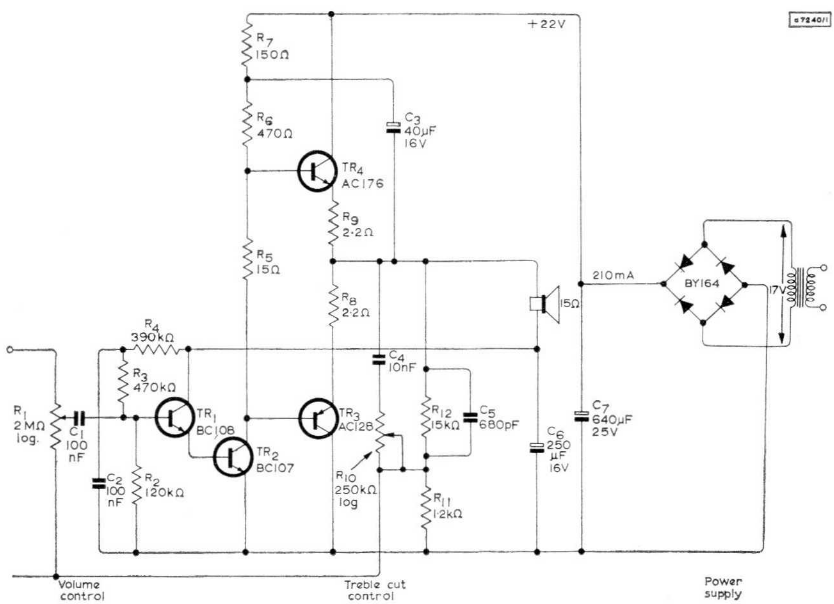


Fig. 38—3W record player amplifier circuit

The loudspeaker and capacitor C_6 are interchanged with respect to their conventional positions to provide a low-voltage supply for the collector of transistor TR_1, and also a point relatively free from a.c. signals from which d.c. feedback can be taken. The output mid-point voltage is compared with the base voltage of transistor TR_1 by the potential divider comprising resistors R_2, R_3 and R_4. Although the voltage is taken from a nominally d.c. point, decoupling is still required to remove low frequencies.

Performance

Output power	3W r.m.s.
Sensitivity for full output	600mV
Input resistance:	
with volume control at maximum	660kΩ
with volume control at minimum	2MΩ
Feedback, depending on position of volume control, up to	25dB
Distortion at 2W output:	
with volume control at maximum	0·4%
with volume control at —10dB	2%
Quiescent current	20mA
Total current at 3W output	210mA

Fig. 39 shows the variation in bass response with volume control setting, due to the variation in input resistance. It should be noted that the circuit given in Fig. 39 representing the pick-up was used to obtain these curves.

Fig. 40 shows the variation of distortion with output power. With the volume control at maximum setting the feedback is high and the distortion low. However, at intermediate settings of the volume control the feedback is reduced, because of the increase in source impedance, and the distortion is higher.

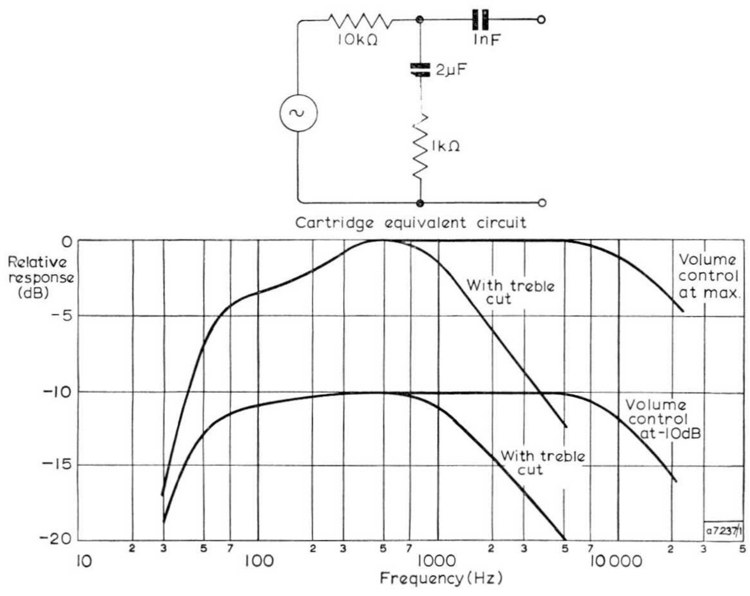


Fig. 39—Frequency response of amplifier

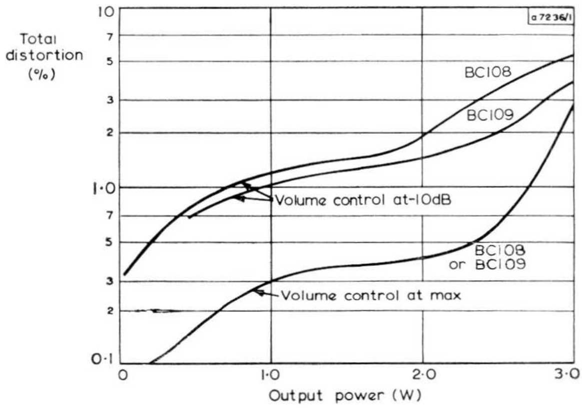


Fig. 40—Variation of total harmonic distortion with output power for basic amplifier and modified circuits

Circuit adapted to include Comprehensive Tone Control

The feedback circuit may be modified to give bass and treble cut and boost controls, as shown in Fig. 41.

The input transistor TR_1 is a BC109, since this device gives a useful improvement in bass response and also in distortion (see Fig. 40). The tone control characteristics with volume control at maximum and at —10dB are shown in Figs. 42 and 43. The value of capacitor C_1 is increased to ensure low-frequency stability with bass boost applied. Capacitor C_9 is included to block d.c. from the bass control, and so prevent noisy operation.

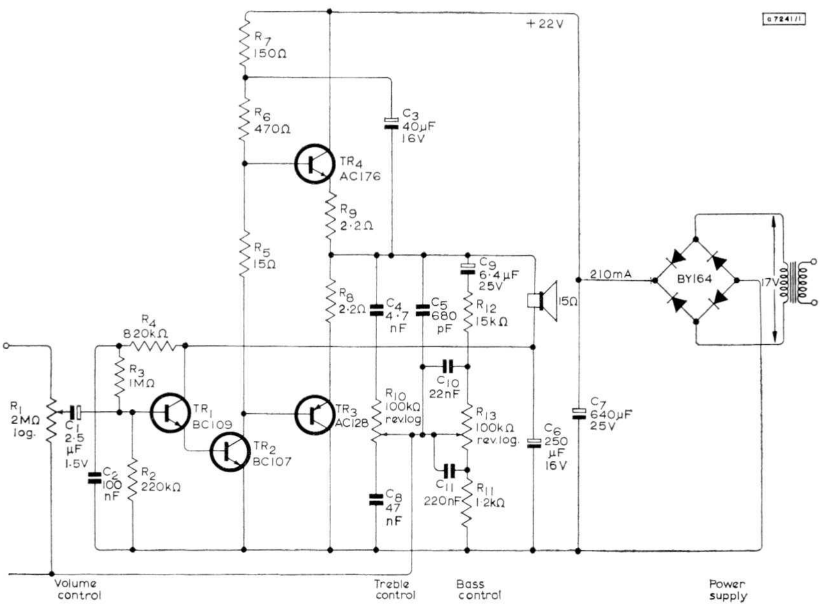


Fig. 41—Amplifier circuit modified to include tone control

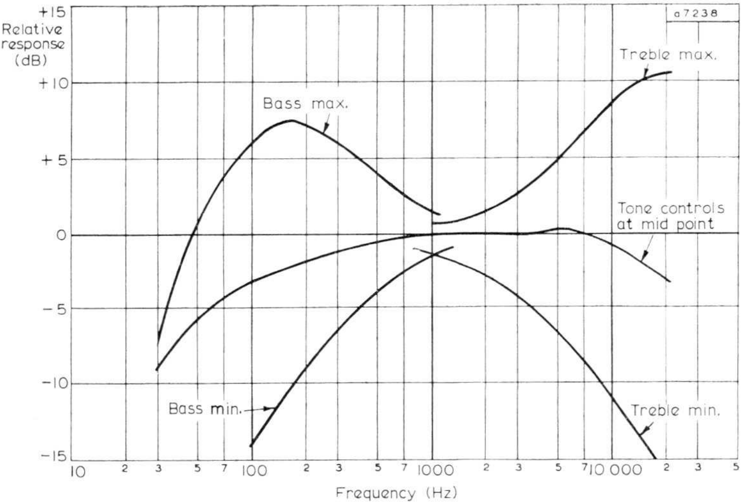


Fig. 42—Tone control characteristics with volume control at maximum

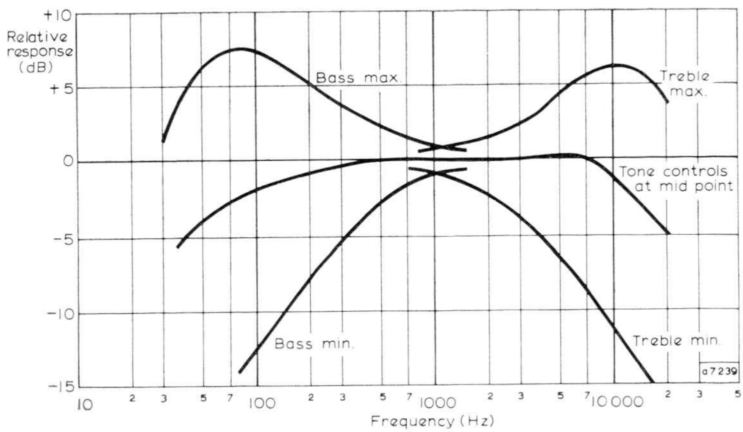


Fig. 43—Tone control characteristics with volume control at —10dB

R.F. AND I.F. STAGES

Attention is drawn to the difficulties associated with the construction and alignment of r.f. and i.f. stages. As stated earlier, access to specialised equipment is essential to those wishing to build these circuits.

A.M. RECEIVER

The r.f. and i.f. stages of a nine-volt medium-wave receiver are shown in Fig. 44. The circuit uses Lockfit transistors, types BF194 and BF195.

Note: The BF194 and BF195 are supplied to receiver manufacturers in sets of three devices. The devices in these sets carry the letters B, C or D added to the normal type number. For replacement or home-construction purposes, these letters should be ignored and the required devices ordered simply by their type numbers, BF194 and BF195.

Circuit Description

Self-oscillating Mixer Stage

The self-oscillating mixer is a conventional circuit, and the BF194 transistor is operated with a collector-emitter voltage of 4·8V, and an emitter current of 1mA. Voltage biasing is used to stabilise the operating point against h_{FE} and V_{BE} spreads; the emitter RC circuit is also necessary to stabilise the oscillator voltage.

The aerial source impedance is 400Ω, and the i.f. load on the mixer is 9kΩ. The working Q-factor of the first i.f. transformer is 80.

First I.F. Stage

The first i.f. transistor is a BF195, and is current-biased from a tap on the detector-diode forward-bias chain. It operates at a collector-emitter voltage of 7·5V, and an emitter current of 1mA.

To ensure good large-signal handling properties, a very large step-down ratio is used between the mixer load and the first i.f. stage. (The signal-handling ability is limited by the amplitude of the signal which can be accommodated at the base of the gain-controlled stage.) Because of the large step-down ratio, the source impedance of the first i.f. stage is very low (21·5Ω). Since this stage drives a load of 50kΩ, it has a very high stability factor (24). However, the gain is sufficiently high, and any increase in load impedance would result in gain spreads caused by variations in output conductance.

Second I.F. Stage

The second i.f. transistor is a BF195, and is current-biased. It operates at a collector-emitter voltage of 7·5V, and an emitter current of 1mA.

The load impedance of the second i.f. stage is 7·5kΩ, and this value ensures maximum power for the a.g.c. line. The source impedance is 530Ω, and therefore the stability factor of the second i.f. stage is 6·5.

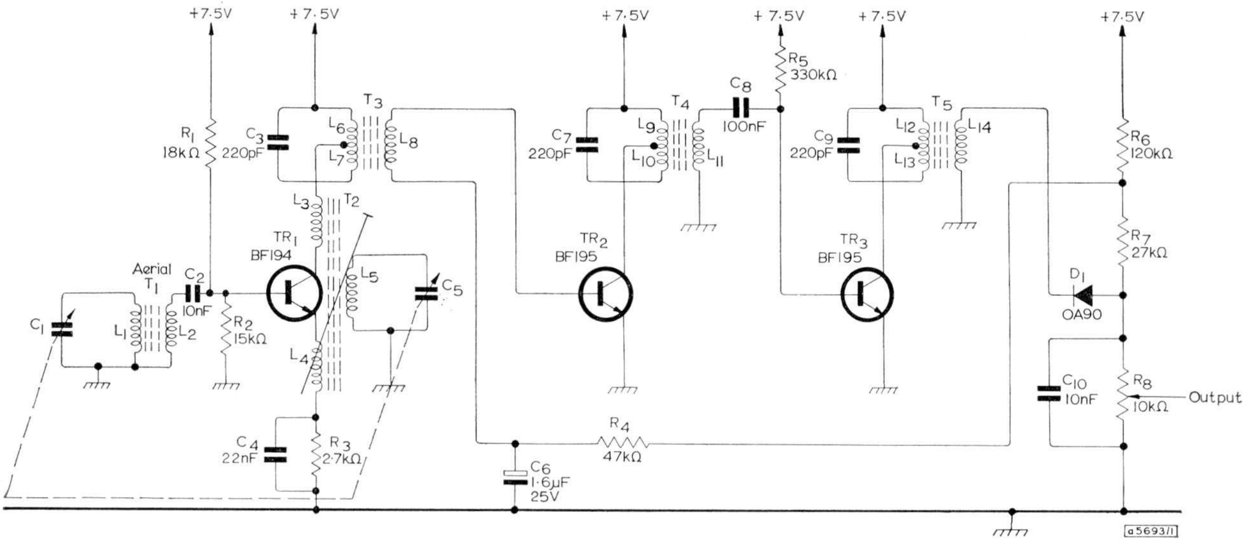


Fig. 44—R.F. and i.f. stages of an a.m. receiver

(Line-up using sets of transistors: TR_1—BF194B, TR_2—BF195C, TR_3—BF195D)

Detector

The OA90 germanium detector diode is forward-biased by 15μA to improve the small-signal efficiency, and to ensure constant loading on the last i.f. transformer.

Performance

A.G.C. range for 6dB change in output (see Fig. 45)	45dB
I.F. bandwidth (−6dB points)	±3·8kHz
Attenuation at 9kHz off-tune	28dB
Gain reduction when the supply voltage is reduced to half its nominal value	15dB
Gain spreads (95% limits) resulting from spreads in h_{FE}, V_{BE}, and 10% resistor tolerances (but excluding coil spreads)	±4dB

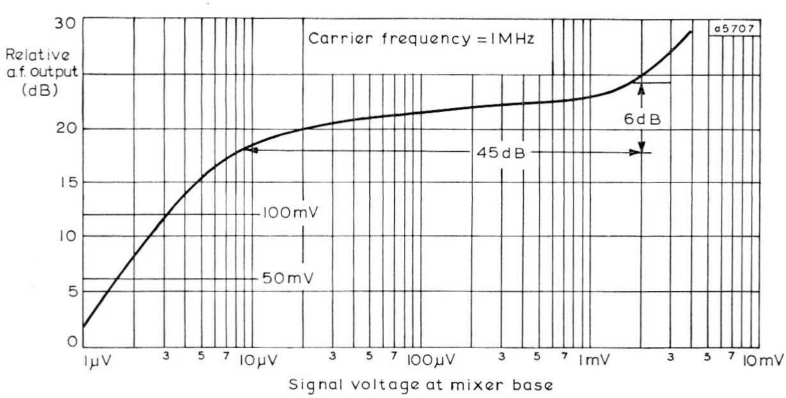


Fig. 45—Automatic gain control characteristic

A.M./F.M. RECEIVER

The r.f. and i.f. stages of an a.m./f.m. receiver operating from a 9V supply are shown in Fig. 47. The design uses three BF195 and two BF194 transistors.

Circuit Description

F.M. Tuner

The f.m. tuner is conventional in design, using a BF195 operated at 1mA as an r.f. amplifier. The self-oscillating mixer uses a BF195 operated at 2mA, the best compromise to obtain minimum oscillator drift with any change in supply voltage (and also good operation at a reduced supply voltage).

A.M. Mixer and A.M./F.M. I.F. Amplifier

The a.m. self-oscillating mixer, TR_3, is voltage-biased using a $3{\cdot}9k\Omega$ emitter resistor in order to obtain stable operation with 100mV drive into the emitter.

The bias arrangement for the a.g.c. stage TR_4 uses d.c. feedback from the collector in order to achieve good gain stability with source voltage variations. However, current biasing of the controlled stage requires some compromise between gain stability, gain spread and a.g.c. voltage; diode D_1 is forward-biased by only $13\mu A$ to achieve good a.g.c. properties. Detector efficiency and distortion are not adversely affected by this low forward-bias current.

The third stage is voltage-biased, to assist in the achievement of a large output voltage swing at the collector.

The f.m. i.f. amplifier is of conventional double-tuned design; the a.m. i.f. amplifier uses single-tuned circuits.

A.M. Performance

6dB bandwidth	±3kHz
Selectivity at 9kHz off-tune	22dB*
Maximum signal handling	110mV/metre
Input change for 6dB change in output	46dB
Oscillator drive at 1MHz	100mV
Lower limit of supply voltage for oscillator operation	2·7V
A.G.C. and signal/noise characteristic	see Fig. 46

*This figure may be increased by increasing the working Q-factor at the expense of the 6dB bandwidth.

M. Performance

3dB bandwidth	±160kHz
Selectivity at 300kHz off tune	43dB
Image rejection ratio	26dB
Limiting sensitivity (3dB below maximum output)	22μV (aerial e.m.f.)
Maximum signal handling	20mV (aerial e.m.f.)
Oscillator drive	200mV
Oscillator drift for 2V change in supply voltage	<25kHz
Oscillator pulling with signal input, at maximum signal	<4kHz
Limiting and signal-to-noise characteristic	see Fig. 48
Total distortion	see Fig. 49
A.M. rejection	see Fig. 50

Transistor and Diode Currents

The quiescent collector currents of the transistors are as follows:

TR_1, TR_3, TR_4, TR_5	1mA
TR_2	2mA

The diode currents are:

D_1	13μA
D_2, D_3	5·3μA

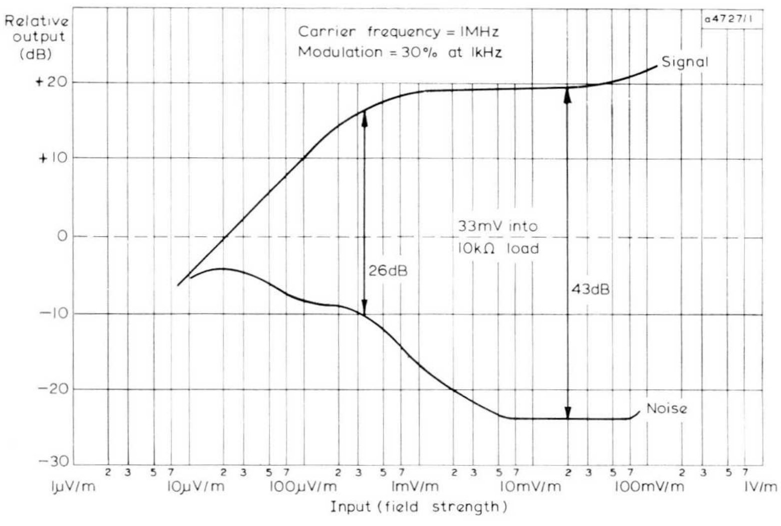
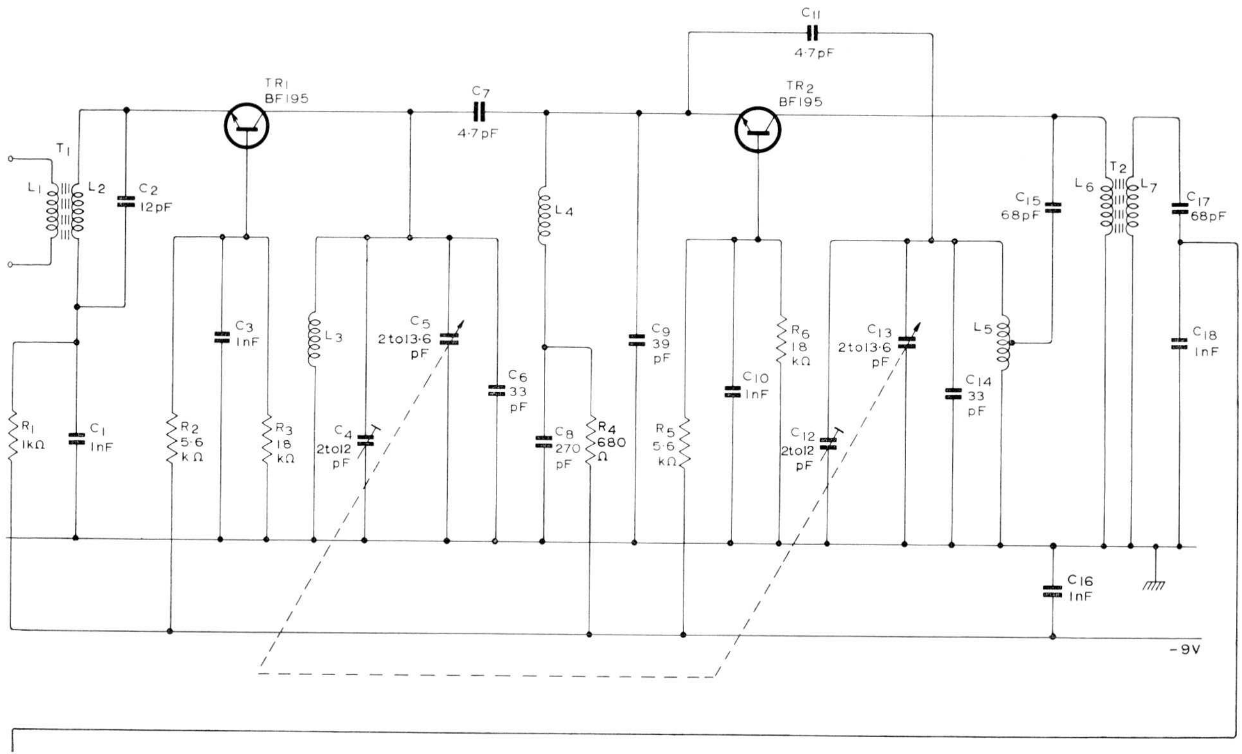
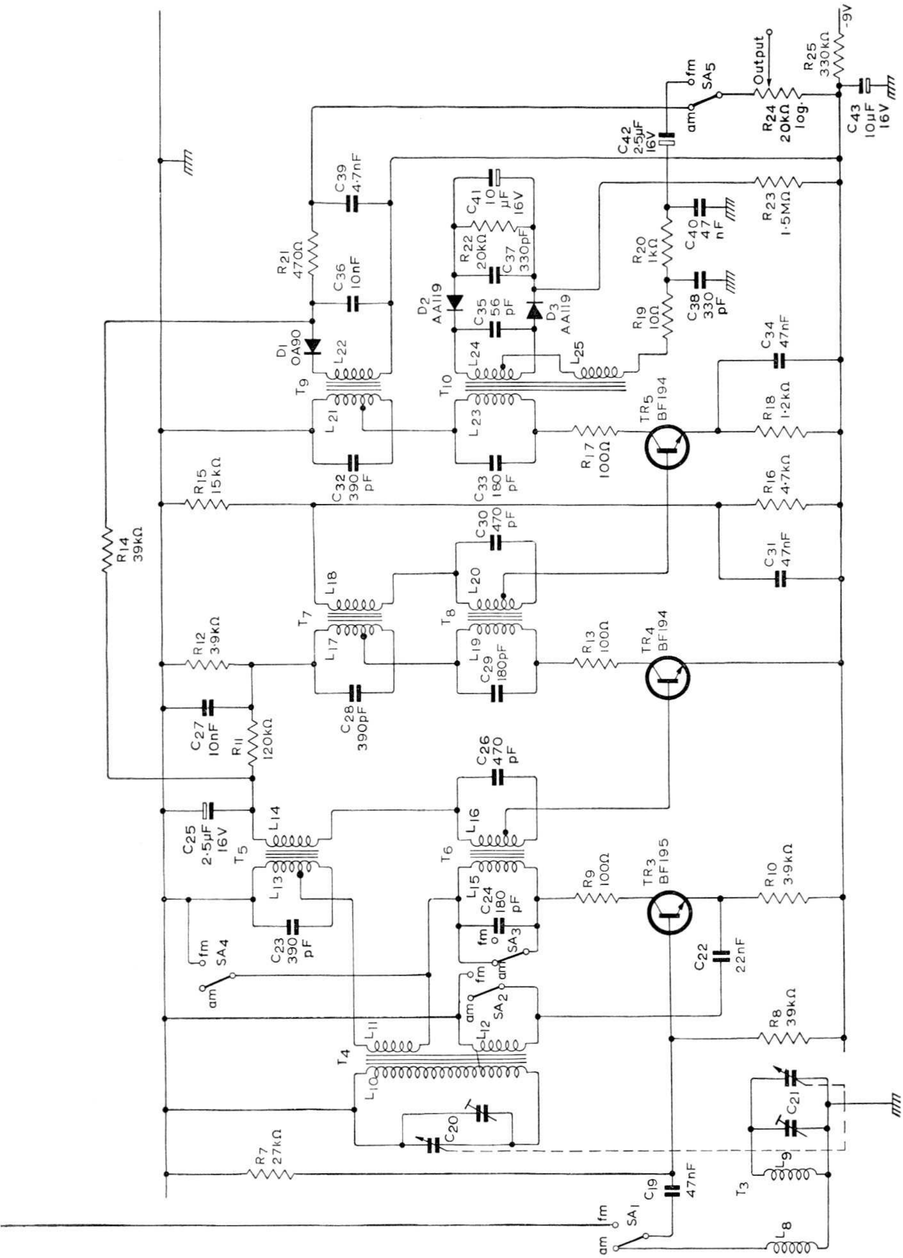


Fig 46—Automatic gain control and signal-to-noise characteristic (a.m.)

Fig. 47—F.M. tuner, r.f. and i.f. stages of an a.m./f.m. receiver





-9V
R25 330kΩ
Output
R24 20kΩ log.
C43 10μF 16V
fm SA5 am
C42 2·5μF 16V
C39 4·7nF
R21 470Ω
C36 10nF
D1 OA90
L22
T9
L21
C32 390 pF
R15 15kΩ
R14 39kΩ
C41 10 μF 16V
R22 20kΩ
C37 330pF
D2 AA119
C35 56 pF
D3 AA119
L24
L25
T10
L23
C33 180 pF
R23 1·5MΩ
R20 1kΩ
C40 47 nF
C38 330 pF
R19 10Ω
C34 47nF
R18 1·2kΩ
R17 100Ω
TR5 BF194
R16 4·7kΩ
C30 470 pF
C31 47nF
L18
L20
T7
T8
L17
L19
C29 180pF
C28 390pF
R12 3·9kΩ
R13 100Ω
TR4 BF194
C27 10nF
R11 120kΩ
C26 470 pF
C25 2·5μF 16V
L14
L16
T5
T6
L13
L15
C24 180 pF
C23 390 pF
R9 100Ω
TR3 BF195
R10 3·9kΩ
fm SA4 am
fm SA3 am
fm am SA2
C22 22nF
L11
L12
R8 39kΩ
T4
L10
C20
C21
L9
T3
R7 27kΩ
C19 47nF
fm SA1 am
L8

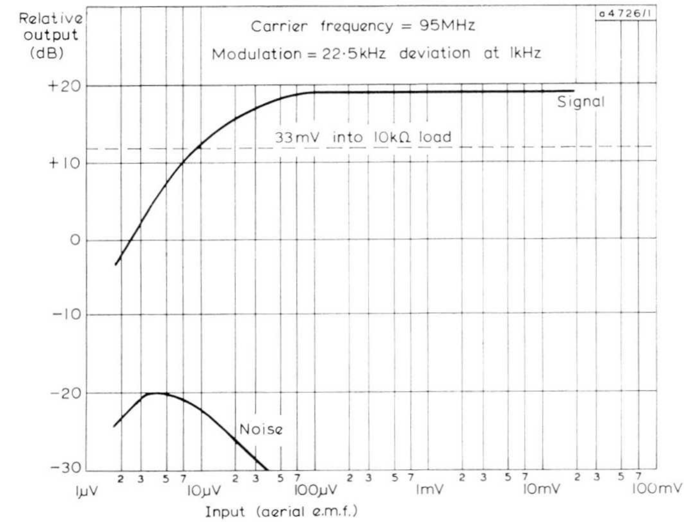


Fig. 48.—Limiting and signal-to-noise characteristic (f.m.)

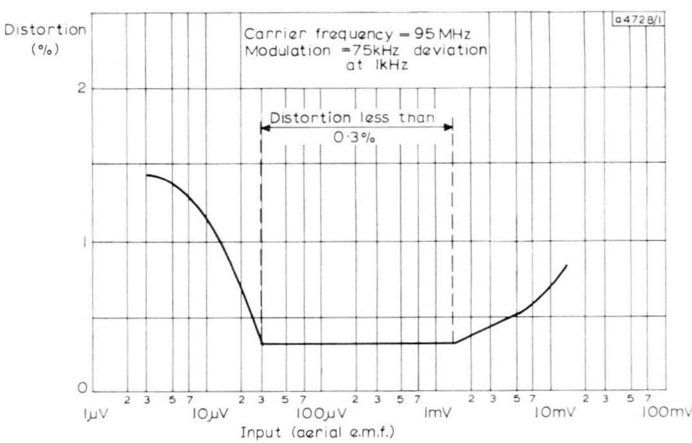


Fig. 49—Variation of audio distortion with input voltage (f.m.)

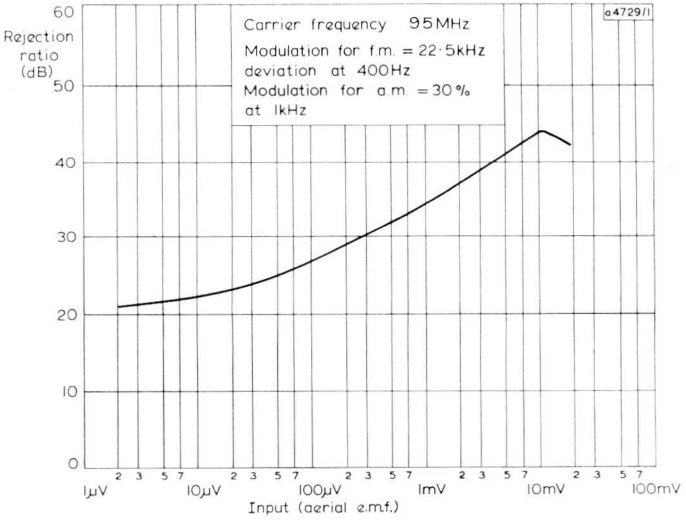


Fig. 50—A.M. rejection ratio

COMPONENT LISTS FOR CIRCUITS IN CHAPTER 4

1W Audio Amplifier (page 35)

Transistors

Circuit reference	Mullard type
TR_1	AC127
TR_2	AC128
TR_3	AC128
TR_4	AC127

Resistors

Tolerance: ±5%
Power rating: $\frac{1}{8}$W

Circuit reference	Value
R_1	2·2kΩ
R_2	15kΩ
R_3	15kΩ
R_4	2·7Ω
R_5	2·2kΩ
R_6	1·5kΩ
R_7	510Ω
R_8 thermistor type VA1040	130Ω
R_9 preset potentiometer	200Ω
R_{10}	39Ω

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	10μF	electrolytic 16V	C426AR/E10
C_2	125μF	electrolytic 10V	C426AR/D125
C_3	400μF	electrolytic 6·4V	C437AR/D400
C_4	4·7nF	polyester	C296AC/A4K7
C_5	320μF	electrolytic 2·5V	C426AR/A320
C_6	320μF	electrolytic 6·4V	C426AR/C320

Heatsinks

For output transistor TR_3 (AC128) a cooling clip is required. For output transistor TR_4 (AC127) a cooling clip mounted on a 1·5mm (16 gauge) bright aluminium heatsink with an area of 12·5cm^2 is required.

3W Audio Amplifier (page 37)

Transistors and Rectifier

Circuit reference	Mullard type
TR_1	BC148 or BC108
TR_2	AC128
TR_3	AC176
TR_4	AC128
Full-wave bridge rectifier	BY164

Resistors

Circuit reference	Value	Tolerance (%)	Power rating (W)
R_1	39kΩ	±5	$\frac{1}{8}$
R_2	120kΩ	±5	$\frac{1}{8}$
R_3	47kΩ	±5	$\frac{1}{8}$
R_4	390Ω	±10	$\frac{1}{8}$
R_5	22Ω	±10	$\frac{1}{8}$
R_6	2·2kΩ	±10	$\frac{1}{8}$
R_7	15Ω	±5	$\frac{1}{8}$
R_8	560Ω	±5	$\frac{1}{2}$
R_9	2·2Ω	±10	$\frac{1}{2}$
R_{10}	2·2Ω	±10	$\frac{1}{2}$

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	220nF	polyester	C296AA/A220K
C_2	25μF	electrolytic 25V	C426AR/F25
C_3	250μF	electrolytic 16V	C437AR/E250
C_4	3·3nF	polyester	C296AC/A3K3
C_5	400μF	electrolytic 16V	C437AR/E400
C_6 (mono)	1250μF	electrolytic 25V	C431BR/F1250
C_6 (stereo)	2000μF	electrolytic 25V	C431BR/F2000

Mains Transformer

The transformer used in this circuit may be obtained from the following manufacturers under the type numbers given:

Colne Electric Limited, Rickmansworth, Herts	20040
Drake Transformers Limited, Billericay, Essex	1471-136
Gardners Transformers Limited, Christchurch, Hants	GR 97180
Parmeko Limited, Aylestone Park, Leicester	P 3201

Heatsinks

The driver and output transistors should be mounted in clips having good thermal contact and fixed to heatsinks of 1·5mm (16 gauge) bright aluminium. The area required for the driver is 12cm^2 and for each output transistor 35cm^2.

10W Audio Amplifier (page 39)

Transistors and Rectifier

Circuit reference	Mullard type
TR_1	BC148 or BC108
TR_2	AC128
TR_3	AD161
TR_4	AD162
Rectifier bridge	$4 \times$ BY126

Resistors

Tolerance: $\pm 5\%$

Power rating: $\frac{1}{8}$W except R_{10}, which should be a 1W component, R_{11} and R_{12}, which should be $\frac{1}{2}$W components

Circuit reference	Value	Circuit reference		Value
R_1	15kΩ	R_7		12Ω
R_2	68kΩ	R_8	thermistor type VA1100	15Ω
R_3	100kΩ	R_9	preset potentiometer	20Ω
R_4	10Ω	R_{10}		220Ω
R_5	1kΩ	R_{11}		0·5Ω
R_6	1·2kΩ	R_{12}		0·5Ω

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	25μF	electrolytic 25V	C426AR/F25
C_2	5μF	electrolytic 16V	C426AR/H5
C_3	400μF	electrolytic 16V	C437AR/E400
C_4	500μF	electrolytic 2·5V	C426AR/A500
C_5	2000μF	electrolytic 16V	C431BR/E2000
C_6	4000μF	electrolytic 25V	C431BR/F4000

Mains Transformer

The transformer used in this circuit may be obtained from the following manufacturers under the type numbers given:

Colne Electric Limited, Rickmansworth, Herts	20041
Drake Transformers Limited, Billericay, Essex	294-139
Gardners Transformers Limited, Christchurch, Hants	GR 97181
Parmeko Limited, Aylestone Park, Leicester	P 3202

Heatsinks

Transistors TR_2, TR_3 and TR_4 should be mounted on a common 1·5mm (16 gauge) bright aluminium heatsink with a minimum area of 100cm^2 or blackened aluminium with an area of 60cm^2. Transistor TR_2 has an isolated can, but TR_3 and TR_4 should both be mounted on mica washers.

