BABANI PRESS No.12

HI-FI, P.A., GUITAR & DISCOTHEQUE AMPLIFIER DESIGN HANDBOOK

BY

B.B.BABANI

INCLUDES CIRCUITS UP TO 1100 WATTS OUTPUT
TREMOLO, VIBRATO AND FUZZ-BOX ETC.

75p

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BABANI PRESS
The Publishing Division of
Babani Trading and Finance Co. Limited
The Grampians,
Shepherds Bush Road,
London W6 7NF

1st Published November 1972

Reprinted, October 73

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BERNARD B. BABANI

I.S.B.N. 0 85934 011 2

Printed by
Jos. Adam, Avenue E. G. Lebon 115, B-1160 Brussels, Belgium
for Babani Press, The Publishing division of Babani Trading
and Finance Co. Ltd., The Grampians, Western Gate,
London, W.6

INTRODUCTION

Dear Reader,

I would like to take this opportunity of thanking the following manufacturers of extremely high grade audio equipment for giving us permission to use data and circuitry from their technical publications and engineering departments.

G.E.C. Marconi Osram Valve Co. Ltd., Mullard Ltd., R.C.A. Ltd., Thorn Radio Valves and Tubes Ltd.

It should be noted, however, that no application should be made to any of these firms for help, further information or data with reference to any of the contents published in this book. Any help required must be sent to the publishers of this book who will then answer to the best of their ability.

We also would like to thank many technical magazines and manufacturers of audio equipment in the continent of Europe and the United States of America who have made available to us vast quantities of data.

NOTES

This book covers the design of amplifiers up to 1100 watts output and most of these circuits are using valves. The reason is that the design of high powered Hi Fi audio amplifiers using transistors with an output of above 100 watts are difficult to construct outside of commercial laboratories etc. and the cost could be extremely high. At a later date we shall publish a handbook covering construction of high powered solid state amplifiers, as soon as we are satisfied that the home constructor can build them without any serious problems arising. Many of the circuits shown for tremolo's, vibrato's and fuzz boxes use transistors as this makes them easy to construct, cheap and portable. They may be connected to valved apparatus without any problems and will work just as well as when operated with transistorized amplifiers of lower output.

SPECIAL NOTE FOR NORTH AMERICAN READERS

In order to help readers of this book in the U.S.A., a table of tube equivalents showing American types that may be used in place of the British tubes shown in the circuits in this book. Where no alternatives are shown, your local parts dealer will, undoubtedly, be able to supply you with a British original

British	U.S.A. Types
Tubes	
B309	12AT7-6061-6201-6679
DH77	6AT6-6BT6-6066
ECC82	12AU7-5814-5963-6067-6189
ECC83	12AX7-12BZ7-12DF7-5751-6057
ECL86	6GW8
EF86	6BK8-6CF8-6267
EL84	6BQ5-7189-7320
EZ90	6BX4-6X4-5993-6202
GZ34	5AR4-5AW4-5CG4-5T4-5U4-5W4
KT33C	25A6-25C6
KT66	6CN5-6L6-5881-5932
KT88	6CA7-6550
N709	6BQ5-7189-7320
U52	5AR4-5AW4-5CG4-5T4-5U4-5W4
U78	6BX4-6X4-5993-6202
U709	6BW4-6CA4
Z729	6BK8-6CF8-6267
Transistors	
BC109	SK3020
Rectifier Diodes	
BY105	SK3051

CONTENTS

CHAPTER 1

TREMOLO AND VIBRATO UNITS

CIRCUITS FOR TREMOLO UNITS

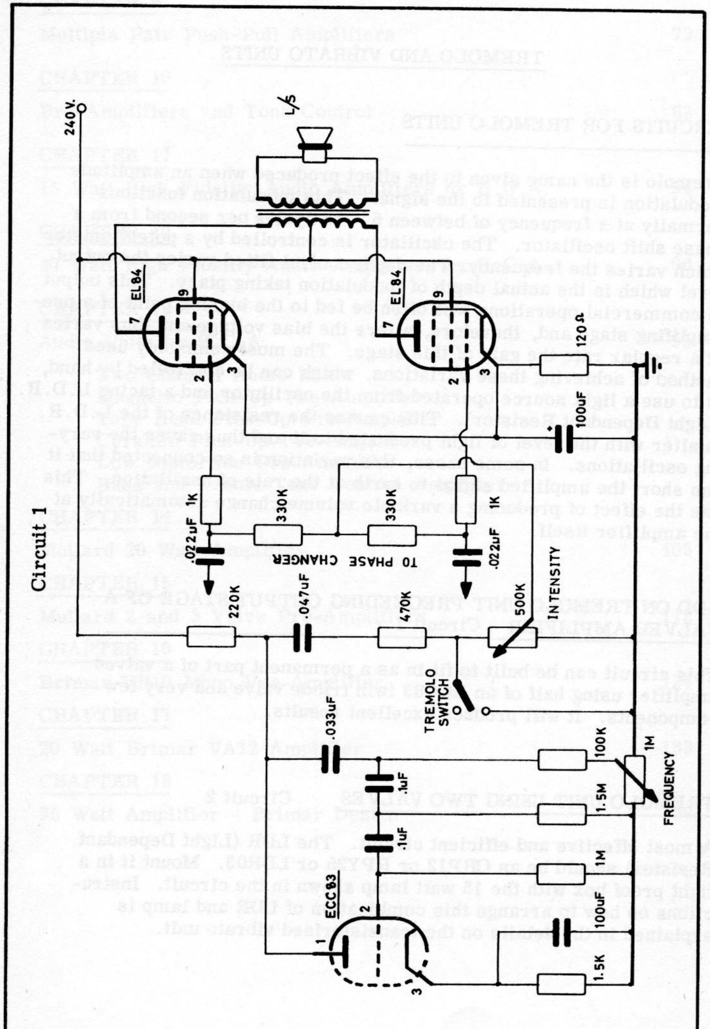
Tremolo is the name given to the effect produced when an amplitude modulation is presented to the signal. This modulation functions normally at a frequency of between 5 - 15 cycles per second from a phase shift oscillator. The oscillator is controlled by a potentiometer which varies the frequency. The depth control fitted varies the output level which is the actual depth of modulation taking place. This output in commercial operations will often be fed to the biasing point of a pre-amplifing stage and, therefore, alters the bias voltage and thus varies at a regular rate the gain of this stage. The most commonly used method of achieving these variations, which can be controlled by hand, is to use a light source operated from the oscillator and a facing L.D.R. (Light Dependant Resistor). This causes the resistance of the L.D.R. to alter with the level of light presented to it and thus cause the varying oscillations. In some cases, this variation is so connected that it can short the amplified signal to earth at the rate of oscillation. This has the effect of producing a variable volume change automatically at the amplifier itself.

ADD ON TREMOLO UNIT PRECEEDING OUTPUT STAGE OF A VALVED AMPLIFIER Circuit 1

This circuit can be built to fit in as a permanent part of a valved amplifier using half of an ECC83 twin triode valve and very few components. It will produce excellent results.

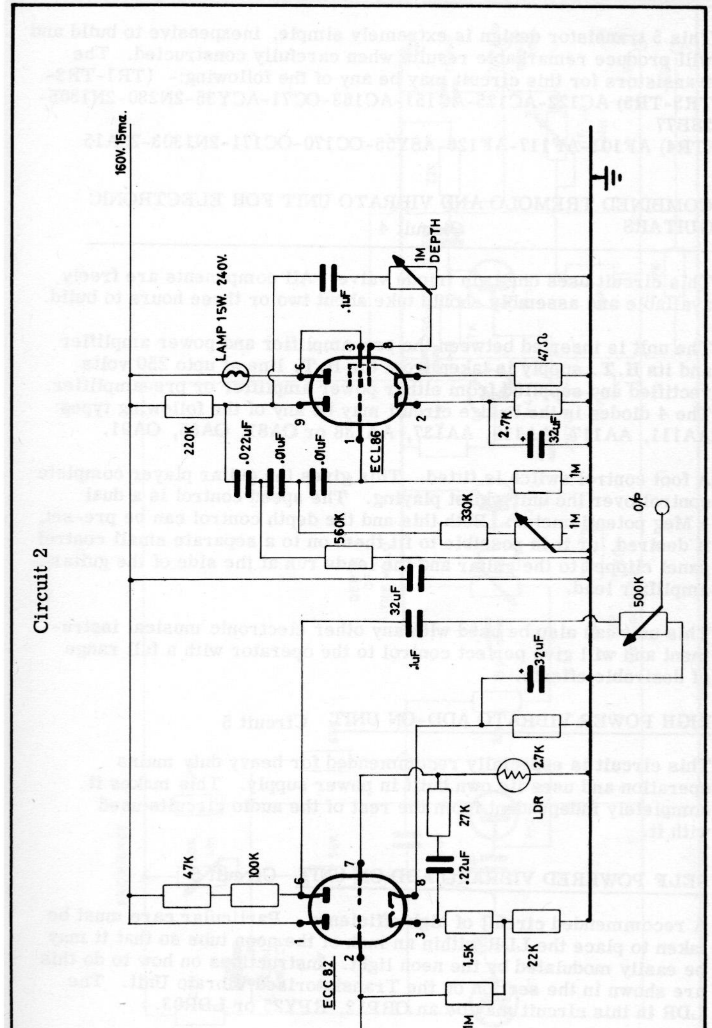
TREMOLO UNIT USING TWO VALVES Circuit 2

A most effective and efficient circuit. The LDR (Light Dependant Resistor) should be an ORP12 or RPY25 or LDR03. Mount it in a light proof box with the 15 watt lamp shown in the circuit. Instructions on how to arrange this combination of LDR and lamp is explained in the details on the transistorised vibrato unit.



Circuit 1

ADD-ON TREMOLO UNIT PRECEDING OUTPUT STAGE OF A VALVED AMPLIFIER



Circuit 2

TRANSISTORISED TREMOLO UNIT   Circuit 3

This 5 transistor design is extremely simple, inexpensive to build and will produce remarkable results when carefully constructed. The transistors for this circuit may be any of the following:- (TR1-TR2-TR3-TR5) AC122-AC125-AC151-AC163-OC71-ACY35-2N280-2N1305-2SB77
(TR4) AF101-AF117-AF126-ASY55-OC170-OC171-2N1303-2SA15

COMBINED TREMOLO AND VIBRATO UNIT FOR ELECTRONIC GUITARS   Circuit 4

This circuit uses one twin triode valve. All components are freely available and assembly should take about two or three hours to build.

The unit is inserted between the pre-amplifier and power amplifier and its H.T. supply is taken from the H.T. line at upto 250 volts rectified and supplied from either power amplifier or pre-amplifier. The 4 diodes in the bridge circuit may be any of the following types: AA111, AA117, AA119, AA137, AA138 or OA81, OA85, OA91.

A foot control switch is fitted. This gives the guitar player complete control over the unit whilst playing. The speed control is a dual 1 Meg potentiometer. Both this and the depth control can be pre-set, if desired, or it is possible to fit them on to a separate small control panel clipped to the guitar and the leads run at the side of the guitar amplifier lead.