Pre-amplifier (page 42)

Transistors

Circuit reference	Mullard type
TR_1	BC149 or BC109
TR_2	BC148 or BC108

Resistors

Tolerance: $\pm 5\%$ except R_2 and R_4, which may be $\pm 10\%$ components
Power rating: $\frac{1}{8}$W

Circuit reference		Value
R_1		see text
R_2		1·8MΩ
R_3		270kΩ
R_4		2MΩ
R_5		47kΩ
R_6		47kΩ
R_7	linear potentiometer	500kΩ
R_8	linear potentiometer	250kΩ
R_9		47kΩ
R_{10}		150kΩ
R_{11}		150kΩ
R_{12}		47kΩ
R_{13}		22kΩ
R_{14}		2·2kΩ
R_{15}		15kΩ
R_{16}	linear potentiometer	50kΩ
R_{17}	logarithmic potentiometer	20kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	10nF	polyester	C296AA/A10K
C_2	10μF	electrolytic 16V	C426AR/E10
C_3	40μF	electrolytic 16V	C426AR/E40
C_4	1·5nF	polyester	C296AC/A1K5
C_5	6·8nF	polyester	C296AC/A6K8
C_6	6·8nF	polyester	C296AC/A6K8
C_7	1·5nF	polyester	C296AC/A1K5
C_8	10μF	electrolytic 16V	C426AR/E10
C_9	10μF	electrolytic 16V	C426AR/E10
C_{10}	10μF	electrolytic 16V	C426AR/E10
C_{11}	10pF	polystyrene	
C_{12}	see text		

3W Record-player Amplifier

Components for Basic Version (page 44)

Transistors and Rectifier

Circuit reference	Mullard type
TR_1	BC108
TR_2	BC107
TR_3	AC128
TR_4	AC176
Full-wave bridge rectifier	BY164

Resistors

Circuit reference	Value	Tolerance (%)	Power rating (W)
R_1 logarithmic potentiometer	2MΩ		
R_2	120kΩ	±5	$\frac{1}{8}$
R_3	470kΩ	±5	$\frac{1}{8}$
R_4	390kΩ	±5	$\frac{1}{8}$
R_5	15Ω	±5	$\frac{1}{8}$
R_6	470Ω	±5	$\frac{1}{2}$
R_7	150Ω	±5	$\frac{1}{2}$
R_8	2·2Ω	±10	$\frac{1}{4}$
R_9	2·2Ω	±10	$\frac{1}{4}$
R_{10} logarithmic potentiometer	250kΩ		
R_{11}	1·2kΩ	±10	$\frac{1}{8}$
R_{12}	15kΩ	±10	$\frac{1}{8}$

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	100nF	polyester	C296AA/A100K
C_2	100nF	polyester	C296AA/A100K
C_3	40μF	electrolytic 16V	C426AR/E40
C_4	10nF	polyester	C296AA/A10K
C_5	680pF	polystyrene	C295AC/680E
C_6	250μF	electrolytic 16V	C437AR/E250
C_7	640μF	electrolytic 25V	C437AR/F640

Mains Transformer

The transformer used in this circuit may be obtained from the following manufacturers under the type numbers given:

Colne Electric Limited, Rickmansworth, Herts	20042
Drake Transformers Limited, Billericay, Essex	351-245
Gardners Transformers Limited, Christchurch, Hants	GR 97182
Parmeko Limited, Aylestone Park, Leicester	P 3203

Components for Version including Comprehensive Tone Control (page 47)

Transistors and Rectifier

Circuit reference	Mullard type
TR_1	BC109
TR_2	BC107
TR_3	AC128
TR_4	AC176
Full-wave bridge rectifier	BY164

Resistors

Circuit reference		Value	Tolerance (%)	Power rating (W)
R_1	logarithmic potentiometer	2MΩ		
R_2		220kΩ	±5	$\frac{1}{8}$
R_3		1MΩ	±5	$\frac{1}{8}$
R_4		820kΩ	±5	$\frac{1}{8}$
R_5		15Ω	±5	$\frac{1}{8}$
R_6		470Ω	±5	$\frac{1}{2}$
R_7		150Ω	±5	$\frac{1}{2}$
R_8		2·2Ω	±10	$\frac{1}{4}$
R_9		2·2Ω	±10	$\frac{1}{4}$
R_{10}	reverse logarithmic potentiometer	100kΩ		
R_{11}		1·2kΩ	±10	$\frac{1}{8}$
R_{12}		15kΩ	±10	$\frac{1}{8}$
R_{13}	reverse logarithmic potentiometer	100kΩ		

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	2·5μF	electrolytic 1·5V	
C_2	100nF	polyester	C296AA/A100K
C_3	40μF	electrolytic 16V	C426AR/E40
C_4	4·7nF	polyester	C296AC/A4K7
C_5	680pF	polystyrene	C295AC/680E
C_6	250μF	electrolytic 16V	C437AR/E250
C_7	640μF	electrolytic 25V	C437AR/F640
C_8	47nF	polyester	C296AA/A47K
C_9	6·4μF	electrolytic 25V	C426AR/F6·4
C_{10}	22nF	polyester	C296AA/A22K
C_{11}	220nF	polyester	C296AA/A220K

Mains Transformer

For suppliers of the transformer, see page 62.

Heatsinks

Each output transistor should be mounted on a 1·5mm (16 gauge) bright aluminium heatsink with an area of 30cm^2. In the circuit shown the driver transistor does not require a heatsink. However, an increase in output power (3W at clipping) may be obtained, by fitting a heat clip on the driver transistor and reducing the value of R_6 to 330Ω to give a higher drive current, with a 12Ω load. (Alternatively, a larger driver transistor, the BFY51, may be used with the same reduction in the value of R_6 and a 12Ω load.)

A.M. Receiver (page 50)

Transistors and Diode

See note on page 49

Circuit reference	Mullard type
TR_1	BF194
TR_2	BF195
TR_3	BF195
D_1	OA90

Resistors

Tolerance: ±5%
Power rating: $\frac{1}{8}$W

Circuit reference	Value	Circuit reference	Value
R_1	18kΩ	R_5	330kΩ
R_2	15kΩ	R_6	120kΩ
R_3	2·7kΩ	R_7	27kΩ
R_4	47kΩ	R_8 logarithmic carbon potentiometer	10kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1 } C_5		foil dielectric variable capacitor	AC0039
C_2	10nF	polyester	C296AA/A10K
C_3	220pF	ceramic plate	C333CH/C220E
C_4	22nF	polyester	C296AA/A22K
C_6	1·6μF	electrolytic 25V	
C_7	220pF	ceramic plate	C333CH/C220E
C_8	100nF	polyester	C296AA/A100K
C_9	220pF	ceramic plate	C333CH/C220E
C_{10}	10nF	polyester	C296AA/A10K

Transformers

Aerial transformer—T_1

L_1 80 turns of 12×46 s.w.g. stranded wire; close-wound to fit Mullard 9·5mm diameter ferrite rod, type FX1247

L_2 $3\frac{1}{2}$ turns of 12×46 s.w.g. stranded wire; interwound with earth end of L_1

Oscillator transformer—T_2

L_3 10 turns of 3×46 s.w.g. stranded wire

L_4 2 turns of 3×46 s.w.g. stranded wire

L_5 100 turns of 3×46 s.w.g. stranded wire; wave-wound on 4·8mm former

First i.f. transformer—T_3

L_6+L_7 42+108 turns of 3×46 s.w.g. stranded wire

L_8 2 turns of 3×46 s.w.g. stranded wire; wave-wound on 4·8mm former; $Q_O \simeq Q_L = 80$ when L_6+L_7 is shunted with 330kΩ resistor

Second i.f. transformer—T_4

L_9+L_{10} 80+70 turns of 3×46 s.w.g. stranded wire

L_{11} 9 turns of 3×46 s.w.g. stranded wire; wound on 4·8mm former; $Q_O = 120$, $Q_L = 95$

Third i.f. transformer—T_5

$L_{12}+L_{13}$ 51+99 turns of 3×46 s.w.g. stranded wire

L_{14} 32 turns of 3×46 s.w.g. stranded wire; wound on 4·8mm former; $Q_O = 120$, $Q_L = 40$

A.M./F.M. Receiver (pages 54 and 55)

Transformers and Coils

Except where otherwise stated, all transformers and coils are wound on 4mm formers with ferrite frames and slugs.

F.M. aerial transformer—T_1

L_1+L_2	2+2 turns of 24 s.w.g. enamelled-copper wire; wound on ferrite twin bead

First f.m. i.f. transformer—T_2

L_6+L_7	14+14 turns of 36 s.w.g. enamelled-copper wire; wound on a 4·4mm diameter former L_7 is critically coupled to L_6 when L_7 is loaded with 600Ω at the capacitive tap

Medium-wave aerial transformer—T_3

L_8	80 turns of 12×46 s.w.g. silk-covered enamelled-copper wire; wound on Mullard ferrite rod, type FX1247 (203×9·5mm); $Q_O=200$
L_9	$3\frac{1}{2}$ turns of 12×46 s.w.g. silk-covered enamelled-copper wire; interwound with earth end of L_8

A.M. oscillator transformer—T_4

L_{10} (primary)	$119\frac{7}{8}$ turns of 6×48 s.w.g. enamelled-copper wire; $Q_O=145$
L_{11} (collector)	$9\frac{1}{2}$ turns of 6×48 s.w.g. enamelled-copper wire
L_{12} (emitter)	$1\frac{7}{8}$ turns of 6×48 s.w.g. enamelled-copper wire; wound on 'baby' coil former;

First a.m. i.f. transformer—T_5

L_{13} (primary)	130 turns of 3×46 s.w.g. enamelled-copper wire; tapped at 41 turns from the earth end; tap impedance=6·5kΩ; $Q_O=75$, $Q_L=70$
L_{14} (secondary)	$5\frac{1}{4}$ turns of 3×46 s.w.g. enamelled-copper wire

Turns ratio, primary tap: secondary=7·8:1

Second f.m. i.f. transformer—T_6

L_{15} (primary)	8 turns of 36 s.w.g. enamelled-copper wire; $Q_O=Q_L=53$
L_{16} (secondary)	6 turns of 36 s.w.g. enamelled-copper wire tapped at $1\frac{3}{4}$ turns from earth end; unloaded tap impedance=118Ω; $Q_O=53$, $Q_L=50$ L_{16} is critically coupled when secondary tap is loaded with 2·2kΩ

Turns ratio, primary:secondary tap=4·5:1

Second a.m. i.f. transformer—T_7

L_{17} (primary)	110 turns of 3×46 s.w.g. enamelled-copper wire tapped at 73 turns from earth end; tap impedance=35·6kΩ; $Q_O=98$, $Q_L=70$
L_{18} (secondary)	10 turns of 3×46 s.w.g. enamelled-copper wire

Turns ratio, primary tap:secondary=6·9:1

Third f.m. i.f. transformer—T_8

L_{19} (primary)	As L_{15} (T_6)
L_{20} (secondary)	As L_{16} (T_6)

Third a.m. i.f. transformer—T_9

L_{21} (primary)	110 turns of 3×46 s.w.g. enamelled-copper wire tapped at 48 turns from earth end; tap impedance = 21·6kΩ; Q_O = 140, Q_L = 50
L_{22} (secondary)	31 turns of 3×46 s.w.g. enamelled-copper wire

Turns ratio, primary tap:secondary = 1·53:1

F.M. ratio detector transformer—T_{10}

L_{23} (primary)	8 turns of 34 s.w.g. enamelled-copper wire; Q_O = 90, Q_L = 58
L_{24} (secondary)	2×7 turns of 36 s.w.g. enamelled-copper wire; Q_O = 75
L_{25} (tertiary)	3½ turns of 36 s.w.g. enamelled-copper wire; overwound on primary at earth end. Primary to secondary coupling is 0·7 with secondary loaded with 10kΩ and tertiary open-circuited.
Coil L_3	3½ turns of 20 s.w.g. tinned-copper wire; spaced one diameter and wound on a 6mm diameter former
Coil L_4	12 turns of 28 s.w.g. enamelled-copper wire; wound on a 4·4mm former with ferrite core
Coil L_5	3½ turns of 20 s.w.g. tinned-copper wire; spaced two diameters and wound on a 6mm diameter former; tapped two turns from the earth end

Transistors and Diodes

Circuit reference	Mullard type	Circuit reference	Mullard type
TR_1	BF195	D_1	OA90
TR_2	BF195	D_2	AA119
TR_3	BF195	D_3	AA119
TR_4	BF194		
TR_5	BF194		

Resistors

Tolerance: ±5% except R_{23}, which may be a ±10% component

Power rating: ⅛W

Circuit reference	Value	Circuit reference	Value
R_1	1kΩ	R_{14}	39kΩ
R_2	5·6kΩ	R_{15}	15kΩ
R_3	18kΩ	R_{16}	4·7kΩ
R_4	680Ω	R_{17}	100Ω
R_5	5·6kΩ	R_{18}	1·2kΩ
R_6	18kΩ	R_{19}	10Ω
R_7	27kΩ	R_{20}	1kΩ
R_8	39kΩ	R_{21}	470Ω
R_9	100Ω	R_{22}	20kΩ
R_{10}	3·9kΩ	R_{23}	1·5MΩ
R_{11}	120kΩ	R_{24} logarithmic potentiometer	20kΩ
R_{12}	3·9kΩ		
R_{13}	100Ω	R_{25}	330kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	1nF	polyester	C296AC/A1K
C_2	12pF	ceramic plate	C333CB/L12E
C_3	1nF	polyester	C296AC/A1K
C_4	2 to 12pF	trimmer	
C_5	2 to 13·6pF	ganged tuning capacitor	
C_{13}	2 to 13·6pF		
C_6	33pF	ceramic plate	C333CC/C33E
C_7	4·7pF	ceramic plate	C333CB/L4E7
C_8	270pF	polystyrene	
C_9	39pF	ceramic plate	C333CC/C39E
C_{10}	1nF	polyester	C296AC/A1K
C_{11}	4·7pF	ceramic plate	C333CB/L4E7
C_{12}	2 to 12pF	trimmer	
C_{14}	33pF	ceramic plate	C333CC/C33E
C_{15}	68pF	ceramic plate	C333CH/C68E
C_{16}	1nF	polyester	C296AC/A1K
C_{17}	68pF	ceramic plate	C333CH/C68E
C_{18}	1nF	polyester	C296AC/A1K
C_{19}	47nF	metallised foil	C280AE/P47K
C_{20}, C_{21}	ganged tuning capacitor, Plessey 'W' type		
C_{22}	22nF	metallised foil	C280AE/P22K
C_{23}	390pF	polystyrene	
C_{24}	180pF	ceramic plate	C333CH/C180E
C_{25}	2·5μF	electrolytic 16V	C426AR/H2·5
C_{26}	470pF	polystyrene	
C_{27}	10nF	metallised foil	C280AE/P10K
C_{28}	390pF	polystyrene	
C_{29}	180pF	ceramic plate	C333CH/C180E
C_{30}	470pF	polystyrene	
C_{31}	47nF	metallised foil	C280AE/P47K
C_{32}	390pF	polystyrene	
C_{33}	180pF	ceramic plate	C333CH/C180E
C_{34}	47nF	metallised foil	C280AE/P47K
C_{35}	56pF	ceramic plate	C333CH/C56E
C_{36}	10nF	metallised foil	C280AE/P10K
C_{37}	330pF	polystyrene	
C_{38}	330pF	polystyrene	
C_{39}	4·7nF	polyester	C296AC/A4K7
C_{40}	47nF	metallised foil	C280AE/P47K
C_{41}	10μF	electrolytic 16V	C426AR/E10
C_{42}	2·5μF	electrolytic 16V	C426AR/H2·5
C_{43}	10μF	electrolytic 16V	C426AR/E10

CHAPTER 5

TAPE RECORDERS

In this chapter, a 4W tape recorder amplifier and a tape pre-amplifier are described, and the modifications necessary for stereo operation are considered. An automatic gain control circuit for use with tape recorders is also included.

The amplifier and pre-amplifier circuits are designed around the Marriott type X/RPS/36 recording head and X/ES/11 erase head; they will not operate with heads of different inductance and bias requirements.

Component lists for the circuits are given at the end of the chapter.

Record level meter 500 Ω, 100 μA for maximum record level

Record/playback switch S_B in record position

Position of equalisation switch S_C	Tape speed (in/s)
e	$7\frac{1}{2}$
f	$3\frac{3}{4}$
g	$1\frac{7}{8}$
h	$\frac{15}{16}$

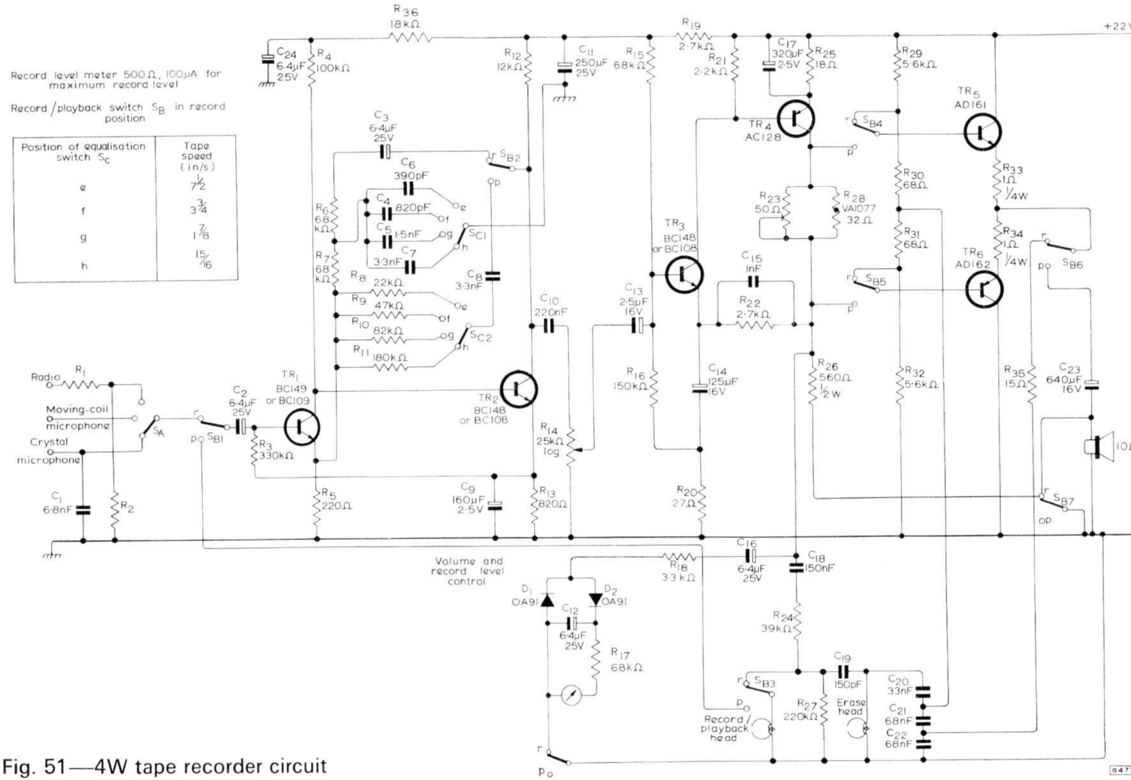


Fig. 51—4W tape recorder circuit

4W TAPE RECORDER

The 4W tape recorder shown in Fig. 51 uses silicon transistors in the first three stages to obtain high sensitivity. The output stage uses a complementary matched pair, transistors AD161 (n-p-n) and AD162 (p-n-p), which are also used for the oscillator during recording. Equalisation is provided for tape speeds of $7\frac{1}{2}$in/s (19·1cm/s), $3\frac{3}{4}$in/s (9·5cm/s), $1\frac{7}{8}$in/s (4·8cm/s) and $\frac{15}{16}$in/s (2·4cm/s), which conforms to C.C.I.R. standards where appropriate.

Circuit Description

Playback

The BC149 or BC109 has been chosen for the input stage because it is particularly suitable for use in audio input stages where low noise is essential. Its collector is directly connected to the base of TR_2; its emitter is connected via an RC network to the collector of TR_2. This RC network provides a path for negative feedback by means of which equalisation is effected. The frequency response curves on playback are shown in Fig. 53.

A d.c. feedback path is also provided from the emitter of TR_2 to the base of TR_1. The d.c. feedback along this path stabilises the working points of TR_1 and TR_2.

The other transistors in the tape recorder form a conventional amplifier. The complementary pair TR_5 and TR_6, is driven by TR_4, which is itself driven by the high-gain BC148 or BC108 transistor. Potentiometer R_{23} is adjusted to give a quiescent output-stage current of 5mA.

Record

Different input circuits are needed to match the different signal sources. For example, the input circuit for a crystal microphone contains a shunt capacitor, C_1, and that for a radio receiver is a potential divider. A moving coil microphone, however, can be connected directly because the amplifier has an input resistance of 200kΩ. During recording, TR_5 and TR_6 operate as an oscillator at 50kHz, the erase head being used as an oscillator coil. The capacitive taps, C_{20}, C_{21} and C_{22}, also provide the necessary conditions for sustained oscillation.

The recording level indicator gives a reading proportional to the peak-to-peak voltage at the collector of TR_4. The full recording level with a sine-wave is 4V r.m.s., which produces a current of 95μA through the meter and 110μA r.m.s. through the recording head.

During recording, the equalisation circuits connected to switch S_C are not used; feedback is via the path C_3, R_6 and R_7, which has been designed to give treble boost. The frequency response curves on record are shown in Fig. 54.

The recording current is taken from the collector of TR_4, and the recording head is a Marriott type X/RPS/36* with the following characteristics:

Quarter-track head:	
Gap	0·0001in
Width	0·043in
Spacing	0·136in
Inductance at 1kHz	70mH
Bias at 50kHz	$>$11V
Recording current for 2dB below 2% distortion	110μA
Output at 1kHz using E.M.I. test tape T.B.T.I.	0·44mV
Output at 2kHz for 110μA recording current	1·07mV

Power Supply

The circuit of a simple power supply suitable for use with the tape recorder is shown in Fig. 52. It contains a mains transformer with a secondary output of 17V. This is rectified by means of a bridge rectifier, type BY164, and smoothed by means of a 640μF capacitor.

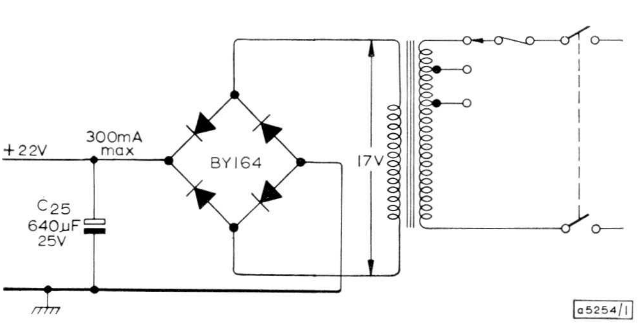


Fig. 52—Power supply circuit for a 4W tape recorder

*Record/playback heads with a larger inductance are not suitable because they would require a larger bias voltage.

Performance

Overall frequency response

tape speed = $7\frac{1}{2}$in/s	55Hz to 20kHz
tape speed = $3\frac{3}{4}$in/s	55Hz to 12kHz
tape speed = $1\frac{7}{8}$in/s	55Hz to 6kHz

Playback

Output power	4W r.m.s
Sensitivity at base of TR_1, at 1kHz and with tape speed of $7\frac{1}{2}$in/s	400μV
Distortion for 4W output	1·2%

Noise relative to 4W, with volume control set so that a signal produced with 110μA recording current gives 4W output; 3dB bandwidth = 30Hz to 15kHz:

tape speed = $7\frac{1}{2}$in/s	—60dB
tape speed = $3\frac{3}{4}$in/s	—57dB
tape speed = $1\frac{7}{8}$in/s	—55dB
tape speed = $\frac{15}{16}$in/s	—50dB

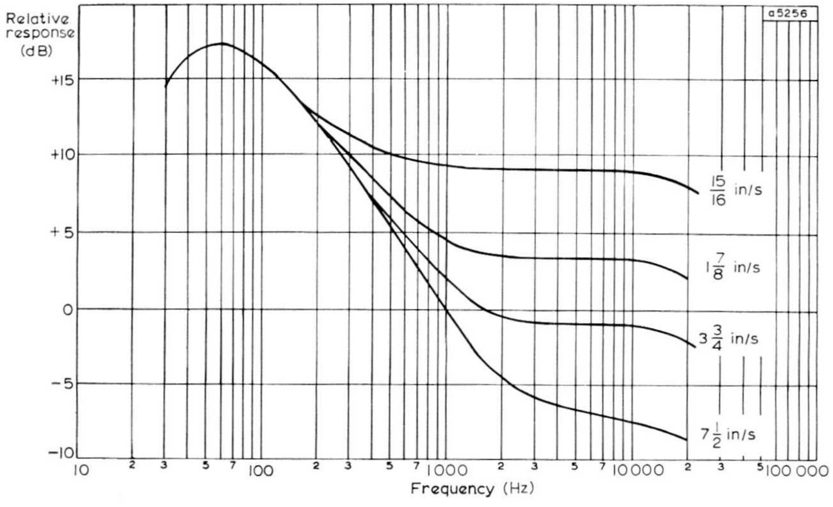


Fig. 53—Playback characteristic

Record

Sensitivity at base of TR_1 for $110\mu A$ recording current, with volume control at maximum setting	$100\mu V$
Distortion at full record level	1%
Recording current level for 4W output, with volume control at maximum setting	$40\mu A$

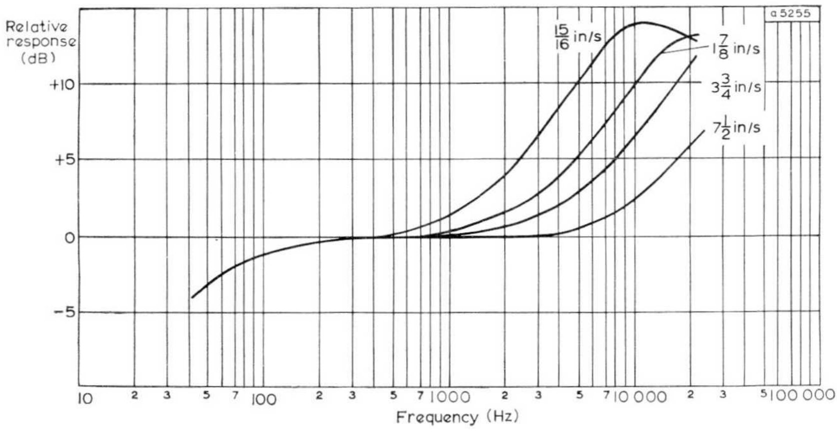


Fig. 54—Record characteristic

TAPE PRE-AMPLIFIER

Circuit Description

Fig. 56 shows a pre-amplifier circuit derived from the tape recorder circuit already described. The oscillator stage in this circuit uses a complementary pair comprising one AC176 and one AC128 transistor.

Performance

The distortion measured at the collector of transistor TR_4 (AC128) in the playback position is shown in Fig. 55. The measurements were made at a speed of $3\frac{3}{4}$in/s, although these characteristics vary little with speed.

At a speed of $7\frac{1}{2}$in/s with the volume control at maximum, the output voltage is 6·55V for an input voltage of 400μV.

The total current drain on playback is 25mA.

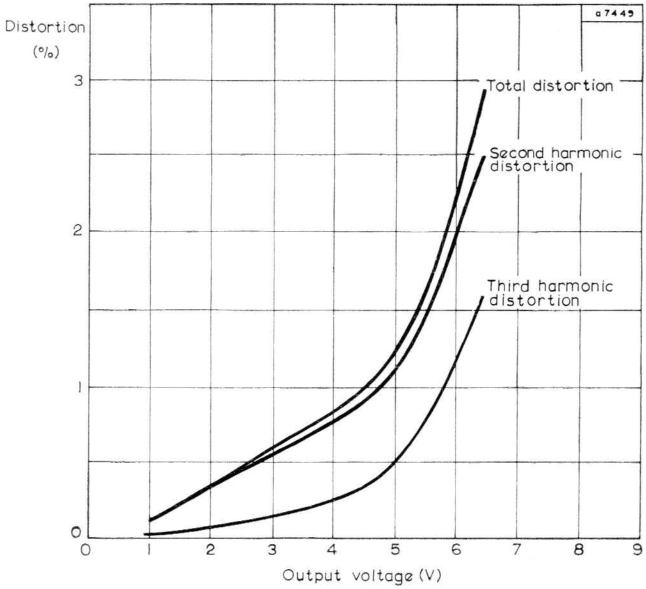


Fig. 55—Variation of harmonic distortion with output voltage for pre-amplifier circuit

Record level meter 500Ω, 100μA for maximum record level

Record/playback switch S_B in record position

Position of equalisation switch S_c	Tape speed (in/s)
e	$7\frac{1}{2}$
f	$3\frac{3}{4}$
g	$1\frac{7}{8}$
h	$\frac{15}{16}$

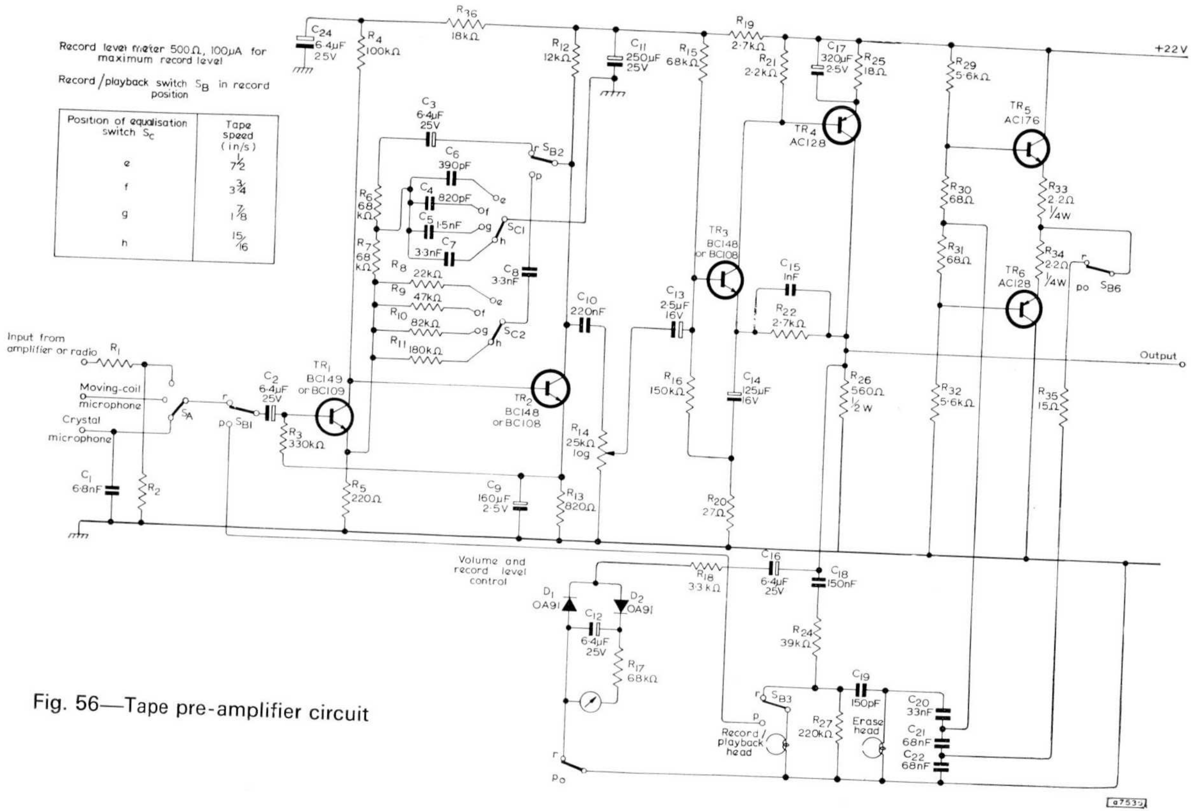


Fig. 56—Tape pre-amplifier circuit

MODIFICATIONS FOR STEREO OPERATION

In a stereo recording system, the bias and erase frequencies of the two channels must be equal, if the production of audio beat frequencies is to be prevented. Therefore, the bias and erase power for the two channels must be derived from one common oscillator, or from two oscillators which have their frequencies locked together. The former arrangement (Fig. 57) can be used for the pre-amplifier of Fig. 56 only if the AC176/AC128 pair is replaced by an AD161/AD162 pair and the emitter resistors reduced accordingly. In the latter arrangement (Fig. 58), two AC176/AC128 pairs will operate satisfactorily.

Bias and erase power from a common oscillator

A circuit that uses a common oscillator for stereo operation is given in Fig. 57. This shows that the two tuned erase heads, which form the load and feedback circuits of one of the complementary pairs of transistors, are in parallel. The bias for the two recording heads, however, is still

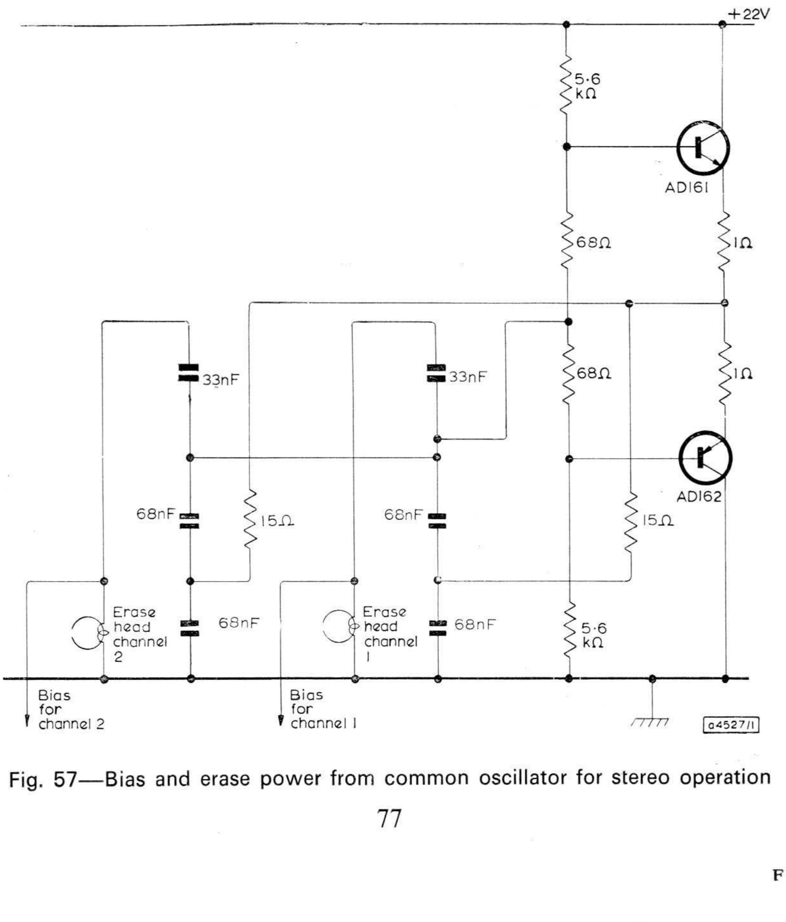


Fig. 57—Bias and erase power from common oscillator for stereo operation

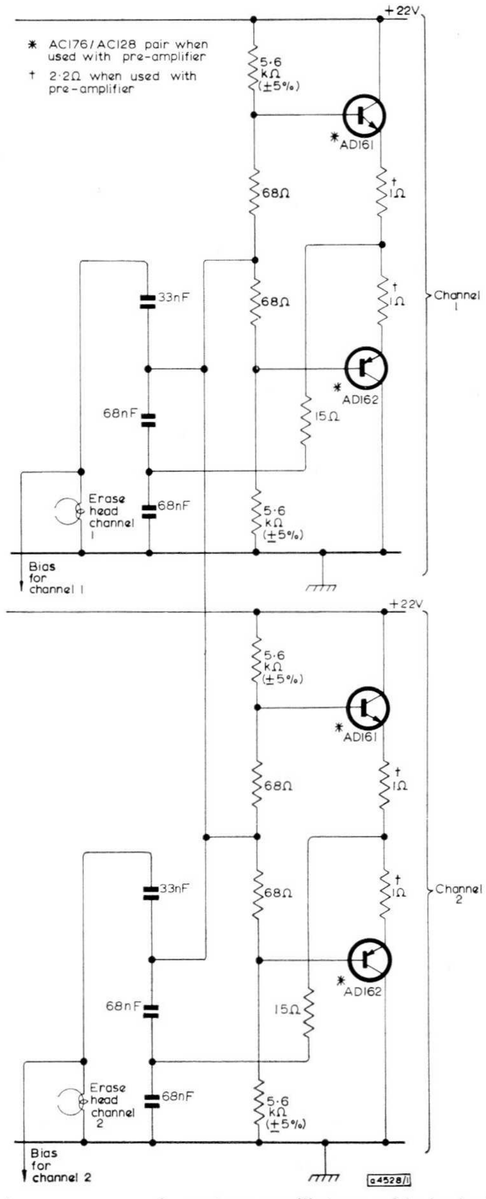


Fig. 58—Bias and erase power from two oscillators with locked frequencies for stereo operation

derived from the erase voltages across the erase heads as in the mono design. This has the advantage that the other pair of complementary transistors used for the output stage during playback may now be used as a monitoring facility.

Bias and erase power from two oscillators with locked frequencies

A circuit that uses two locked identical oscillators for stereo operation is given in Fig. 58. Locking of the oscillators is achieved by connecting together the two feedback paths.

Recording level indicator

Fig. 59 shows a recording level indicator for stereo operation. The meter gives a reading proportional to the peak-to-peak voltage at the collector of the driver transistor in the channel which has the highest recording level. For an input of 4V r.m.s. (this corresponds to the r.m.s. voltage at the collector of the driver transistor when distortion due to tape saturation becomes significant) the d.c. current through the meter is 95μA.

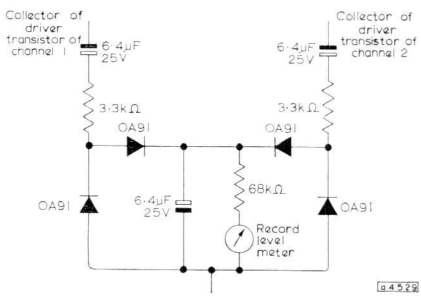


Fig. 59—Recording level indicator for stereo operation

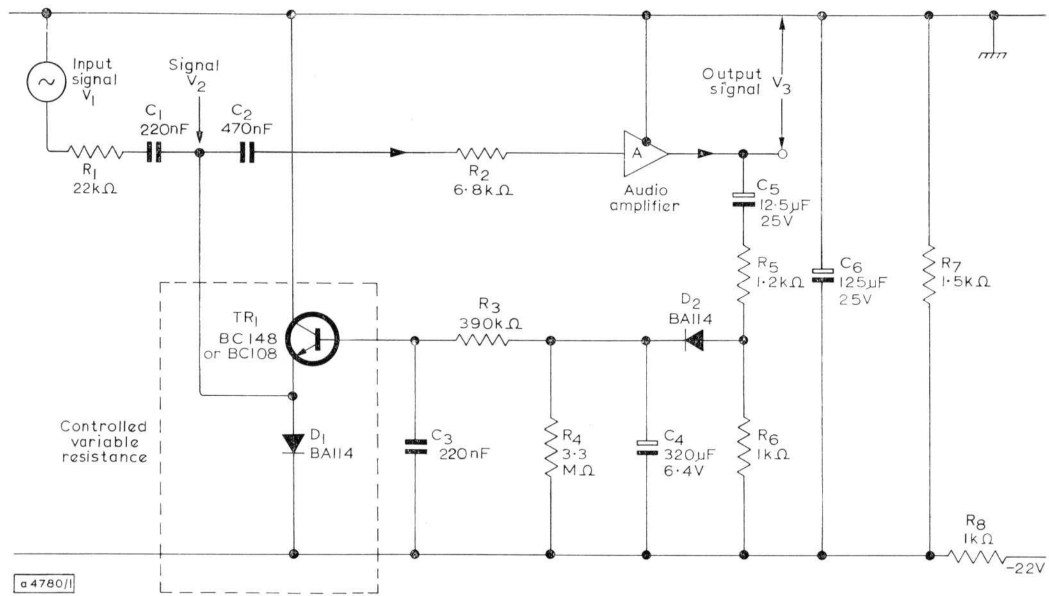


Fig. 60—Automatic gain control circuit

A.G.C. CIRCUIT FOR TAPE RECORDERS

When recording on tape, the current in the recording head must not exceed the value at which tape saturation occurs if severe distortion is to be avoided. This can be prevented by monitoring the current by means of a meter or magic eye, and manually adjusting the recording level. Another method is to use a circuit that electronically reduces the gain of the amplifier when the optimum recording level is exceeded.