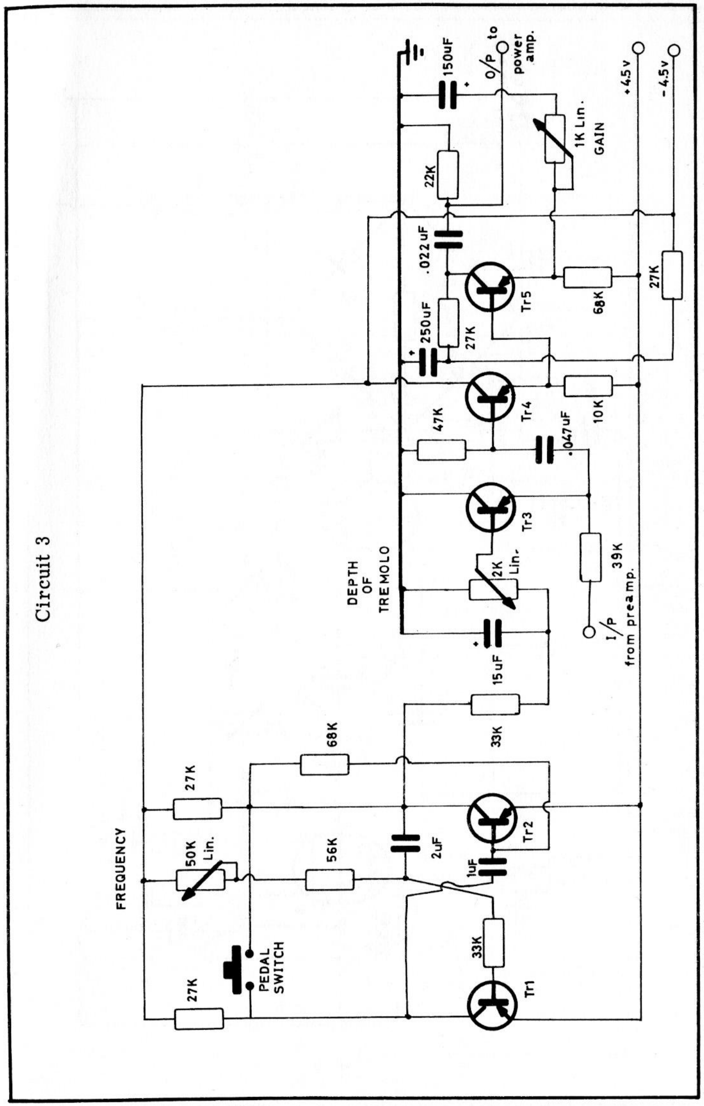
This unit can also be used with any other electronic musical instrument and will give perfect control to the operator with a full range of desirable effects.

HIGH POWER VIBRATO ADD-ON UNIT   Circuit 5

This circuit is especially recommended for heavy duty mains operation and uses its own built in power supply. This makes it completely independent from the rest of the audio circuits used with it.

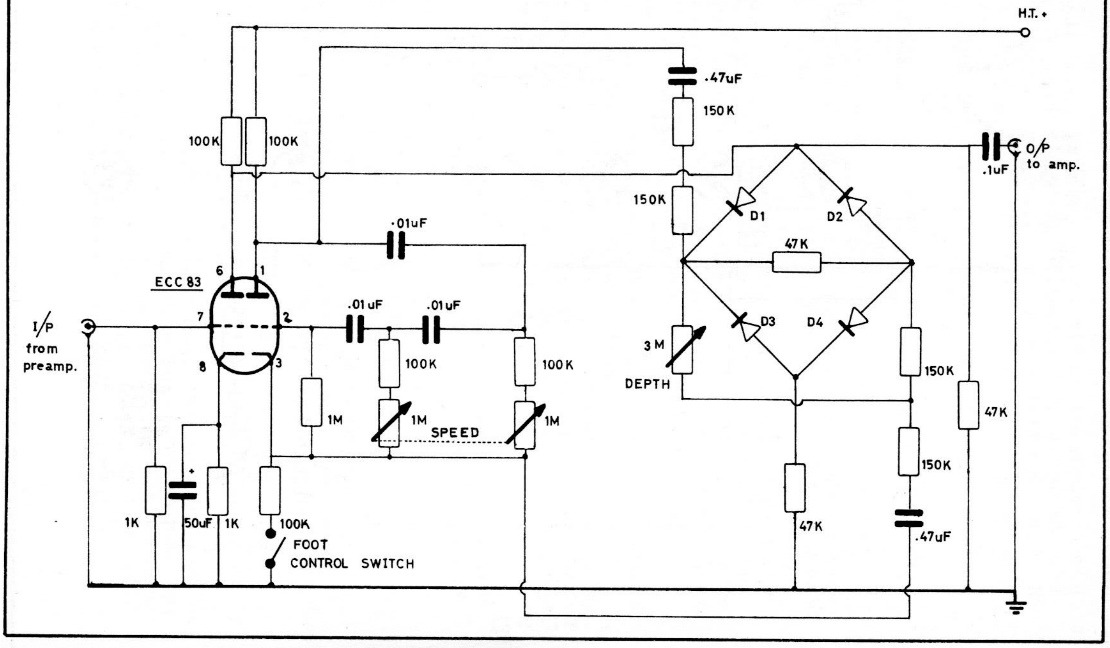
SELF POWERED VIBRATO ADD ON UNIT   Circuit 6

A recommended circuit of high efficiency. Particular care must be taken to place the LDR within an inch of the neon tube so that it may be easily modulated by the neon light. Instructions on how to do this are shown in the section on the Transistorised Vibrato Unit. The LDR in this circuit may be an ORP12, RPY25 or LDR03.



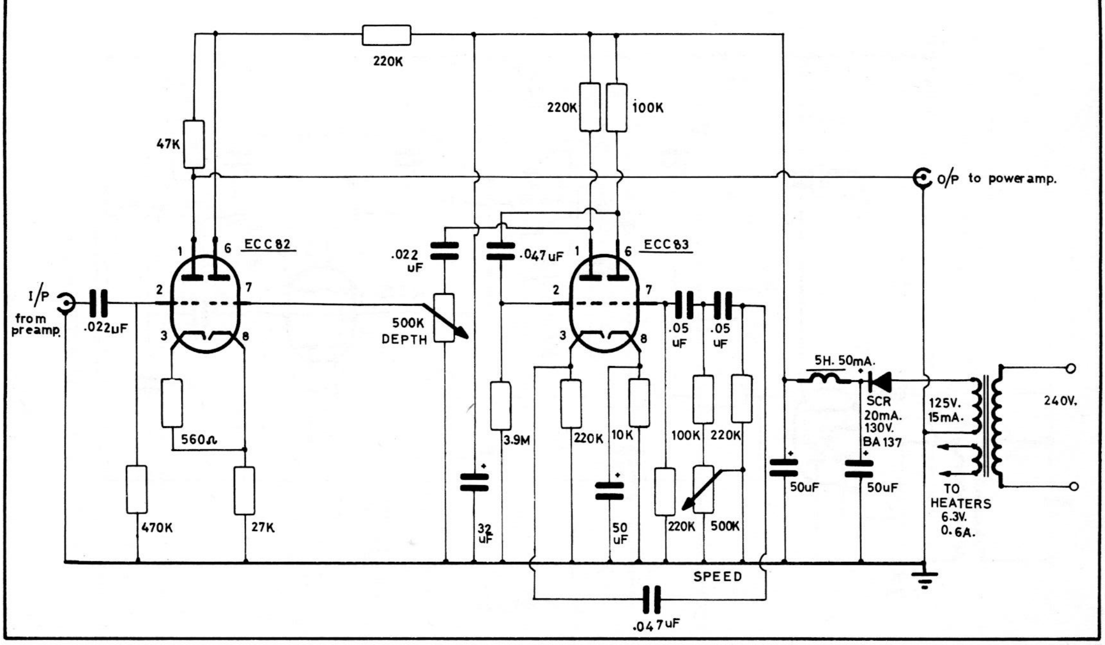
TRANSISTORISED TREMOLO UNIT

Circuit 4

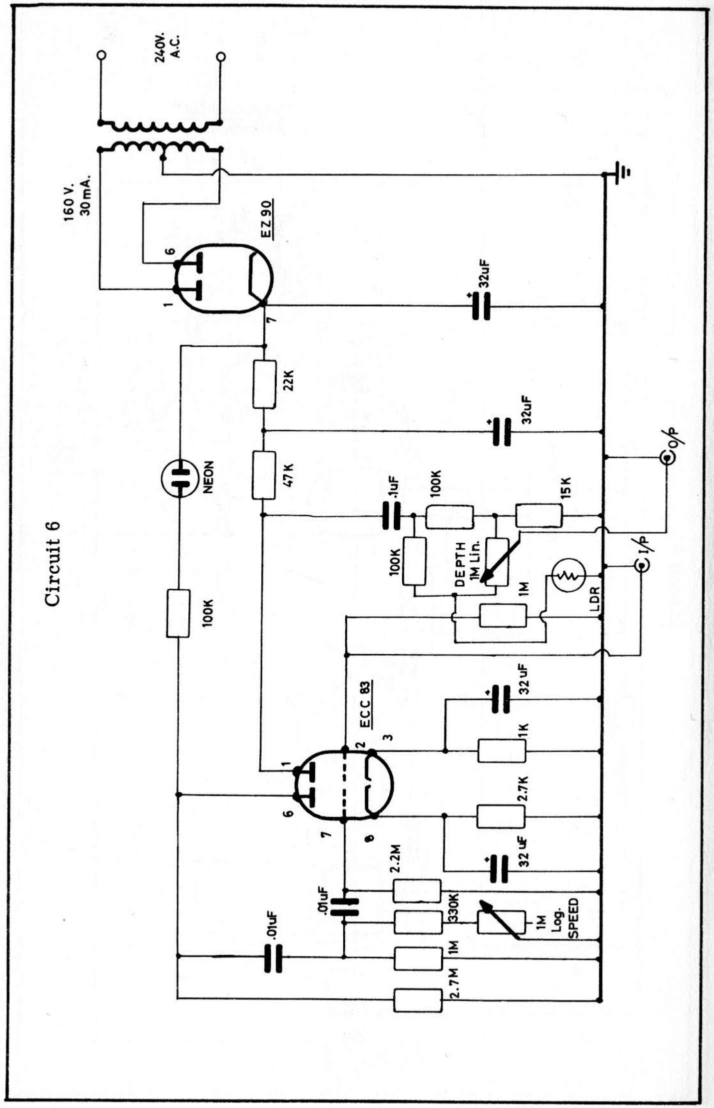


COMBINED TREMOLO AND VIBRATO UNIT FOR ELECTRONIC GUITARS

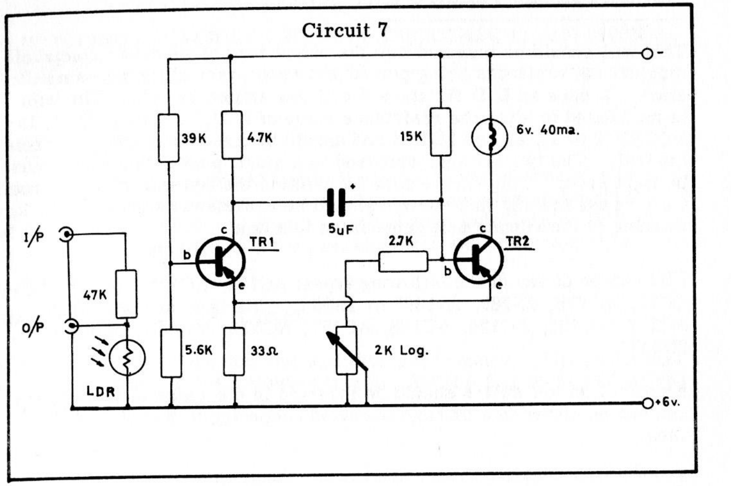
Circuit 5



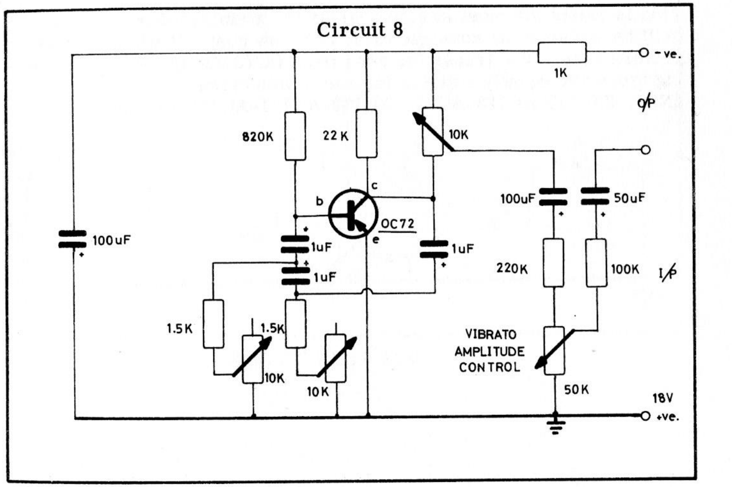
HIGH POWER VIBRATO ADD-ON UNIT



SELF POWERED VIBRATO ADD-ON UNIT



TRANSISTORISED VIBRATO UNIT



SIMPLE VIBRATO UNIT WITH SINGLE TRANSISTOR

TRANSISTORISED VIBRATO UNIT   Circuit 7

This unit should be inserted into the output lead of whatever electronic musical instrument is being played and must preceed the pre-amplifier stage. It uses an L.D.R. and a 6 volt low amperage bulb. The bulb is modulated to alter the resistance value of L.D.R. The L.D.R. is an ORP12 or RPY25 or LDR03 and should be fixed about one inch from the bulb. The two are then enclosed in a small plastic box which must be light proof. This can be done by painting the box with black enamel. Care in assembling this portion should be exercised as the perfect working of the vibrato unit depends on this box.