To ensure that the distortion is recorded for only a short time, the gain control must respond very rapidly to the overload signal. When the gain has been reduced, however, the recovery time must be long, for otherwise volume compression will result, and loud and soft passages of music will be recorded with equal loudness.

A gain control circuit that uses a minimum of components and can produce an attenuation of 40dB within 150 milliseconds is shown in Fig. 60. The recovery rate does not exceed 6dB per minute. With a faster recovery rate, the same circuit arrangement is suitable as a volume compressor in other applications.

Circuit Description

Gain control is achieved by means of an attenuator containing a diode. Controlling the direct current through the diode controls its slope resistance, and, consequently, the amount of attenuation.

The severe harmonic distortion caused by the exponential characteristic of one diode can be reduced by using two diodes in a reverse-parallel arrangement. In the circuit in Fig. 60, the base-emitter junction of a transistor is used as one of the diodes. Because the transistor is needed as a d.c. amplifier, this arrangement saves the expense of one diode.

Part of the amplifier output is rectified and used to provide the gain control signal. The time constants of the rectifier circuit are such as to give the high rate of attenuation and the long recovery time.

In Fig. 60 unit A is the audio amplifier and R_2 represents its input resistance, which should be about 7kΩ.

The variable attenuator comprises R_1, C_1, D_1 and part of the transistor TR_1. Because C_3 is a short circuit to a.c. signals, the base-emitter junction of the transistor is, in effect, in reverse-parallel with D_1 for all a.c. signals.

Resistors R_5 and R_6 form a fixed attenuator by means of which a constant fraction of the output is fed back to the control circuit. The feedback voltage is rectified by D_2, charges C_4, and is applied through R_3 to the base of TR_1. Hence, the voltage across C_4 controls the current through TR_1 and D_1.

With an input signal well below that which causes tape saturation, the voltage across C_4 is low. Therefore, the transistor current is also low and the slope resistances of it and the diode are high. Consequently, the shunting effect that the diode and transistor have on the input to the amplifier is negligible.

As the input signal is increased, the voltage across C_4 and the current through TR_1, will also increase. Eventually, the transistor current will reach a value which produces a significant reduction in the slope resistance of D_1. The voltage drop across R_1 will be increased and there will be an appreciable attenuation of the signal to the amplifier. This occurs when the amplifier has an output of approximately 2V r.m.s.

A further increase in the input produces another rise in the voltage across C_4. The consequent increase in transistor current now produces a rapid reduction in the diode slope resistance. The shunting effect of the transistor and diode on the input to the amplifier is greatly increased, and V_2 is much less than V_1.

Once attenuation has started, voltages V_2 and V_3 remain practically constant. Over the full range of attenuation, 40dB, they increase slightly because of slight increases in V_{BE}, the forward voltage of D_1, at higher currents (of the order of 300μA), and the base current through R_3.

The attack time is largely determined by the time constant of C_4 and the source impedance of the charging circuit, which is R_5 in parallel with R_6. For an attack time of about 200ms, the source impedance should be approximately 600Ω.

The recovery time is determined by the time constant of C_4 and its discharge circuit—R_4 in parallel with R_3 and the slope resistances of TR_1 and D_1. Over a wide range of the attenuation, R_3 is much larger than the slope resistances of the diode and transistor. The rate of discharge of C_4 is then controlled by the value of R_3, and, consequently, the rate of change of attenuation is constant. At very low transistor currents, the slope resistances of the diode and the transistor greatly exceed R_3, and the rate of discharge of C_4 via the transistor is very low. This tends to make the rate of change of attenuation also very low. Under these conditions, however, R_4 provides an alternative discharge path. Hence, the rate of change of attenuation has a low limit that is determined by the value of R_4. When R_3 is large with respect to the slope resistances of the diode and the transistor, the effect of R_4 is negligible.

The transistor is a BC148 or BC108 n-p-n planar type, which, with its low leakage current and high gain, makes this simple arrangement possible. The silicon diodes also are used because of their low leakage currents.

In Fig. 60, the gain is controlled by attenuating the signal before it reaches the audio amplifier. This has the disadvantage that, when the signal is attenuated, the noise originating in the first transistor of the amplifier is unaffected. A better arrangement would be to have the attenuation after the first stage of the amplifier, thus reducing noise and signal simultaneously. This arrangement is shown in Fig. 61. A high voltage is required to operate the attenuating circuit; this will probably be available only at the output of the stage feeding the recording head.

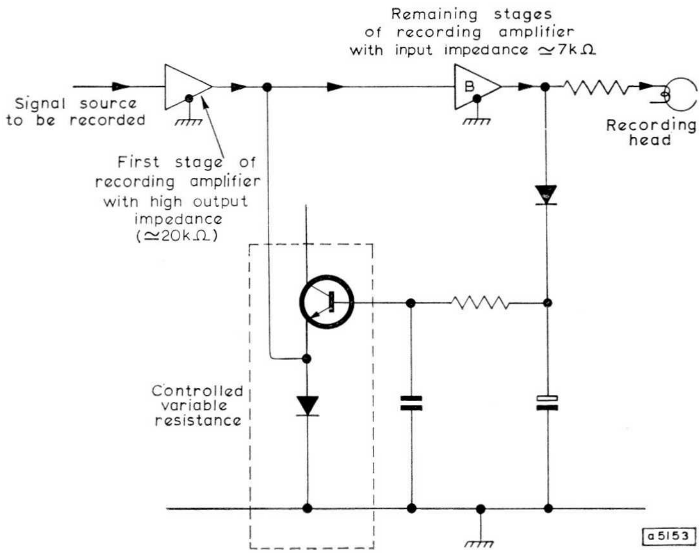


Fig. 61—Incorporation of gain control circuit in a tape amplifier

Use of Electrolytic Capacitors in Timing Circuits

The maximum leakage currents quoted in published data on electrolytic capacitors are too high for use in long time-constant circuits. Experience suggests, however, that the leakage currents of capacitors taken off the shelf will be low enough or will soon fall to a suitably low value. Therefore, electrolytic capacitors of a suitable type are readily available. At the worst, a tape recorder which has been stored for a very long time will have an initial recovery time slightly shorter than normal, but this will rapidly approach the normal value.

Performance

Where the circuit controls the gain, the signal V_2 is not maintained exactly constant nor completely independent of the amount of attenuation.

This is shown in Fig. 62 by the input/output characteristics of the circuit with high-gain and low-gain transistors. When a low-gain transistor is used, an attenuation of 40dB is accompanied by an increase of 7dB in the output; with a high-gain transistor, the corresponding increase in output is 4·5dB.

The change of attenuation with time is shown in Fig. 63. This diagram shows that the rate of change of attenuation is substantially constant over much of the range. It does not exceed 6dB per minute and is almost independent of the gain of the transistor.

The variation of total harmonic distortion of V_2 with the amount of attenuation for various values of V_2 is given in Fig. 64. These curves show that V_2 should not exceed 4mV r.m.s. if the total harmonic distortion is not to exceed 2%.

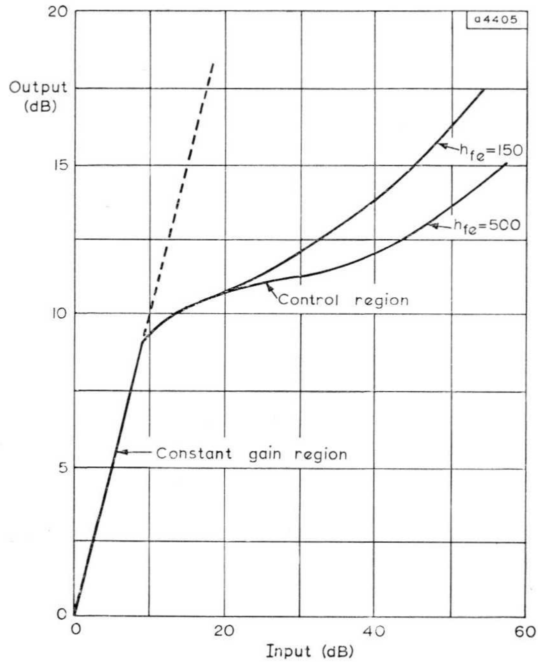


Fig. 62—Variation of output with input for high-gain and low-gain transistors

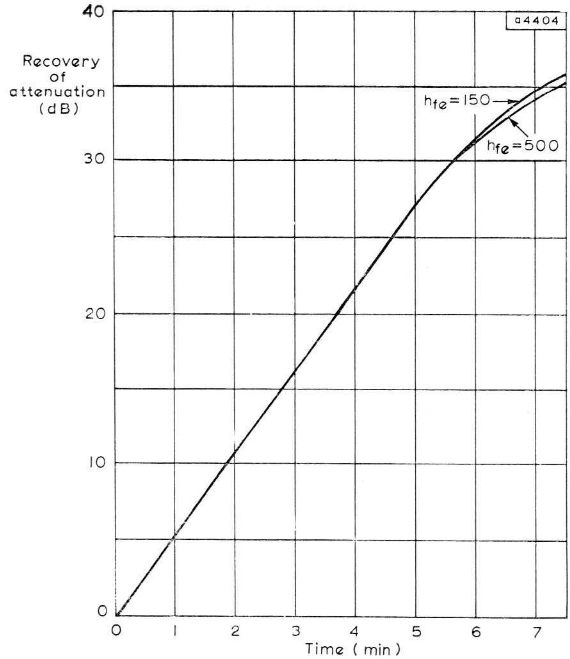


Fig. 63—Decrease of attenuation with time for high-gain and low-gain transistors

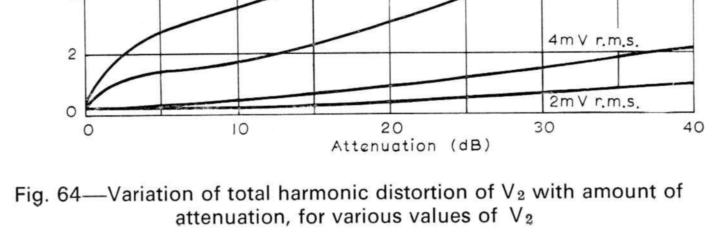


Fig. 64—Variation of total harmonic distortion of V_2 with amount of attenuation, for various values of V_2

COMPONENT LISTS FOR CIRCUITS IN CHAPTER 5

4W Tape Recorder (page 70)

Heads

Record/playback head	Marriot X/RPS/36
Erase head	Marriot X/ES/11

Transistors and Diodes

Circuit reference	Mullard type
TR_1	BC149 or BC109
TR_2	BC148 or BC108
TR_3	BC148 or BC108
TR_4	AC128
TR_5	AD161
TR_6	AD162
D_1	OA91
D_2	OA91
Full-wave bridge rectifier	BY164

Resistors

Tolerance: $\pm 5\%$ except R_{33} and R_{34}, which may be $\pm 10\%$ components
Power rating: $\frac{1}{8}$W except R_{26}, which should be a $\frac{1}{2}$W component, R_{33} and R_{34}, which should be $\frac{1}{4}$W components

Circuit reference	Value	Circuit reference	Value
R_1 }	chosen to	R_{20}	27Ω
R_2 }	suit input	R_{21}	2·2kΩ
R_3	330kΩ	R_{22}	2·7kΩ
R_4	100kΩ	R_{23} preset potentiometer	50Ω
R_5	220kΩ		
R_6	68kΩ	R_{24}	39kΩ
R_7	68kΩ	R_{25}	18Ω
R_8	22kΩ	R_{26}	560Ω
R_9	47kΩ	R_{27}	220kΩ
R_{10}	82kΩ	R_{28} thermistor type VA1077	32Ω
R_{11}	180kΩ		
R_{12}	12kΩ	R_{29}	5·6kΩ
R_{13}	820Ω	R_{30}	68Ω
R_{14} logarithmic potentiometer	25kΩ	R_{31}	68Ω
		R_{32}	5·6kΩ
R_{15}	68kΩ	R_{33}	1Ω
R_{16}	150kΩ	R_{34}	1Ω
R_{17}	68kΩ	R_{35}	15Ω
R_{18}	3·3kΩ	R_{36}	18kΩ
R_{19}	2·7kΩ		

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	6·8nF	polyester	C296AC/A6K8
C_2	6·4μF	electrolytic 25V	C426AR/F6·4
C_3	6·4μF	electrolytic 25V	C426AR/F6·4
C_4	820pF	polystyrene	C295AC/820E
C_5	1·5nF	polyester	C296AC/A1K5
C_6	390pF	polystyrene	
C_7	3·3nF	polyester	C296AC/A3K3
C_8	3·3nF	polyester	C296AC/A3K3
C_9	160μF	electrolytic 2·5V	C426AR/A160
C_{10}	220nF	polyester	C296AA/A220K
C_{11}	250μF	electrolytic 25V	C437AR/F250
C_{12}	6·4μF	electrolytic 25V	C426AR/F6·4
C_{13}	2·5μF	electrolytic 16V	C426AR/H2·5
C_{14}	125μF	electrolytic 16V	C426AR/E125
C_{15}	1nF	polyester	C296AC/A1K
C_{16}	6·4μF	electrolytic 25V	C426AR/F6·4
C_{17}	320μF	electrolytic 2·5V	C426AR/A320
C_{18}	150nF	polyester	C296AA/A150K
C_{19}	150pF	ceramic plate	C333CH/C150E
C_{20}	33nF	polyester	C296AA/A33K
C_{21}	68nF	polyester	C296AA/A68K
C_{22}	68nF	polyester	C296AA/A68K
C_{23}	640μF	electrolytic 16V	C437AR/E640
C_{24}	6·4μF	electrolytic 25V	C426AR/F6·4
C_{25}	640μF	electrolytic 25V	C437AR/F640

Mains Transformer

The transformer used in this circuit may be obtained from the following manufacturers under the type numbers given:

Colne Electric Limited, Rickmansworth, Herts	20042
Drake Transformers Limited, Billericay, Essex	351-245
Gardners Transformers Limited, Christchurch, Hants	GR 97182
Parmeko Limited, Aylestone Park, Leicester	P 3203

Heatsinks

Each output transistor should be mounted on a 1·5mm (16 gauge) bright aluminium heatsink with an area of 26cm². The driver transistor TR_4 should be mounted by means of a heat clip onto a 1·5mm bright aluminium heatsink with an area of 13cm².

Tape Pre-amplifier (page 76)

As for the tape recorder amplifier (page 86) except:

Transistors

TR_5	AC176
TR_6	AC128

Resistors

R_{23}	omitted
R_{28}	omitted
R_{33}	2·2Ω
R_{34}	2·2Ω

Capacitors

C_{23}	omitted

A.G.C. Circuit (page 80)

Transistor and Diode

Circuit reference	Mullard type
TR_1	BC148 or BC108
D_1	BA114

Resistors

Tolerance: ±5% except R_4, which may be a ±10% component
Power rating: $\frac{1}{8}$W

Circuit reference	Value
R_1	22kΩ
R_2	see text
R_3	390kΩ
R_4	3·3MΩ
R_5	1·2kΩ
R_6	1kΩ
R_7	1·5kΩ
R_8	1kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	220nF	polyester	C296AA/A220K
C_2	470nF	polyester	C296AA/A470K
C_3	220nF	polyester	C296AA/A220K
C_4	320μF	electrolytic 6·4V	C426AR/C320
C_5	12·5μF	electrolytic 25V	C426AR/F12·5
C_6	125μF	electrolytic 25V	C426AR/E125

CHAPTER 6

CAR RADIOS

The chapter is divided into two sections comprising:

1. Two a.f. amplifiers, a 5W class A amplifier and a 6W class B amplifier, suitable for use in a car radio;
2. R.F. and i.f. stages for a car radio.

Attention is again drawn to the difficulties associated with the construction and alignment of r.f. and i.f. stages, as stated earlier. Access to specialised equipment is essential to those wishing to build these circuits.

Component lists for the circuits are given at the end of the chapter.

5W CLASS A AMPLIFIER

The circuit of a 5W class A audio amplifier for use in a car radio is given in Fig. 65. It contains two BC148 or BC108 transistors and one BD121 transistor, and is designed to operate from a nominal 14V supply.

Circuit Description

Direct coupling is used between stages, and the base-emitter voltage of TR_1 serves as a reference for the emitter voltage of TR_3. Consequently, the emitter voltage of TR_1 controls the direct current through TR_3 and determines its operating point.

Variable resistor R_{11} is used to set the quiescent current—900mA— of the output stage. The loudspeaker is driven by means of a centre-tapped choke which incorporates a small winding to provide feedback to the first stage. The choke has an inductance of 30mH and a d.c. resistance of 1Ω. The voltage ratio of choke to secondary is 13:1. Resistor R_{10} and capacitor C_5 limit the frequency response of the feedback loop and amplifier.

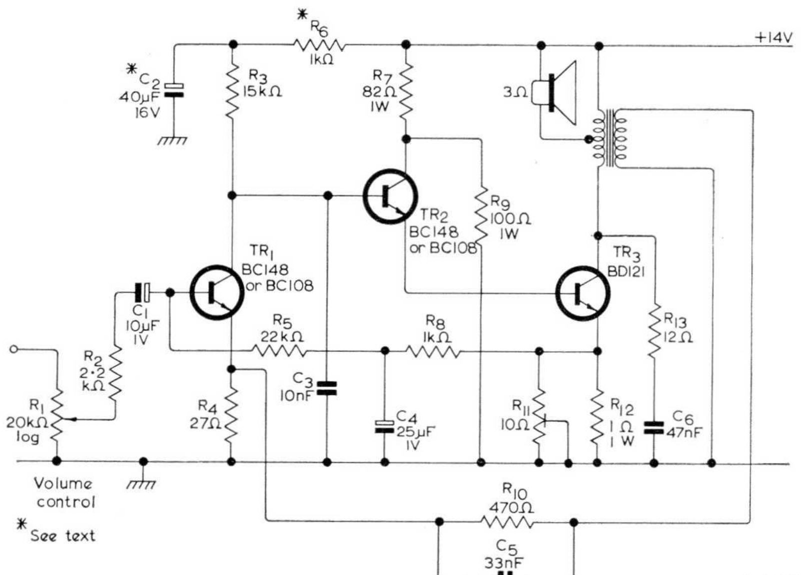


Fig. 65—5W class A audio amplifier circuit for use in a car radio

Resistor R_{13} and capacitor C_6 are included to ensure that the load remains resistive, and hence the circuit stable, at high frequencies. As low-frequency stability depends upon the source impedance, a 2·2kΩ resistor, R_2, is connected in series with the wiper of the volume control R_1.

Decoupling between resistors R_3 and R_7 is effected by resistor R_6 and capacitor C_2. The values given for R_6 and C_2 in the circuit diagram may have to be modified as they depend upon the decoupling at the supply input. The voltage drop across R_6 should not exceed 1V; if it does, the value of R_3 must be reduced.

Performance

Output power with symmetrical clipping	5W r.m.s.
Sensitivity (input for 5W output with volume control set for maximum volume)	43mV
Frequency response (—3dB points)	50Hz to 10kHz
Input resistance (including 2·2kΩ resistor but excluding volume control)	13kΩ
Negative feedback	12dB
Distortion at 5W output (see Fig. 66)	2·5%

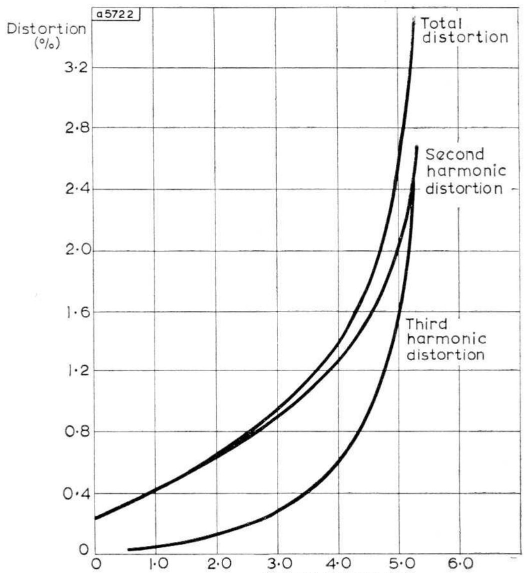


Fig. 66—Variation of harmonic distortion with output power

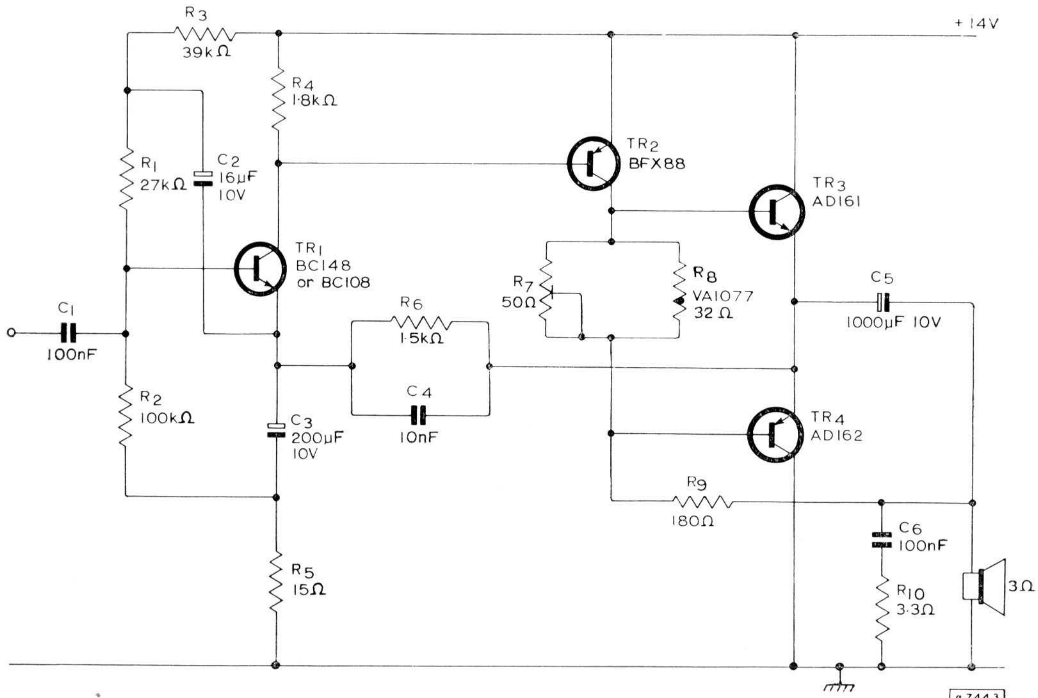


Fig. 67—6W class B audio amplifier circuit for use in a car radio

6W CLASS B AMPLIFIER

The circuit of a 6W class B audio amplifier for use in a car radio is shown in Fig. 67. The amplifier contains an AD161/AD162 complementary output pair, a BFX88 driver transistor, and a BC148 or BC108 transistor as the first-stage amplifier.

Circuit Description

The amplifier is a conventional four-transistor circuit with one exception: the decoupling capacitor C_2 in the input stage is returned to the emitter of transistor TR_1 instead of to chassis. The purpose of this is two-fold: firstly, to neutralise the effect of ripple on the h.t. line, and secondly, to bootstrap the base resistor, thereby increasing the input resistance.

Performance

Output power	6W r.m.s.
Sensitivity for full output	52mV
Frequency response at 3W (—3dB points)	74Hz to 10·6kHz
Input resistance at 1kHz	100kΩ
Feedback	20dB
Distortion at 6W output (see Fig. 68)	1·2%
Quiescent current in output stage only	30mA
Total current at 6W output	700mA

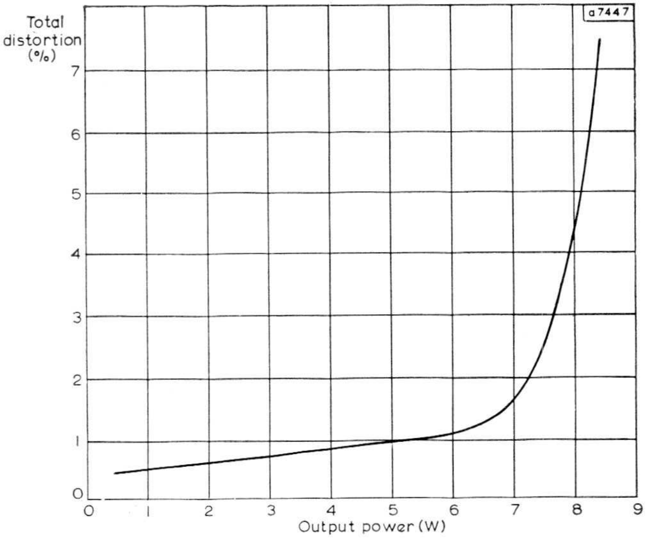


Fig. 68—Variation of total harmonic distortion with output power

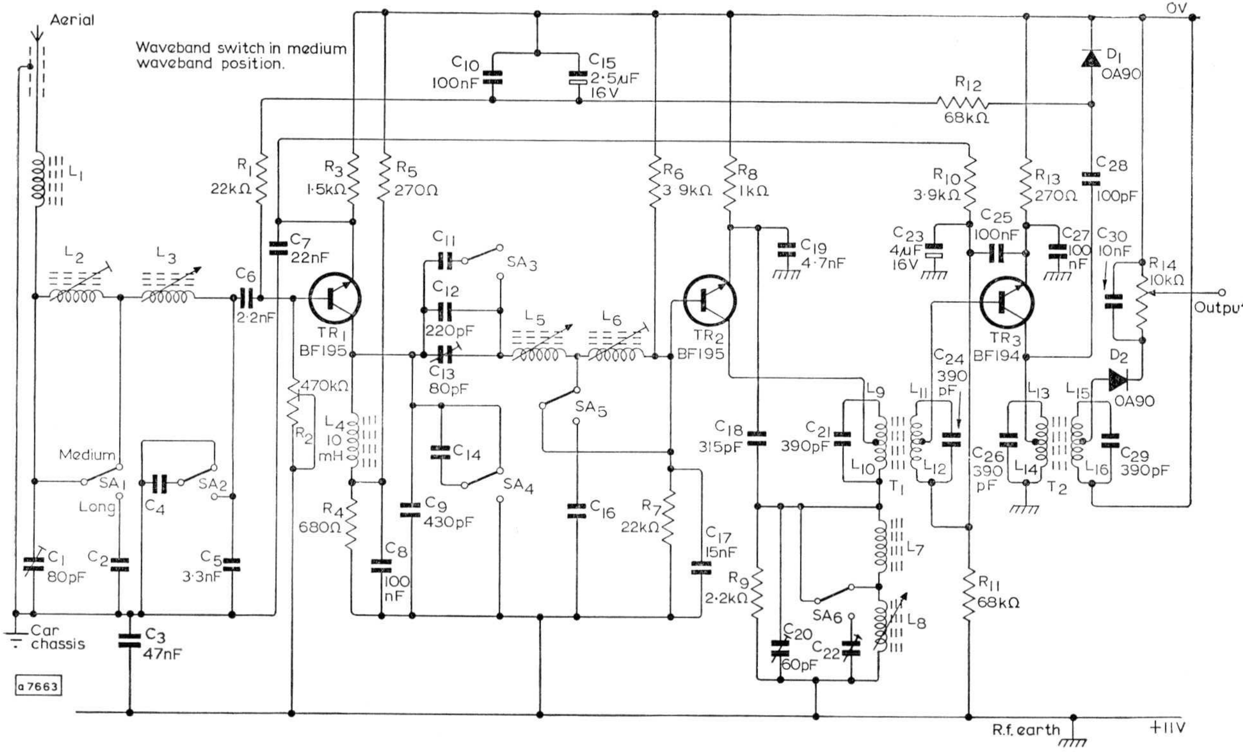


Fig. 69—R.F. and i.f. stages of a car radio receiver

R.F. AND I.F. STAGES

The r.f. and i.f. stages of a medium- and long-wave car radio are shown in Fig. 69. The circuit is designed to operate from an 11V supply and uses two BF195 transistors and one BF194 transistor.

Circuit Description

R.F Amplifier and Self-oscillating Mixer Stages

The design of the r.f. amplifier stage is necessarily one of compromise between factors such as gain, noise performance, signal handling and avoidance of mixer blocking.

The wide range of a.g.c. required of a car radio can be satisfactorily met by applying full control to the r.f. amplifier. The maximum signal handling can be achieved by ensuring that the base of the r.f. amplifier is ultimately reverse-biased by the a.g.c. circuit.

A serious limitation encountered when using an r.f. amplifier in front of a self-oscillating mixer is r.f. blocking of the oscillator. If the receiver is tuned rapidly to a strong carrier (for example, by a push button), a very large signal can appear at the mixer base before the a.g.c. voltage has time to build up. Consequently the oscillator may be prevented from operating, and the receiver will be blocked. In this design, it was found necessary to limit the maximum input to the mixer stage (a) by the choice of tapping ratio between the r.f. amplifier collector and the mixer base, and (b) by reducing the r.f. amplifier collector voltage. These requirements limit the r.f. gain to a value considerably lower than could otherwise be achieved with the BF195.

A conventional car-radio mixer is used, comprising a capacitively-tapped Colpitts oscillator, and using a BF195 transistor. The gain of the stage is limited by the oscillator voltage developed across the primary of the first i.f. transformer T_1 on the long waveband.

I.F. Stage

The increase in stage gain, theoretically possible when transistors with low feedback capacitance are used, is obtained only if the stage gain is not determined by the a.g.c. line damping, which limits the collector loading of the i.f. stage. In the past, the limitation in loading has necessitated a rather high value of base source impedance. This has the disadvantage of giving considerable spreads and reduced selectivity, which can be overcome only by some sacrifice in gain. With the BF194, however, it is possible to realise this increase in gain quite simply, by using a high value of preset potentiometer (R_2) as one of the bias resistors of the r.f. amplifier. This allows a high value of collector load to be used in the i.f. stage.

The temperature effects with silicon planar transistors are small, and virtually the only requirement for stabilisation is that the spreads in quiescent current should be small.

Detector

The a.f. amplifiers recommended for use with this circuit have input impedances greater than 10kΩ. A high value of detector load resistance (10kΩ potentiometer R_{14}) can therefore be used. This allows the diode to work at a higher voltage level, and results in increased diode efficiency and about 2 to 3dB improvement in sensitivity.

Performance

Sensitivity for 7mV output at detector, at 1MHz with 30% modulation	1·4μV
Signal-to-noise ratio with 15pF/60pF aerial:	
1·4μV aerial e.m.f.	5dB
10μV aerial e.m.f.	20dB
A.G.C. range for 10dB change in output (see Fig. 70)	90dB
Maximum signal handling*	
30% modulation	2V
80% modulation	1V
Oscillator drive voltage, for medium waveband:	
oscillator frequency = 2·11MHz	100mV
oscillator frequency = 1·01MHz	120mV

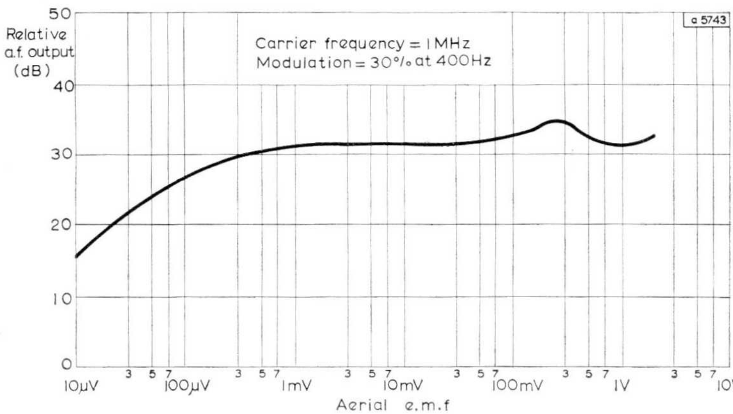


Fig. 70—Automatic gain control characteristic

*No mixer blocking occurs at any signal level

COMPONENT LISTS FOR CIRCUITS IN CHAPTER 6

5W Class A Amplifier (page 90)

Transistors

Circuit reference	Mullard type
TR_1	BC148 or BC108
TR_2	BC148 or BC108
TR_3	BD121

Resistors

Circuit reference	Value	Tolerance (%)	Power rating (W)
R_1 logarithmic potentiometer	20kΩ		
R_2	2·2kΩ	±20	$\frac{1}{8}$
R_3	15kΩ	±20	$\frac{1}{8}$
R_4	27Ω	±10	$\frac{1}{8}$
R_5	22kΩ	±20	$\frac{1}{8}$
R_6	1kΩ	±20	$\frac{1}{8}$
R_7	82Ω	±10	1
R_8	1kΩ	±20	$\frac{1}{8}$
R_9	100Ω	±10	1
R_{10}	470Ω	±10	$\frac{1}{8}$
R_{11} preset potentiometer	10Ω		
R_{12}	1Ω	±20	1
R_{13}	12Ω	±20	$\frac{1}{8}$

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	10μF	electrolytic 1V	C426AR/E10
C_2	40μF	electrolytic 16V	C426AR/E40
C_3	10nF	polyester	C296AA/A10K
C_4	25μF	electrolytic 1V	C426AR/C25
C_5	33nF	polyester	C296AA/A33K
C_6	47nF	polyester	C296AA/A47K

Heatsinks

The values of R_7 and R_9 ensure that the maximum permissible junction temperature of transistor TR_2 is not exceeded and so no heat clip is required. The output transistor, however, has a maximum dissipation of 11W; therefore at an ambient temperature of 60°C, a 1·5mm (16 gauge) bright aluminium heatsink with an area of 50cm^2 is required.

6W Class B Amplifier (page 94)

Transistors

Circuit reference	Mullard type
TR_1	BC148 or BC108
TR_2	BFX88
TR_3	AD161
TR_4	AD162

Resistors

Circuit reference	Value	Tolerance (%)	Power rating (W)
R_1	27kΩ	±5	$\frac{1}{8}$
R_2	100kΩ	±5	$\frac{1}{8}$
R_3	39kΩ	±5	$\frac{1}{8}$
R_4	1·8kΩ	±10	$\frac{1}{8}$
R_5	15Ω	±5	$\frac{1}{8}$
R_6	1·5kΩ	±10	$\frac{1}{8}$
R_7 preset potentiometer	50Ω		
R_8 thermistor type VA1077	32Ω		
R_9	180Ω	±10	$\frac{1}{2}$
R_{10}	3·3Ω	±10	$\frac{1}{8}$

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	100nF	polyester	C296AA/A100K
C_2	16μF	electrolytic 10V	C426AR/D16
C_3	200μF	electrolytic 10V	C426AR/D200
C_4	10nF	polyester	C296AA/A10K
C_5	1000μF	electrolytic 10V	C437AR/D1000
C_6	100nF	polyester	C296AA/A100K

Heatsinks

The driver dissipation is only 350mW, so no heatsink is required for this transistor. If mica washers are used, each output transistor should be mounted on a 1·5mm (16 gauge) bright aluminium heatsink with an area of 16cm^2.

R.F. and I.F. Stages (page 94)

Transformers

Both i.f. transformers are double-tuned and critically coupled; they are wound on 4·8mm diameter formers, with ferrite slugs and frames.

First i.f. transformer—T_1

L_9+L_{10}	80+30 turns of 6×46 s.w.g. litz wire; *Z_{do} = 50kΩ, Q_o = 140
$L_{11}+L_{12}$	5+105 turns of 6×46 s.w.g. litz wire; Z_{do} = 235Ω, Q_o = 150

Second i.f. transformer—T_2

$L_{13}+L_{14}$	60+50 turns of 6×64 s.w.g. litz wire; Z_{do} = 35kΩ, Q_o = 150
$L_{15}+L_{16}$	35+75 turns of 6×64 s.w.g. litz wire; Z_{do} = 10kΩ, Q_o = 150

Coils for which no details are given are part of the permeability tuner.

Transistors and Diodes

Circuit reference	Mullard type
TR_1	BF195
TR_2	BF195
TR_3	BF194
D_1	OA90
D_2	OA90

Resistors

Tolerance: ±5%

Power rating: $\frac{1}{8}$W

Circuit reference		Value
R_1		22kΩ
R_2	carbon trimming potentiometer	470kΩ
R_3		1·5kΩ
R_4		680Ω
R_5		270Ω
R_6		3·9kΩ
R_7		22kΩ
R_8		1kΩ
R_9		2·2kΩ
R_{10}		3·9kΩ
R_{11}		68kΩ
R_{12}		68kΩ
R_{13}		270Ω
R_{14}	logarithmic carbon potentiometer	10kΩ

*Z_{do} = coil impedance with the other coil open-circuited.

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	80pF	trimmer	
C_3	47nF	polyester	C296AA/A47K
C_5	3·3nF	polyester	C296AC/A3K3
C_6	2·2nF	polyester	C296AC/A2K2
C_7	22nF	polyester	C296AA/A22K
C_8	100nF	polyester	C296AA/A100K
C_9	430pF	polystyrene	
C_{10}	100nF	polyester	C296AA/A100K
C_{12}	220pF	polystyrene	
C_{13}	80pF	trimmer	
C_{15}	2·5μF	electrolytic 16V	C426AR/H2·5
C_{17}	15nF	polyester	C296AA/A15K
C_{18}	315pF	polystyrene	
C_{19}	4·7nF	polyester	C296AC/A4K7
C_{20}	60pF	trimmer	808 01011
C_{21}	390pF	polystyrene	
C_{23}	4μF	electrolytic 16V	C426AR/G4
C_{24}	390pF	polystyrene	
C_{25}	100nF	polyester	C296AA/A100K
C_{26}	390pF	polystyrene	
C_{27}	100nF	polyester	C296AA/A100K
C_{28}	100pF	ceramic plate	C333CH/C100E
C_{29}	390pF	polystyrene	
C_{30}	10nF	polyester	C296AA/A10K

Capacitors for which no values are given are part of the permeability tuner.

CHAPTER 7

HIGH-QUALITY AUDIO EQUIPMENT

In this chapter audio amplifiers with output powers of 10W and 25W are described, together with a pre-amplifier suitable for use with either amplifier. The chapter includes auxiliary audio amplifiers, control circuits and filters.

In the construction of high-quality equipment attention must be paid to component layout; the recommendations made in this chapter should be carefully followed.

Components lists for the circuits are given at the end of the chapter.

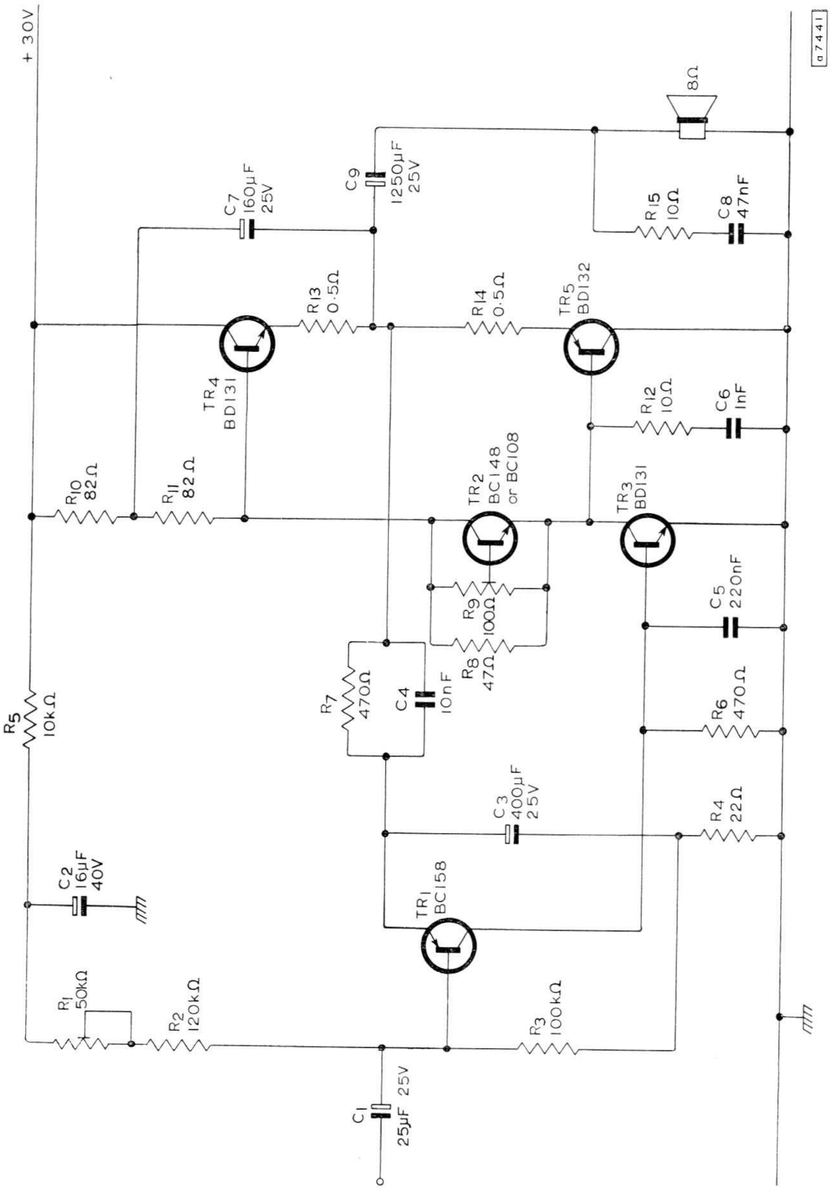


Fig. 71—10W high-quality audio amplifier circuit

10W AUDIO AMPLIFIER

The circuit for a 10W high-quality audio amplifier is shown in Fig. 71. The use of a complementary pair of output transistors gives a much simpler circuit than the quasi-complementary circuit usually employed in class B hi-fi amplifiers. In addition, the greater symmetry of the circuit gives lower distortion, particularly in the crossover region.

Circuit Description

The circuit uses a BD131/BD132 complementary pair to deliver an output power of 10W at a distortion level of less than 0·1 %. The familiar four-transistor, directly-coupled configuration is used, but an additional transistor is included to stabilise the quiescent current of the output transistors. Preset potentiometer R_1 sets the mid-point voltage for symmetrical clipping.

Power Supply

Fig. 72 shows a suitable power supply. The supply requirements are:

Total quiescent current for one channel	120mA
Full load current for one channel (at 30V)	600mA
Supply voltage	30V
Maximum supply voltage	40V

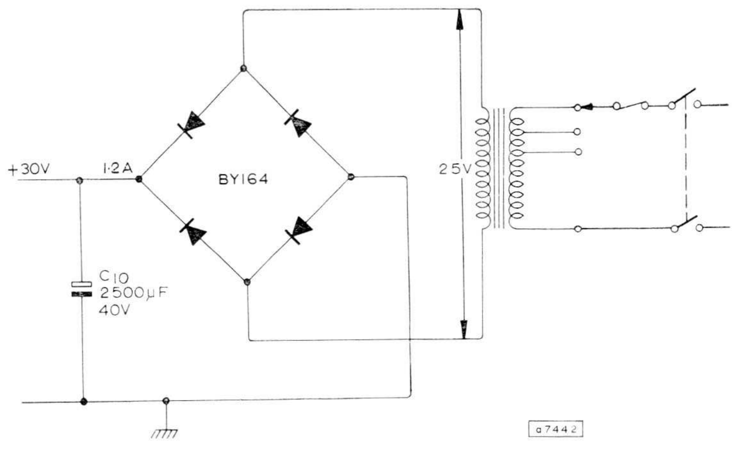


Fig. 72—Power supply circuit for 10W amplifier

Performance

Output power (with 30V supply)	10W r.m.s.
Sensitivity for full output	430mV
Frequency response at 2W output (–3dB points)	20Hz to 35kHz
Input resistance	90kΩ
Distortion at 10W output (see Fig. 73)	$< 0{\cdot}1\%$
Quiescent current of output transistors	15mA

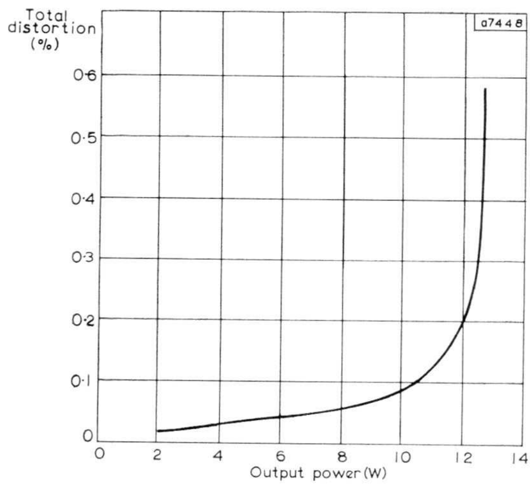


Fig. 73—Variation of total harmonic distortion with output power

25W AUDIO AMPLIFIER

The circuit of a 25W high-quality amplifier suitable for use as one channel of a stereo system is shown in Fig. 74. In this circuit, two matched BDY20 output transistors are driven by a matched complementary pair consisting of one BFX29 and one BFX84. Diodes D_1, D_2 and D_3 are included to ensure that distortion in the crossover region—that is, at low output powers—is very low.

Circuit Description

The voltage at the mid-point between the output transistors is set by means of the preset potentiometer R_1 to give symmetrical clipping. Transistor TR_3 provides bias for the output stage as well as thermal and voltage compensation. The quiescent current of the output stage is set to 35 to 40mA, or that of the whole power amplifier to 45 to 50mA, by means of R_9.

High-frequency stabilisation, necessitated by the high value of feedback, is effected by C_4, C_5, C_6, C_9, R_{14} and R_{23}.

Power Supply

The power supply requirements are:

Total quiescent current for one channel only	45 to 50mA
Supply voltage	50V
Maximum supply voltage under quiescent conditions	60V
Total current taken from the supply for one channel only:	
for 25W output into 8Ω load	0·8A
for 25W output into 4Ω load	1·1A

For short-circuit protection, a 1·5A fuse is required in the supply to each channel.

The circuit for a simple power supply is shown in Fig. 75.

Performance

Output power (8Ω load, 50V supply)	25W r.m.s.
Sensitivity for 25W output	200mV
Frequency response (–3dB points)	10Hz to 30kHz
Input resistance	200kΩ
Distortion at 25W output (see Fig. 76)	0·2%

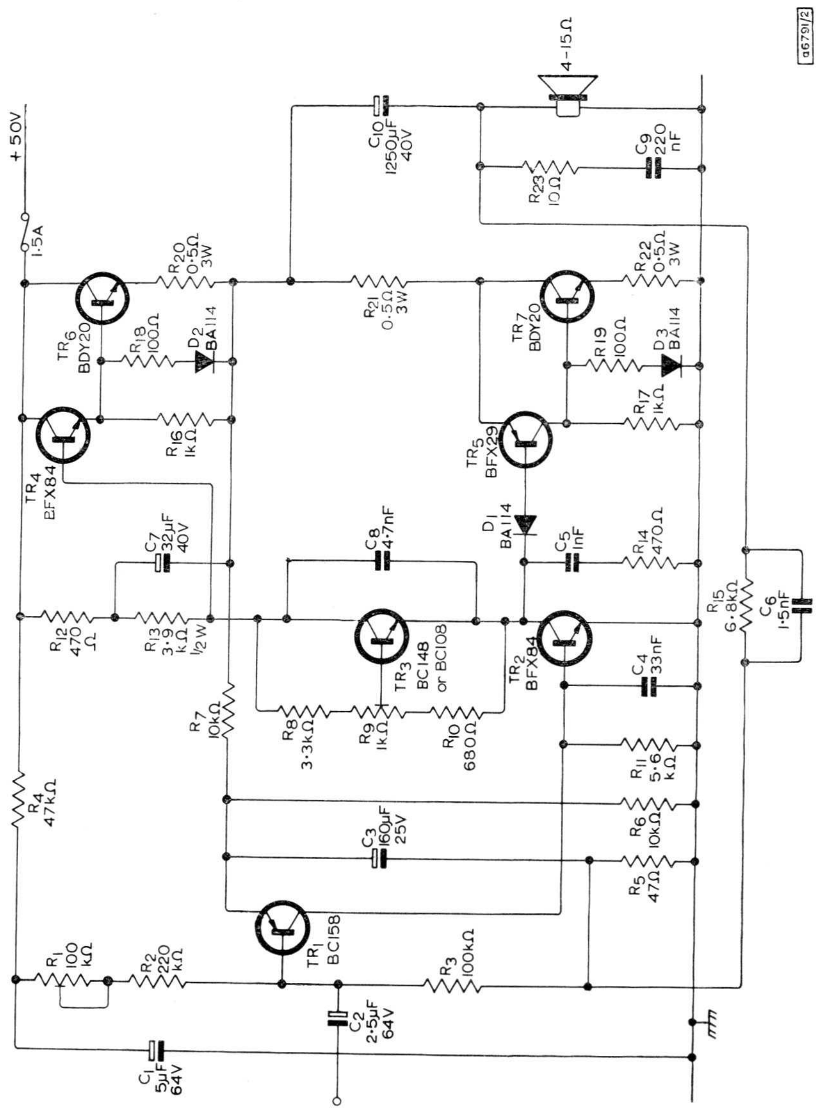


Fig. 74—25W high-quality audio amplifier circuit

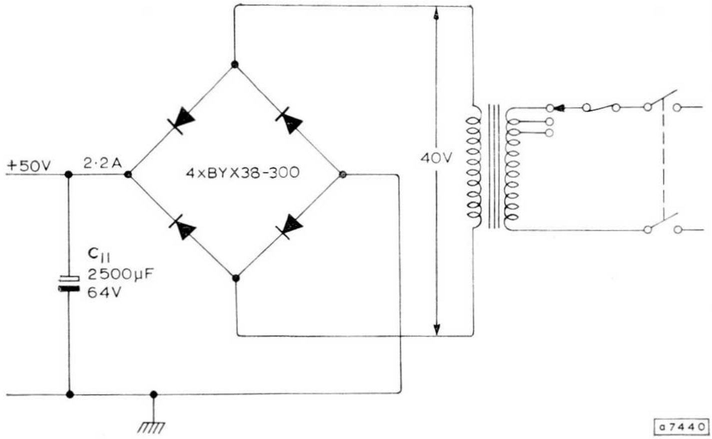


Fig. 75—Power supply circuit for 25W amplifier

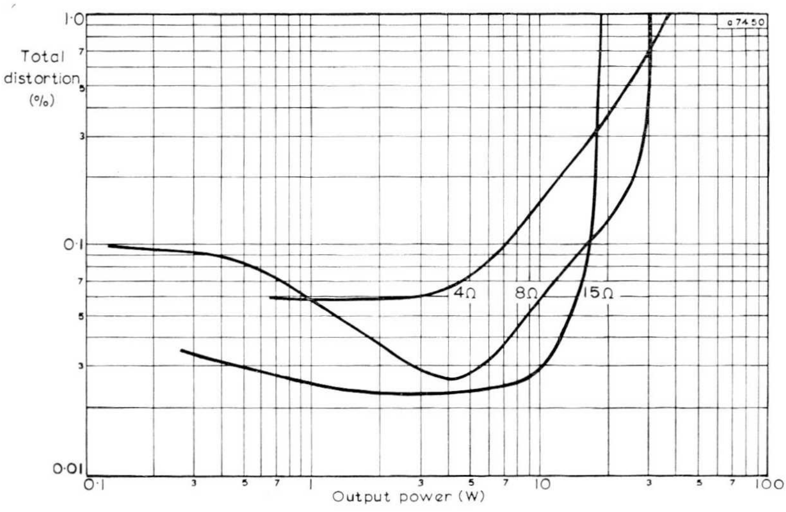
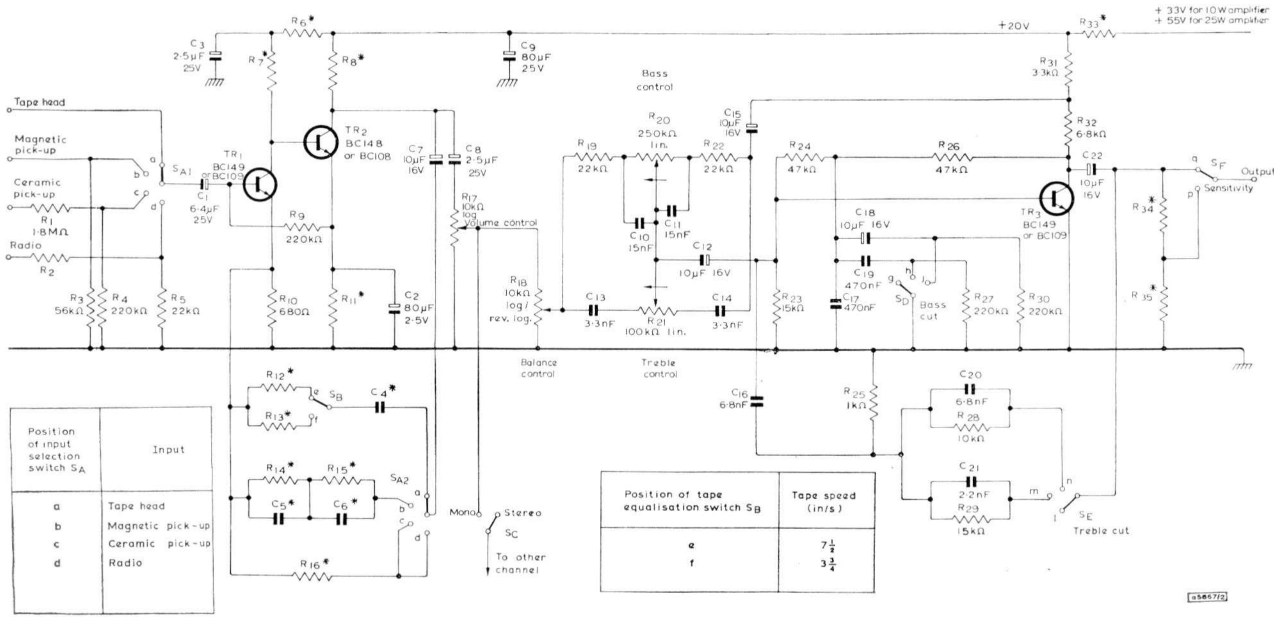


Fig. 76—Variation of total harmonic distortion with output power
for three load impedances



Position of input selection switch S_A	Input
a	Tape head
b	Magnetic pick-up
c	Ceramic pick-up
d	Radio

Position of tape equalisation switch S_B	Tape speed (in/s)
e	$7\frac{1}{2}$
f	$3\frac{3}{4}$

Fig. 77—Pre-amplifier circuit for use with 10W and 25W amplifiers
*The values of these components depend on the power amplifier to be used; see components list. Switch S_F is omitted in a pre-amplifier to be used with a 10W amplifier

PRE-AMPLIFIER FOR USE WITH THE 10W AND 25W AUDIO AMPLIFIERS

A pre-amplifier for use with the 10W and 25W audio amplifiers is shown in Fig. 77.

Circuit Description

Transistors TR_1 and TR_2 form a directly coupled pair, the base voltage of TR_1 being derived from the emitter of TR_2. Equalisation for tape head, magnetic pick-up, ceramic pick-up and radio inputs is obtained by feedback from the collector of TR_2 to the emitter of TR_1. The equalisation characteristics are shown in Figs. 78 and 79.

No value is given for R_2, the input resistor for the radio position, because this must be chosen to suit the available signal. The sensitivity at the base of TR_1 is 7·5mV for 200mV output.

Transistor TR_3 is used in a feedback tone control circuit which gives a voltage gain of three times. A circuit with linear potentiometers has been used because more boost can be obtained, and also because of the difficulty in obtaining and matching reverse-logarithmic potentiometers. A BC149 or BC109 is used because a low-noise device is required in the first stage after the volume control. The tone control characteristics are shown in Fig. 80.

The working point of TR_3 is fixed by R_{23}, R_{24} and R_{26}, a.c. feedback being removed by decoupling capacitors, C_{17}, C_{18} or C_{19}. By switching these capacitors, bass cut is obtained, and the effect on the pre-amplifier and 25W power amplifier combined is shown in Fig. 81. Networks can also be switched in by S_E to give treble cut as shown in Fig. 82.

Resistors R_{34}, R_{35} and Switch S_F are required in the pre-amplifier for use with the 25W amplifier only. When a high-sensitivity speaker is used with the 25W circuit in quiet listening conditions, the noise originating in TR_3 is perceptible even when a low-noise transistor is used. Since the full sensitivity will not be required under these conditions, a switched attenuator is used at the output of the pre-amplifier, reducing the sensitivity and noise by 12dB.

Performance

Sensitivity at 1kHz for 200mV (25W version) or 400mV (10W version) output:	
ceramic pick-up	170mV
magnetic pick-up	4mV
tape head with tape speed of 7½in/s	6·5mV
Total distortion for 200mV output with volume control at maximum	$< 0{\cdot}02\%$

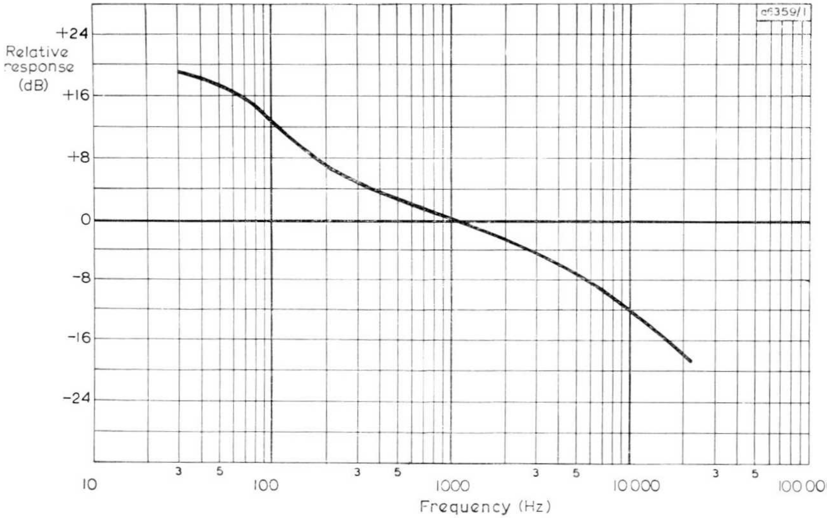


Fig. 78—Equalisation of pre-amplifier with magnetic pick-up input

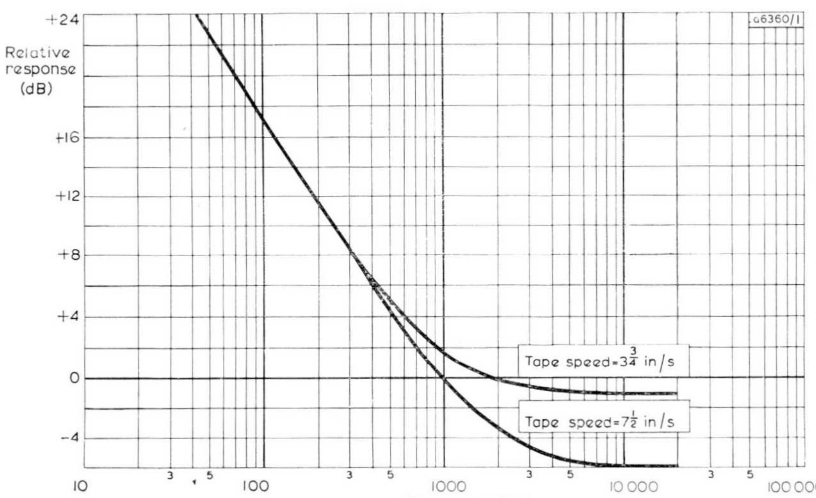


Fig. 79—Equalisation of pre-amplifier with tape head input

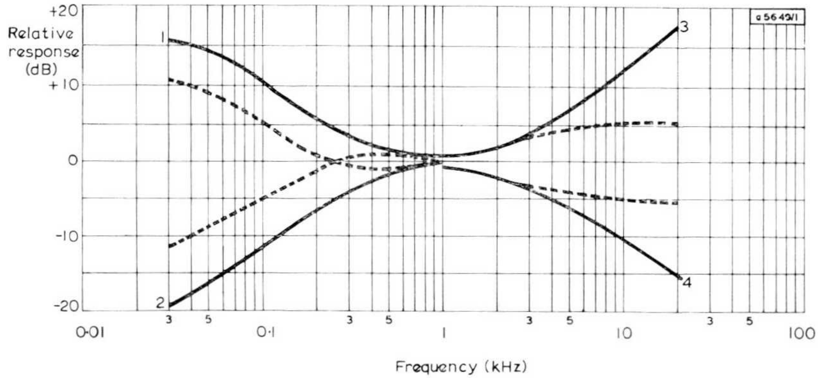


Fig. 80—Tone control characteristics of pre-amplifier
Curve 1: maximum bass boost, treble flat
Curve 2: maximum bass cut, treble flat
Curve 3: maximum treble boost, bass flat
Curve 4: maximum treble cut; bass flat
The dotted curves represent intermediate positions of the controls

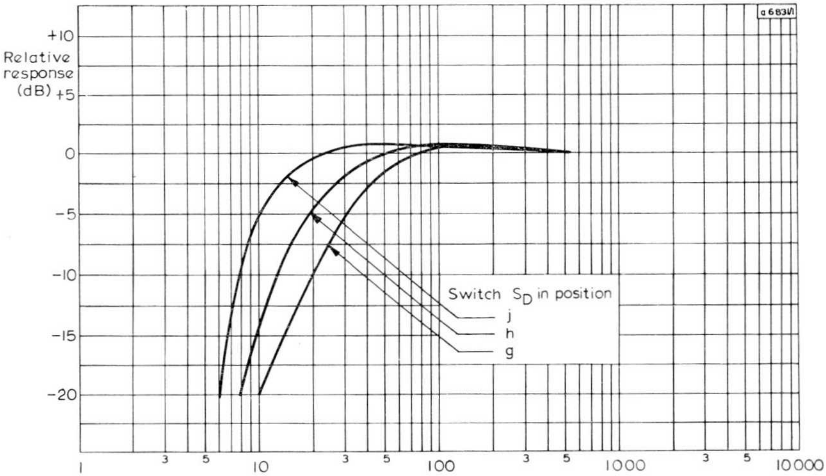


Fig. 81—Low-frequency response of pre-amplifier and 25W amplifier

Performance (continued)

Total distortion with volume control at minimum, measured at the volume control	see Fig. 83
Signal-to-noise ratio for 200mV output, for all inputs	>60dB
Total current:	
25W version	2mA
10W version	3mA

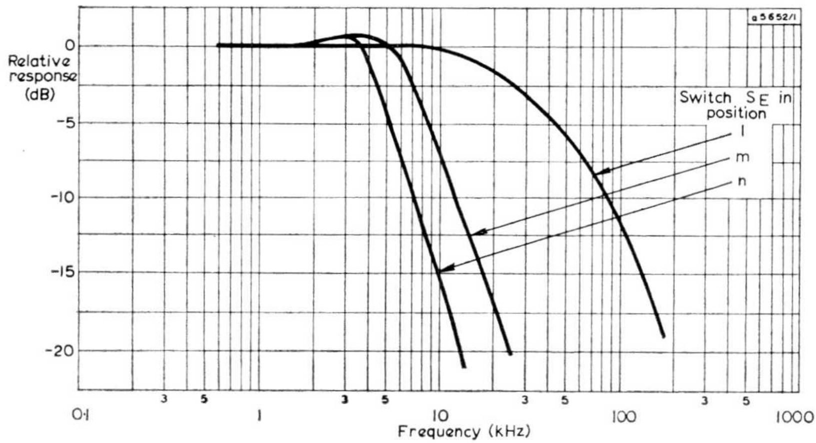


Fig. 82—High-frequency response of pre-amplifier and 25W amplifier

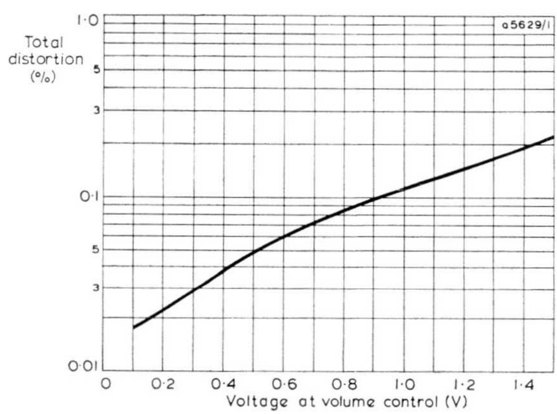


Fig. 83—Total pre-amplifier distortion (measured at the volume control) plotted against the voltage at the volume control

LAYOUT OF HIGH-QUALITY AUDIO CIRCUITRY

The circuits described in this chapter are capable of high-quality reproduction, but only if they are carefully laid out and constructed. The relative positions of input, output and power supply, and the earthing and screening arrangements are of particular importance, and it is suggested that layouts should be designed with reference to the code of good practice outlined here.

General Layout

It should be remembered that the input sensitivity of an amplifier is typically 3mV at 1kHz, on the magnetic pick-up position, and the output voltage is of the order of 20V, implying a voltage gain approaching 10^4. It is therefore essential to keep the output separated and screened from the input. The magnetic field from the mains transformer may cause hum, so the transformer should be as remote as possible from the input (this is dealt with more fully below).

Earthing

Currents of several amps magnitude circulate in the power supply and output stages. It is important that no wiring carrying these currents is included in the input circuit, otherwise hum or instability, due to the small but significant resistance of the wires, will result. The paths of currents in the output stage and power supply are shown in Fig. 84.

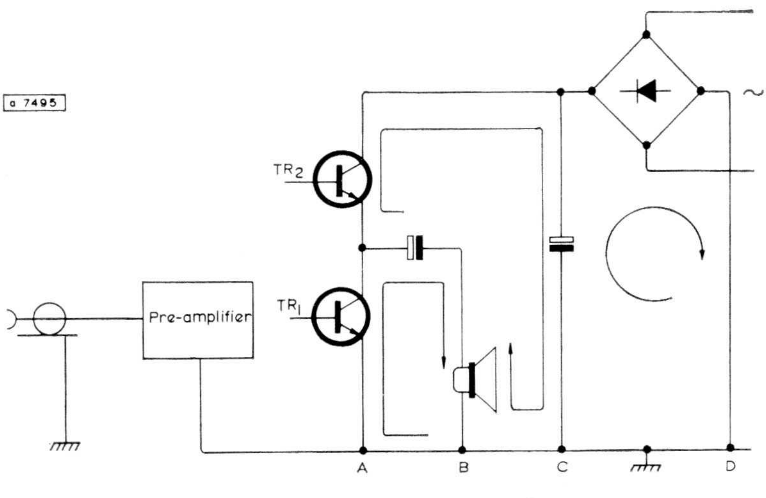


Fig. 84—Current paths in power supply and output stage

Between points A and B there is a voltage due to TR_1; between B and C a voltage due to TR_2; and between C and D a voltage due to the power supply. The input can be earthed at A, but not at B, C, or D. In practice it is usual to combine A, B, C and D in one common earthing point.

The earthing arrangement for stereo amplifiers is considerably more complex than for mono, since the single power supply and common earth for the two signal inputs make it more difficult to avoid earth loops. The arrangement recommended for the 25W amplifier is shown in Fig. 85.

Point E is the common earthing point. The voltage across the wire AE is effectively in series with the input to the power amplifier. This does not cause any trouble provided the wire is short, because the sensitivity at the input of the power amplifier is 200mV and the output-stage current is taken to the earth point separately.

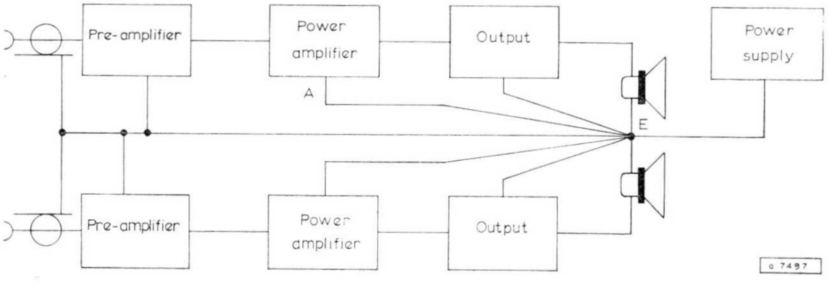


Fig. 85—Earthing arrangement for a stereo amplifier

Magnetic Fields

The principal source of magnetic fields in an amplifier is the mains transformer. Hum due to this cause will be at a frequency of 150Hz because the hysteresis of the core results in the magnetic field being predominantly third harmonic. The magnetic field will induce a voltage in any circuit loop.

The input loop of a pre-amplifier is shown in Fig. 86. The area of this loop should be minimised, and it may also be necessary to screen the mains transformer by making the parts of the chassis surrounding it from mild steel rather than aluminium.

The magnetic field associated with the currents in the output stage may occasionally cause trouble. Again, the remedy is to minimise the area of the circuit loops in the output stage by, for example, running emitter and collector leads close together, in order to minimise radiation.

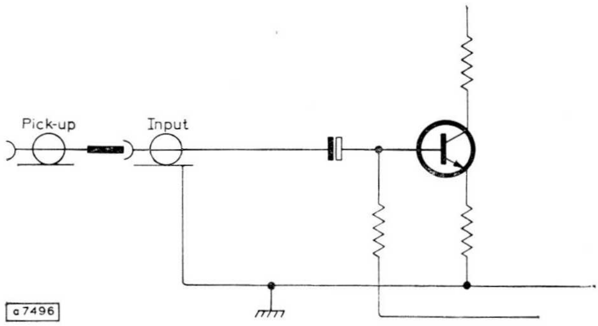


Fig. 86—Input loop of a pre-amplifier

Hum

There are many possible causes of hum and only the more probable are indicated here. A clue to the cause of the hum may often be obtained by determining its frequency.

50Hz hum may be due to stray pick-up caused by poor screening, earth loops or television break-through. The last is recognisable by the distinctive buzz of the field sync pulses, and also by the fact that it is usually accompanied by the sound channel.

100Hz hum originates in the power supply and is often indicative of poor decoupling in the h.t. line or incorrect connection of the earth leads (see Page 114).

150Hz hum comes from the magnetic field of the transformer and has already been discussed.

Resistor Noise

Every resistor has inherent thermal noise which could only be eliminated by refrigerating the component. However, certain resistors, notably carbon composition types, give an excess noise which is proportional to the voltage across them. Although this excess noise does not usually contribute significantly to the noise in the transistor circuit it is advisable to use high-stability resistors for the collector load and base bias resistors of the input stage.

AUXILIARY HIGH-QUALITY CIRCUITS

In the design of the auxiliary a.f. amplifiers, control circuits and filters which follow, use has been made of the technique, common in audio engineering, of having a high input impedance and a low output impedance. All the circuits can therefore be connected together irrespective of impedance. This means that a uniform supply voltage is necessary for all the circuits, and a value of 18V has been chosen. The supply voltage can vary by $\pm 15\%$ without any appreciable change in the performance of the circuits.

AUXILIARY HIGH-QUALITY AMPLIFIERS

The auxiliary amplifiers described in this section are derived from the basic voltage amplifier shown in Fig. 87. To achieve stabilisation against temperature variations, two d.c. feedback loops are used: one from the emitter of the second transistor to the base of the first, and the other from the collector of the second transistor to the emitter of the first. The values of the components for circuits with voltage gains A_V of 10, 20, 30 and 40dB are shown in Table 2. The voltages in the circuit, and the input and output impedance values Z_{in} and Z_{out}, are shown in Table 3.

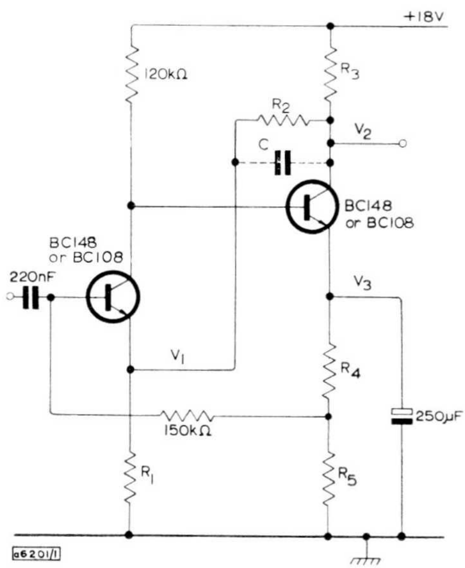


Fig. 87—Simple audio amplifier circuit. Component values are given in Table 2

TABLE 2
Component Values

Circuit reference	A_v = 10dB	A_v = 20dB	A_v = 30dB	A_v = 40dB
R_1	4·7kΩ	1·5kΩ	1·5kΩ	1kΩ
R_2	12kΩ	15kΩ	56kΩ	180kΩ
R_3	1·8kΩ	2·2kΩ	2·2kΩ	2·2kΩ
R_4	470Ω	560Ω	330Ω	680Ω
R_5	1·2kΩ	470Ω	270Ω	220Ω
C	—	—	—	10pF

TABLE 3
Voltages and Impedances

Voltage or impedance	A_v = 10dB	A_v = 20dB	A_v = 30dB	A_v = 40dB
V_1	3·4V	0·97V	0·4V	0·15V
V_2	10·8V	9·3V	9·3V	9·7 V
V_3	5·6V	3·55V	2·3V	3·4V
Z_{in}	145kΩ	140kΩ	135kΩ	110kΩ
Z_{out}	63Ω	140Ω	260Ω	700Ω

Figs. 88 to 91 show for the four circuits the total distortion as a function of the output voltage at three frequencies, and the noise voltage at the output as a function of the generator resistance at the input. The total distortion for all four amplifiers remains below 0·1% for output voltages up to 1V at 1kHz, and below 1% for output voltages up to 3V. The noise voltage referred to the input in all four amplifiers is less than 1μV. The frequency response (–3dB points) of all amplifiers is from 20Hz to 20kHz.