TR1 can be of any of the following types: AC122, AC125, AC151, AC163, OC71, ACY35, 2N280, 2N1305 or 2SB77. TR2 is either an AC105, AC117, AC121, AC124, AC128, AC153, ACY20, 2N1141, 2N1142 or 2N2391.

A control on/off switch should be inserted in the negative battery line and can be either foot or hand operated according to the wish of the user.

SIMPLE VIBRATO UNIT WITH SINGLE TRANSISTOR   Circuit 8

This is one of the most simple designs for a vibrato unit that can be built by the amateur constructor at very low cost. It will give good performance. The transistor used is an OC72 but any of the following types are equally suitable for use in this circuit:- NKT212-2N281-2N1305-AC128-AC131-AC132-AC151-AC153-ACY36.

CHAPTER 2

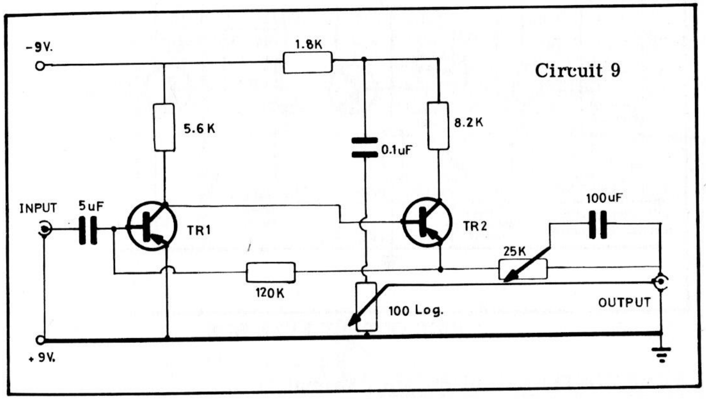
ELECTRONIC ACCESSORY UNITS FOR DISCOTHEQUE AND MUSICAL GROUP USE

FUZZ BOXES

Fuzz boxes are a recent addition to the growing list of special effect gadgets used with electronic music. The unit operates by squaring off the sound wave form to give a very distorted output which is then clipped before being fed into the pre-amplifier and power amplifier.

SIMPLE FUZZ BOX   Circuit 9

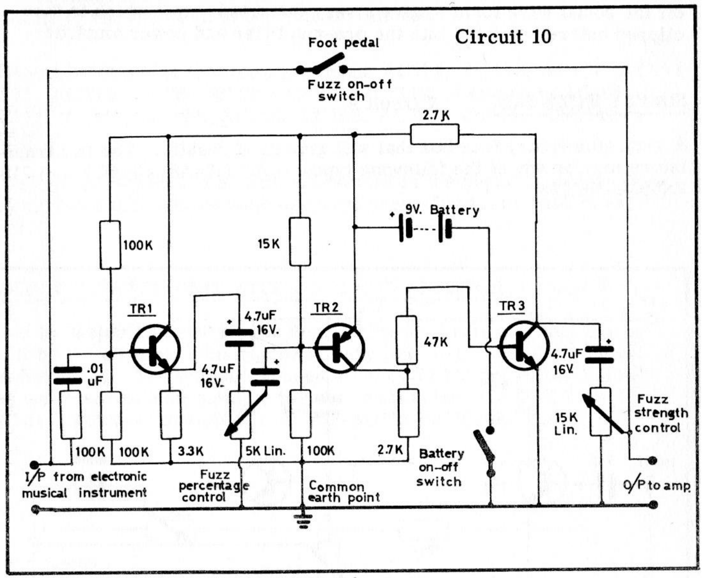
A very elementary fuzz box that will give good results. The two transistors may be any of the following types:- ACY17-ASY48-AC117-AC184 2N652A-NKT237



SIMPLE FUZZ BOX

HIGH OUTPUT FUZZ BOX Circuit 10

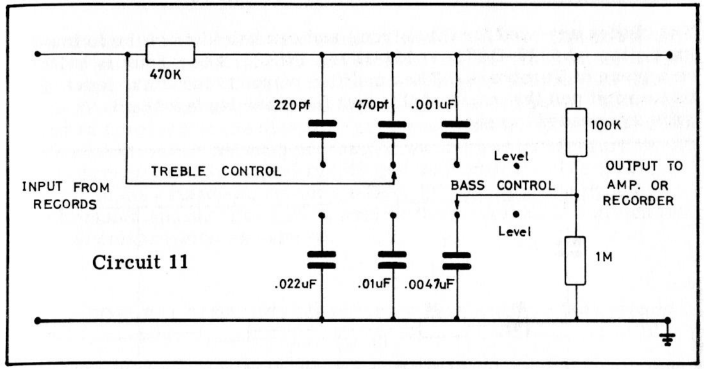
This design is the result of research work by R.C.A. and other well known manufacturers of audio equipment and is most effective in its performance. The transistors used are as follows:- (TR1) SK3020-BC107B (TR2) SK3005-OC45-AF127 (TR3) SK3020-BC107B. Alternatives are shown to the original in case any one has difficulty in obtaining the R.C.A. types SK3020 and SK3005 but it must be noted that the original circuit was designed around these 2 types of R.C.A. transistors.



HIGH OUTPUT FUZZ BOX

TONE CONTROL CORRECTOR Circuit 11

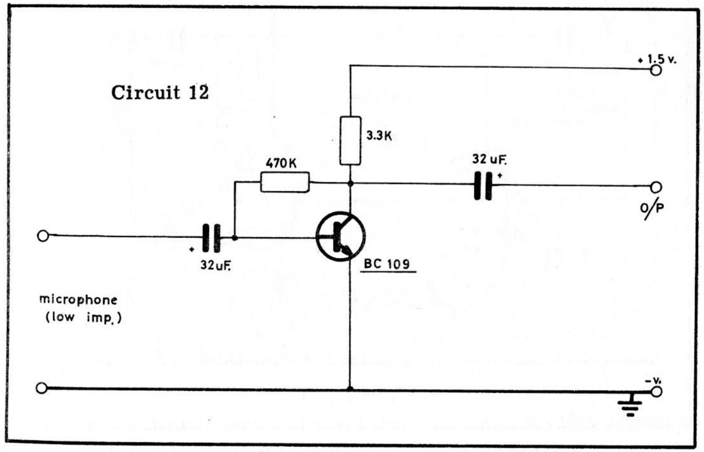
This easily built gadget allows the operator to decide how much treble or bass shall be recorded on to tape or into the amplifier via inputs from either microphone, records or electronic music. This is particularly effective in dealing with old pressings which are worn or scratched and can very often extend the life of an old record which would normally be destroyed.



TONE CONTROL CORRECTOR

GENERAL PURPOSE TRANSISTORISED MICROPHONE PRE-AMPLIFIER Circuit 12

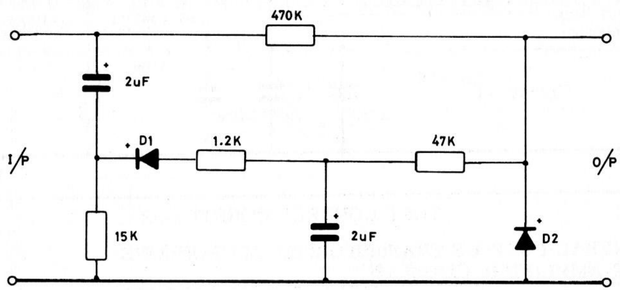
This design is of Spanish origin and very easy and cheap to build. Notice that the transistor used is a n-p-n type but if one has a p-n-p, such as an OC71 or OC72, this may be used. Simply reverse the battery polarity and the 32 mf capacitors should be wired with polarity in reverse to that shown in the circuit.



GENERAL PURPOSE TRANSISTORISED MICROPHONE PRE-AMPLIFIER

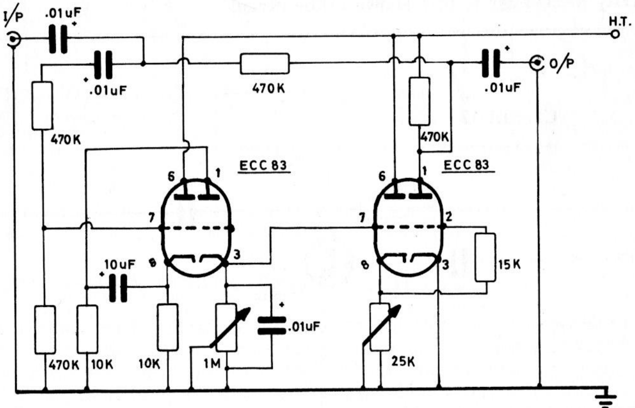
SUB-MINIATURE AUDIO COMPRESSOR Circuit 13

Two diodes are used for this circuit and can be either of the following types:- AA119-OA71-AA113-IN541-IN542. The circuit is built on a piece of Veroboard. The amplifier output is fed to the input of this circuit and the output of this unit feeds the loudspeaker. Two units are needed for stereo.



Circuit 13

SUB-MINIATURE AUDIO COMPRESSOR



Circuit 14

ELECTRONIC COMPRESSOR LIMITER WITH VALVES

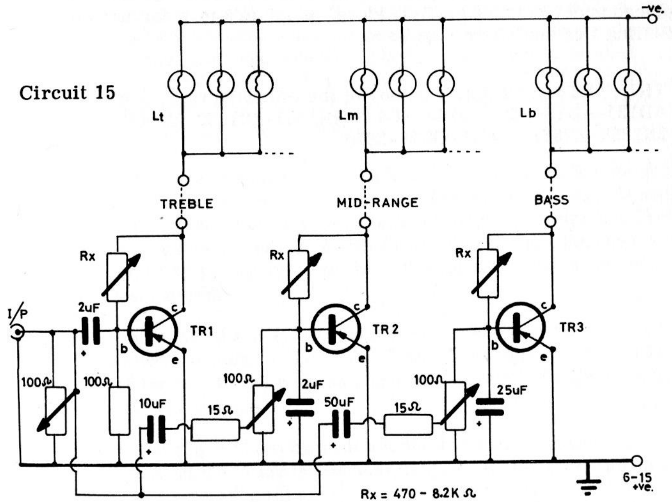
VALVED ELECTRONIC COMPRESSOR/LIMITER Circuit 14

This design produces a slow expansion rate but a very fast compression. This is particularly important in recording work. The unit uses 2 twin-triode valves and can be built quite cheaply. It draws its H.T. (high tension) from any rectified supply at about 150 volts. This may be taken from either the amplifier or pre-amplifier. This unit should be placed in circuit between pre-amplifier and main amplifier. None of the component values used in this unit are critical. The 1 Meg control in the cathode circuit of the first ECC83 allows variation of compression speed. The 25K control in the second ECC83 alters the level of compression as desired.

PSYCHEDELIC COLOURED LIGHT UNIT FOR AUDIO AMPLIFIERS Circuit 15

Many discotheques and even private audio enthusiasts will like a simple circuit which can produce coloured lights which literally dance when music is being played through an amplifier. The circuit is transistorised, inexpensive and extremely easy to build and produces superb

Circuit 15



PSYCHEDELIC COLOURED LIGHT UNIT

results with both small and large amplifiers. The unit is connected in parallel to any amplifier at the loudspeaker output point, and does not interfere with the amplifier's performance or the quality of reproduction from the loudspeakers. The circuit is not difficult to construct and the cost of the parts are extremely low.

In order to set up this unit it will be necessary to set a value for the three resistors marked Rx on the drawing and this will be somewhere between 500 - 10,000 ohms. It is recommeded that each of these be a 10K ohms linear potentiometer and their values adjusted by connecting the unit to the amplifier which must be switched on but with volume control at minimum so that the different coloured lamps each show a low glawing light which can just be seen under ordinary room lighting conditions. Once this adjustment has been made it should not be altered when the amplifier has its volume turned up and music fed through it so that the lights commence to dance.