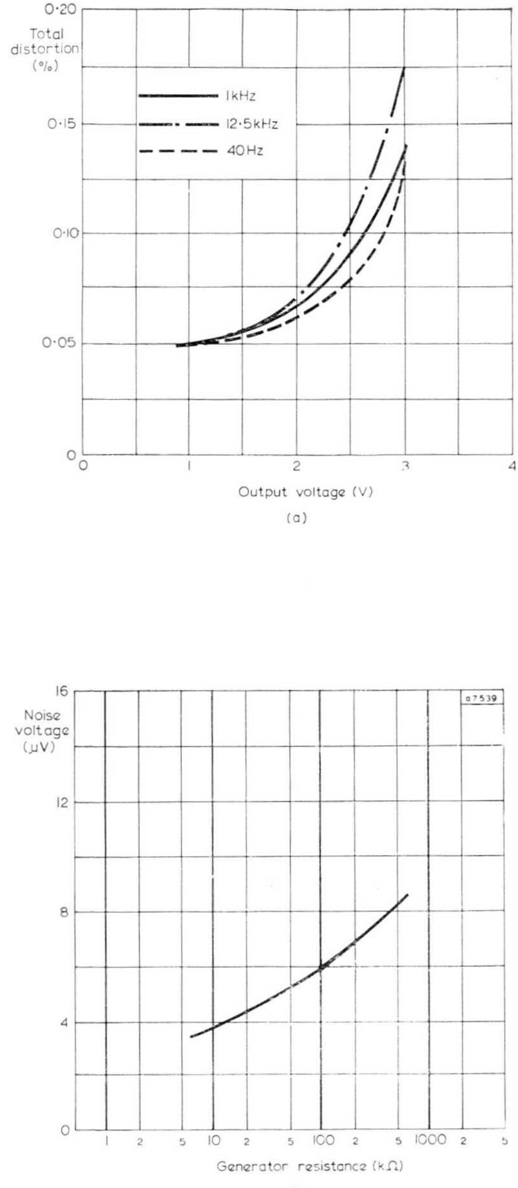
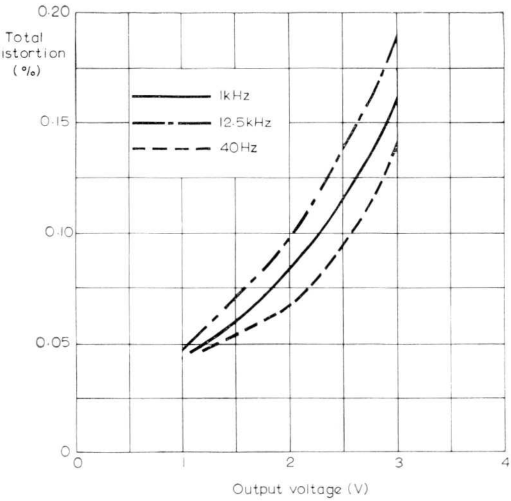
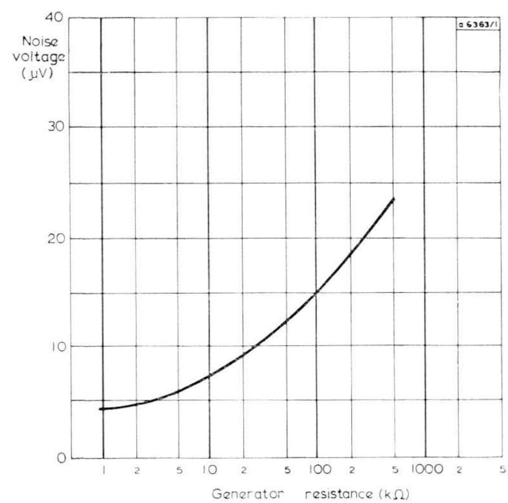


Fig. 88—Total distortion and noise voltage at output of 10dB amplifier

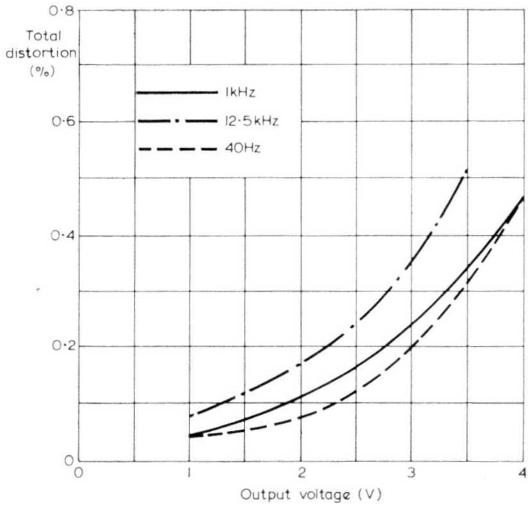


(a)

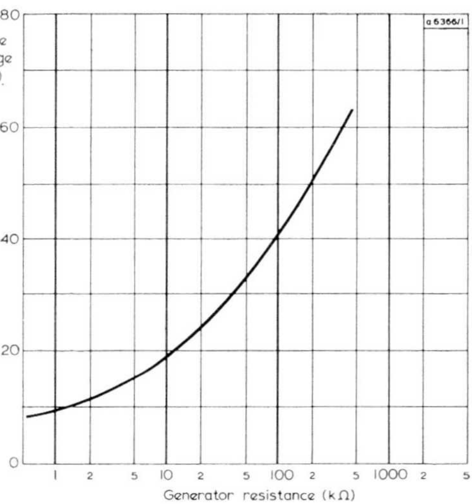


(b)

Fig. 89—Total distortion and noise voltage at output of 20dB amplifier

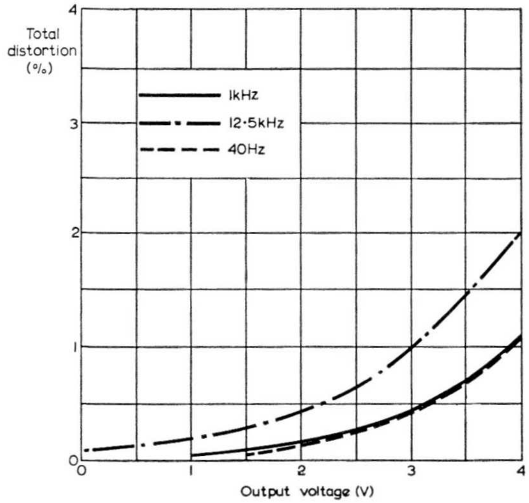


(a)

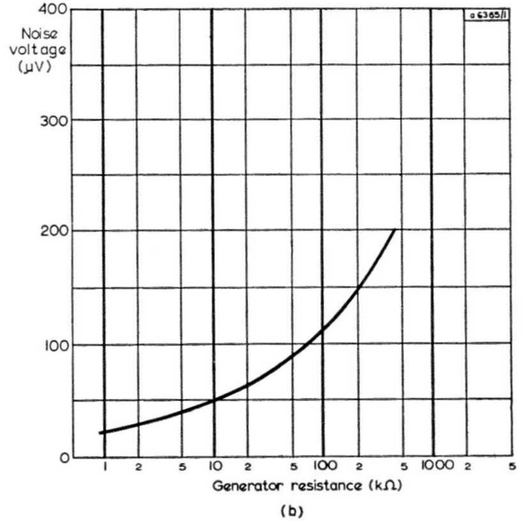


(b)

Fig. 90—Total distortion and noise voltage at output of 30dB amplifier



(a)



(b)

Fig. 91—Total distortion and noise voltage at output of 40dB amplifier

Buffer Amplifier

The circuit of a two-stage buffer amplifier is shown in Fig. 92. The first stage works in the common-emitter configuration with a large amount of feedback, while the second stage is an emitter follower. This ensures a high input impedance of 3·6MΩ, and a low output impedance of 250Ω. The voltage gain is unity, and the frequency response (–3dB points) is from 20Hz to 20kHz. Fig. 93 shows the total distortion and the noise voltage at the output. The distortion remains below 0·5% with output voltages up to 2·5V.

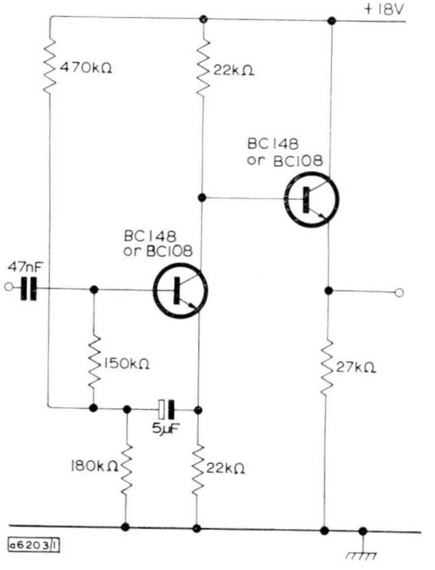
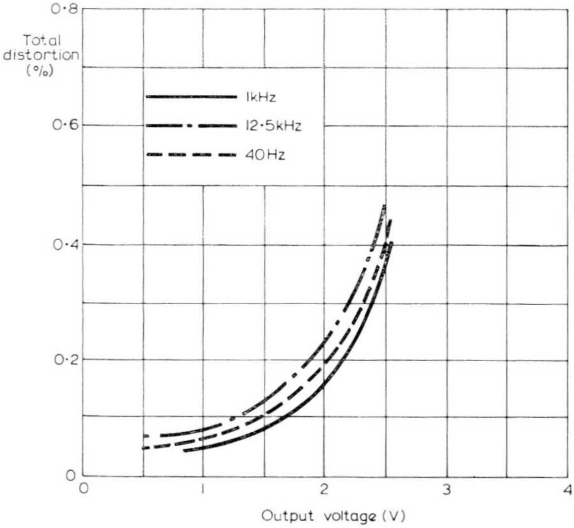
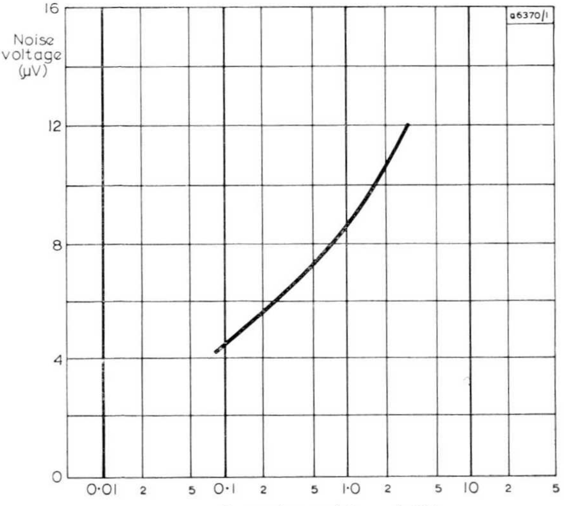


Fig. 92—Buffer amplifier circuit



(a)



(b)

Fig. 93—Total distortion and noise voltage at output of buffer amplifier

Microphone Amplifier

The circuit of a microphone amplifier with a voltage gain adjustable between 13 and 40dB by varying the feedback is shown in Fig. 94. The total distortion for the limiting values of the voltage gain is shown in Fig. 95. With an output voltage of 2V, the distortion is 0·75% for a gain of 40dB, and 0·15% for a gain of 13dB. The values of noise voltage correspond to those of the 10dB and 40dB amplifiers of Fig. 87. The input and output impedances, and the frequency response (–3dB points) are shown in Table 4.

TABLE 4
Impedances and Frequency Response

Impedance or frequency	$A_v = 13dB$	$A_v = 40dB$
Z_{in}	145kΩ	120kΩ
Z_{out}	47Ω	120Ω
f_{lower}	< 20Hz	< 20Hz
f_{upper}	≫ 20kHz	≫ 20kHz

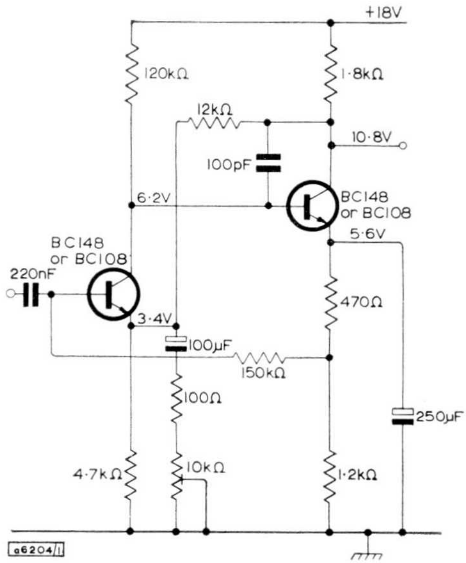
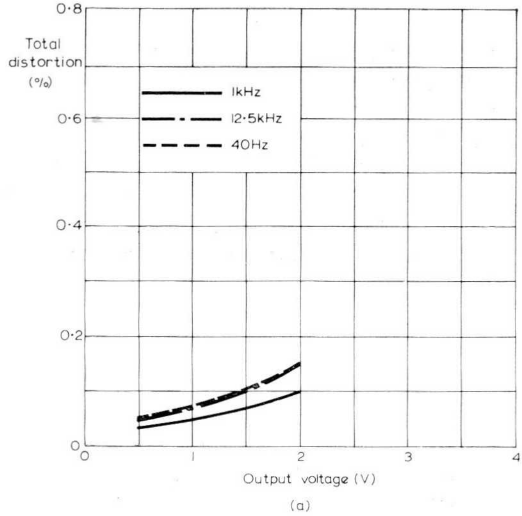


Fig. 94—Microphone amplifier circuit



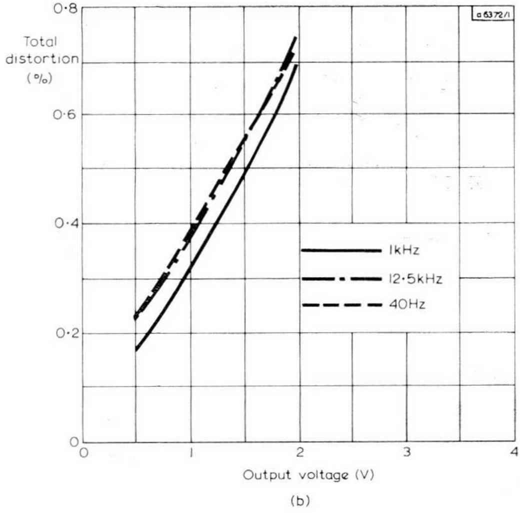


Fig. 95—Total distortion of microphone amplifier with voltage gain of (a) 13dB and (b) 40dB

Mixer Amplifier

Fig. 96 shows the circuit of a mixer amplifier, with two inputs feeding two transistors with a common collector load resistor. An emitter follower stage ensures a low output impedance of 70Ω, and the input impedance is 2·5MΩ. For both inputs, the voltage gain is unity.

Fig. 97 shows the total distortion of the mixer amplifier with a signal on one input and the other input short-circuited. The distortion is 0·5% for an output voltage of 2V, and falls to 0·1% for outputs less than 0·5V. For distortion of 0·5%, the voltage of each input must not exceed 1V, otherwise with full modulation of both inputs there will be overloading.

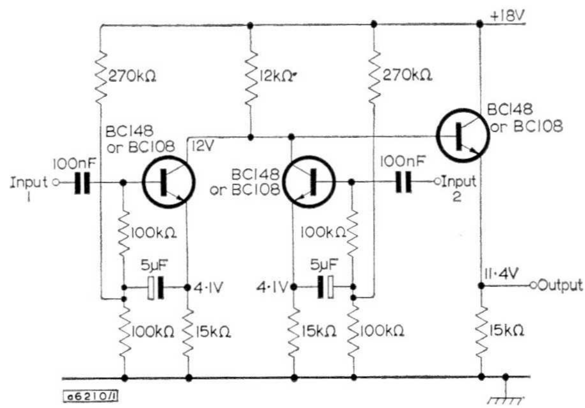


Fig. 96—Mixer amplifier circuit with two inputs

Amplifier with High Voltage Output

The circuit of an amplifier with a voltage gain of 20dB and designed for a maximum output voltage of 10V is shown in Fig. 98. To achieve this high output voltage, it is necessary for the supply voltage to be 45V instead of the standard 18V. The total distortion at 1kHz and maximum output voltage is 0·11% (Fig. 99). The frequency response (–3dB points) is from 20Hz to 20kHz, and the input and output impedances are 140kΩ and 200Ω respectively.

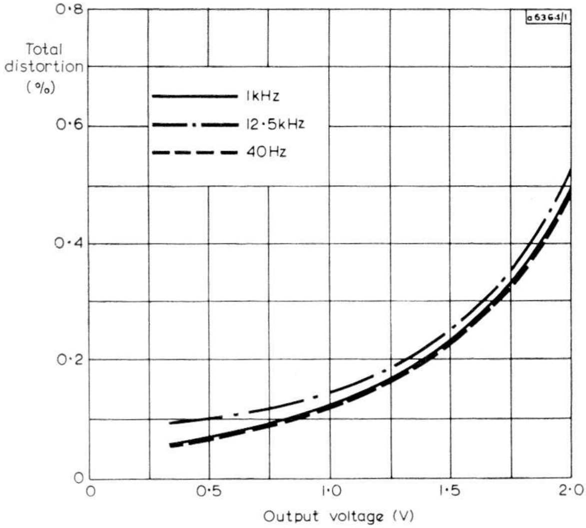


Fig. 97—Total distortion of mixer amplifier

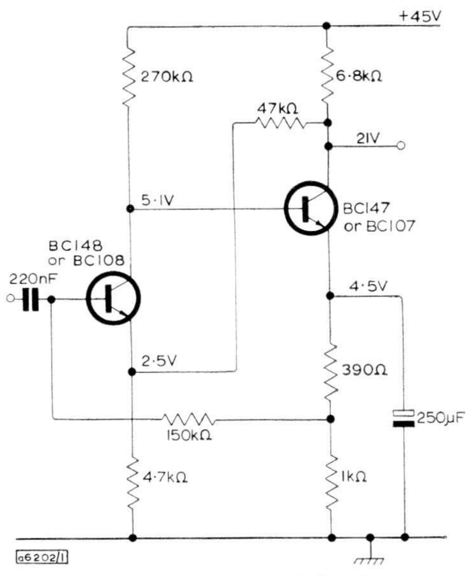


Fig. 98—Amplifier with high voltage output

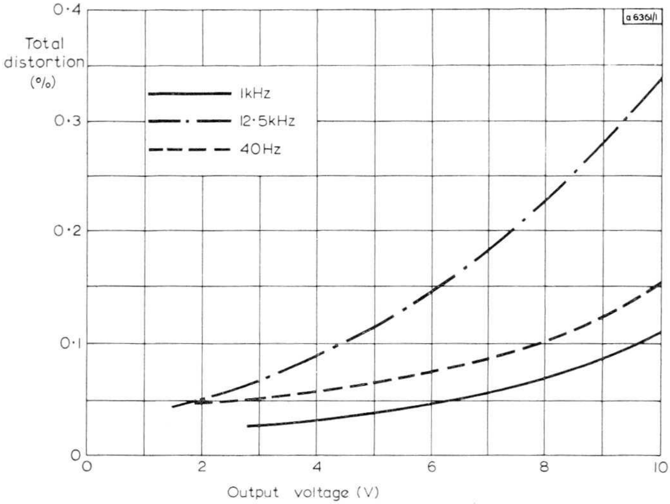


Fig. 99—Total distortion of amplifier shown in Fig. 98

AUXILIARY HIGH-QUALITY CONTROL CIRCUITS

Balance Control

The balance control of Fig. 100 makes it possible in stereophonic systems to vary the voltage gain in both channels by 6dB in opposite directions. The controlling variable resistor is inserted in the feedback circuit. The average gain is 23·4dB, and Fig. 101 shows the total distortion for minimum and maximum gain. The differences are slight because of the large amount of feedback applied. The noise voltage corresponds to that of the 20dB amplifier of Fig. 87. The frequency response (–3dB points) is from 20Hz to 20kHz, while the input and output impedances are 140kΩ and 85Ω respectively.

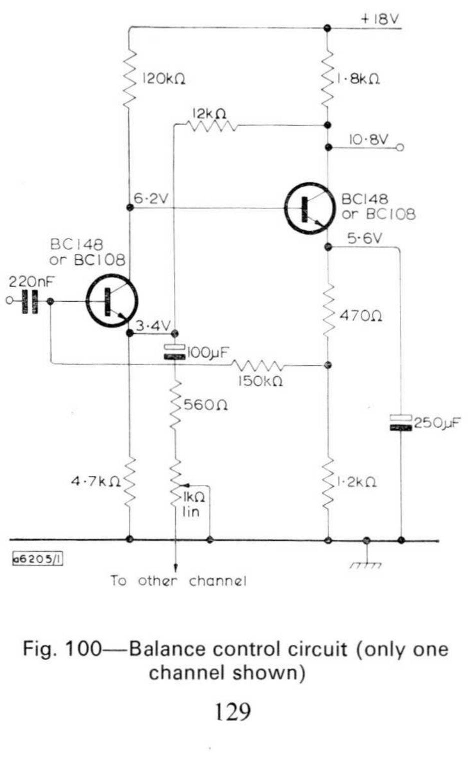
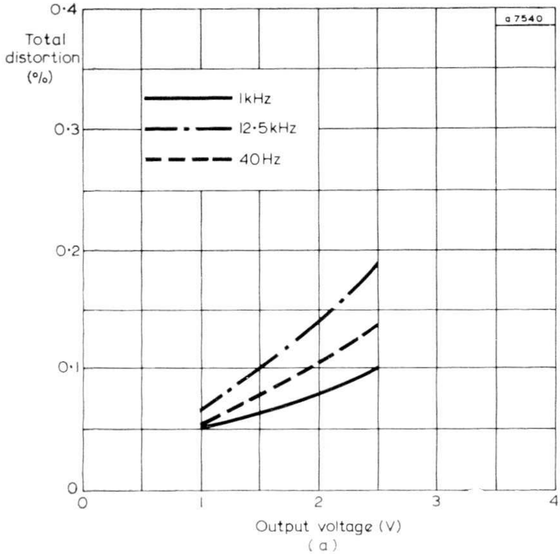


Fig. 100—Balance control circuit (only one channel shown)



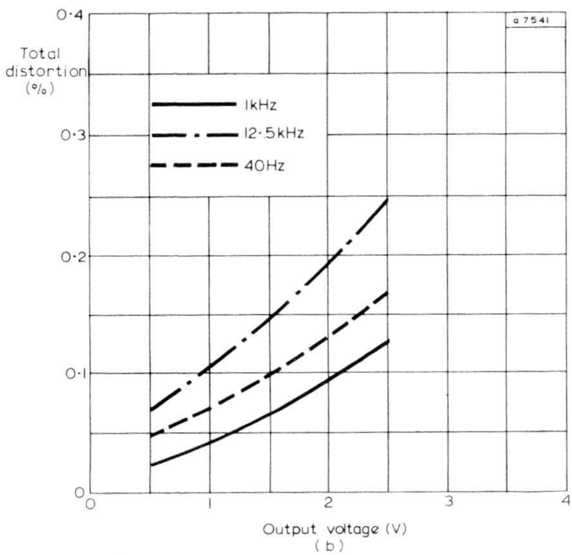


Fig. 101—Total distortion of balance control for (a) minimum and (b) maximum gain

Active Tone Control

The active tone control circuit of Fig. 102, unlike the well-known frequency-dependent voltage-divider circuits, operates with a frequency-dependent feedback network between the collector and base of the transistor. The tone control characteristics are shown in Fig. 103. The range of control extends from −22 to +19·5dB at 30Hz, and from −19 to +19·5dB at 20kHz. The flat frequency response is obtained when the variable resistors are in the physically central position. The voltage gain is then 0·91. Fig. 104 shows the total distortion as a function of output voltage for frequencies of 40Hz, 1kHz, and 12·5kHz, measured with the controls in the flat position. For small input voltages (less than 250mV), the total distortion remains below 0·1%, and for an output voltage of 2V it rises to 0·85% at 12·5kHz. The input and output impedances at 1kHz are 40kΩ and 180Ω respectively.

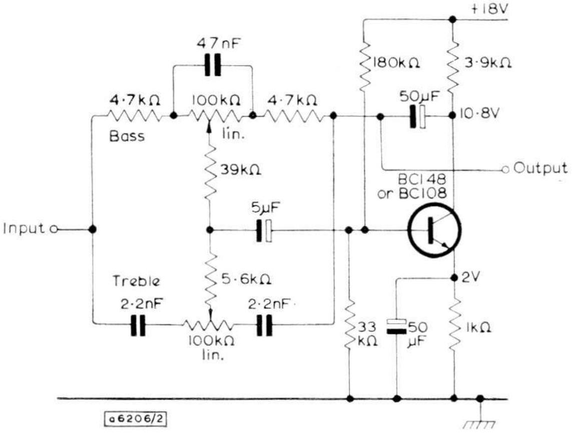


Fig. 102—Active tone control circuit

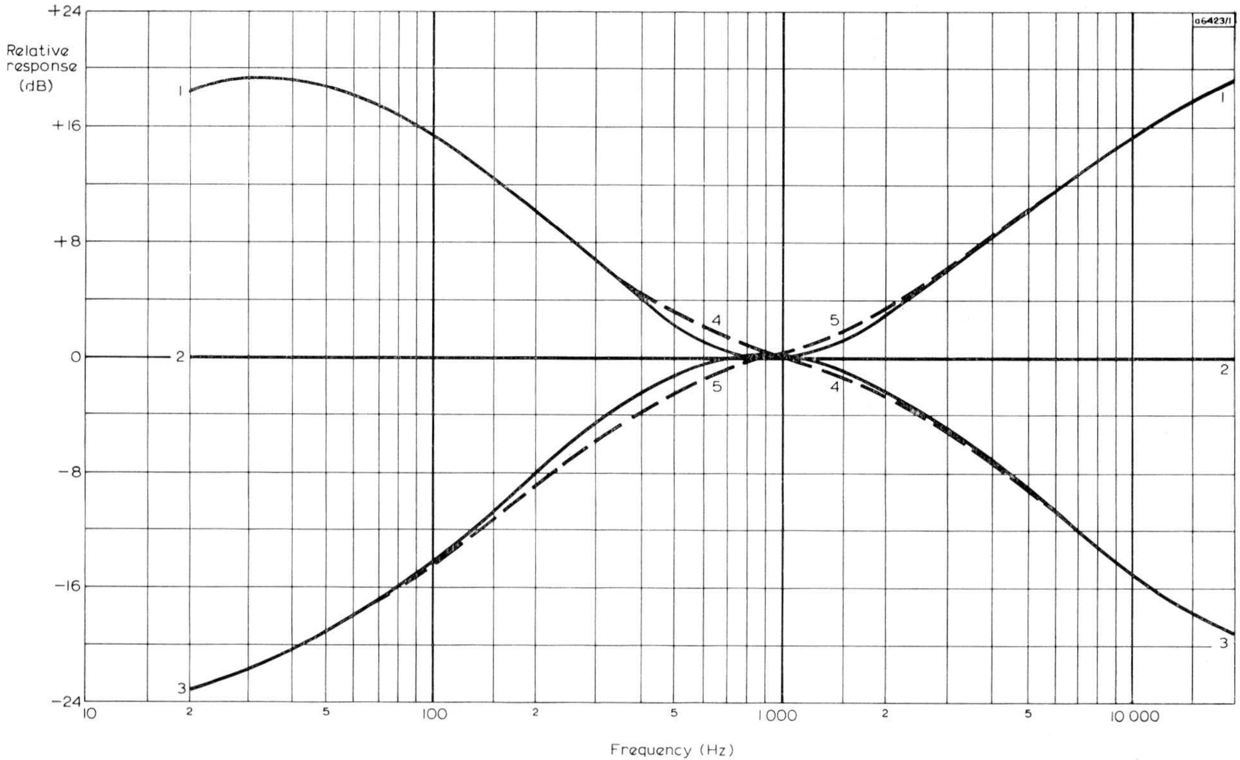


Fig. 103—Characteristics of active tone control circuit

Curve 1: maximum bass boost, maximum treble boost
Curve 2: linear frequency response (controls flat)
Curve 3: maximum bass cut, maximum treble cut
Curve 4: maximum bass boost, maximum treble cut
Curve 5: maximum bass cut, maximum treble boost

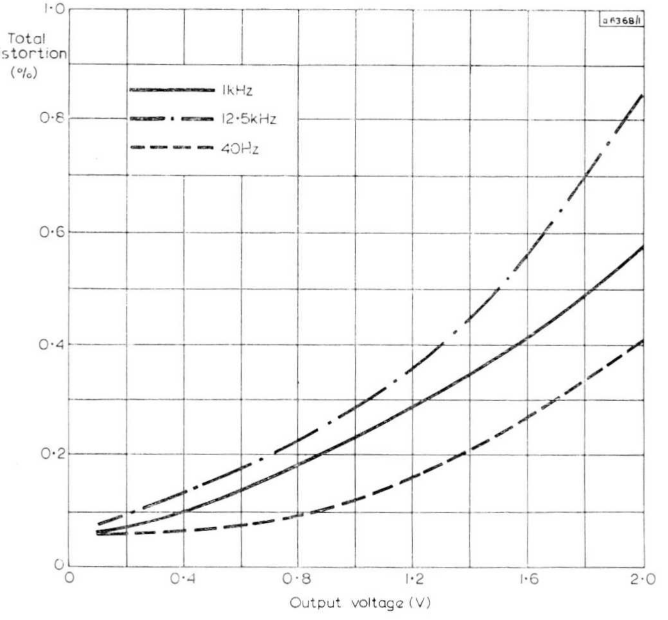


Fig. 104—Total distortion of active tone control

Sound Source Width Control

Fig. 105 shows a circuit by which the width of the sound source in a stereophonic system can be varied continuously. For this, part of the signal voltage of one channel is added to the second channel. The width control can be adjusted continuously between in-phase crosstalk of 100% (corresponding to mono operation) and anti-phase crosstalk of 24%. Greater anti-phase crosstalk is not necessary since the sound impression will 'fall apart' for greater values. A 5kΩ preset potentiometer is used to balance the two channels. The voltage gain of the circuit is 0·5. Input and output impedances are 750kΩ and 47Ω respectively, and the frequency response (–3dB points) is from 20Hz to 20kHz.

Fig. 106 shows the total distortion of the circuit as a function of output voltage for three frequencies.

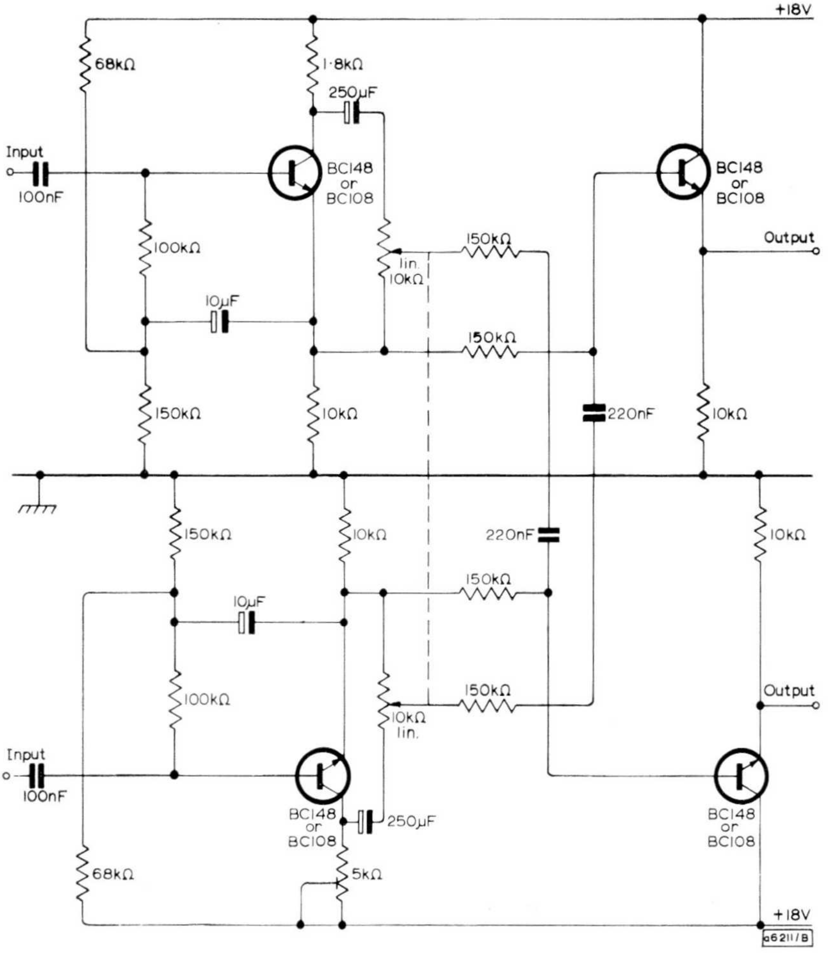


Fig. 105—Sound source width control circuit

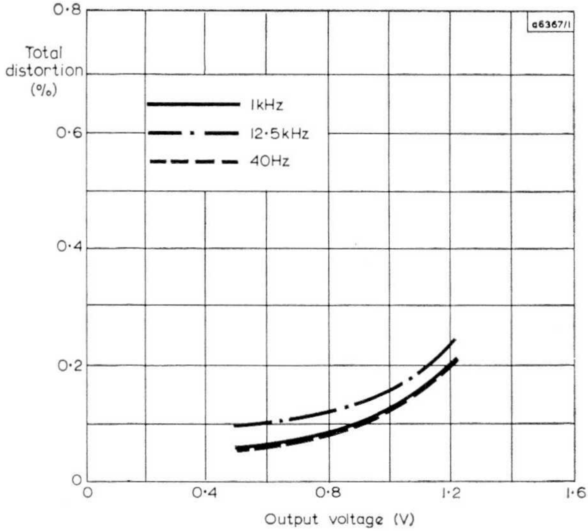


Fig. 106—Total distortion of sound source width control.

AUXILIARY HIGH-QUALITY FILTERS

Low-pass/High-pass Filter

The circuit of a low-pass/high-pass filter is shown in Fig. 107. It consists of two RC networks connected in series with a buffer amplifier, the circuit of which is given in Fig. 92. The adjustable frequency characteristics are shown in Fig. 108. The following frequency limits can be selected:

lower frequency 40, 80, 160, and 270Hz;
upper frequency 11, 9, 4·5, and 3·2kHz.

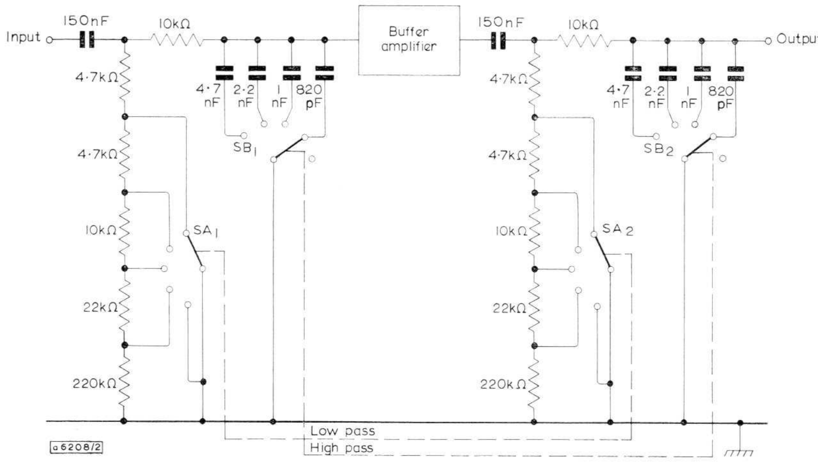


Fig. 107—Low-pass/high-pass filter circuit

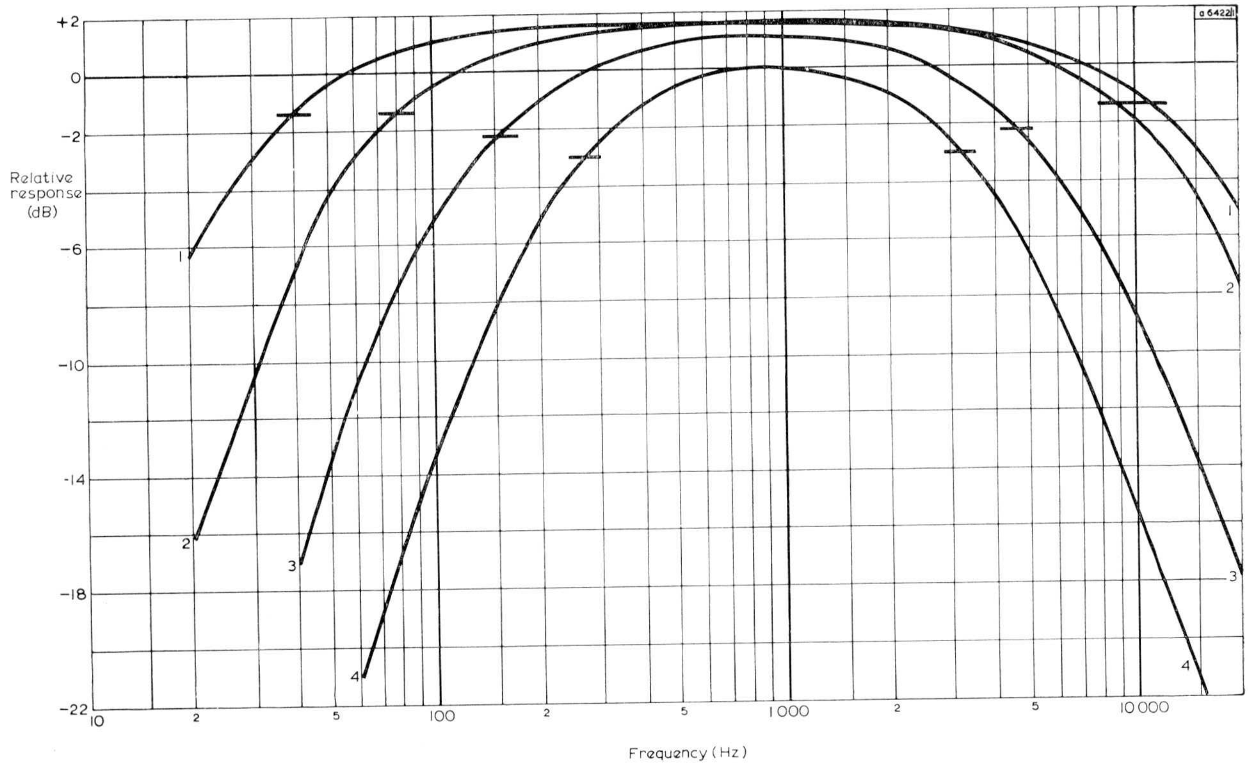


Fig. 108—Frequency characteristics of low-pass/high-pass filter
Curve 1: f_{c1} = 40Hz, f_{c2} = 11kHz
Curve 2: f_{c1} = 80Hz, f_{c2} = 9kHz
Curve 3: f_{c1} = 160Hz, f_{c2} = 4·5kHz
Curve 4: f_{c1} = 270Hz, f_{c2} = 3·2kHz
(f_{c1} = lower cut-off frequency; f_{c2} = upper cut-off frequency)

Noise and Rumble Filter

The circuit of a noise and rumble filter is shown in Fig. 109. Bass and treble cut are produced by an RC network connected between two emitter followers, and a feedback loop from the output to the input through a second RC network. In this way, a high slope of about 13dB per octave is achieved. The frequency limit of the rumble filter is fixed at 45Hz, and the noise filter can be switched to limits of 16, 12, and 7kHz. The resulting frequency characteristics are shown in Fig. 110. The voltage gain is 0·95, and the total distortion at 1kHz and an output voltage of 2V is 0·35%, falling to less than 0·1% at 1V. The input and output impedances are 1·7MΩ and 450Ω respectively.