The light unit may be operated with a supply voltage from 6 - 15 volts thus enabling it to be used if necessary from car batteries when connected to mobile amplifiers. When operating on a 6 volt supply, the bulbs used for the lights should be 2 volts each at 50 to 70 ma. When used with a 12 volt supply 6 volt, 60 - 200 ma bulbs are suitable. The bulbs in each of the three sound ranges should be either coloured with a heat proof translucent paint of the desired colours or they may be placed under coloured bulb holders, which are manufactured by Bulgin, Belling Lee and other suppliers.

TR1, TR2 and TR3 may be any of the following types:- AD130-AD131-AD138/50-AD140-AD149-2N1331-2N1333-2N1435-2N2836-2N2869-2N3611-OC16-OC26-2SB83.

CHAPTER 3
AMPLIFIERS OF 12 TO 14 WATTS

3-1. A 14W Ultra-Linear Amplifier.—This class of amplifier is probably the most popular for use in domestic equipment. As it is frequently required to work from a diversity of signal sources—radio tuners, record players, tape recorders, microphones etc., the domestic amplifier is generally used in conjunction with auxiliary pre-amplifying and equalisation equipment. Consequently, a slight departure is made here by describing a complete amplifier with tone controls and, in addition, alternative input selection and equalisation units. Higher power amplifiers suitable for domestic use are described in Chapter 4 and separate pre-amplifier and tone control circuits are discussed in Chapter 9.

The main amplifier circuit is given in fig. 3-3 and the two types of input unit appear in figs. 3-4 and 3-5. This design is based on the very popular " G.E.C. 912-PLUS " amplifier. The full output of 12-14W (the precise value depending upon the quality of the output transformer) is obtained with less than 1% total harmonic distortion over the frequency range of 30c/s to 20kc/s. An input of 50mV will give 12W output.

Comprehensive tone controls are built into the main amplifier and both the alternative input selection units provide switching for radio input, four different record characteristics and microphone. The pre-amplifiers described in Chapter 9 are very suitable for combining with the G.E.C. 912-PLUS when the flexibility of a separate pre-amplifier and tone control unit is desired. Details are given at the end of this chapter.

One of the units contains passive networks (i.e. there is no amplification) for equalising the four record characteristics and is intended for use with crystal pickups and microphones. The alternative unit is a single-valve pre-amplifier for which details are given not only for equalising the four record characteristics from different types of pickup but also for matching radio and microphone outputs to the main amplifier.

The passive input unit (fig. 3-4) has a sensitivity of 150mV on the " record " positions of the selector switch. The " radio " and " microphone " inputs are switched straight through to the main amplifier and consequently the original sensitivity of 50mV is retained.

The pre-amplifier (fig. 3-5) can be made specifically to suit the requirements of the particular pick-up, radio and microphone used, but maximum sensitivities are : 10mV for the " record " positions of the selector switch, 1mV for " microphone," and, nominally, 50mV for " radio." The required input impedance and sensitivity at any particular position of the input selector switch will depend upon the pickup or microphone used. Table 3-II gives appropriate alternative values for the resistors which determine these characteristics.

The first stage consists of a Z729 low-noise pentode followed by the tone control networks, described later. Next, the first section of a B309 double triode* is a typical voltage

*A B719 may be used ;

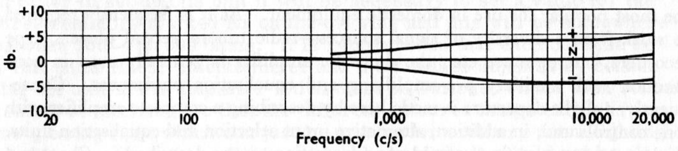


Fig. 3-1. 12W ultra-linear amplifier : effect of the " presence " control, with bass and treble controls at the level positions (mid-travel).

amplifier with negative feedback applied to the cathode circuit from the output transformer secondary. The feedback network contains an additional tone control (" presence ") as described later. The second half of the B309 is a phase-splitter which feeds the N709 ultra-linear output stage, the characteristics of which are given in Table 3-I. The output transformer is tapped at 20% to 40% of the turns of each half-primary from the centre tap. The power supply is of conventional design with a U709 feeding a capacitance-input filter.

Bass Loudness.—This control is continuously variable, giving linear response at the centre, bass attenuation anticlockwise, and bass accentuation clockwise.

Treble.—This is a switched control providing a linear position, two degrees of attenuation and one degree of boost.

Treble Slope.—A continuously variable control which alters the slope of the high frequency cut selected by the treble switch.

Presence.—This switch, the effect of which is shown in fig. 3-1, is designed to vary the listener's apparent position when listening to music. It introduces a frequency-selective network into the negative feedback loop.

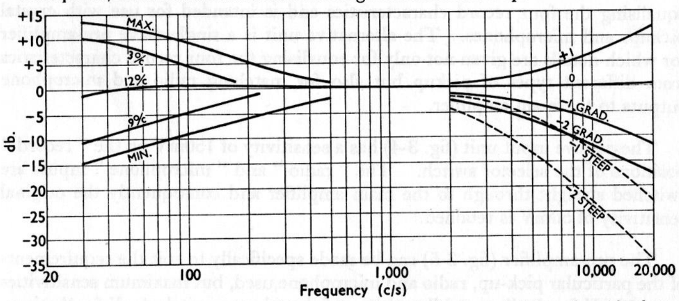


Fig. 3-2. 12W ultra-linear amplifier : the effect of the bass, treble and treble slope controls. The right-hand solid curves illustrate the four positions of the treble switch. The dotted curves show how the two treble cut curves are modified by fully rotating the treble slope control. The position of each dotted curve will alter as treble slope is decreased until, when it is at minimum, the treble response is that shown by the corresponding solid curve.

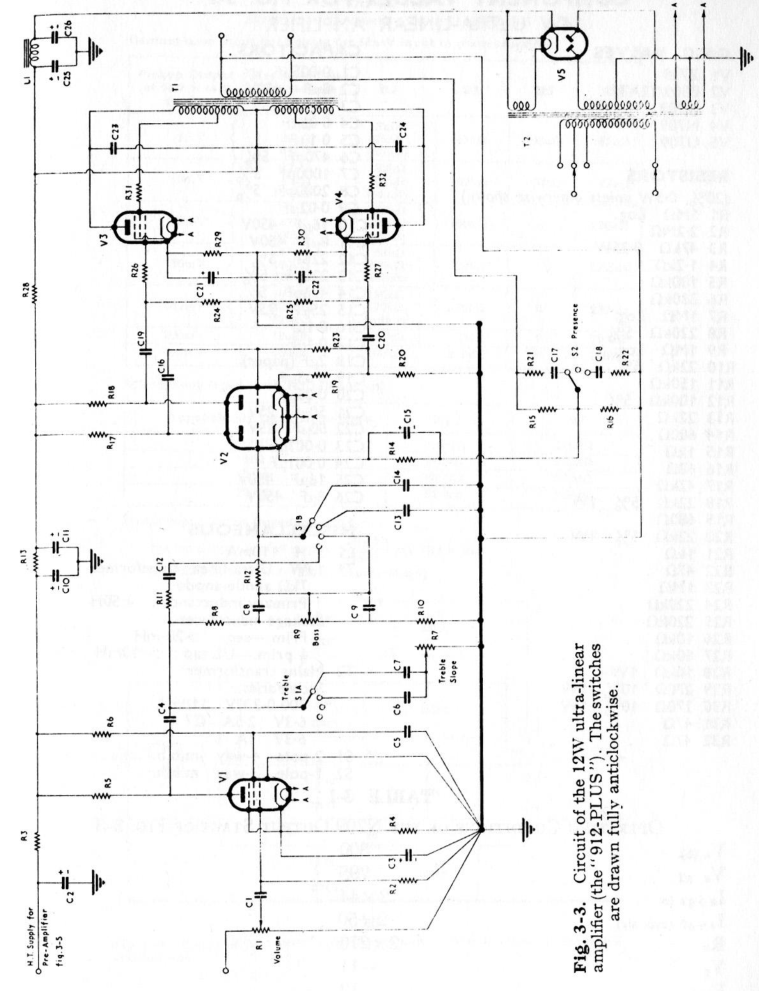


Fig. 3-3. Circuit of the 12W ultra-linear amplifier (the "912-PLUS"). The switches are drawn fully anticlockwise.

COMPONENT VALUES FOR FIG. 3-3

14W ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 Z729
V2 B309/12AT7
V3 N709
V4 N709
V5 U709

RESISTORS

(20%, 0·5W *unless otherwise shown*)

R1 1MΩ Log.
R2 2·2MΩ
R3 47kΩ 0·25W
R4 1·2kΩ
R5 100kΩ
R6 330kΩ
R7 1MΩ Log.
R8 220kΩ 5%
R9 1MΩ Log.
R10 22kΩ 5%
R11 150kΩ
R12 100kΩ 5%
R13 22kΩ
R14 680Ω
R15 1kΩ
R16 68Ω
R17 47kΩ
R18 22kΩ 5% 1W
R19 680Ω
R20 22kΩ 5% 1W
R21 1kΩ
R22 47Ω
R23 1MΩ
R24 220kΩ
R25 220kΩ
R26 10kΩ
R27 10kΩ
R28 10kΩ 1W
R29 270Ω 10% 1W
R30 270Ω 10% 1W
R31 47Ω
R32 47Ω

CAPACITORS

C1 0·005μF
C2 8μF 450V
C3 25μF 25V
C4 0·05μF
C5 0·1μF
C6 470pF 5%
C7 1000pF 5%
C8 2000pF 5%
C9 0·02μF
C10 16μF 450V
C11 8μF 450V
C12 22pF 5%
C13 220pF 5%
C14 470pF 5%
C15 25μF 25V
C16 0·1μF
C17 0·1μF
C18 2μF (paper)
C19 0·05μF
C20 0·05μF
C21 50μF 25V
C22 50μF 25V
C23 0·001μF
C24 0·001μF
C25 16μF 450V
C26 8μF 450V

MISCELLANEOUS

L1 10H 110mA
T1 14W Ultra-linear transformer
7kΩ anode-anode
Primary inductance : ≮50H
Leakage inductances :
Prim.—sec. : ≯20 mH
½ prim.—UL tap : ≯ 10mH
T2 Mains transformer
Secondaries :
300-0-300V 110mA
6·3V 2·5A CT
6·3V 1A
S1 2-pole 4-way m.b.b.
S2 1-pole 3-way m.b.b.

TABLE 3-I

OPERATING CONDITIONS OF THE N709 OUTPUT STAGE OF FIG. 3-3

$V_{a\ (b)}$	300	V
$V_{a,\ g2}$	295	V
$I_{a+g2\ (o)}$	2×44	mA
$I_{a+g2\ (max\ sig)}$	2×50	mA
R_k	2×270	Ω
V_g	−11	V
P_{out}	12	W
$R_{L\ (a-a)}$	7	kΩ
z_{out} (approx.)	2	kΩ
D	<1	%
$V_{in\ (rms)}$ (to first stage)	50	mV

TABLE 3-II

ALTERNATIVE COMPONENT VALUES FOR FIG. 3-5

Gramophone Input (all figures for 50mV input to main amplifier)

Pickup Output (at 4cm/s as a guide)	Required Load Impedance	R2	R3	R5	R19
5mV	10kΩ 22kΩ 47kΩ	15kΩ 47kΩ Infinity	47kΩ	100kΩ	47kΩ
10mV	22kΩ 47kΩ 100kΩ	33kΩ 100kΩ Infinity	100kΩ	100kΩ	47kΩ
20mV	22kΩ 47kΩ 100kΩ	33kΩ 100kΩ Infinity	100kΩ	0	220kΩ
50mV	47kΩ 100kΩ 220kΩ	68kΩ 220kΩ Infinity	220kΩ	0	220kΩ
100mV	100kΩ 220kΩ 470kΩ	150kΩ 470kΩ Infinity	470kΩ	0	220kΩ
200mV	100kΩ	100kΩ	1MΩ	0	220kΩ
500mV	100kΩ	100kΩ	2·2MΩ	0	220kΩ

Microphone Input (1mV for full output)

Approximate Input Impedance	R1	R7
2·2MΩ	Infinity	2·2MΩ
1MΩ	2·2MΩ	2·2MΩ
470kΩ	680kΩ	2·2MΩ
220kΩ	220kΩ	2·2MΩ

Radio Input (Input impedance = R4)

Required input for full output $= \frac{50 \times R4}{220}$ mV (R4 in kΩ).

Standard value for R4 = 220kΩ (input = 50mV).

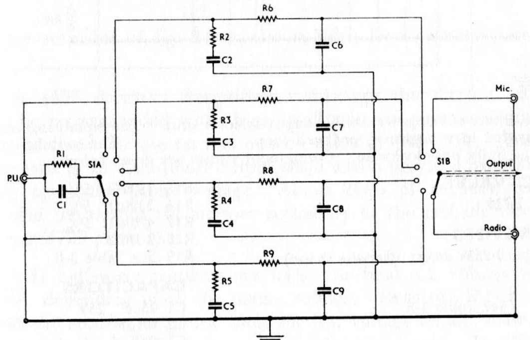


Fig. 3-4. Circuit of the passive input selector unit. Both switches are shown fully anticlockwise.

RESISTORS
(10%, 0·25W)

R1, R2, R3	470kΩ
R4, R5, R6	150kΩ
R7, R8, R9	470kΩ

CAPACITORS

C1	470pF	10%
C2	0·002μF	25%
C3	0·005μF	25%
C4	0·005μF	25%
C5	0·01μF	25%
C6	220pF	10%
C7	100pF	10%
C8	100pF	10%
C9	33pF	10%

SWITCH
S1 2-pole 6-way m.b.b.

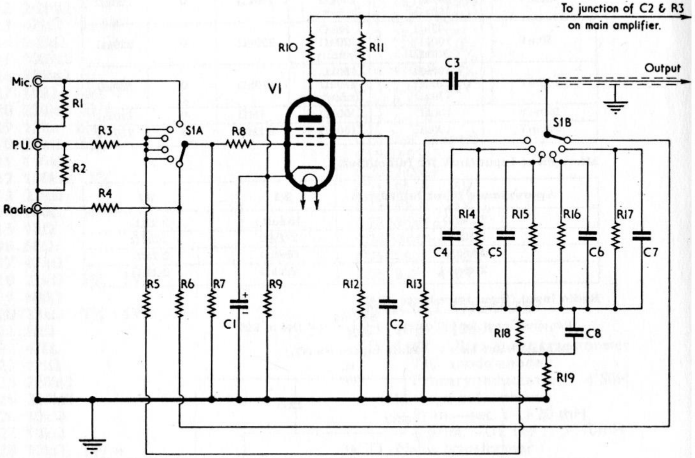


Fig. 3-5. Circuit of the pre-amplifier input selector unit. The equalising components are incorporated in a negative feedback loop The switches are shown fully anticlockwise. Component values are given below.

G.E.C. VALVE
V1 Z729

RESISTORS
(20%, 0·25W unless otherwise shown)

*R1, R2, R3, R4, R5	See Table 3-II		
R6	220kΩ		
R7	2·2MΩ		
R8	2·2kΩ		
*R9	3·3kΩ	5%	
*R10	220kΩ	5%	0·5W
*R11	470kΩ	5%	
*R12	330kΩ	5%	
R13	220kΩ		
R14	220kΩ	5%	
R15	330kΩ	5%	
R16	330kΩ	5%	
R17	470kΩ	5%	
R18	3·3MΩ		
R19	See Table 3-II		

**High Stability*

CAPACITORS

C1	25μF	25V
C2	0·1μF	
C3	0·05μF	
C4	330pF	10%
C5	150pF	10%
C6	68pF	10%
C7	33pF	10%
C8	1000pF	10%

SWITCH
S1 2-pole 6-way m.b.b.

Using a Separate Pre-Amplifier.—The 912-PLUS may be converted for use with the pre-amplifier described in Section 9-3 (fig. 9-3) in the following way:

Referring to fig. 3-3, the connection to the grid of the left-hand section of V2 is removed so that the whole of the circuit to the left-hand side of V2 is discarded. The " presence " components are removed from the feedback loop, leaving only the series feed resistor R15 and the lower part of the V2 cathode resistance R16. This results in a normal application of negative feedback from the output transformer secondary to the cathode circuit of the first valve of what is now the main amplifier.

The h.t. line is terminated at C10 and C11, R13 is removed and C10, C11 are strapped to give a total capacitance of 24μF. The h.t. series resistor R28 is reduced to 3·3kΩ (1W) and the h.t. connection to the pre-amplifier is taken from the junction of C11 and R28. No additional decoupling is required in the pre-amplifier h.t. line. The heater supply for the pre-amplifier should consist of a separate pair of twisted wires running direct to the mains transformer.

The pre-amplifier connections should be made through an octal socket mounted on the main amplifier chassis near V2 and, to prevent damage to V2 by disconnecting the pre-amplifier when the power is switched on, a grid leak of 1MΩ should be connected between the input grid of V2 and earth.

This combined equipment will have somewhat less overall distortion than the normal 912-PLUS, the second harmonic distortion being, in fact, as low as 0·3%.

The pre-amplifier of fig. 9-2 is also suitable for operation with the 912-PLUS

3-2. A 14W d.c./a.c. Amplifier.—Although the d.c./a.c. amplifier is hardly to be recommended when high quality is the main consideration, fig. 3-6 shows a circuit for use on d.c. or a.c. mains of 190-250V. Two KT33C valves are used in an ultra-linear output stage which provides from 7W to 14W depending upon the mains voltage. About 20db of negative feedback is applied from the output transformer secondary to the cathode circuit of the first valve, a Z729 pentode.

Two U31 half-wave rectifiers are used, the final h.t. voltage being from 180 to 240, depending upon the mains voltage. Resistors R23 and R25, in series with the rectifier cathodes, limit the h.t. voltage on a.c. mains to about the same as that obtained from d.c. mains of the same value. In addition these resistors limit the rectifier peak current and equalise the flow through the two valves. The power supply circuit in fig. 3-6 on 190-250V a.c. or d.c. mains but, for continuous operation on 240-260V a.c., R23 and R25 may be increased to 180Ω each in order to limit the anode and screen dissipation of the output valves.

For d.c. operation only, the rectifiers, resistors R23, R25 and the reservoir capacitor C14 may be omitted. Without rectifiers there is no protection for the electrolytic capacitors against reversed polarity of the mains and paper types should be substituted. A lower capacitance of 4μF will be adequate for C8 and C11 on most d.c. mains.

The current in the heater chain is controlled by a 303 barretter for mains voltages of 190-250 but for continuous operation above 240V a Type 304 should be substituted.

On low mains voltages, the resistances of L1 and T1 are important and should be as low as possible.

Table 3-III gives the operating data for the output stage of this amplifier.

COMPONENT VALUES FOR FIG. 3-6

KT33C 14W DC/AC ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 Z729
V2 L63/6J5
V3 KT33C
V4 KT33C
V5 U31
V6 U31
V7 Barretter 303

RESISTORS

(20%, 0·5W *unless otherwise shown*)

R1 500kΩ
R2 250√speech coil impedance
R3 2·2kΩ
R4 47Ω
R5 220kΩ
R6 1MΩ
R7 15kΩ
R8 2·2MΩ
R9 15kΩ 1W } matched to 5%
R10 15kΩ 1W } matched to 5%
R11 1·5kΩ
R12 220kΩ
R13 220kΩ
R14 15kΩ
R15 15kΩ
R16 220Ω 5% 1W
R17 220Ω 5% 1W
R18 15kΩ 1W
R19 100Ω
R20 100Ω
R21 470—1500Ω
R22 470—1500Ω
R23 100Ω 5W
R24 62Ω 5W
R25 100Ω 5W

CAPACITORS

C1 0·02μF
C2 25μF 25V
C3 0·1μF
C4 4μF 350V
C5 0·05μF
C6 0·1μF
C7 0·1μF
C8 32μF 350V
C9 25μF 25V
C10 25μF 25V
C11 32μF 350V
C12 1000pF
C13 1000pF
C14 32μF 450V

MISCELLANEOUS

L1 5—10H 200mA
T1 14W Ultra-linear transformer
3·2kΩ anode-anode
Primary inductance : ≮25H
Leakage inductances :
Prim.—sec. : ≯5mH
½ prim—UL tap : ≯5mH

TABLE 3-III

OPERATING CONDITIONS OF THE KT33C OUTPUT STAGE OF FIG. 3-6

$V_{a\ (b)}$ (approx)	235	V
$V_{a,\ g2}$ (approx)	220	V
$I_{a+g2\ (o)}$	2×70	mA
$I_{a+g2\ (max\ sig)}$	2×75	mA
$p_{a+g2\ (o)}$	2×15	W
$p_{a+g2\ (max\ sig)}$ (approx)	2×10	W
R_k	2×220	Ω
V_g	−15	V
P_{out}	14	W
$R_{L\ (a\text{-}a)}$	3·2	kΩ
z_{out} (approx)	3	kΩ
D	<0·5	%
$V_{in\ (rms)}$ (to first stage)	0·5	V

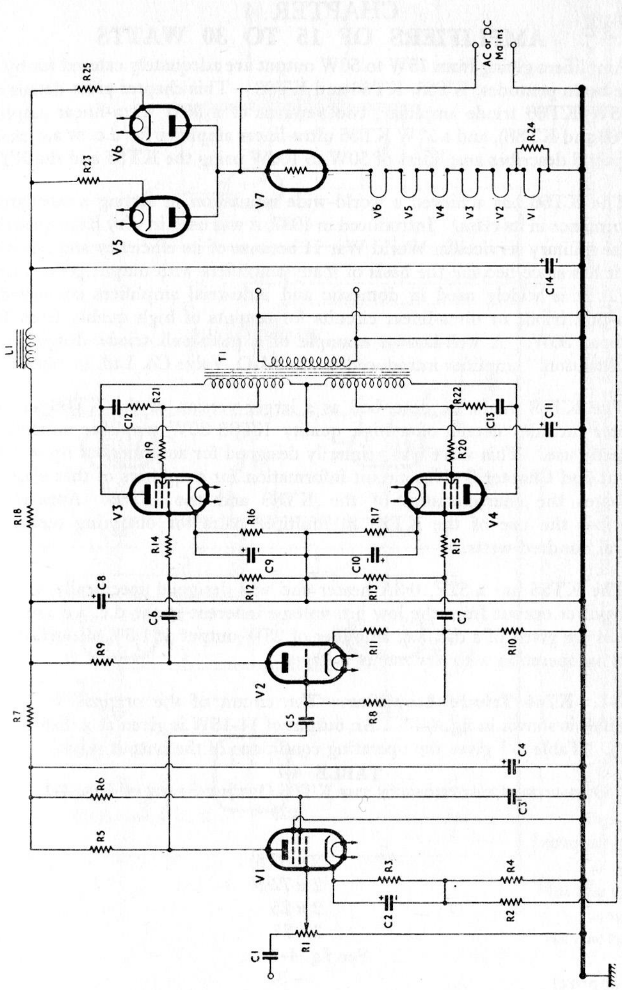


Fig. 3-6. Circuit of the d.c./a.c. 14W amplifier.

CHAPTER 4
AMPLIFIERS OF 15 TO 30 WATTS

Amplifiers giving from 15W to 50W output are adequately catered for by the three beam pentodes, KT55, KT66 and KT88. This chapter gives details of a 14-15W KT66 triode amplifier, two versions of a 30W ultra-linear amplifier (KT66 and KT88), and a 25W KT55 ultra-linear amplifier for d.c. or a.c. mains. Chapter 5 describes amplifiers of 50W to 100W using the KT55 and the KT88.

The KT66 has achieved a world-wide reputation in setting a standard of performance in its class. Introduced in 1937, it was used in very large quantities by the military services in World War II because of its efficiency and reliability and it has since become the basis of many amplifiers with output powers up to 50W. It is widely used in domestic and industrial amplifiers connected in push-pull triode or ultra-linear circuits for outputs of high quality from 15W to about 30W. A well-known example of a push-pull triode design is the " Williamson " amplifier introduced by the M-O. Valve Co. Ltd. in 1947.

The KT88 might be described as a larger version of the KT66 and this chapter includes details of a high quality KT88 30W amplifier suitable for domestic use. This valve was primarily designed for amplifiers of up to 100W output and Chapter 5 gives circuit information for amplifiers of that class and compares the characteristics of the KT88 and the KT66. Appendix A describes the use of the KT88 in multiple pairs for obtaining outputs of several hundred watts.