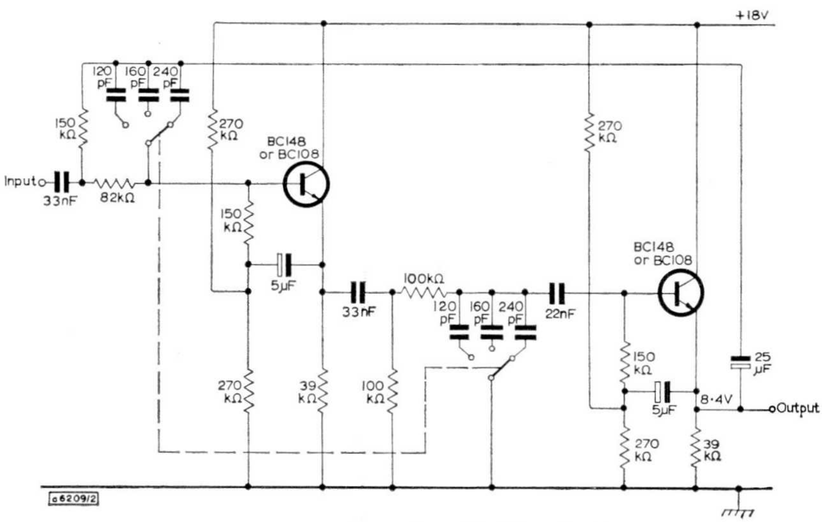


Fig. 109—Noise and rumble filter circuit

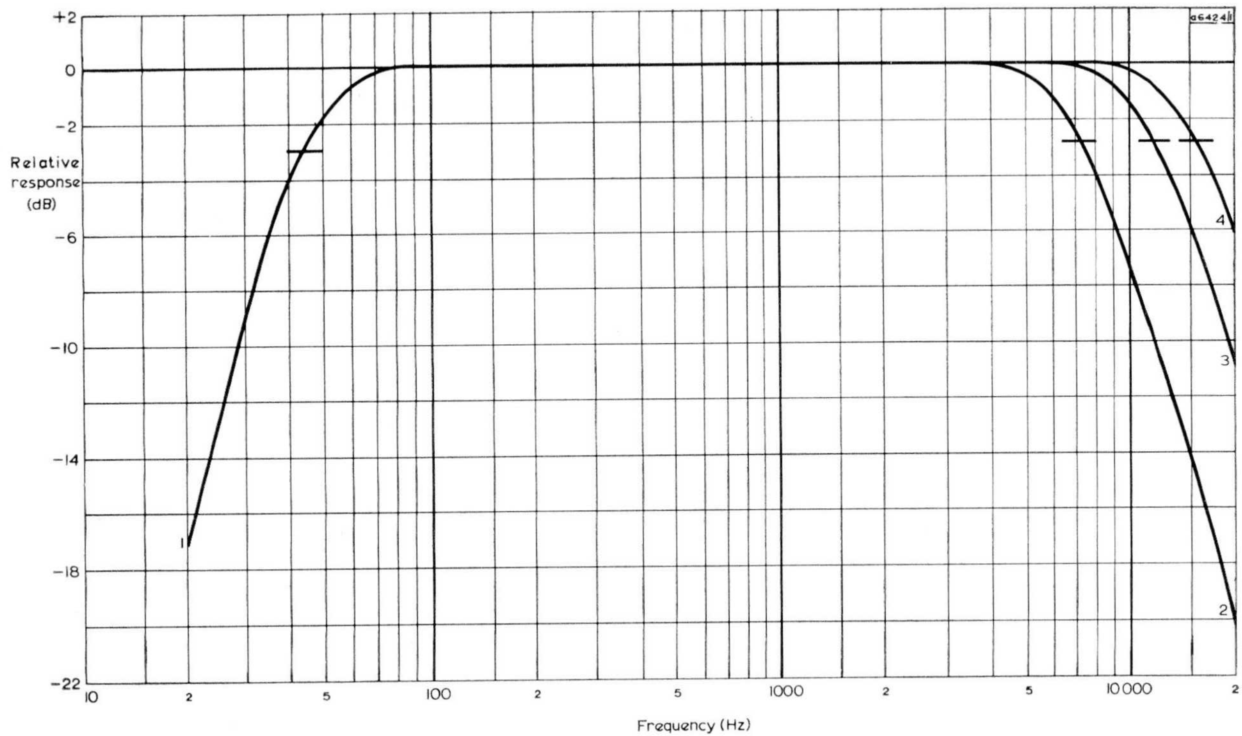


Fig. 110—Frequency characteristics of noise and rumble filter
For all curves $f_{c1} = 45$Hz
Curve 2: $f_{c2} = 7$kHz
Curve 3: $f_{c2} = 12$kHz
Curve 4: $f_{c2} = 16$kHz

COMPONENTS FOR CIRCUITS IN CHAPTER 7

10W Audio Amplifier (page 102)

Transistors and Rectifier

Circuit reference	Mullard type	Circuit reference		Mullard type
TR_1	BC158	TR_4	matched pair	BD131
TR_2	BC148 or BC108	TR_5	matched pair	BD132
TR_3	BD131	Full-wave bridge rectifier		BY164

Resistors

Tolerance: $\pm 10\%$ except R_2 and R_3, which should be $\pm 5\%$ components
Power rating: $\frac{1}{8}$W except R_{10}, which should be a 2W component, R_{11}, R_{13} and R_{14}, which should be 1W components

Circuit reference	Value	Circuit reference	Value
R_1 preset potentiometer	50kΩ	R_9 preset potentiometer	100Ω
R_2	120kΩ	R_{10}	82Ω
R_3	100kΩ	R_{11}	82Ω
R_4	22Ω	R_{12}	10Ω
R_5	10kΩ	R_{13}	0·5Ω
R_6	470Ω	R_{14}	0·5Ω
R_7	470Ω	R_{15}	10Ω
R_8	47Ω		

Resistors R_2, R_3, R_5 and R_6 should be high-stability components.

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	25μF	electrolytic 25V	C426AR/F25
C_2	16μF	electrolytic 40V	C426AR/G16
C_3	400μF	electrolytic 25V	C437AR/F400
C_4	10nF	polyester	C296AA/A10K
C_5	220nF	polyester	C296AA/A220K
C_6	1nF	polyester	C296AC/A1K
C_7	160μF	electrolytic 25V	C437AR/F160
C_8	47nF	polyester	C296AA/A47K
C_9	1250μF	electrolytic 25V	C431BR/F1250
C_{10}	2500μF	electrolytic 40V	C431BR/G2500

Mains Transformer

The transformer used in this circuit may be obtained from the following manufacturers under the type numbers given:

Colne Electric Limited, Rickmansworth, Herts	20043
Drake Transformers Limited, Billericay, Essex	294-140
Gardners Transformers Limited, Christchurch, Hants	GR 97183
Parmeko Limited, Aylestone Park, Leicester	P 3204

Heatsinks

The dissipation in the driver transistor may reach 1W and a 1·5mm (16 gauge) bright aluminium heatsink with an area of 10cm^2 should be used for this transistor. Each output transistor should be mounted on a 1·5mm bright aluminium heatsink with an area of 15cm^2.

25W Audio Amplifier (page 74)

Transistors and Diodes

Circuit reference		Mullard type
TR_1		BC158
TR_2		BFX84
TR_3		BC148 or BC108
TR_4	matched pair	BFX84
TR_5		BFX29
TR_6	matched pair	BDY20
TR_7		BDY20
D_1		BA114
D_2		BA114
D_3		BA114
Bridge rectifier		4×BYX38-300

Resistors

Tolerance: ±10% except R_2 and R_3, which should be ±5% components.
Power rating: $\frac{1}{8}$W except R_{13}, which should be a $\frac{1}{2}$W component, R_{20}, R_{21} and R_{22}, which should be 3W components.

Circuit reference	Value	Circuit reference	Value
R_1 preset potentiometer	100kΩ	R_{13}	3·9kΩ
R_2	220kΩ	R_{14}	470Ω
R_3	100kΩ	R_{15}	6·8kΩ
R_4	47kΩ	R_{16}	1kΩ
R_5	47Ω	R_{17}	1kΩ
R_6	10kΩ	R_{18}	100Ω
R_7	10kΩ	R_{19}	100Ω
R_8	3·3kΩ	R_{20}	0·5Ω
R_9 preset potentiometer	1kΩ	R_{21}	0·5Ω
R_{10}	680Ω	R_{22}	0·5Ω
R_{11}	5·6kΩ	R_{23}	10Ω
R_{12}	470Ω		

Resistors R_2, R_3, R_4 and R_{11} should be high-stability components.

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	5μF	electrolytic 64V	C426AR/H5
C_2	2·5μF	electrolytic 64V	C426AR/H2·5
C_3	160μF	electrolytic 25V	C437AR/F160
C_4	33nF	polyester	C296AA/A33K
C_5	1nF	polyester	C296AC/A1K
C_6	1·5nF	polyester	C296AC/A1K5
C_7	32μF	electrolytic 40V	C426AR/G32
C_8	4·7nF	polyester	C296AC/A4K7
C_9	220nF	polyester	C296AA/A220K
C_{10}	1250μF	electrolytic 40V	C431BR/G1250
C_{11}	2500μF	electrolytic 64V	C431BR/H2500

Mains Transformer

The transformer used in this circuit may be obtained from the following manufacturers under the type numbers given:

Colne Electric Limited, Rickmansworth, Herts	20044
Drake Transformers Limited, Billericay, Essex	786-174
Gardners Transformers Limited, Christchurch, Hants	GR 97184
Parmeko Limited, Aylestone Park, Leicester	P 3205

Heatsinks

With a 4Ω load, the dissipation in each output transistor can be 18W. The maximum permissible junction temperature of the BDY20 is 200°C and the thermal resistance between junction and mounting base is 1·5degC/W. Therefore, with a mounting which has a thermal resistance of 0·5degC/W, and a maximum ambient temperature of 45°C, the thermal resistance of the heatsink must not exceed 6·5degC/W. This can be achieved with 80cm² of 2mm (14 gauge) bright aluminium or 50cm² of 2mm blackened aluminium for each transistor. As the dissipation of each driver transistor may be 1W, the driver transistors should be mounted on small heatsinks, such as the Redpoint 5F heatsink.

Pre-amplifier (page 109)

Transistors

Circuit reference	Mullard type
TR_1	BC149 or BC109
TR_2	BC148 or BC108
TR_3	BC149 or BC109

Resistors

Tolerance: $\pm 10\%$

Power rating: $\frac{1}{8}$W

Circuit reference	Value 10W circuit	Value	Value 25W circuit
R_1		1·8MΩ	
R_2		see text	
R_3		56kΩ	
R_4		220kΩ	
R_5		22kΩ	
R_6	100kΩ		220kΩ
R_7	100kΩ		220kΩ
R_8	12kΩ		18kΩ
R_9		220kΩ	
R_{10}		680Ω	
R_{11}	560Ω		820Ω
R_{12}	6·8kΩ		3·3kΩ
R_{13}	12kΩ		6·8kΩ
R_{14}	22kΩ		10kΩ
R_{15}	560kΩ		150kΩ
R_{16}	10kΩ		4·7kΩ
R_{17} logarithmic potentiometer		10kΩ	
R_{18} logarithmic/reverse logarithmic potentiometer		10kΩ	
R_{19}		22kΩ	
R_{20} linear potentiometer		250kΩ	
R_{21} linear potentiometer		100kΩ	
R_{22}		22kΩ	
R_{23}		15kΩ	
R_{24}		47kΩ	
R_{25}		1kΩ	
R_{26}		47kΩ	
R_{27}		220kΩ	
R_{28}		10kΩ	
R_{29}		15kΩ	
R_{30}		220kΩ	
R_{31}		3·3kΩ	
R_{32}		6·8kΩ	
R_{33}	3·9kΩ		18kΩ
R_{34}	—		15kΩ
R_{35}	—		4·7kΩ

Resistors R_6, R_7, R_8 and R_9 should be high-stability components.

Capacitors

Circuit reference	Value 10W circuit	Value 25W circuit	Description	Mullard type 10W circuit	Mullard type 25W circuit
C_1	6·4μF		electrolytic 25V	C426AR/F6·4	
C_2	80μF		electrolytic 2·5V	C426AR/A80	
C_3	2·5μF		electrolytic 25V	C426AR/H2·5	
C_4	10nF	22nF	metallised foil	C281AB/A10K	C281AB/A22K
C_5	3·3nF	6·8nF	polyester	C296AC/A3K3	C296AC/A6K8
C_6	10nF	22nF	metallised foil	C281AB/A10K	C281AB/A22K
C_7	10μF		electrolytic 16V	C426AR/E10	
C_8	2·5μF		electrolytic 25V	C426AR/H2·5	
C_9	80μF		electrolytic 25V	C426AR/F80	
C_{10}	15nF		metallised foil	C281AB/A15K	
C_{11}	15nF		metallised foil	C281AB/A15K	
C_{12}	10μF		electrolytic 16V	C426AR/E10	
C_{13}	3·3nF		polyester	C296AC/A3K3	
C_{14}	3·3nF		polyester	C296AC/A3K3	
C_{15}	10μF		electrolytic 16V	C426AR/E10	
C_{16}	6·8nF		polyester	C296AC/A6K8	
C_{17}	470nF		metallised foil	C281AB/A470K	
C_{18}	10μF		electrolytic 16V	C426AR/E10	
C_{19}	470nF		metallised foil	C281AB/A470K	
C_{20}	6·8nF		polyester	C296AC/A6K8	
C_{21}	2·2nF		polyester	C296AC/A2K2	
C_{22}	10μF		electrolytic 16V	C426AR/E10	

CHAPTER 8

HIGH-QUALITY F.M. TUNERS

Two f.m. tuners are described, the first using a BF200 silicon planar transistor in the r.f. stage and the second using a BFW10 field-effect transistor. The tuners may be used with the f.m. stages of the i.f. amplifier described in Chapter 4 (page 52).

A comparison of the two tuners will reveal that the circuits give very similar performances. However, it should be noted that the f.e.t. tuner achieves this performance at the expense of a relatively high supply voltage and high operating current in the r.f. stage. In addition, the field-effect transistor must be selected for a value of I_{DSS} (drain current at zero source-gate voltage) of 8 to 20mA.

The chapter also includes a stereo decoder circuit for reception of stereophonic broadcasts.

The difficulties associated with the construction and lining-up of r.f. and i.f. amplifiers must again be stressed. Only those constructors with previous experience in this type of circuitry and with access to specialised equipment should consider building the circuits described in this chapter.

Component lists for the circuits are given at the end of the chapter.

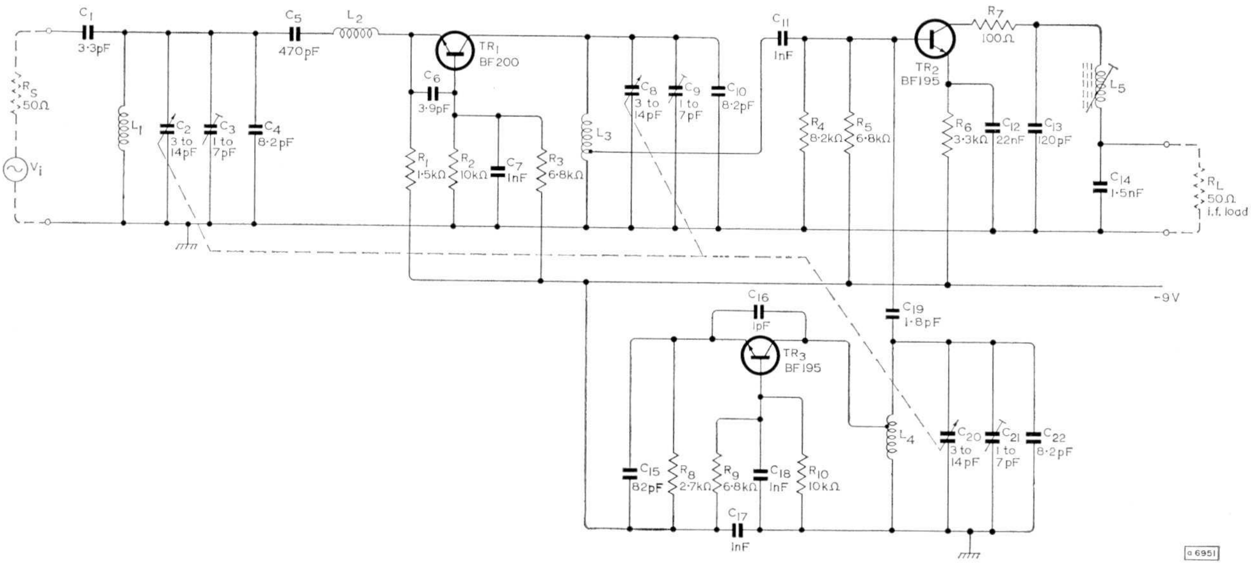


Fig. 111—Three-ganged, capacitively-tuned f.m. tuner circuit

F.M. TUNER USING BF200 SILICON PLANAR TRANSISTOR

High-quality f.m. reception calls for a tuner having a low noise factor and excellent signal-handling ability. These objectives can be achieved by exploiting the inherently good signal-handling capability of the BF200 in a circuit configuration favouring optimum noise performance.

This high-performance tuner uses a tuned aerial circuit matched to the r.f. transistor by a fixed inductance. By this means it is possible to retain the benefits of a low-loss selective aerial circuit, which can be approximately power matched, and, through an impedance transform, to allow the transistor to be matched for optimum noise performance. The transistor is then required to handle less signal for a given noise performance.

Circuit Description

The complete circuit of a three-ganged capacitively-tuned f.m. tuner is given in Fig. 111.

R.F. Stage

The BF200 transistor, TR_1, operates in the common-base configuration at a collector current of 2mA. The optimum source impedance for minimum noise is equivalent to the parallel combination of 125Ω resistance and 500Ω inductance, giving a typical noise factor of 2dB.

The emitter resistor R_1, inductance L_2 and capacitor C_6 cause some additional losses which result in a typical noise factor for the tuner of about 4·5dB. Capacitor C_6 prevents parasitic oscillation at very high frequencies and must be mounted between the base and emitter leads of the BF200 and as closely as possible to them.

The low feedback capacitance of the BF200 allows the collector to be connected to the top of the tuned interstage circuit. The collector output damping of the BF200 is negligible.

The transducer gain of the r.f. stage is about 16dB at a 3dB bandwidth of the loaded interstage circuit of 1MHz. The value of Q_O of this tuned circuit is 200.

Mixer Stage

The mixer transistor, TR_2, is a Lockfit transistor, type BF195, and operates in the common-emitter configuration at a collector current of 1mA. At this current and with a source impedance of about 100Ω, the contribution of the mixer stage to the total noise factor is small—approximately 0·5dB.

By keeping the oscillator signal at the base of TR_2 to as low a value as possible—about 50mV—harmonics of the oscillator frequency at the input to the mixer are virtually eliminated, while sufficient conversion gain is maintained.

Because the base of TR_2 has an inductive tap on the interstage circuit, no additional 10·7MHz i.f. series filter is necessary between the base and the emitter. To prevent i.f. parasitic oscillations, a 100Ω resistor, R_7, is connected in series with the collector.

The value of Q_O of the i.f. tuned circuit is about 150 and the value of Q_L is 50. The mixer gain is about 18dB.

Oscillator Stage

The oscillator transistor, TR_3, is also a BF195, and operates in the common-base configuration at a current of 1mA. The coupling between the top of the oscillator coil and the base of the mixer is obtained via capacitor C_{19}. The low value of this capacitance—1·8pF—guarantees that large aerial signals have practically no effect on the oscillator circuit. The oscillator frequency is below the aerial frequency.

Performance

In obtaining the performance data given in this section, the aerial and i.f. trimmer capacitors were adjusted at each frequency to give no padding error at the measuring point.

Transducer Gain

Transducer gain is defined as the ratio of the power delivered to the load to the power supplied into the stage if it were power-matched to the source, that is if the input impedance were equal to the source impedance.

$$\text{Transducer gain} = \frac{V_O^2}{R_L} \cdot \frac{4R_S}{V_S^2}$$

where V_O = voltage developed across load
R_L = load impedance
R_S = source impedance of generator
V_S = e.m.f. of generator

The values of transducer gain at three tuning frequencies are given in Table 5. A curve of transducer gain at a tuning frequency of 98MHz plotted against supply voltage is given in Fig. 112.

TABLE 5

Measured Values of Transducer Gain

Tuning Frequency (MHz)	Transducer Gain (dB)
90	32
98	35
105	35

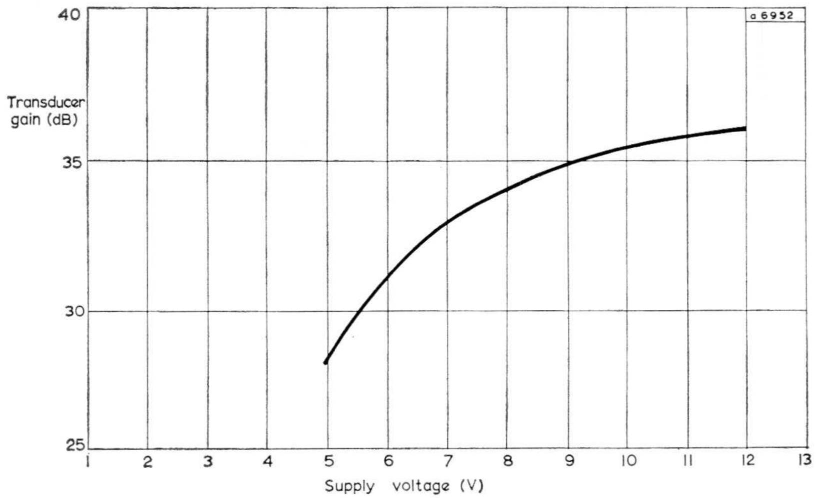


Fig. 112—Variation of transducer gain with supply voltage at a tuning frequency of 98MHz

Noise Factor

The values of noise factor at three tuning frequencies are given in Table 6.

TABLE 6

Measured Values of Noise Factor

Tuning Frequency (MHz)	Noise Factor (dB)
90	4·7
98	4·5
105	4·3

Repeat-spot Suppression

Repeat spots are caused by harmonics of the oscillator signal mixing with those of a strong aerial signal resulting in a signal having the intermediate frequency. For example, if an unwanted aerial signal of 88MHz is present, its second harmonic (176MHz) mixes with the second harmonic (165·3MHz) of an oscillator signal of 82·65MHz to give 10·7MHz—the intermediate frequency. The 82·65MHz oscillator signal corresponds to an aerial signal of 93·35MHz and so the 88MHz aerial signal will be received when the dial is set to 93·35MHz.

Repeat-spot suppression is sometimes known as half i.f. rejection. The values of repeat-spot suppression are given for two input levels at each of three tuning frequencies in Table 7. The value of R_S is 50Ω throughout. A graph of repeat-spot suppression against input voltage at a tuning frequency of 98MHz is given in Fig. 113. From this graph it can be seen that the repeat-spot suppression improves by 1dB for a decrease in signal of 2dB.

TABLE 7
Measured Values of Repeat-spot Suppression

Tuning Frequency (MHz)	Input Voltage (μV)	Repeat-spot Suppression (dB)
90	40	73
90	1	89
98	40	71
98	1	87
105	40	70
105	1	86

Double-beat Suppression

Double beats occur if two strong aerial signals have frequencies or harmonics which combine with the oscillator signal, or its harmonics, to give the intermediate frequency. For example, if two strong unwanted aerial signals with frequencies of 94MHz and 96·7MHz are present, and the dial is set to 100·7MHz, the audio signals associated with the two strong signals will be heard superimposed upon each other. This is because the dial setting of 100·7MHz corresponds to an oscillator frequency of 90MHz and the second harmonic of this oscillator signal (180MHz) combines with the two strong unwanted aerial signals (94+96·7—180) to give 10·7MHz—the intermediate frequency.

The values of double-beat suppression given in Table 8 were measured with two identical interference signals at 95 and 97·5MHz. Double-beat reception is possible at the tuning frequencies 92·5, 100 and 101·6MHz and at these frequencies, measurements of double-beat suppression for a required signal level of 40μV are given. The value of R_S is 50Ω. A graph of double-beat suppression against input voltage is given in Fig. 114.

TABLE 8
Measured Values of Double-beat Suppression

Tuning Frequency (MHz)	Double-beat Suppression (dB)
92·5	63
100	66
101·6	65

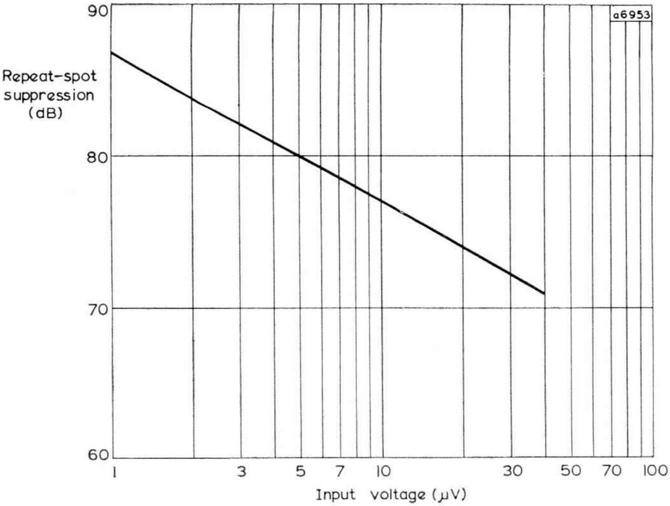


Fig. 113—Variation of repeat-spot suppression with input voltage at a tuning frequency of 98MHz

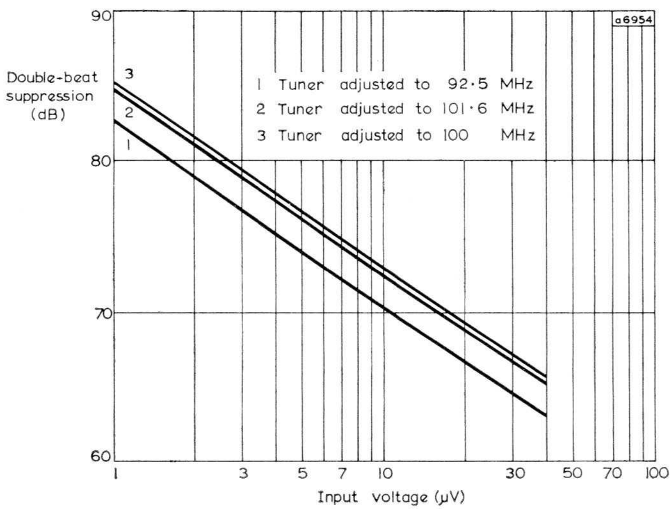


Fig. 114—Variation of double beat suppression with input voltage

Continuous-beat Suppression

Continuous-beats occur when two strong aerial signals have frequencies, or harmonics, which combine to give the intermediate frequency. If continuous beats occur, they occur regardless of the dial setting. For example, if two strong aerial signals are present with frequencies of 90MHz and 95·35MHz, their second harmonics (180MHz and 190·7MHz respectively) combine to produce 10·7MHz—the intermediate frequency.

The continuous-beat suppression, measured at a tuning frequency of 98MHz, is 67dB at input voltage levels of 40μV for both signals.

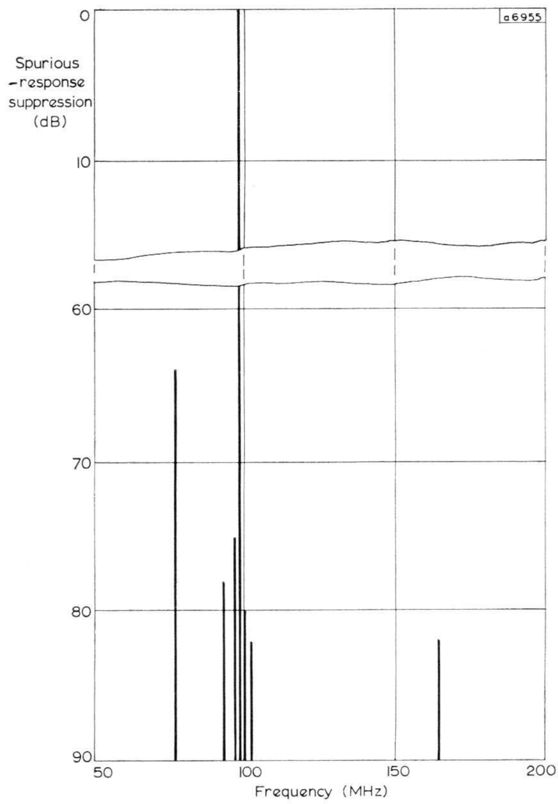


Fig. 115—Histogram of spurious-response suppression

Spurious-response Suppression

The one-signal spurious-response suppression is given in histogram form in Fig. 115. The tuner frequency is 98MHz and the reference level of the required input signal is 8μV. The spikes in Fig. 115 show that at a certain input frequency, an i.f. output signal occurs. The input signal level at this frequency must be many times higher to obtain the same i.f. output signal as for a wanted input signal of 8μV. The ratio of the required interference input signal level and the required input signal of 8μV is given as spurious-response suppression. This ratio is indicated by the top level of the spikes in Fig. 115.

Image Rejection

The image rejection is determined by the selectivity of the aerial at a frequency different from the tuned frequency, and by interstage circuit quality factors. In the tuner described, the image-rejection frequency is $2 \times 10{\cdot}7$MHz below the tuning frequency. The values of image rejection at three tuning frequencies are given in Table 9.

TABLE 9
Measured Values of Image Rejection

Tuning Frequency (MHz)	Image Rejection (dB)
90	63
98	64
105	63

I.F. Rejection

The values of i.f. rejection at three tuning frequencies are given in Table 10.

TABLE 10
Measured Values of I.F. Rejection

Tuning Frequency (MHz)	I.F. Rejection (dB)
90	64
98	63
105	61

Signal Handling

When the value of R_S is 50Ω, signals of more than 1V can be handled. The shift of the oscillator frequency is not more than 20kHz for signals up to 1V.

Oscillator Voltage on Mixer

The oscillator voltage at the base of the mixer is kept low. The values of oscillator voltage at three tuning frequencies are given in Table 11.

TABLE 11
Measured Values of Oscillator Voltage at Base of Mixer

Tuning Frequency (MHz)	Oscillator Voltage (mV)
90	35
98	50
105	60

Frequency Shift of Oscillator

The oscillator frequency shift due to supply voltage variations is about 20kHz per 2V supply voltage variation.

Oscillator Radiation

When the value of R_S is 50Ω, the oscillator radiation at the input terminals of the tuner is 300μV.

F.M. TUNER USING BFW10 FIELD-EFFECT TRANSISTOR

The circuit for a high-quality, two-ganged, capacitively-tuned f.m. tuner is shown in Fig. 116. The tuner uses a BFW10 field-effect transistor in the r.f. input stage, and BF195 transistors in the mixer and oscillator stages. Because it has a square-law input characteristic, the field-effect transistor is particularly useful in this circuit, where a high degree of spurious-response suppression is required. The low noise factor and high input impedance of the device are also advantageous in this type of circuit.

Circuit Description

R.F. Stage

The BFW10 field-effect transistor operates at a typical drain current of 10mA and typical drain-source voltage of 6V. The optimum source impedance for minimum noise is equivalent to the parallel combination of 910Ω resistance and 370Ω inductance, giving a typical noise factor of 2·5dB.

The aerial input circuit is transformer-coupled to the input of the field-effect transistor, the unloaded quality factor, Q_O, of the aerial transformer being about 80 at a secondary inductance of 330nH.

The transducer gain of the r.f. stage is about 8·5dB at 100MHz. The value of Q_O for the interstage circuit is 200.

The d.c. stabilisation is obtained by a voltage divider for the gate electrode and a source resistance of value 680Ω. A limitation of this design is that the minimum value of I_{DSS} for the field-effect transistor is 8mA because the gate of the transistor may not be positive with respect to the source.

Mixer and Oscillator Stages

Comparison of the circuit diagram of this tuner with that of the planar-transistor version shown on page 146 will show that the mixer and oscillator stages of the two circuits are almost identical. The description given on pages 147 and 148 applies equally to this tuner.

Performance

For definitions of terms transducer gain, repeat-spot suppression, double-beat suppression, continuous-beat suppression, spurious-response suppression and image rejection, see pages 148 to 153.

Transducer Gain

The values of transducer gain measured at three tuning frequencies are given in Table 12. In Fig. 117 the transducer gain at a tuning frequency of 98MHz is plotted as a function of supply voltage.

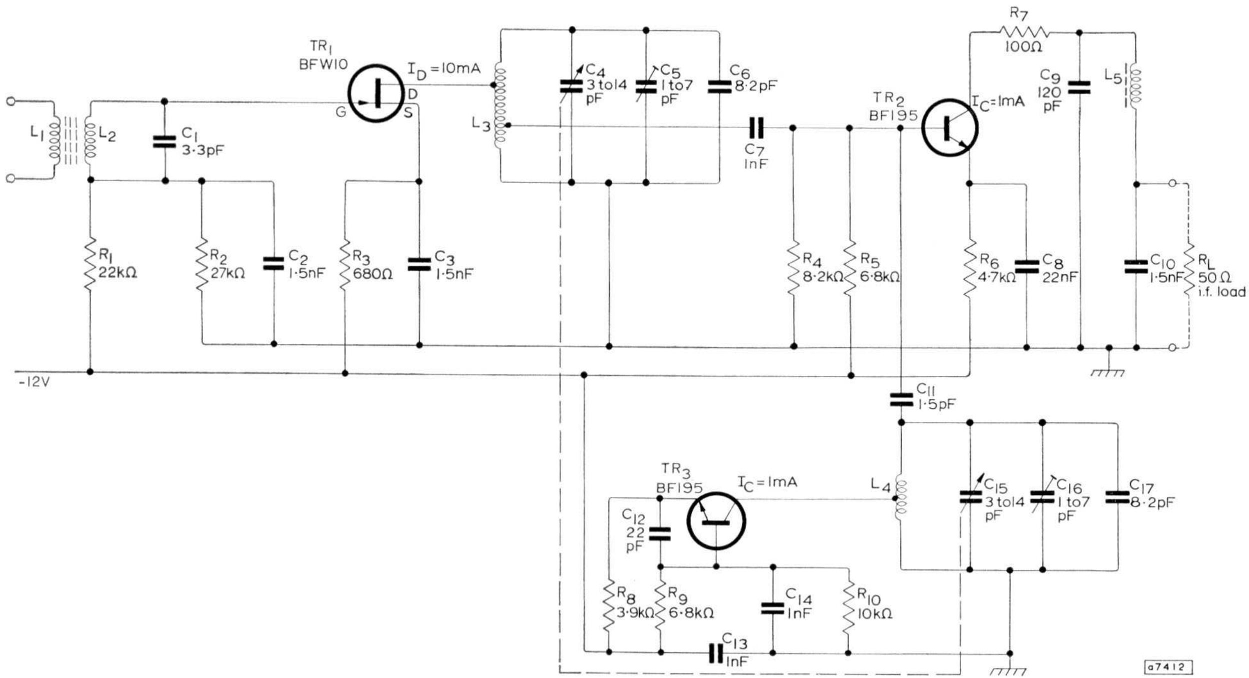


Fig. 116—Two-ganged, capacitively-tuned f.m. tuner circuit

TABLE 12
Measured Values of Transducer Gain

Tuning Frequency (MHz)	Transducer Gain (dB)
90	26
98	30
106	29

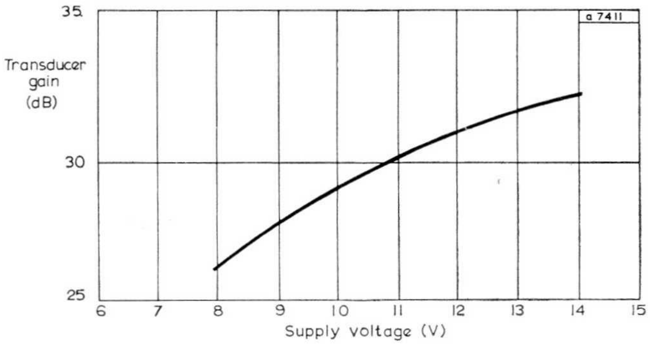


Fig. 117—Variation of transducer gain with supply voltage at a tuning frequency of 98MHz

Noise Factor

The values of noise factor measured at three tuning frequencies are given in Table 13.

TABLE 13
Measured Values of Noise Factor

Tuning Frequency (MHz)	Noise Factor (dB)
90	6·9
98	4·2
106	4·1

In this circuit, the noise contribution of the mixer stage rises when the power gain of the r.f. stage decreases. Because of the selectivity of the aerial input circuit, the power gain of the r.f. stage at 90MHz is 3dB below that at 98MHz, hence the increase in noise factor at the lower frequency.

Repeat-spot Suppression

The values of repeat-spot suppression at three tuning frequencies and an aerial signal level of 100μV are given in Table 14.

TABLE 14

Measured Values of Repeat-spot Suppression

Tuning Frequency (MHz)	Repeat-spot Suppression (dB)
90	68
98	61
106	58

For an aerial signal level of 1μV the repeat-spot suppression would be about 80dB.

Double-beat Suppression

The values of double-beat suppression at three tuning frequencies are given in Table 15.

TABLE 15

Measured Value of Double-beat Suppression

Tuning Frequency (MHz)	Double-beat Suppression (dB)
92·5	60
100	57
101·6	59

Continuous-beat Suppression

The values of continuous-beat suppression at three tuning frequencies with both aerial signals at a level of 100μV are given in Table 16.

TABLE 16

Measured Values of Continuous-beat Suppression

Tuning Frequency (MHz)	Continuous-beat Suppression (dB)
90	65
98	60
106	56

Spurious-response Suppression

The one-signal spurious-response suppression is given in histogram form in Fig. 118. The tuning frequency is 98MHz. The reference level of the required input signal is 20μV.

Image Rejection

The values of image rejection at three tuning frequencies is given in Table 17.

TABLE 17

Measured Values of Image Rejection

Tuning Frequency (MHz)	Image Rejection (dB)
90	42
98	43
106	38

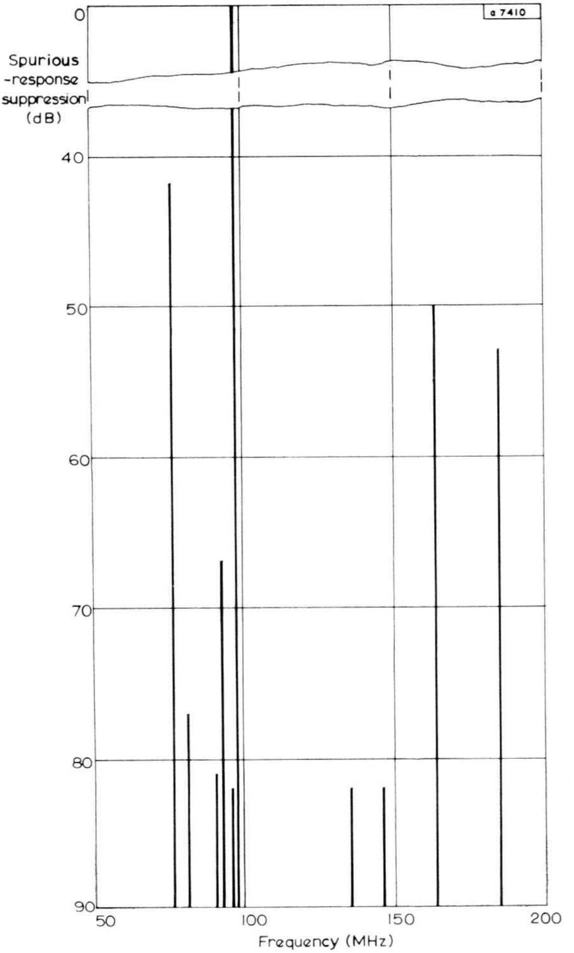


Fig. 118—Histogram of spurious-response suppression

I.F. Rejection

The values of i.f. rejection at three tuning frequencies are given in Table 18.

TABLE 18
Measured Value of I.F. Rejection

Tuning Frequency (MHz)	I.F. Rejection (dB)
90	76
98	76
106	75

Signal Handling

With a value of source impedance of 300Ω, signals of more than 1V can be handled. The shift of oscillator frequency is not more than 20kHz for signals up to 1V.

Oscillator Voltage on Mixer

The oscillator voltages at the base of the mixer at three tuning frequencies are given in Table 19.

TABLE 19
Oscillator Voltage at Base of Mixer

Tuning Frequency (MHz)	Oscillator Voltage (mV)
90	40
98	60
106	80

Frequency Shift of Oscillator

The frequency shift of the oscillator due to a supply voltage variation of 1V is less than 20kHz. The lower limit of supply voltage for oscillator operation is 4V.

STEREO DECODER

Stereo Transmissions

Stereophonic programmes are regularly broadcast on the v.h.f. transmissions of BBC Radio Three. For these transmissions to be received stereophonically, certain equipment in addition to the normal v.h.f. receiver is required. This additional equipment consists of two separate audio amplifiers and loudspeakers to amplify the left- and right-hand channels and some means of separating—or decoding—the channels.

The system the BBC, in common with other European broadcasting authorities, uses is the pilot-tone (Zenith-G.E.) system, with a 50μs pre-emphasis time-constant instead of the 75μs used in the U.S.A. In this system, the left- and right-hand channels are encoded at the transmitter by a multiplexing process and transmitted on a single wavelength. The channels are decoded after detection in the receiver. The system is compatible; that is, listeners with unadapted v.h.f. receivers can receive an acceptable monophonic representation of the stereophonic broadcast.

The sum of the left- and right-hand channels provides a satisfactory monophonic signal which can be transmitted and received in the normal way. For stereophonic reception, however, additional information is necessary to enable the stereo receiver to separate the left- and right-hand signals. In the Zenith-G.E. system, this extra information consists of a 'difference' signal—left minus right—which is transmitted as the upper and lower sidebands of a 38kHz subcarrier. The 38kHz subcarrier is itself totally suppressed.

To ensure synchronisation between the decoder in the receiver and the encoder at the transmitter, a low-level synchronising signal, or pilot tone, is also transmitted. The frequency of the pilot tone is 19kHz.

Circuit Description

The circuit diagram of a decoder is shown in Fig. 119. The pilot tone is extracted from the multiplex input by means of a 19kHz circuit immediately following the discriminator output. The pilot tone is then amplified by the first transistor and applied to the base of the second transistor. The second transistor is biased to operate in class B; that is, only the positive half-cycles of the amplified pilot tone are amplified. The distorted waveform resulting from this type of amplification is very rich in second harmonic content, and this second harmonic (38kHz) is extracted by means of a tuned circuit at the collector of the second transistor. The peak-to-peak amplitude of the 38kHz signal, between the collector end of the coil and the tap on the coil, is limited to approximately twice the supply voltage; hence its amplitude is constant regardless of input level once a certain value has been exceeded.

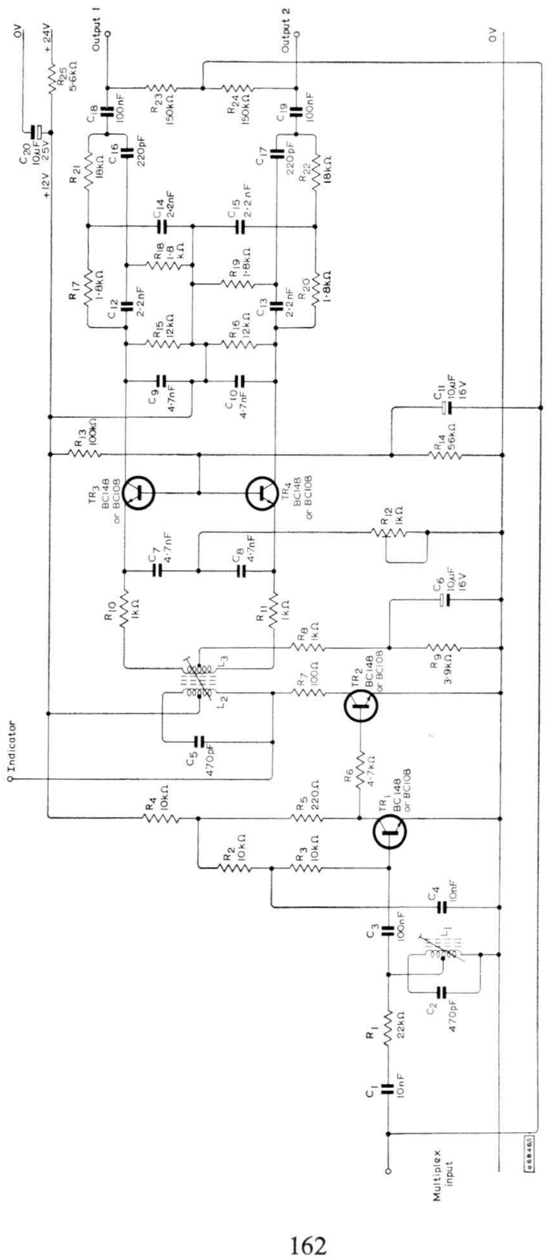


Fig. 119—Stereo decoder circuit

A centre-tapped secondary winding applies antiphase switching signals to the emitters of two transistors and the complete multiplex information is applied to their bases. With this arrangement, synchronous detection to obtain the 'difference' signal, along with 'matrixing' of the 'sum' and 'difference' signals, occurs and the two separated left- and right-hand signals appear—one at each collector. With a simple decoder operating on the principle described, only a limited separation is obtainable, so arrangements have been made in the design of the circuit shown in Fig. 119 to improve the separation. This is done by deliberately introducing a controlled amount of crosstalk into the two outputs in antiphase to the inherent crosstalk, thereby cancelling the crosstalk.

Conventional de-emphasis occurs in the collector circuits of the two switching transistors but this is insufficient to remove the 38kHz waveform from the output. As the 38kHz waveform is of sufficient amplitude to constitute a nuisance, in that it may give rise to distortion in any following audio amplifier or to beating with the oscillator frequency of a tape recorder, it is removed by passing each output through a parallel-T filter tuned to 38kHz. This does have a slight adverse effect on the de-emphasis but it is kept within acceptable limits.

In the absence of the 38kHz switching waveform at the emitters of the two switching transistors, they are biased to operate in class A and any signal applied to their bases is reproduced at their collectors. The circuit is therefore compatible, requiring no switching between mono and stereo operation.

The input impedance of the decoder is approximately 15kΩ and the decoder may be fed directly from a semiconductor discriminator capable of providing a loaded output with a peak-to-peak amplitude of approximately 750mV. It may also be fed from a typical valve discriminator, in which case the discriminator output should be attenuated by means of a series resistor shunted by a capacitor. Typical values for the resistor and capacitor are 3·3kΩ to 33kΩ and a few hundred picofarads. Such an arrangement not only provides the correct input level at all multiplex frequencies but also reduces the discriminator loading to an acceptable value.

Each of the audio amplifiers into which the outputs of the decoder are fed must have an input impedance of not less than 50kΩ.

Stereo Indicator

A stereo indicator gives a visual indication when a stereo transmission is in progress. It is also essential in the alignment of the decoder if only a limited amount of test equipment is available. A stereo indicator circuit using a single transistor is shown in Fig. 120.

This stereo indicator circuit operates by detecting the 38kHz signal generated when the 19kHz pilot tone is passed through the frequency-doubling circuit in the decoder. This 38kHz signal is fed through a limiting resistor and blocking capacitor to a shunt rectifier. The voltage derived from this shunt rectifier is applied to the base of a transistor. The transistor remains cut off when the signal is absent and is turned on when it is present. Therefore, the bulb in the collector circuit glows when a stereophonic broadcast is in progress but remains extinguished when a monophonic transmission is being received.

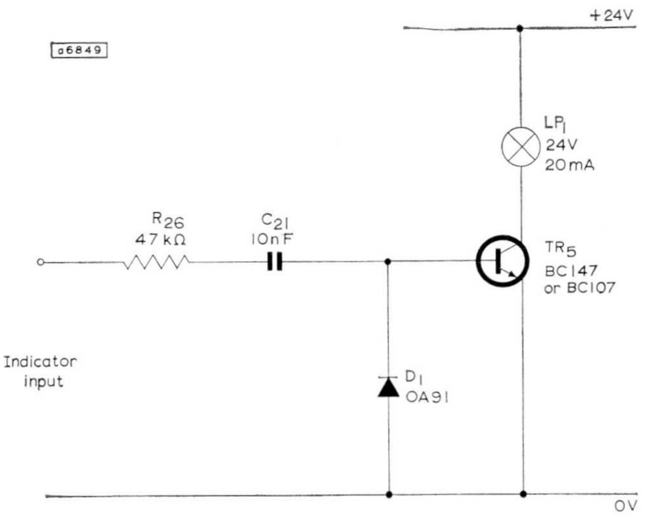


Fig. 120—Stereo indicator circuit

Alignment of the Decoder

To assist those people who do not have access to a wide range of accurate test equipment, a simple alignment procedure using the BBC transmissions has been devised. Tones are transmitted by the BBC outside normal listening hours to enable the channels to be identified and to assist in alignment. Details of the broadcasts are given in Appendix 2. For the initial alignment, however, any stereo broadcast can be used. The alignment procedure is as follows.

First, remove the de-emphasis component network from the discriminator output.

Connect a potentiometer with a value of at least 50kΩ across the discriminator output and connect the decoder input to the potentiometer slider. With the potentiometer set for maximum output, tune the receiver to a stereo broadcast on Radio Three.

Adjust the 19kHz and 38kHz coil slugs to give maximum output at the 'Indicator' terminal (maximum brightness of the indicator lamp). Adjust the input potentiometer setting until the indicator lamp just glows, then adjust the coil slugs again to give maximum brightness. Keep repeating this procedure until adjusting the cores either way results in the lamp getting dimmer. The two coils are now aligned. If tolerances in coil manufacture result in difficulty being experienced in this adjustment, especially of the 38kHz coil, because the slugs are at one end of their travel, replace the 470pF capacitor across the coil by a 390pF capacitor.

When the coils are aligned, connect the discriminator output directly to the decoder input and discard the input potentiometer.

Adjust the preset potentiometer in the decoder circuit to its mid-position. The decoder should give a reasonable degree of separation, but further adjustment using the test-tone transmissions is necessary, as follows.

Connect the two outputs to two audio amplifiers and identify the left- and right-hand channels. When the left-hand channel only is being transmitted, disconnect the left-hand amplifier and turn the 38kHz slug anti-clockwise (when viewed from the top) until the output from the right-hand loudspeaker is a minimum. Anti-clockwise rotation is necessary because a false minimum is encountered if the slug is rotated clockwise.

Reconnect the left-hand amplifier.

When the right-hand channel only is being transmitted, disconnect the right-hand amplifier and adjust the preset potentiometer from its mid-point position to give minimum output from the left-hand loudspeaker.

Reconnect the right-hand amplifier.

The decoder is now correctly aligned.

Performance

The channel separation of the decoder is shown graphically in Fig. 121. The signal-to-noise ratio is well above the acceptable level although the signal-to-noise ratio of a stereo broadcast is inherently slightly lower than that of a mono broadcast. For good stereo reproduction, the i.f. bandwidth of the tuner unit should be not less than 180kHz.

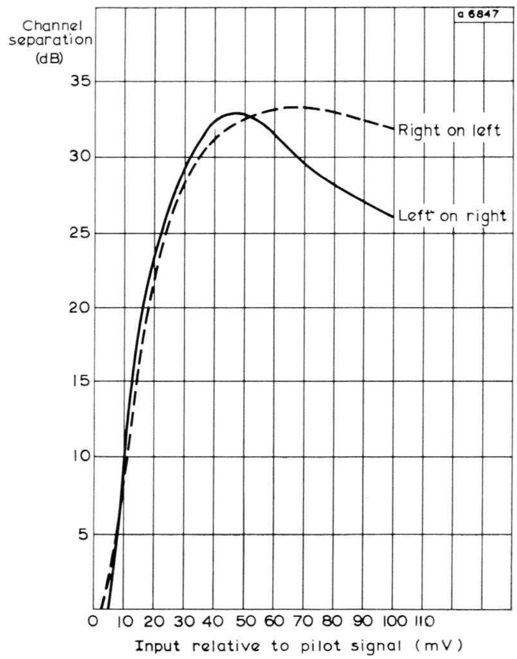


Fig. 121—Channel separation characteristics

Aerial Requirements for Stereo Reception

One common cause of poor results from an f.m. stereo receiver is an inadequate aerial. Satisfactory stereo results cannot be expected unless a really good signal is fed into the tuner.

Unlike the other signals in the complex f.m. stereo signal, the 19kHz signal is of constant amplitude and does not vary with the modulation. Therefore, it is a measure of the signal strength fed to the f.m. tuner. Because its amplitude is much less than that of the other signals, the 19kHz signal is also nearer the noise level. When it is weak the separation between the two a.f. signals is unlikely to be great enough or constant, and the stereophonic effect will be lost.

The output circuit or i.f. filtering circuit of the ratio detector in an f.m. mono receiver requires an a.f. bandwidth of about 15kHz. In an f.m. stereo receiver, however, an a.f. bandwidth of 53kHz is required. Consequently, if all other factors are equal, the signal-to-noise ratio of a stereo receiver is not so good as that of a mono receiver. Therefore, before fitting a stereo decoder, it would be wise to check that the strength of the received signal is sufficient to ensure that the deterioration in signal-to-noise ratio does not make good stereo reception an impossibility.

This check may be made by means of the simple circuits shown in Fig. 122. The circuit in Fig. 122(a) will produce an attenuation of about 12dB. If this is connected between the aerial feeder and the tuner aerial sockets when receiving mono, there should be little audible difference. Any increase in the background noise (hiss, clicks and pops) indicates that a more efficient aerial or a re-siting of the aerial is needed.

If the test is satisfactory, two of the circuit in Fig. 122(a) in series, or one of the circuit in Fig. 122(b), should be connected between the aerial feeder and the tuner aerial sockets. If again there is no marked increase in the background noise, the aerial is quite adequate for f.m. stereo. However, if there is a marked increase in the background noise, the aerial system may not be good enough for f.m. stereo. An aerial array or a new position for the aerial may be needed.

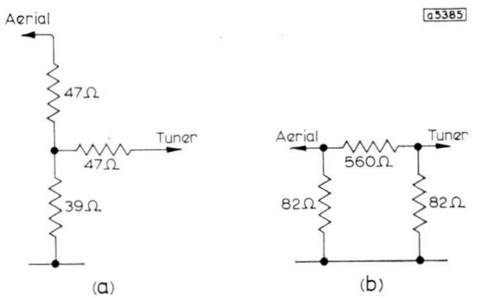


Fig. 122—Circuits for checking signal-to-noise ratio

COMPONENT LISTS FOR CIRCUITS IN CHAPTER 8

F.M. Tuner using BF200 Transistor (page 146)

Transistors

Circuit reference	Mullard type
TR_1	BF200
TR_2	BF195
TR_3	BF195

Resistors

Tolerance: ±10%
Power rating: $\frac{1}{8}$W

Circuit reference	Value	Circuit reference	Value
R_1	1·5kΩ	R_6	3·3kΩ
R_2	10kΩ	R_7	100Ω
R_3	6·8kΩ	R_8	2·7kΩ
R_4	8·2kΩ	R_9	6·8kΩ
R_5	6·8kΩ	R_{10}	10kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	3·3pF	tubular ceramic	
C_2	3 to 14pF	ganged tuning capacitor	
C_8	3 to 14pF	ganged tuning capacitor	
C_{20}	3 to 14pF	ganged tuning capacitor	
C_3	1 to 7pF	trimmer	
C_4	8·2pF	ceramic plate	C333CB/L8E2
C_5	470pF	tubular ceramic	
C_6	3·9pF	ceramic plate	C333CB/L3E9
C_7	1nF	polystyrene	C295AC/1K
C_9	1 to 7pF	trimmer	
C_{10}	8·2pF	ceramic plate	C333CB/L8E2
C_{11}	1nF	polystyrene	C295AC/1K
C_{12}	22nF	metallised foil	C280AE/P22K
C_{13}	120pF	ceramic plate	C333CH/C120E
C_{14}	1·5nF	polystyrene	C295AA/1K5
C_{15}	82pF	ceramic plate	C333CH/C82E
C_{16}	1pF	tubular ceramic	
C_{17}	1nF	polystyrene	C295AC/1K
C_{18}	1nF	polystyrene	C295AC/1K
C_{19}	1·8pF	tubular ceramic	
C_{21}	1 to 7pF	trimmer	
C_{22}	8·2pF	ceramic plate	C333CB/L8E2

Coils

All the coils are wound with enamelled-copper wire and all except L_5 are air-cored. The i.f. coil, L_5, follows normal design practice to match the input of the amplifier.

Circuit reference	L_1	L_2	L_3	L_4	L_5
Q_O	200		200	200	150
		150			
Q_L	50		100	—	—
No. of turns	4	22	4	4½	—
Wire diam. (mm)	1	0·2	1	1	—
Pitch (mm)	2	close	2	2	—
Inside diam. (mm)	8	4	8	8	—
No. of turns between tap and chassis end of coil	—	—	¾	3½	—

F.M. Tuner using BFW10 Field-effect Transistor (page 156)

Transistors

Circuit reference	Mullard type
TR_1	BFW10
TR_2	BF195
TR_3	BF195

Resistors

Tolerance: $\pm 10\%$
Power rating: $\frac{1}{8}$W

Circuit reference	Value
R_1	22kΩ
R_2	27kΩ
R_3	680Ω
R_4	8·2kΩ
R_5	6·8kΩ
R_6	4·7kΩ
R_7	100Ω
R_8	3·9kΩ
R_9	6·8kΩ
R_{10}	10kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	3·3pF	tubular ceramic	
C_2	1·5nF	tubular ceramic	
C_3	1·5nF	tubular ceramic	
C_4	3 to 14pF	ganged tuning	
C_{15}	3 to 14pF	capacitor	
C_5	1 to 7pF	trimmer	
C_6	8·2pF	ceramic plate	C333CB/L8E2
C_7	1nF	polystyrene	C295AC/1K
C_8	22nF	metallised foil	C280AE/P22K
C_9	120pF	ceramic plate	C333CH/C120E
C_{10}	1·5nF	polystyrene	C295AA/1K5
C_{11}	1·5pF	tubular ceramic	
C_{12}	22pF	tubular ceramic	
C_{13}	1nF	polystyrene	C295AC/1K
C_{14}	1nF	polystyrene	C295AC/1K
C_{16}	1 to 7pF	trimmer	
C_{17}	8·2pF	ceramic plate	C333CB/L8E2

Coils

All the coils are wound with enamelled-copper wire. L_1 and L_2 are wound on powdered iron; L_3 and L_4 are air-cored. The i.f. coil, L_5, follows normal design practice to match the input of the amplifier.

Circuit reference	L_1	L_2	L_3	L_4	L_5
Q_O	80	80	200		150
Q_L	4	4	100		
No. of turns	3	4	4	$4\frac{1}{2}$	
Wire diam. (mm)	0·15	0·15	1	1	
Pitch (mm)			2	2	
Inside diam. (mm)			8	8	
No. of turns between tap and chassis end of coil			drain 1 mixer $\frac{1}{2}$	$3\frac{1}{2}$	

Stereo Decoder and Indicator (pages 162 and 164)

Transistors and Diode

Circuit reference	Mullard type
TR_1	BC148 or BC108
TR_2	BC148 or BC108
TR_3	BC148 or BC108
TR_4	BC148 or BC108
TR_5	BC147 or BC107
D_1	OA91

Resistors

Tolerance: $\pm 10\%$ except R_{17} to R_{22} which should be $\pm 5\%$ components.
Power rating: $\frac{1}{8}$W

Circuit reference	Value	Circuit reference	Value
R_1	22kΩ	R_{14}	56kΩ
R_2	10kΩ	R_{15}	12kΩ
R_3	10kΩ	R_{16}	12kΩ
R_4	10kΩ	R_{17}	1·8kΩ
R_5	220Ω	R_{18}	1·8kΩ
R_6	4·7kΩ	R_{19}	1·8kΩ
R_7	100Ω	R_{20}	1·8kΩ
R_8	1kΩ	R_{21}	18kΩ
R_9	3·9kΩ	R_{22}	18kΩ
R_{10}	1kΩ	R_{23}	150kΩ
R_{11}	1kΩ	R_{24}	150kΩ
R_{12} preset potentiometer	1kΩ	R_{25}	5·6kΩ
R_{13}	100kΩ	R_{26}	47kΩ

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	10nF	metallised foil	C280AE/P10K
C_2	470pF	polystyrene or silvered-mica	
C_3	100nF	metallised foil	C280AE/P100K
C_4	10nF	metallised foil	C280AE/P10K
C_5	470pF	polystyrene or silvered-mica	
C_6	10μF	electrolytic 16V	C426AR/E10
C_7	4·7nF	polyester	C296AC/A4K7
C_8	4·7nF	polyester	C296AC/A4K7
C_9	4·7nF	polyester	C296AC/A4K7
C_{10}	4·7nF	polyester	C296AC/A4K7
C_{11}	10μF	electrolytic 16V	C426AR/E10
C_{12}	2·2nF	polystyrene	C295AA/2K2
C_{13}	2·2nF	polystyrene	C295AA/2K2
C_{14}	2·2nF	polystyrene	C295AA/2K2
C_{15}	2·2nF	polystyrene	C295AA/2K2
C_{16}	220pF	ceramic plate	C333CH/C220E
C_{17}	220pF	ceramic plate	C333CH/C220E
C_{18}	100nF	metallised foil	C280AE/P100K
C_{19}	100nF	metallised foil	C280AE/P100K
C_{20}	10μF	electrolytic 25V	C426AR/H10
C_{21}	10nF	metallised foil	C280AE/P10K

Capacitors C_2 and C_5 should be high-quality polystyrene or silvered-mica types.
Capacitors C_{12} to C_{17} should be $\pm 5\%$ components.

Coils

Circuit reference	Mullard type
L_1	WF2949
L_2/L_3	WF2951

For coil details see Figs. 123 and 124.

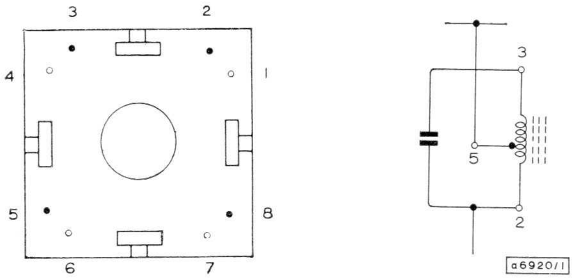


Fig. 123—19kHz coil, L_1 (Mullard type WF2949), viewed from underside. Pin 8 is the coil clamp and should be connected to earth

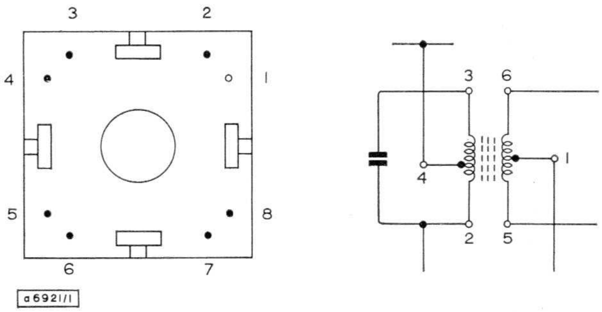


Fig. 124—38kHz coil, L_2/L_3 (Mullard type WF2951), viewed from underside. Pin 8 is the coil clamp and should be connected to earth

Stereo Indicator Bulb LPI

A 24V, 20mA indicator bulb is required, a suitable component being the Electroniques Limited type SGF9/D/GN/24, available from Edinburgh Way, Harlow, Essex.

CHAPTER 9

TEST EQUIPMENT

Items of test equipment which will be found useful in commissioning and maintaining audio and radio circuitry include a voltmeter, ohmmeter, milliammeter, output power meter, a.f. signal generator, i.f. and r.f. signal generator with provision for modulation at audio frequencies, oscilloscope, transistor tester and stereo balance meter.

It is beyond the scope of this book to give designs for the complex equipment required for aligning and testing h.f. circuitry, and it is recommended that professionally built equipment should be obtained for this purpose.

OSCILLOSCOPE

The reader is referred to the following sources for design information on a simple oscilloscope and auxiliary circuitry:

"The Wireless World Oscilloscope", available from Wireless World, Dorset House, Stamford Street, London, S.E.1;

"Oscilloscope Trace Doubler" by W. Kemp, published in The Radio Constructor, Vol. 22, No. 4, Nov. 1968;

"Wide-band Oscilloscope Amplifier" by G. Sowersby, published in The Radio Constructor, Vol. 22, No. 5, Dec. 1968.

The last two are available from Data Publications Limited, 57 Maida Vale, London, N.9.

VOLTMETER

It is essential that the voltmeter does not alter the conditions in the circuit when measurements are being made. The voltmeter should therefore have a resistance of at least 20kΩ/V. A voltage coverage of 2·5 to 100V d.c. is required. A valve voltmeter is very suitable for the measurements if precautions are taken to prevent earth-loop currents damaging the transistors.

The simplest approach to the testing of the directly-coupled amplifiers described in this book is to measure voltage levels throughout the circuit. A conventional four-transistor circuit is used here as an example, and the voltage levels resulting from various transistor failures are given in Table 20.

TABLE 20

Voltage Levels

Transistor Fault	Voltage at point				General Performance
	A	B	C	D	
TR_2 o–c	high	no voltage	high	slightly higher than normal	no output
TR_2 s–c	low	twice normal voltage	low	slightly higher than normal	no output
TR_3 o–c	near-chassis	low	high	low	no output
TR_4 o–c	normal	normal	normal	normal	high distortion
TR_3 or TR_4 s–c	indeterminate				emitter resistors burn out

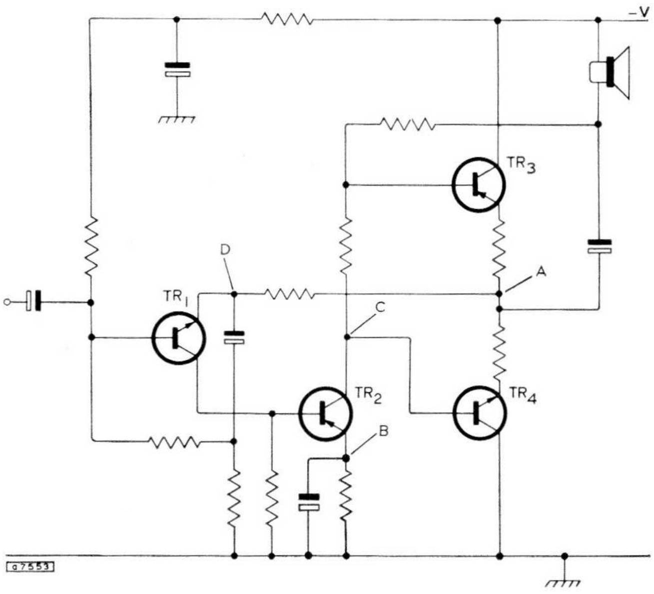


Fig. 125—Direct-coupled amplifier with voltage test points

OHMMETER

The voltage at the output terminals of the ohmmeter should not exceed 1·5V otherwise there is a possibility of the transistors and miniature electrolytic capacitors in the circuit under test being damaged. In all cases when resistance measurements are made, the equipment should be switched off.

If only one component is suspect this can be isolated from the circuit by unsoldering the connection at one end and then testing with the ohmmeter. However, if components have to be checked while they are still in the circuit there are two methods available to prevent the 'diode elements' of the transistors affecting the measurements. In the first method the base connection of the transistor is unsoldered so that the transistor is effectively disconnected from the circuit. The resistance measurements may then be made, although allowance must be made for other resistors shunting the component under test. The second method uses the voltage of the ohmmeter to 'reverse bias' the diode elements of the transistor so that the transistor does not conduct.

The basic circuit of a series-connected ohmmeter is shown in Fig. 126 in which both the colour and the polarity of the output terminals are shown. To provide the reverse bias for a p-n-p transistor, the positive ohmmeter terminal (coloured black) is connected to the base (n-type material) and the negative ohmmeter terminal (coloured red) connected to the emitter or collector (p-type material). The connections for measuring the values of the base bias resistors of a p-n-p transistor are shown in Fig. 127; for an n-p-n transistor the connections are reversed. If the stage being checked is not isolated from the rest of the circuit the other stages will appear as a resistance in parallel with that being measured. The measured value of resistance will therefore be lower than the actual resistance.

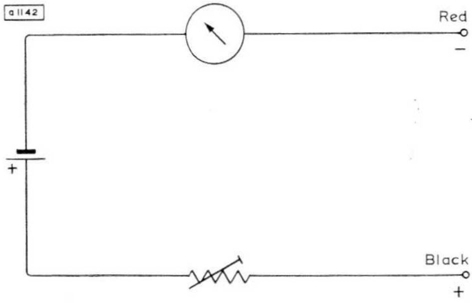


Fig. 126—Circuit of series-connected ohmmeter

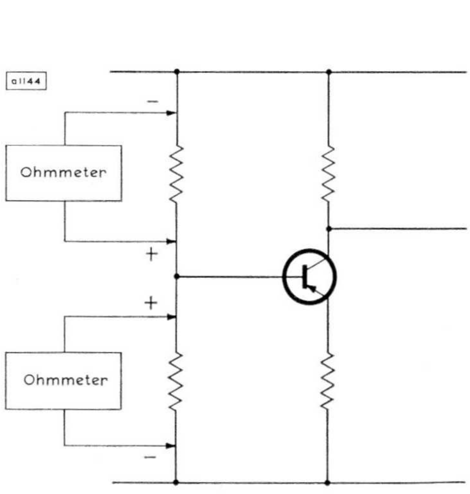


Fig. 127—Method of checking base bias resistors with ohmmeter

MILLIAMMETER

A milliammeter with a full-scale deflection of 100mA may be used to measure the currents in the pre-amplifier and driver stages of the amplifiers described, but the output transistors in some of the circuits draw currents in excess of 1A.

OUTPUT-POWER METER

An output-power meter can be used for checking a.f. output power. The loudspeaker should be replaced by the output-power meter which is switched to the loudspeaker impedance. The meter then indicates the output power directly.

A voltmeter can also be used conveniently for this measurement if an output-power meter is not available. The loudspeaker should be replaced by a resistor corresponding to the loudspeaker impedance and the power level calculated as the voltage drop across the resistor.

AUDIO-FREQUENCY SIGNAL GENERATOR

The signal generator shown in Fig. 128 is a Wien-bridge oscillator which will supply an output voltage of 1V r.m.s. over a frequency range 15Hz to 200kHz.

Circuit Description

The circuit uses a thermistor R_5 as an amplitude control device, and the output is essentially independent of small changes in supply voltage and ambient temperature. The oscillator uses only three transistors and, apart from the frequency-determining capacitors, only one capacitor. The circuit operates from a 9V supply, although little difference in performance is obtained with supply voltages from 6 to 10V. The current consumption is:

6V supply	6·3mA
9V supply	11·5mA
10V supply	13mA

Performance

The frequency coverage is 15Hz to 200kHz in four ranges:

Range 1	15Hz to 200Hz
Range 2	150Hz to 2kHz
Range 3	1·5kHz to 20kHz
Range 4	15kHz to 200kHz

The approximate harmonic distortion at 1kHz is:

second harmonic	0·1%
third harmonic	0·05%

At frequencies above 100kHz the frequency calibration does not

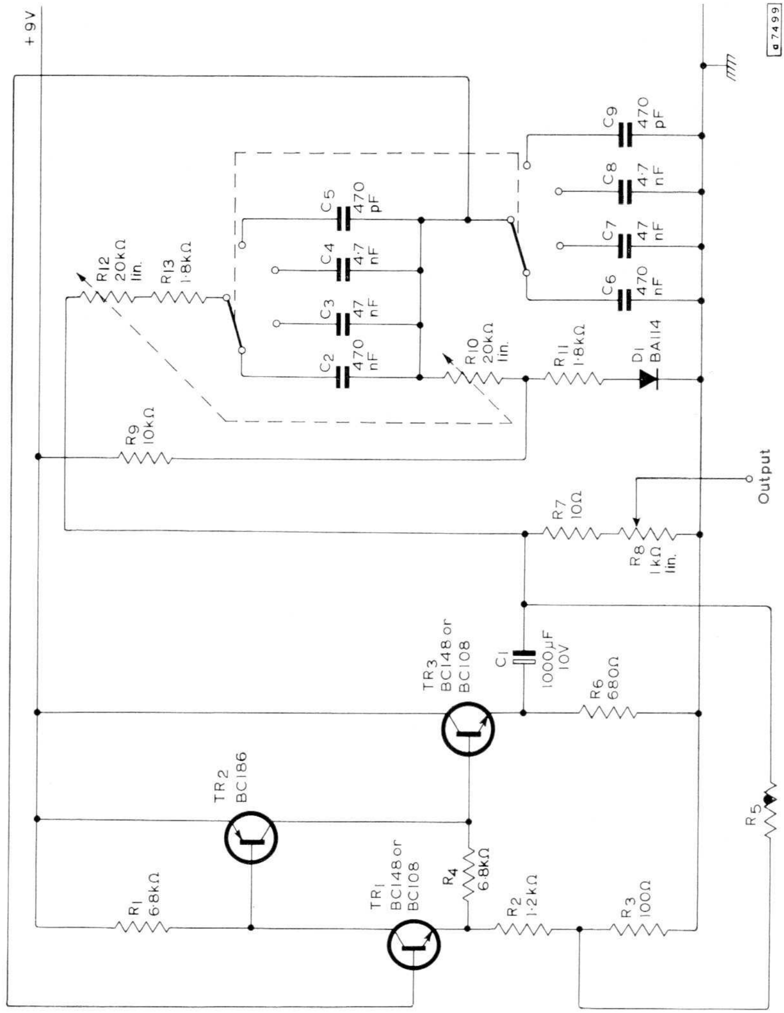


Fig. 128—A.F. signal generator circuit

coincide with that for lower ranges, because of stray capacitances, and the output voltage is slightly higher.