The KT55 has a 52V, 0·3A heater and was designed specifically to give a high-power output from the low h.t. voltage inherent in the d.c./a.c. amplifier. Details are given of a d.c./a.c. amplifier of 25W output at 1·5% distortion with notes on operation with low mains voltages.

4-1. KT66 Triode Amplifier.—The circuit of the original Williamson amplifier is shown in fig. 4-1*. An output of 14-15W is given at a distortion of 0·1%. Table 4-I gives the operating conditions of the output stage.

TABLE 4-I

OPERATING CONDITIONS OF THE KT66 OUTPUT STAGE OF FIG. 4-1

$V_{a\ (b)}$	450	V
$V_{a,\ g2}$ (approx)	410	V
$I_{a+g2\ (o)}$	2×62·5	mA
$I_{a+g2\ (max\ sig)}$	2×72·5	mA
$p_{a+g2\ (o)}$	2×25	W
$p_{a+g2\ (max\ sig)}$	2×21	W
R_k	See fig. 4-1	
V_g (approx)	−37	V
P_{out}	15	W
$R_{L(a-a)}$	10	kΩ
z_{out}	300	Ω
D	<0·1	%
$V_{in\ (pk)}$ (approx) (to first stage)	1·9	V

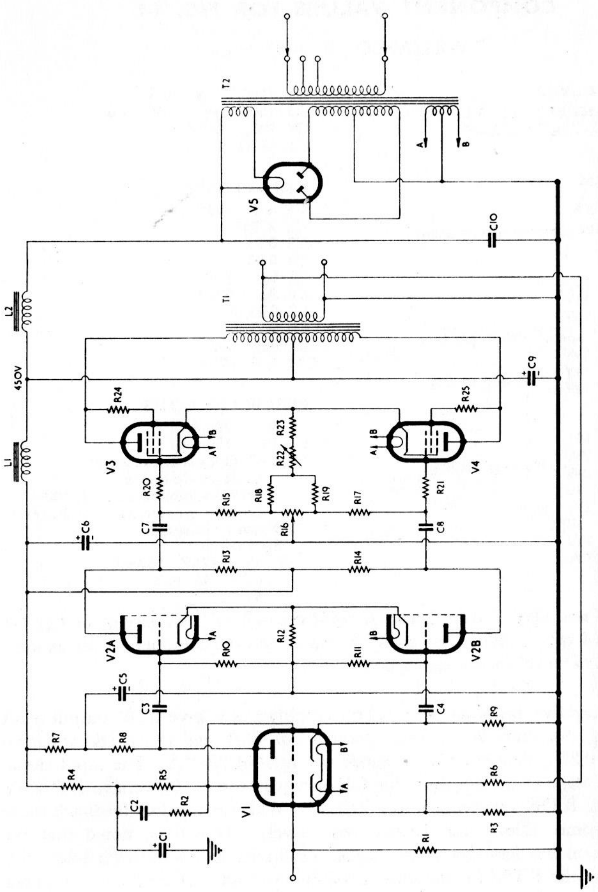


Fig. 4-1. Circuit of the " Williamson " amplifier. R22 is adjusted for a total cathode current of 125mA for the two output valves. The output stage is balanced by connecting a d.c. voltmeter across the whole of the output transformer primary and adjusting R16 for zero reading. C2 and R2 increase the margin of stability at high frequencies.

COMPONENT VALUES FOR FIG. 4-1

" WILLIAMSON " AMPLIFIER

G.E.C. VALVES

V1 B65/6SN7 or 2× L63/6J5
V2 B65/6SN7 or 2× L63/6J5
V3 KT66
V4 KT66
V5 U52/5U4 or U54

RESISTORS

(20%, 0·25W *unless otherwise shown*)

R1 1MΩ
R2 4·7kΩ
R3 470Ω 10%
R4 33kΩ 1W
R5 47kΩ 1W
R6 1200√ speech coil impedance
R7 22kΩ 1W
R8 22kΩ 1W } Matched to 5%
R9 22kΩ 1W }
R10 470kΩ
R11 470kΩ
R12 390Ω 10%
R13 47kΩ 2W } Matched to 5%
R14 47kΩ 2W }
R15 100kΩ 10%
R16 100Ω 2W w.w.
R17 100kΩ 10%
R18 100Ω 1W
R19 100Ω 1W
R20 1kΩ
R21 1kΩ
R22 100Ω 2W w.w.
R23 150Ω 5% 3W w.w.
R24 100Ω 0·5W
R25 100Ω 0·5W

CAPACITORS

C1 8μF 500V
C2 200pF
C3 0·05μF
C4 0·05μF
C5 8μF 500V
C6 8μF 500V
C7 0·25μF
C8 0·25μF
C9 8μF 500V
C10 8μF 600V

MISCELLANEOUS

L1 30H 20mA
L2 10H 150mA
T1 14W Output transformer
10kΩ anode-anode
Primary inductance : ≮ 100H
Leakage inductance : ≯ 30mH
T2 Mains transformer
Secondaries :
425-0-425V 150mA
6·3V 4A CT
5V 3A

4-2. Two 30W Ultra-Linear Amplifiers.—The basic circuit of fig. 4-4 may be used with either KT66 or KT88 valves and the components list on page 43 gives suitable values for each type of valve.

With negative feedback, the KT66 amplifier will give 32W output with about 0·5% distortion at an anode potential of 400V and the KT88 will give 32W with 0·25% distortion at an anode potential of 335V.* The input signal to the first stage of the amplifier for full output in the KT66 version is 600mV whereas the KT88 version requires 500mV. With no negative feedback these figures become 120mV and 100mV respectively. It will be noted that the KT88 version has less distortion, higher sensitivity and requires a lower h.t. voltage than the KT66 for the same maximum output. If feedback is omitted the KT88 amplifier has only 1% distortion at 32W output as against the 2% of the KT66 version. As the KT88 valves are conservatively run in this circuit they will have a long life.

The output stage is preceded by a conventional double triode voltage amplifier which is fed by a triode phase-splitter comprising one half of a further double triode. The other half of this valve is the input stage voltage amplifier, which is

*The KT88 output stage is, in fact, operated here almost in Class A, hence the reduced distortion compared with the KT66 version.

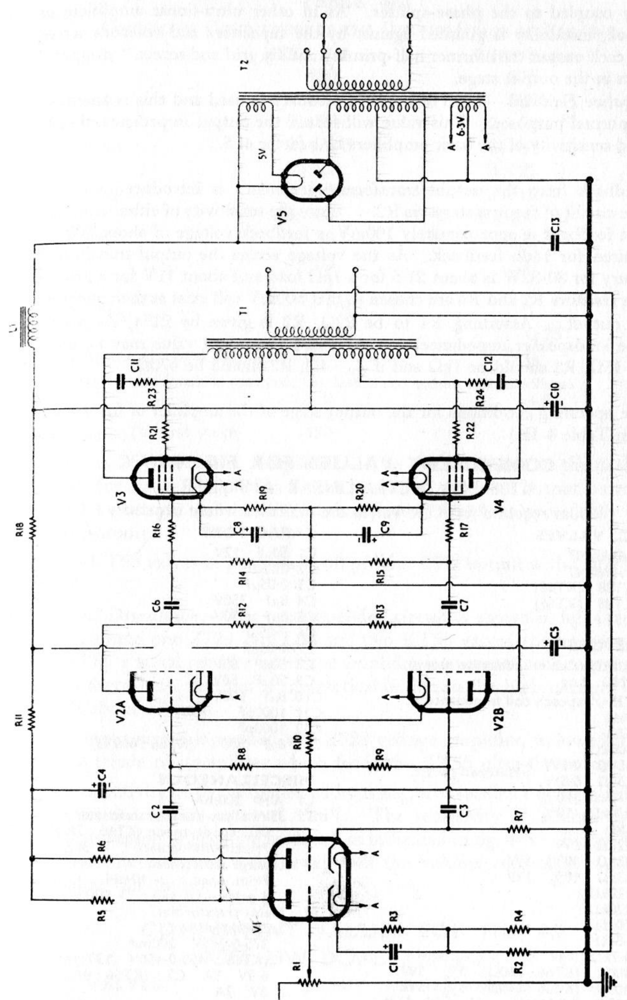


Fig. 4-4. Circuit of a 30W ultra-linear amplifier using KT66 or KT88 valves in the output stage. The component values and circuit conditions differ : see the components list opposite.

directly coupled to the phase-splitter. As in other ultra-linear amplifiers in this book, instability is guarded against by the capacitors and resistors across part of each output transformer half-primary and by grid and screen " stopper " resistors in the output stage.

Negative Feedback.—14db of negative feedback is used and this is adequate for all normal purposes. This value will reduce the output impedance, distortion and sensitivity of the basic amplifiers by a factor of 5.

Feedback from the output transformer secondary is introduced into the cathode circuit of the first stage via R2. Since the sensitivity of either amplifier without feedback is approximately 100mV, a feedback voltage of about 500mV is required for 14db feedback. As the voltage across the output transformer secondary for 30-32W is about 21·5 for a 15Ω load and about 11V for a load of 4Ω, the resistors R2 and R4 are chosen so that 500mV will exist at their junction at full output. Assuming R4 to be 22Ω, R2 is given by $225\sqrt{Z_o}$ (where Z_o=the loudspeaker impedance) and the nearest standard value may be used. If Z_o=15Ω, R2 should be 1kΩ and if Z_o=4Ω, R2 should be 470Ω.

The operating conditions for the output stage of the amplifier of fig. 4-4 are given in Table 4-II.

COMPONENT VALUES FOR FIG. 4-4
KT88 30W ULTRA-LINEAR AMPLIFIER
Values required for KT66 valves are indicated where necessary

G.E.C. VALVES

V1 B65/6SN7
V2 B65/6SN7
V3 KT88 (KT66)
V4 KT88 (KT66)
V5 U54

RESISTORS

(*20%, 0·25W unless otherwise shown*)

R1 1MΩ Log.
*R2 225$\sqrt{\text{speech coil impedance}}$
R3 1kΩ
R4 47Ω
R5 100kΩ
R6 15kΩ 0·5W } Matched to 5%
R7 15kΩ 0·5W }
R8 470kΩ 10%
R9 470kΩ 10%
R10 1kΩ
R11 22kΩ 1W
R12 33kΩ 10% 1W
R13 33kΩ 10% 1W
R14 220kΩ
R15 220kΩ
R16 10kΩ
R17 10kΩ
R18 4·7kΩ 1W
R19 400Ω (KT66 : 560Ω) 5% 5W
R20 400Ω (KT66 : 560Ω) 5% 5W
R21 270Ω 0·5W
R22 270Ω 0·5W
R23 470—1500Ω 0·5W
R24 470—1500Ω 0·5W

**For 14db feedback.*

CAPACITORS

C1 50μF 12V
C2 0·05μF
C3 0·05μF
C4 8μF 350V
C5 8μF 450V
C6 0·05μF
C7 0·05μF
C8 50μF 50V
C9 50μF 50V
C10 8μF 500V
C11 1000pF
C12 1000pF
C13 8μF 500V (KT66 : 600V†)

MISCELLANEOUS

L1 10H 200mA
T1 35W Ultra-linear transformer
6kΩ anode-anode (KT66 : 7kΩ)
‡Primary Inductance : ≮ 50H
Leakage inductances :
‡Prim.—sec. : ≯ 10mH
‡½ prim.—UL tap : ≯ 10mH
T2 Mains transformer
Secondaries :
375-0-375V 200mA
(KT66 : 450-0-450V 150mA)
6·3V 5A CT (KT66 : 4A)
5V 3A

†Or two 16μF 350V in series.

‡With these values, R23, R24 and C11, C12 may be omitted.