Components

Transistors and Diode

Circuit reference	Mullard type
TR_1	BC148 or BC108
TR_2	BC186
TR_3	BC148 or BC108
D_1	BA114

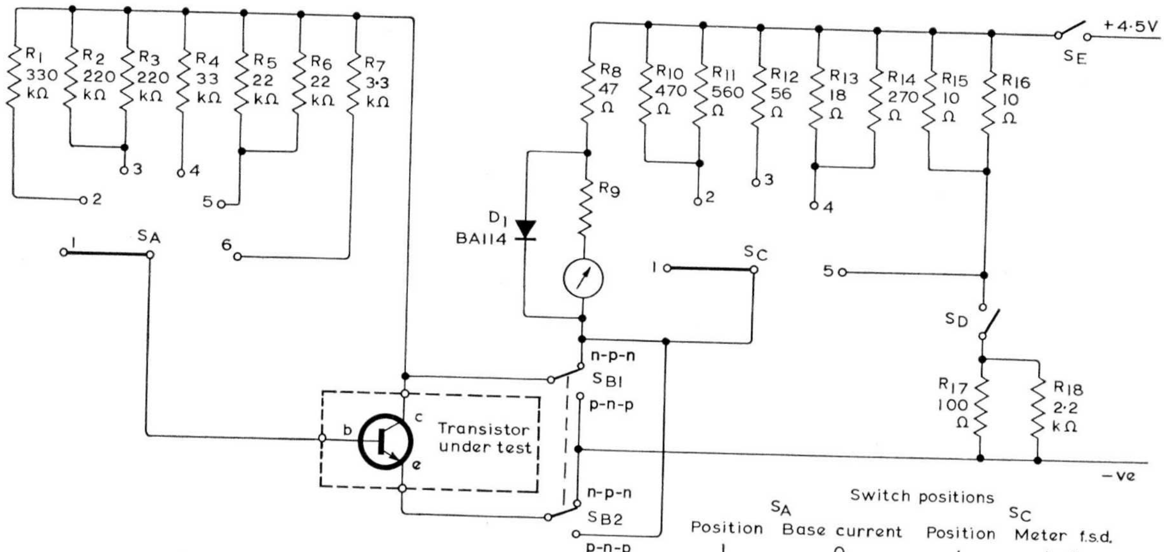
Resistors

Tolerance: ±5%
Power rating: $\frac{1}{8}$W

Circuit reference	Value	
R_1	6·8kΩ	
R_2	1·2kΩ	
R_3	100Ω	
R_4	6·8kΩ	
R_5 thermistor STC type R53		
R_6	680Ω	
R_7	10Ω	
R_8 linear potentiometer	1kΩ	
R_9	10kΩ	
R_{10} linear potentiometer	20kΩ	matched to within 2dB
R_{12} linear potentiometer	20kΩ	(standard stereo matching)
R_{11}	1·8kΩ	
R_{13}	1·8kΩ	

Capacitors

Circuit reference	Value	Description	Mullard type
C_1	1000μF	electrolytic 10V	C437AR/D1000
C_2	470nF	metallised foil	C280AE/A470K
C_3	47nF	metallised foil	C280AE/P47K
C_4	4·7nF	metallised foil	
C_5	470pF	metallised foil	
C_6	470nF	metallised foil	C280AE/A470K
C_7	47nF	metallised foil	C280AE/P47K
C_8	4·7nF	metallised foil	
C_9	470pF	metallised foil	



Switch positions

SA Position	SA Base current	SC Position	SC Meter f.s.d.
1	0	1	1mA
2	10μA	2	3mA
3	30μA	3	10mA
4	100uA	4	30mA
5	300μA	5	100mA
6	1mA		

a7500

Fig. 129—Transistor tester circuit

TRANSISTOR TESTER

The transistor tester shown in Fig. 129 uses a 1mA meter with suitable shunts to measure the collector current of the transistor under test. The base current of the transistor may be switched to zero for measurement of collector-emitter reverse leakage current, I_{CEO}, or to 10μA, 30μA, 100μA, 300μA or 1mA for measurement of static forward current transfer ratio, h_{FE}.

Circuit Description

A meter with 1mA f.s.d. is used and R_9 is chosen to give a total resistance of 450Ω. Switch S_A selects base resistors R_1 to R_7, whose common connection is taken to the collector to avoid the necessity for an additional pole on n-p-n/p-n-p selector switch S_B. Shunt resistors R_{10} to R_{16} are switched in by S_C to give collector current ranges 0 to 1mA, 0 to 3mA, 0 to 10mA, 0 to 30mA and 0 to 100mA. Switch S_D and resistors R_{17} and R_{18} provide a battery check facility. Diode D_1 provides meter protection, limiting the current through the meter to 2mA.

Switches S_D and S_E should be spring-loaded to the 'off' position to prevent current drain from the battery when no measurement is being made. Switches S_A and S_B should be break-before-make, and switch S_C should be make-before-break.

Measurement of I_{CEO} and h_{FE}

Switch S_A is set to position 1, S_B to the correct polarity and S_C to a high range. The transistor is inserted into the socket and with S_E depressed to the 'on' position, S_C is turned until a reading is obtained. For germanium transistors this current, I_{CEO}, should be about 0·5 to 2mA, although for large transistors it may be higher. For silicon types, I_{CEO} should be negligible. If a high reading is obtained the device is probably faulty (short-circuited collector to emitter).

To measure gain, switch S_A is turned to the position corresponding to the required base current, S_C is set to a high range and S_E is again depressed. If the collector current obtained is I_C and the base current is I_B, the gain of the transistor is given by:

$$h_{FE}=\frac{I_C-I_{CEO}}{I_B}$$

If I_{CEO} is small enough to be neglected, h_{FE} can be read directly on the meter, but only for base currents of 10μA, 100μA and 1mA. For other values of base current the gain must be computed using the formula given above.

If no increase in collector current is obtained when the base current is switched in, it may be assumed that the transistor is open-circuit.

Battery Check

The internal battery should be checked periodically. With no transistor in the socket, S_C is switched to the 100mA position and S_D and S_E are depressed. A current of 45mA is drawn from a 4·5V battery, so the check is made on full load.

Accuracy

The accuracy of the circuit is determined by the resistor tolerance—5% components are recommended—and the battery voltage. Accuracy also depends on the type of transistor under test and its gain.

The base resistors are returned to the collector of the transistor, and with this arrangement the base current is dependent to a small degree on the voltage drop in the meter circuit. In addition, with germanium transistors the base current will be higher than the value selected, due to the low value of V_{BE}.

With switch S_A in the 10μA position, for example, for a silicon transistor the base current will be 10μA with the meter reading at full scale. With the meter reading at one-third of full scale the base current will be 11μA. A similar error will occur on all ranges. For a germanium transistor the corresponding base current with meter reading at full scale will be 11·5μA, and 12·5μA at one-third of full scale.

For most purposes the error for silicon devices may be neglected. For germanium devices it should be remembered that the base current is approximately 20 percent higher than the selected value.

Components

Diode

Circuit Reference	Mullard type
D_1	BA114

Resistors

Tolerance: ±5%
Power rating: $\frac{1}{8}$W

Circuit reference	Value	Circuit reference	Value
R_1	330kΩ	R_{10}	470Ω
R_2	220kΩ	R_{11}	560Ω
R_3	220kΩ	R_{12}	56Ω
R_4	33kΩ	R_{13}	18Ω
R_5	22kΩ	R_{14}	270Ω
R_6	22kΩ	R_{15}	10Ω
R_7	3·3kΩ	R_{16}	10Ω
R_8	47Ω	R_{17}	100Ω
R_9	see text	R_{18}	2·2kΩ

Suitable meters are manufactured by:
Sifam Electrical Instrument Company Limited,
Woodland Road, Torquay, Devon.
Taylor Electrical Instruments Limited,
Montrose Avenue, Slough, Bucks.

BALANCE METER

A simple meter circuit which may be used to balance the two outputs in a stereo system is shown in Fig. 130.

Components C_1 and C_2 damp the meter movement. Resistor R_3 is selected to suit the current rating of the centre-zero meter and the amplifier output (typically 10kΩ for a 1mA movement). The voltage rating of the electrolytic capacitors should be half the supply voltage of the system with which the meter is used.

Suitable meters are manufactured by:
Sifam Electrical Instrument Company Limited,
Woodland Road, Torquay, Devon.
Taylor Electrical Instruments Limited,
Montrose Avenue, Slough, Bucks.

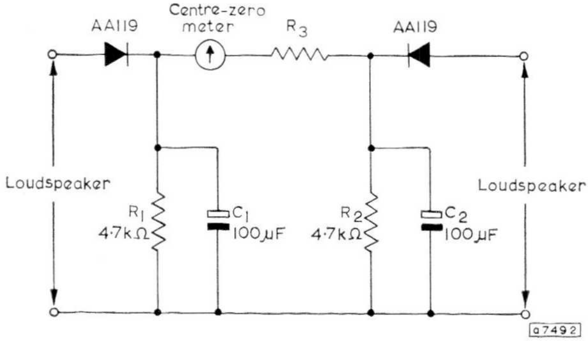


Fig. 130—Balance meter circuit

PREVENTION OF EARTH-LOOP CURRENTS

Transistors may be destroyed by earth-loop currents if the chassis of test equipment is not earthed. The chassis may attain an a.c. potential of 50 to 100V through electrostatic coupling between the primary winding and the core of the mains transformer, as illustrated in Fig. 131. If, for example, the equipment under test is connected to the output terminals of an unearthed signal generator, this electrostatic potential is transferred to the chassis of the equipment, as shown in Fig. 131b. An isolating capacitor in the test leads will not prevent the transfer of the potential because this is an electrostatic effect. If an earthed soldering iron were now applied to the base of a transistor in the circuit, a relatively large current would flow to earth through the transistor and so destroy it.

The formation of this electrostatic potential can be prevented by earthing the chassis of the equipment and test instrument. An isolating capacitor, usually 1μF, should be connected in the lead between a signal generator and the equipment to prevent the flow of direct circulating currents.

Transistors can also be damaged by the use of an unearthed soldering iron. It is possible for the bit, through a breakdown of the insulation between element and bit, to attain a potential of a few volts. When the iron is used on a circuit the potential of the bit may drive a relatively large current through the transistors. Damage due to the defect can be prevented by ensuring that the soldering iron is earthed. However, it should be remembered that when an earthed soldering iron is used the power supply to the circuit should be switched off during soldering.

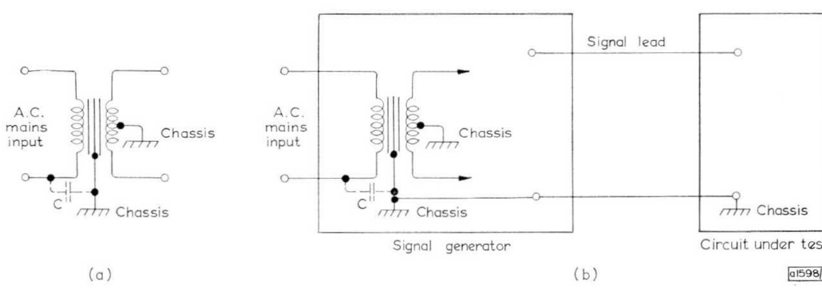


Fig. 131—Diagram showing (a) how the potential of an unearthed chassis is raised and (b) how the potential is transferred to other unearthed chassis

APPENDIX 1

BIASING ARRANGEMENTS FOR H.F. CIRCUITS

Three biasing arrangements used in h.f. circuits—voltage biasing with emitter feedback, current biasing with collector feedback and simple current biasing —were briefly considered in Chapter 2. The three arrangements are now described in detail, and for each of them an example is worked so that their effectiveness can be compared. In the examples, a transistor with a nominal value of h_{FE} of 100 (h_{FE} spread, 50 to 150) and a nominal V_{BE} of 0·7V (V_{BE} spread, 0·65 to 0·75V) is operating from a supply voltage V_S of 7V; the required emitter current is 1mA.

VOLTAGE BIASING WITH EMITTER FEEDBACK

The circuit of a voltage biasing arrangement is given in Fig. 132. In this circuit the voltage V_B is given by

$$V_B = \frac{R_2 V_S}{R_1 + R_2}$$

provided that the voltage developed across R_2 by the flow of base current is neglected.

The voltage V_E is given by

$$V_E = V_B - V_{BE}$$

and therefore

$$V_E = \frac{R_2 V_S}{R_1 + R_2} - V_{BE}.$$

The operating current, I_E, is therefore given by

$$I_E = \frac{R_2 V_S}{R_E(R_1 + R_2)} - \frac{V_{BE}}{R_E}. \qquad \ldots(1)$$

For the effects of spreads and temperature variations of V_{BE} to be negligible, the voltage V_E must be large compared with the changes in V_{BE}. Furthermore, R_E must also be large if changes in I_E caused by variation of V_S are to be small. In practice, a voltage drop, V_E, up to 3V is required with silicon transistors to prevent disproportionately large changes in operating current with changes in supply voltage, when large variations in V_S have to be accommodated. The corresponding value of V_E for germanium transistors is about 1V.

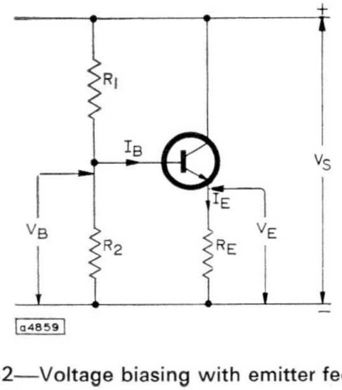


Fig. 132—Voltage biasing with emitter feedback

It has been assumed that the voltage developed across resistor R_2 by the base current I_B is negligible and, if this is true, the operating current I_E is independent of h_{FE}, and therefore of spreads in h_{FE}. If this voltage is significant, however, the operating current is given by

$$I_E = \frac{\dfrac{V_S R_2}{R_1 + R_2} - V_{BE}}{\dfrac{R_1 R_2}{(R_1 + R_2)(1 + h_{FE})} + R_E} \qquad \ldots(2)$$

and the current is then dependent upon h_{FE}.

In the example considered the values of R_1 and R_2 are both 15kΩ and the value of R_E is 2·7kΩ. The nominal collector-emitter voltage is 4·3V, and the actual nominal emitter current is 1·01mA.

Substitution of these values in Eq. 2 gives maximum and minimum values of I_E, resulting from the spread of h_{FE}, of 1·02mA and 0·985mA. Thus, the spread in I_E resulting from the h_{FE} spread is $+1\%$, $-1{\cdot}6\%$.

The maximum and minimum values of I_E, resulting from the spread in V_{BE}, are 1·025mA and 0·99mA. Thus, the spread in I_E resulting from the spread in V_{BE} is $\pm 1{\cdot}7\%$.

If V_S falls from 7V to 3·5V, the nominal emitter current falls from 1·01mA to 0·38mA; that is, I_E falls by 62%.

Eq. 2 reverts to the simple equation (Eq.1) if the following condition is fulfilled

$$R_E \gg \frac{R_1\, R_2}{(R_1 + R_2)(1 + h_{FE})}.$$

This equation can be written

$$I_E R_E \gg \frac{R_1 R_2}{(R_1 + R_2)} \times \frac{I_E}{(1 + h_{FE})}$$

or

$$V_E \gg \frac{R_1 R_2 I_B}{(R_1 + R_2)}.$$

Therefore, for the operating current of the circuit shown in Fig. 132 to be substantially independent of h_{FE}, the voltage developed across the emitter resistor must be very much greater than the voltage developed, by the flow of base current, across the d.c. source resistance at the base of the transistor. The d.c. source resistance at the base of the transistor is formed by the parallel combination of R_1 and R_2.

CURRENT BIASING WITH COLLECTOR FEEDBACK

In the current-biased circuit shown in Fig. 133 the voltage V_C is given by

$$V_C = V_S - R_1 I_E .$$

Since

$$I_E = (1 + h_{FE}) I_B$$

then

$$V_C = V_S - I_B R_I (1 + h_{FE}).$$

The base current I_B is also given by

$$I_B = \frac{V_C - V_{BE}}{R_2} .$$

Hence, if V_C is eliminated, the base current is given by

$$I_B = \frac{V_S - V_{BE}}{R_2 + R_1 (1 + h_{FE})} .$$

Consequently, the operating current is given by

$$I_E = \frac{(V_S - V_{BE}) (1 + h_{FE})}{R_2 + R_1 (1 + h_{FE})} . \qquad \ldots(3)$$

In the example considered, the value of R_1 is 2·7kΩ and the value of R_2 is 330kΩ. The nominal value of V_{CE} is 4·28V, and the actual nominal emitter current is 1·05mA.

Substitution of these values in Eq.3 gives maximum and minimum values of emitter current, caused by the spread in h_{FE}, of 1·29mA and 0·635mA. Thus the spread in I_E caused by the spread in h_{FE} is +23%, − 35%.

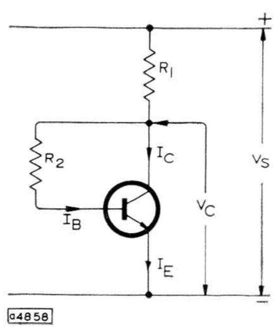


Fig. 133—Current biasing with collector feedback

The variation in I_E with spreads and temperature variations in h_{FE} can be reduced only by applying the following condition:

$$R_1 (1 + h_{FE}) \gg R_2.$$

This lowers the available collector voltage swing and is possible only with large supply voltages.

The maximum and minimum values of emitter current, caused by the spread in V_{BE}, are 1·04mA and 1·06mA. Thus, the spread in I_E caused by the spread in V_{BE} is $\pm 0{\cdot}95\%$.

If V_S falls from 7V to 3·5V, the nominal emitter current falls from 1·05mA to 0·47mA, that is, I_E falls by 55%.

In this arrangement there is a spread in the collector voltage resulting from the spread in h_{FE}. In the extreme case, when most of the available supply voltage is developed across R_1 (when the spread in operating current due to the spread in h_{FE} is very low) the spread in collector voltage is directly proportional to the spread in h_{FE}.

SIMPLE CURRENT BIASING

The simplest method of biasing, using only one resistor, is shown in Fig. 134. The current I_B is given by

$$I_B = \frac{V_S - V_{BE}}{R_1}.$$

Thus, the emitter current is given by

$$I_E = \frac{V_S - V_{BE}}{R_1} (1 + h_{FE}). \qquad \ldots(4)$$

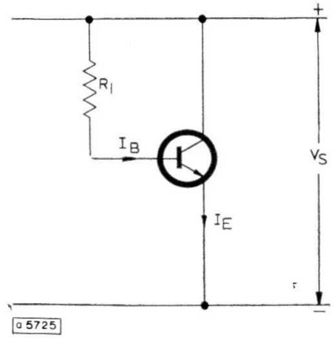


Fig. 134—Simple current biasing

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In the example considered the value of R_1 is 680kΩ. The nominal value of V_{CE} is 7V (equal to V_S), and the actual nominal value of I_E is 0·935mA.

Substitution of these values in Eq. 4 gives maximum and minimum values of I_E, resulting from the spread in h_{FE}, of 1·4mA and 0·475mA. Thus the spread in I_E resulting from the spread in h_{FE} is ±49%. Because the emitter current spread resulting from the spread in h_{FE} is high, simple current biasing is suitable only for transistors with narrow h_{FE} spreads.

The maximum and minimum values of I_E resulting from the spread in V_{BE} are 0·944mA and 0·928mA. Thus the spread in I_E resulting from the spread in V_{BE} is ±0·86%.

If V_S falls from 7V to 3·5V, the nominal emitter current falls from 0·935mA to 0·415mA; that is, I_E falls by 56%. However, because there are no resistors in either the collector or the emitter lead, the collector-emitter voltage is equal to the supply voltage. Thus, a higher-value series resistor can be used in the supply line (shared between the r.f. and i.f. stages), if the collector-emitter voltage in this method is to be the same as that in the other two biasing methods discussed. The higher the value of this series resistor the less dependent the emitter current will be on supply voltage.

APPENDIX 2

BBC TEST-TONE TRANSMISSIONS

Details of test-tone transmissions are reproduced in Table 21 from a BBC publication*. The BBC reserves the right to change all or any part of the schedule.

The tones shown in the table are transmitted on Wednesdays and Saturdays. On other days, a 250Hz tone is transmitted in the left channel only from about four minutes after the end of the last programme on Radio Three until 23.55. This test may be interrupted from time to time.

The schedule given in the table is subject to variation in accordance with programme requirements and essential transmission tests. The zero level reference corresponds to 40% of the maximum level of modulation applied to either stereophonic channel before pre-emphasis. All tests are transmitted with pre-emphasis.

Periods of tone lasting several minutes are interrupted momentarily at one minute intervals.

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*The British Broadcasting Corporation, Information Sheet No. 1605 (1), September, 1968.

TABLE 21

BBC Test-tone Transmissions

Time	Left Channel (A)	Right Channel (B)
23.30	250Hz at zero level	440Hz at zero level
23.32	440Hz at zero level	440Hz at zero level, antiphase to left channel
23.35	440Hz at +8dB	440Hz at +8dB, antiphase to left channel
23.37	440Hz at +8dB	440Hz at +8dB, in phase with left channel
23.39	250Hz at +8dB	440Hz at +8dB
23.40	250Hz at zero level	Nothing
23.44	Nothing	440Hz at zero level
23.47.20 approx.	Tone sequence at −4dB: 60Hz, 900Hz, 5kHz, 10kHz. This sequence is repeated.	Nothing
23.48.20 approx.	Nothing	Tone sequences as for left channel at 23.47.20
23.49.20	250Hz at zero level	Nothing
23.51	Nothing	Nothing
23.53	Reversion to monophonic transmission	

APPENDIX 3

CHARTS AND NOMOGRAMS

This appendix contains miscellaneous charts and nomograms relevant to radio and audio system design:

Chart 1	Fletcher-Munson curves
Chart 2	Every-day noise levels
Chart 3	Frequency range of musical instruments
Chart 4	Radio frequency and wavelength
Chart 5	Decibel charts
Chart 6	Reactance chart

CHARTS 1, 2 and 3

These three charts are of interest in that they give some indication of the performance required of an audio reproduction system in terms of the characteristics of the human ear and the frequency ranges of musical instruments.

Chart 1 shows equal-loudness contours for the ear, the sound intensity required to produce a sensation of equal loudness at varying frequencies being plotted for various levels of loudness from just audible to the point where the sound is "felt" rather than heard. It will be seen that the ear can respond to all frequencies in the range 15Hz to 20kHz, and to an intensity range of about 120 to 130dB in the region of 3kHz. However, the ear is relatively insensitive to both high and low frequencies in the audio band, especially at low intensity levels. The sensitivity of the ear remains constant with varying intensity above a level of about 80dB and between about 70Hz and 6kHz, but is bass-deficient at lower intensity levels.

Chart 2 relates loudness levels expressed in phons to some every-day noises.

Chart 3 shows the frequency range of a selection of musical instruments, male and female speech. The solid line indicates the actual tone range of the instrument and the dotted line accompanying overtone and action noises.

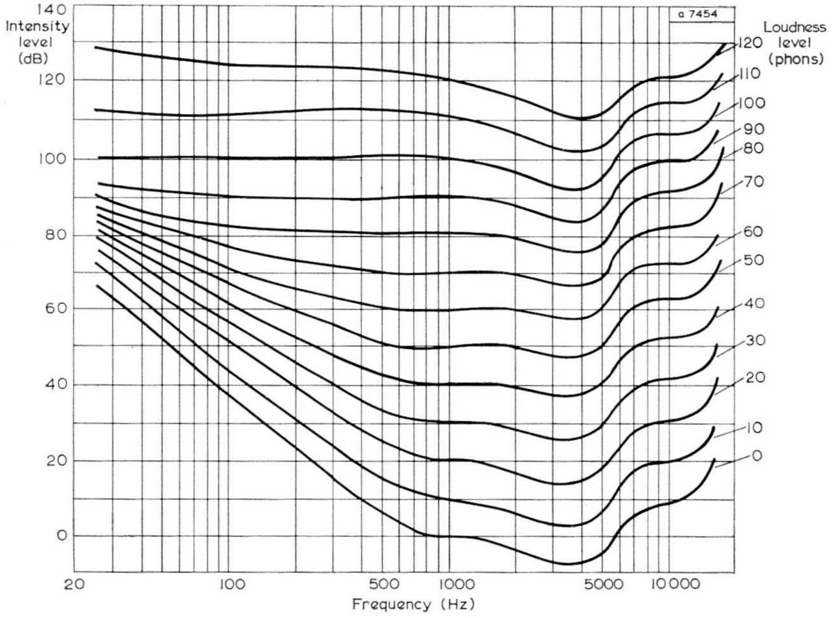


CHART 1. FLETCHER-MUNSON CURVES

Noise	Loudness level in phons
Pneumatic drill	110
Underground train	100
Very loud radio	90
Suburban train	80
Average factory	70
Quiet car	60
Average office	50
Public library	40
Audience noise during concert	30
Whisper	20
Night time in country	10
Soundproof room	approximate zero reference level

CHART 2. EVERY-DAY NOISE LEVELS

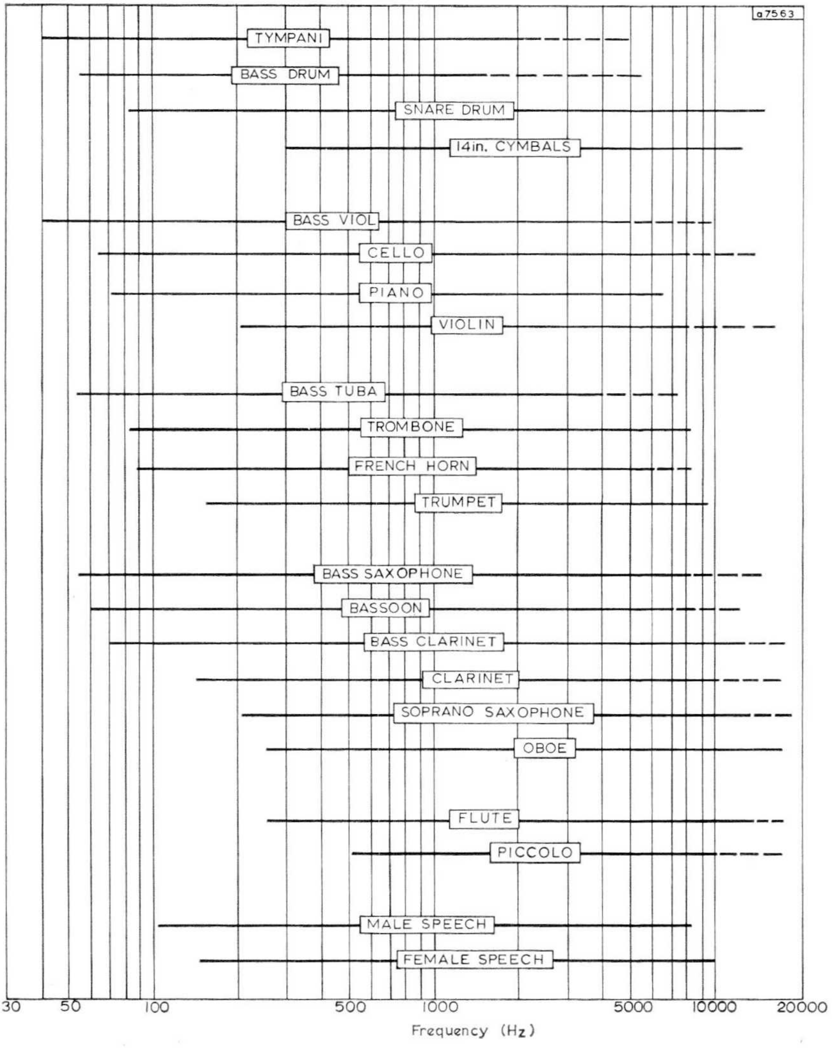


CHART 3. FREQUENCY RANGES OF MUSICAL INSTRUMENTS

CHART 4 RADIO FREQUENCY AND WAVELENGTH

The chart is self-explanatory, values of frequency being tabulated against corresponding values of wavelength. The example will illustrate the use of the chart.

Example: What is the wavelength corresponding to a frequency of 70MHz?

Opposite a figure of 7MHz on the scale a wavelength of 42.9 metres is shown. Multiplication by the correct factors for Band 5 gives a wavelength of 4.29 metres for a frequency of 70MHz.

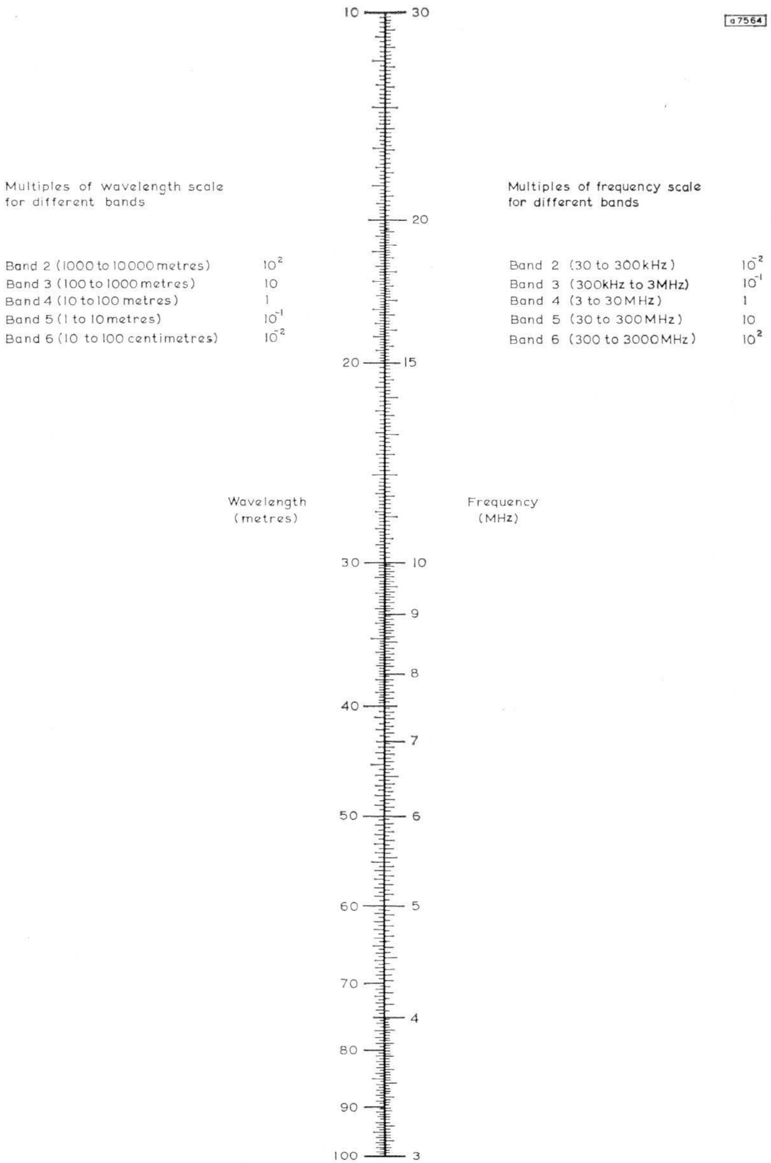


CHART 4. RADIO FREQUENCY AND WAVELENGTH

CHART 5 DECIBELS UP (a) and DOWN (b)

The formula expressing the relationship between power ratio R and decibels dB is:

$$dB = 10 \log_{10} R$$

The formula expressing the relationship between voltage or current ratio, R_e or R_i respectively, and decibels dB, is:

$$dB = 20 \log_{10} R_e$$

$$\text{or} \quad dB = 20 \log_{10} R_i$$

Two points should be remembered when expressing gain in decibels. Firstly, a decibel is a ratio, so gain can only be expressed in decibels when a reference level is understood. Secondly, although voltage and current gains may be expressed in decibels, the decibel is essentially a measure of power ratio. The gain of an amplifier may be given in decibels only if the input and output impedances are equal or some allowance is made for the difference between them. The voltage or current gain of an amplifier is given by the expression:

$$\text{Gain (dB)} = 20 \log_{10} R_e + 10 \log_{10} \frac{Z_1}{Z_2}$$

$$\text{or } 20 \log_{10} R_i + 10 \log_{10} \frac{Z_1}{Z_2}$$

where Z_1 and Z_2 are the input and output impedances of the amplifier.

The use of the charts is illustrated by two examples.

Example 1: What is the power ratio corresponding to 30dB up?

The power ratio is read directly from the 'power' scale (Chart 5a) and is 1000.

Example 2: What is the equivalent in decibels of a voltage ratio of 0·07?

The voltage ratio is located on the 'voltage and current' line (Chart 5b) and the equivalent in dB is read directly as —23dB

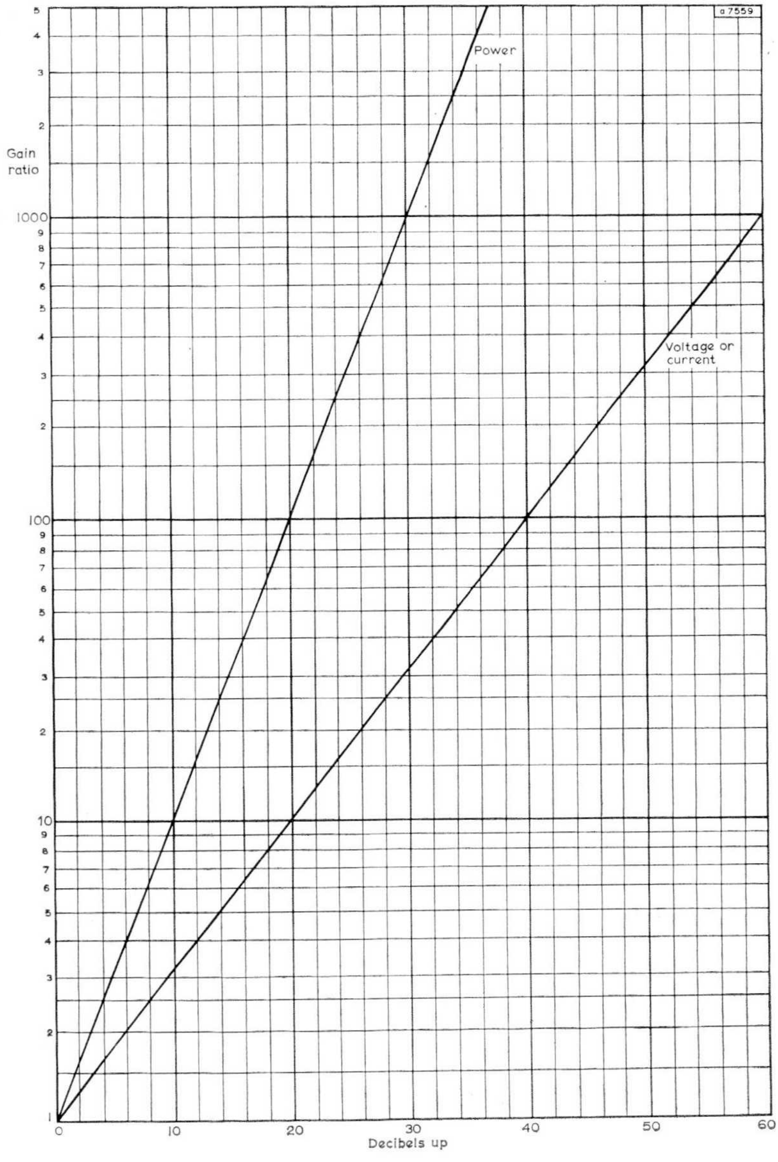


CHART 5a. DECIBELS UP

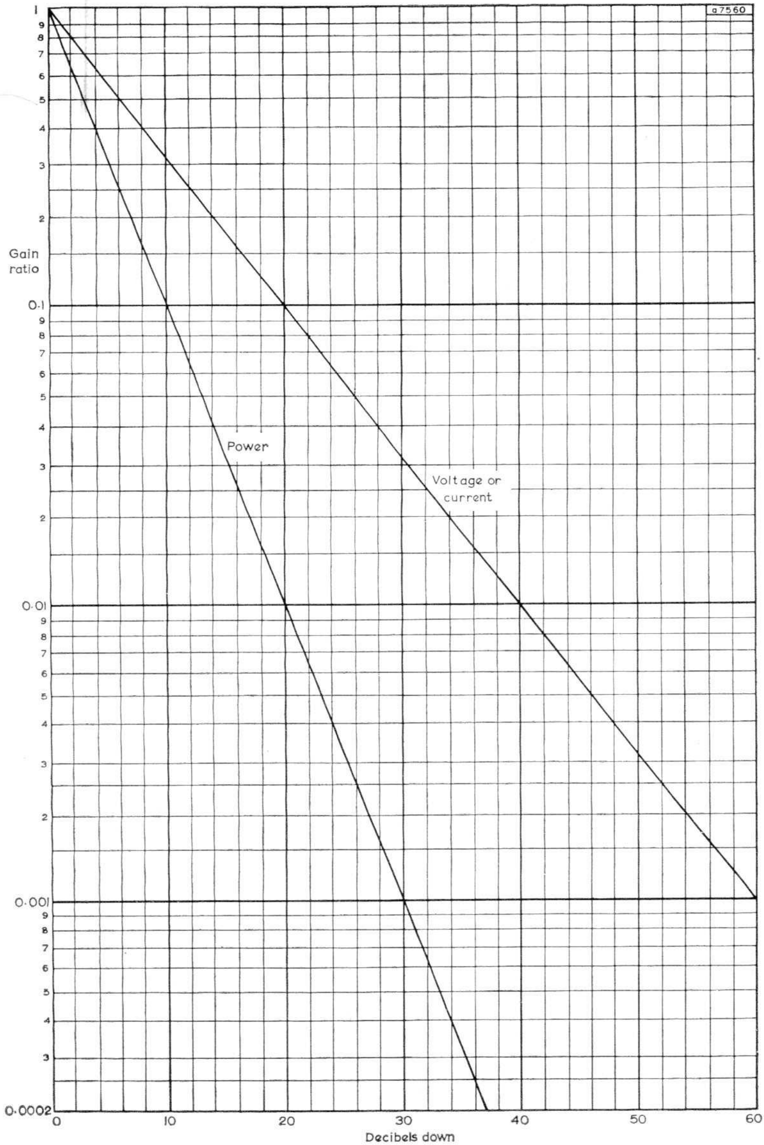


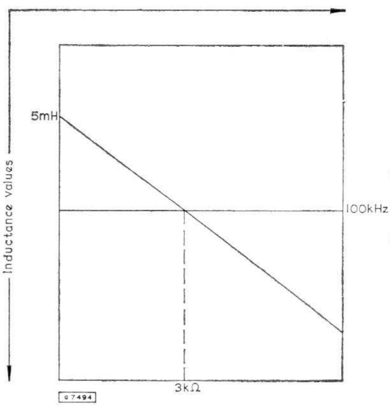
CHART 5b. DECIBELS DOWN

CHART 6 REACTANCE CHART

This chart is used to find the reactance of inductors and capacitors.

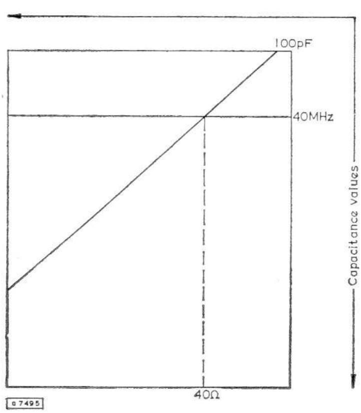
Values of inductance are represented by the lines with negative slope; the inductance scale is on the top and left of the chart. Values of capacitance are represented by the lines with positive slope; the capacitance scale is on the top and right of the chart. Values of frequency and reactance are represented by horizontal and vertical lines respectively. The following examples will illustrate the use of the chart.

Example 1: What is the reactance of a 5mH inductor at a frequency of 100kHz?



The intersection of the line representing 5mH with the 100kHz horizontal gives a reactance value of 3kΩ.

Example 2: What is the reactance of a 100pF capacitor at a frequency of 40MHz?



The intersection of the line representing 100pF with the 40MHz horizontal give a reactance value of 40Ω.