TABLE 4-II

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 4-4

	KT66 Valves	KT88 Valves	
V_a (b)	450	375	V
$V_{a,\ g2}$	400	335	V
I_{a+g2} (o)	2×62·5	2×80	mA
I_{a+g2} (max sig)	2×72·5	2×85	mA
p_{a+g2} (o)	2×25	2×27	W
p_{a+g2} (max sig)	2×15	2×12	W
R_k	2×560	2×400	Ω
V_g (approx)	−36	−32	V
P_{out}	32	30	W
R_L (a−a)	7	5	kΩ
Z_{out}	1·8	1	kΩ
D	0·5	0·25	%
V_{in} (rms) (approx) (to first stage)	600	500	mV

If negative feedback is omitted, the last three values are as follows :

Z_{out}	9	4·5	kΩ
D	2	1	%
V_{in} (rms) (approx) (to first stage)	120	100	mV

4-3. A 25W d.c./a.c. Amplifier.—The main problem in the design of an amplifier for d.c./a.c. operation lies in providing adequate output power with a limited h.t. voltage, and the KT55 beam pentode has been specifically designed for this purpose.

Two KT55 valves in push-pull will provide 25W output with a mains supply of 225V.

Circuit Description.—The recommended circuit is shown in fig. 4-5 and it is designed round one Z729, one L63 and two KT55 valves, the h.t. supply being provided by a metal rectifier when the amplifier is operated on a.c. mains. The use of a thermionic rectifier is impracticable due to the high current required, which is about 275mA.

The input signal is applied to a Z729 voltage amplifier, followed by a conventional triode phase splitter which feeds the KT55 ultra-linear output stage.

The sensitivity of the amplifier without negative feedback is high, full output being obtained for an input of 55mV. The sensitivity is reduced to 300mV by the application of negative feedback as indicated in fig. 4-5. Before feedback is applied, the hum and noise level with the volume control at maximum is 55db below full output.

COMPONENT VALUES FOR FIG. 4-5
KT55 25W DC/AC AMPLIFIER

G.E.C. VALVES

V1 Z729
V2 L63/6J5
V3 KT55
V4 KT55
V5 Barretter 303

R22 Thermistor CZ1 or TH1
R23 63Ω 5% 5W
R24 470-1500Ω 0·5W
R25 470-1500Ω 0·5W

RESISTORS

(20%, 0·25W unless otherwise shown)

R1 1MΩ Log.
*R2 $300\sqrt{\text{speech coil impedance}}$
R3 2·2kΩ
R4 22Ω
R5 220kΩ 10% 0·5W
R6 1MΩ 10%
R7 10kΩ
R8 1MΩ
R9 22kΩ 0·5W } matched to 5%
R10 22kΩ 0·5W }
R11 1·5kΩ
R12 220kΩ
R13 220kΩ
R14 10kΩ
R15 10kΩ
R16 185Ω 5% 5W w.w.
R17 185Ω 5% 5W w.w.
R18 10kΩ 10% 1W
R19 47Ω
R20 47Ω
R21 15+15Ω 10% 10W w.w.

**For 14 db feedback*

CAPACITORS

C1 0·01μF
C2 50μF 12V
C3 0·1μF
C4 50 pF
C5 0·01μF
C6 0·1μF
C7 0·1μF
C8 16μF 350V
C9 50μF 25V
C10 50μF 25V
C11 200μF 275V
C12 1000pF
C13 1000pF
C14 100μF 275V

MISCELLANEOUS

L1 1·5-2H 300mA 75Ω
T1 25W Ultra-linear transformer
2kΩ anode-anode
Primary Inductance : ≮ 10H
Leakage inductances :
Prim.—sec. : ≯ 20mH
½ prim.—UL tap : ≯10mH
MR1 250V 275mA (G.E.C. 13H16XG)

Separate bias resistors are essential in the output stage in view of the high mutual conductance. The 50μF cathode bypass capacitors give a loss of 6db at 50 c/s. They may be increased to 250μF when this loss is objectionable.

The output transformer is tapped at 40% of the turns on each half-primary from the centre tap. The small capacitors, C12, C13 and resistors R24, R25 are required with some output transformers to prevent the possibility of spurious oscillation.

Negative feedback is applied over three stages, about 14db giving a satisfactory reduction in distortion and output impedance without introducing the danger of instability. It is probable that more could be applied with high-quality output transformers but with this value of feedback an output of 25W at 0·25% distortion is obtained. R7 and C4, in the grid circuit of the phase splitter, assist in ensuring stability by reducing the loop gain at ultrasonic frequencies—they result in a loss of 6db at 20kc/s. (See Appendix B, page 119).

A low impedance power supply is obtained by the use of components primarily designed for television receivers. L1, which should have an inductance of about 2H and a d.c. resistance of about 75Ω, is used with two large-value electrolytic capacitors C11 and C14. On d.c. the mains adjusting resistor is not in circuit and the metal rectifier protects the capacitors against reversed polarity. The heater current of 0·3A is controlled by a barretter and thermistor.

The circuit of fig. 4-5 may be changed to pentode operation by simply connecting R19 and R20 to the output transformer centre tap. This will, of course, result in somewhat higher distortion. Compared with pentode operation the ultra-linear circuit reduces the output impedance from 9kΩ to 2·35kΩ, giving unity ratio with the anode-to-anode load. The distortion is 1·5% compared with the 2% of pentode operation. The advantages of ultra-linear operation are gained at the expense of a slight fall in the overall sensitivity of the amplifier, that is, it will require a slightly larger input signal to give the same output.

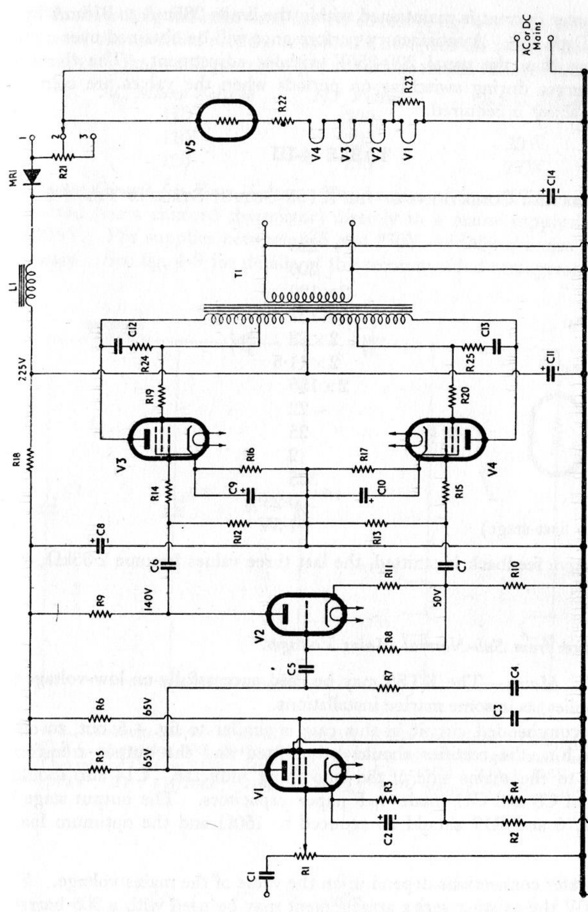


Fig. 4-5. Circuit of the 25W d.c./a.c. amplifier. R7 and C4 increase the margin of stability at high frequencies

The heater current is maintained within the limits 285mA to 315mA by the barretter Type 303. A satisfactory performance will be obtained over a rather wider range than the usual 200-250V without adjustment. The thermistor prevents surges during switching-on periods when the valves are cold. No thermistor shunt is required.

TABLE 4-III

OPERATING CONDITIONS OF THE KT55 OUTPUT STAGE OF FIG. 4-5

$V_{a\ (b)}$	225	V
$V_{a,\ g2}$	200	V
$I_{a+g2\ (o)}$	2×120	mA
$I_{a+g2\ (max\ sig)}$	2×127	mA
$p_{a+g2\ (o)}$	2×23	W
$p_{a+g2\ (max\ sig)}$	2×11·5	W
R_k	2×185	Ω
V_g	−22	V
P_{out}	25	W
$R_{L\ (a-a)}$	2	kΩ
Z_{out}	325	Ω
D	0·25%	%
$V_{in\ (rms)}$ (to first stage)	1·5V	mV

If negative feedback is omitted, the last three values become 2·35kΩ, 1·5% and 300mV.

Operation from Sub-Normal Mains Voltages.

Low d.c. Mains.—The KT55 may be used successfully on low-voltage d.c. mains supplies, as in some marine installations.

The recommended circuit in this case is similar to fig. 4-5 but, to obtain maximum h.t., the rectifier should be omitted and the output transformer connected to the mains side of the smoothing inductor. C14 also should be omitted and C8 and C11 made 4μF paper capacitors. The output stage bias resistors R16 and R17 should be reduced to 150Ω and the optimum load to 1·5kΩ.

The heater connections depend upon the value of the mains voltage. From 150 to 200V the existing series arrangement may be used with a 305 barretter, which is replaced by a suitable resistor for 125-150V. Below 125V two chains will be required, the KT55 valves being connected in series in one chain with a small resistor, if necessary, and a 303, 304 or 305 barretter in series with the remaining valves in the second chain. A suitable tapped resistor may be used, if preferred, instead of the barretter.

Low a.c. Mains.—For low a.c. mains voltages an auto-transformer may be used for the h.t. supply, with the heaters connected as for d.c. mains. An alternative method of obtaining the h.t. supply from low a.c. mains is shown in fig. 4-8. A voltage doubler circuit is used to give an h.t. of 220V from 110V a.c.

The smoothing inductor used for the circuit of fig. 4-8 has a lower resistance (20Ω) and the specified rectifier is specially designed for voltage doubling.

The h.t. voltage and power output at various mains voltages is as follows :

AC Mains Voltage	*HT Voltage*	*Output Power*
110	200	19W
110	220	23W
120	235	27W

Heater Circuit for Four Valves.—The heaters of four KT55 valves may be connected (via a shunted thermistor) directly to a mains supply between 200 and 215V. For supplies between 215 and 250V, additional series resistance is necessary. See fig. 4-9 for details of the recommended arrangement.

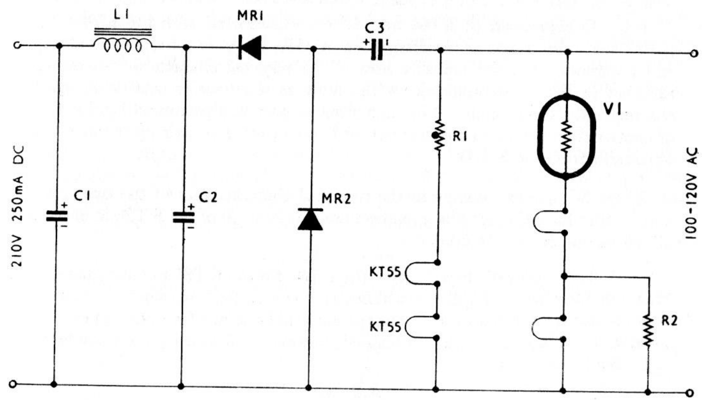


Fig. 4-8. Voltage doubler circuit for the KT55 amplifier. Component values : V1 : G.E.C. Barretter 303, 304 or 305 (depending upon voltage across heater chain) ; R1 : Thermistor CZ1 or TH1 ; R2 : 60Ω 5W ; C1 and C3 : 200μF 275V ; C2 : 100μF 275V ; L1 : 1H 20Ω ; MR1 and MR2 : 250V 300mA (G.E.C. 13D8XG).

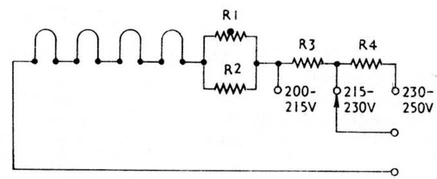


Fig. 4-9. Heater circuit for four KT55 valves. Component values : R1 : Thermistor CZ1 or TH1 ; R2 : 750Ω 10% 2W ; R3 and R4 : 50Ω 5% 10W.

CHAPTER 5

AMPLIFIERS OF 50 TO 100 WATTS

Although the amplifiers described earlier in this book, giving outputs up to about 50W, are adequate for domestic purposes and small public address equipment, the demand often arises for a 50W amplifier with a conservative rating and with the possibility of increasing its output to 100W by modifying the circuit and operating conditions. For such amplifiers the KT88 beam pentode is particularly suitable.

This chapter discusses the use of the KT88 in a design for a 50W ultra-linear amplifier with cathode bias and in a 100W fixed bias ultra-linear amplifier.

Details are also given of a relatively simple and inexpensive a.c. amplifier, with KT55 valves in the output stage, which gives 50W output at 5% distortion.

5-1. Comparison of KT66 and KT88.—Compared with the KT66, the KT88 has an increased anode dissipation of 35W, a higher mutual conductance and a cathode of larger emissive area. The physical characteristics are also different in that it is mounted on the more modern wafer octal base which eliminates the glass pinch. The higher anode potential permitted by this type of construction results in a power output from a push-pull pair up to twice that obtainable from the KT66.

Table 5-I overleaf compares the principal characteristics of the two valves and, as they have the same base connections, initial trials of the KT88 in existing KT66 equipment are facilitated.

It is usually possible to substitute the KT88 for the KT66 in most push-pull triode or ultra-linear amplifiers without circuit modification, since the cathode bias resistor is required to be about the same (500-600Ω) for both valves. No great increase in output will be obtained, however, unless the circuit conditions are suitably modified.

TABLE 5-I

COMPARISON OF KT66 AND KT88

	KT66	*KT88*	
V_h	6·3	6·3	V
I_h	1·27	1·8	A
$V_{a\ (max)}$	500	600	V
$V_{g2\ (max)}$	400	600	V
$p_{a\ (max)}$	25	35	W
$p_{g2\ (max)}$	3·5	6	W
g_m	6·3	11	mA/V
*P_{out} (cathode bias, ultra-linear)	32	50	W
*P_{out} (fixed bias, ultra-linear)	50	100	W
P_{out} (cathode bias, triode-connected)	14	27	W
Overall length	135	120	mm
Seated length	120	105	mm
Diameter	52	52	mm

*Class AB1 Push-Pull.

5-2. KT88 50W Ultra-Linear Amplifier.—The circuit of a complete amplifier with a KT88 ultra-linear output stage giving 50W output at 0·2% distortion is shown in fig. 5-2*. The design follows the practice for other amplifiers in this book but also incorporates the capacitors C8 and C9 shunted by R12 and R13 to ensure stability at frequencies below the cut-off frequency of the output transformer. The desirability of incorporating these components is discussed more fully in Appendix B (page 119). Also shown in fig. 5-2 are networks for reducing the amount of overshoot and consequent "ringing" in the output transformer. These are C6, R10 and C7, R11 between the first two stages. The method of introducing these networks is a variation of the recommendations given in Appendix B for eliminating high-frequency instability in feedback amplifier circuits.

In other respects, the amplifier follows conventional practice. The first double triode, V1, is arranged as a self-balancing floating paraphase phase inverter which feeds V2, the following push-pull voltage amplifier. Potentiometer R22 allows the signal input to the output stage to be adjusted for output stage dynamic balance, as explained on page 8.

The power supply incorporates a thermistor in the h.t. output line in order to reduce the surge from the directly heated rectifier while the remaining valves are warming up. The performance of this amplifier is illustrated in fig. 5-3 and the output stage characteristics are given in Table 5-II below :

TABLE 5-II

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 5-2

$V_{a\ (b)}$	500	V
$V_{a,\ g2}$	425	V
$I_{a+g2\ (o)}$	2×87	mA
$I_{a+g2\ (max\ sig)}$	2×100	mA
$p_{a+g2\ (o)}$	2×40	W
$p_{a+g2\ (max\ sig)}$	2×18	W
R_k	2×525	Ω
V_g (approx)	−50	V
P_{out}	50	W
$R_{L\ (a-a)}$	5	kΩ
D	0·2%	%
$V_{in\ (rms)}$ (to first stage)	500	mV

COMPONENT VALUES FOR FIG. 5-2

KT88 50W ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 B339/12AX7
V2 B329/12AU7
V3 KT88
V4 KT88
V5 U52/5U4

RESISTORS

(20%, 0·5W *unless otherwise shown*)
R1 1MΩ
R2 3·3kΩ
R3 100Ω
R4 3·3kΩ
R5 1MΩ
R6 220kΩ 10%
R7 220kΩ 10%
R8 1MΩ
R9 1MΩ
R10 10kΩ

R11 10kΩ
R12 4·7MΩ
R13 4·7MΩ
R14 100kΩ
R15 470kΩ
R16 470kΩ
R17 10kΩ
R18 680Ω
R19 10kΩ
R20 4·7kΩ
R21 33kΩ 2W
*R22 25kΩ 4W
R23 33kΩ 2W
R24 220kΩ
R25 220kΩ
R26 10kΩ
R27 10kΩ
R28 525Ω 5% 6W
R29 525Ω 5% 6W
R30 100Ω
R31 100Ω
R32 100kΩ 1W
R33 100kΩ 1W
R34 Thermistor CZ6
†R35 6·8kΩ
†R36 6·8kΩ

Optional potentiometer for dynamically balancing the output stage. If omitted, R21 and R23 should be 47kΩ, 2%, 2W.

CAPACITORS

C1 8μF 500V
C2 50μF 12V
C3 50μF 12V
C4 0·25μF
C5 0·25μF
C6 470pF
C7 470pF
C8 0·005μF
C9 0·005μF
C10 0·5μF
C11 0·5μF
C12 50μF 100V
C13 50μF 100V
C14 24μF 350V
C15 24μF 350V
C16 4μF 750V
†C17 8μF 500V

MISCELLANEOUS

L1 5H 250mA
T1 50W Ultra-linear transformer
5kΩ anode-anode
Primary inductance : ≮30H
Leakage inductances :
Prim.—sec. : ≯ 10mH
½ prim.—UL tap : ≯ 10mH
T2 Mains transformer
Secondaries:
500-0-500V 250mA
6·3V 5A CT
6·3V 1-2A CT (for pre-amp.)
5V 3A
F1 1A delayed fuse

†R35, R36 and C17, shown dotted in fig. 5-2, provide the decoupled h.t. feed for the pre-amplifier of fig. 9-3 which was designed to work with this amplifier.

5-3 KT55 50W Fixed Bias Ultra-Linear Amplifier.—Although the KT55 was designed primarily for d.c./a.c. amplifiers, its characteristics lend themselves well to the design of a 50W a.c. mains amplifier. This circuit, illustrated in fig. 5-5, provides an economical method of obtaining 50W output. The use of fixed bias in the ultra-linear arrangement enables a pair of KT55 valves to give 50W output at about 5% distortion with an h.t. potential of only 300V.

There is a large change in anode current between the quiescent and full output conditions and a power supply of very low impedance is necessary. The circuit diagram includes details of a single-phase half-wave rectifier power supply having an impedance of 100Ω over the working range. The grid bias voltage is rectified by a small rectifier from the 52V heater supply for the KT55 valves.

The h.t. supply is derived from an auto-transformer which provides 255V r.m.s. for the metal rectifier MR1. Two other windings of 52V and 6·3V supply the heaters of the KT55 valves and earlier valves, respectively, and a small metal rectifier MR2 provides the grid bias potential for the output stage. The range of negative bias required is from 50V to 60V.

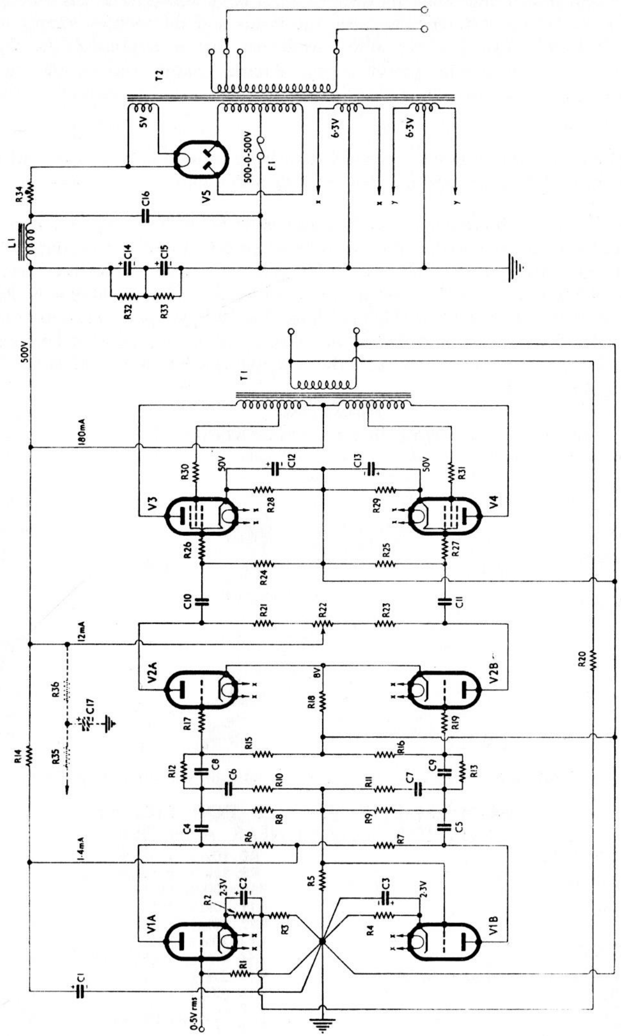


Fig. 5-2. Circuit of the KT88 50W amplifier (the " 88-50 "). The components shown in broken line provide a decoupled h.t. feed for the pre-amplifier of fig. 9-3

The auto-transformer does not need to be of large size since it has only to cater for the voltage difference between the mains and the rectifier supply of 255V. It is this feature of the auto-transformer that is responsible for the excellent regulation of the power supply circuit. Satisfactory results are obtained from a transformer similar in size to those used in domestic radio receivers.

The capacitance-input filter contains two large capacitors of 160μF and a low-resistance 1H smoothing inductor and provides a hum-free d.c. supply.

The correct quiescent current of 70-75mA is set for each valve on the meter in the cathode circuit by adjusting potentiometers R8 and R9. Resistors R17 and R18 are meter shunts, the value of which will depend upon the meter characteristics and the type of meter circuit adopted. The potentiometer R5 in the anode circuit of the first stage is adjusted at 90% of maximum output to give equal cathode currents in the output stage for obtaining dynamic balance. Resistor R19, in the bias line, reduces the potential at C8 to 60V from the 70V it would otherwise be.

TABLE 5-III

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 5-5

$V_{a,\ g2\ (o)}$	325	V
$V_{a,\ g2\ (max\ sig)}$	300	V
$I_{a+g2\ (o)}$	2×75	mA
$I_{a+g2\ (max\ sig)}$	2×165	mA
$p_{a+g2\ (o)}$	2×25	W
$p_{a+g2\ (max\ sig)}$	2×12·5	W
V_{g1}	−50 to −55	V
P_{out}	50	W
$R_{L\ (a-a)}$	2·25	kΩ
z_{out}	2·7	kΩ
*D	5	%
$V_{in\ (rms)}$ (to first stage)	2×2·5	V

*Assuming the cathode currents are balanced at 90% of maximum output by R5

COMPONENT VALUES FOR FIG. 5-5
KT55 50W ULTRA-LINEAR AMPLIFIER

G.E.C. VALVES

V1 B65/6SN7
V2 KT55
V3 KT55

RESISTORS

(20%, 0·25W unless otherwise shown)

R1 1MΩ
R2 1MΩ
R3 1kΩ
R4 33kΩ 1W
R5 20kΩ w.w. preset
R6 33kΩ 1W
R7 100kΩ 10%
R8 10kΩ w.w. preset
R9 10kΩ w.w. preset
R10 100kΩ 10%
R11 10kΩ 10% 1W
R12 10kΩ
R13 10kΩ
R14 4·7kΩ 10% 1W
R15 270Ω
R16 270Ω
R17 } Meter shunts
R18 }
R19 150Ω
R20 470-1500Ω 0·5W
R21 470-1500Ω 0·5W

CAPACITORS

C1 16μF 350V
C2 0·1μF
C3 0·1μF
C4 160μF 450V
C5 1000pF
C6 1000pF
C7 160μF 450V
C8 32μF 100V

MISCELLANEOUS

L1 1H 350mA 20Ω
T1 50W Ultra-linear transformer
2·25kΩ anode-anode
Primary inductance : ≯ 12H
Leakage inductances :
Prim.—sec. : ≯ 10mH
½ prim.—UL tap : ≯ 10mH
T2 Mains auto-transformer (see text)
MR1 275V 350mA (G.E.C. Z13H17XG)
MR2 60V 10mA (G.E.C. Z11H4X)
S1 1-pole 3-way

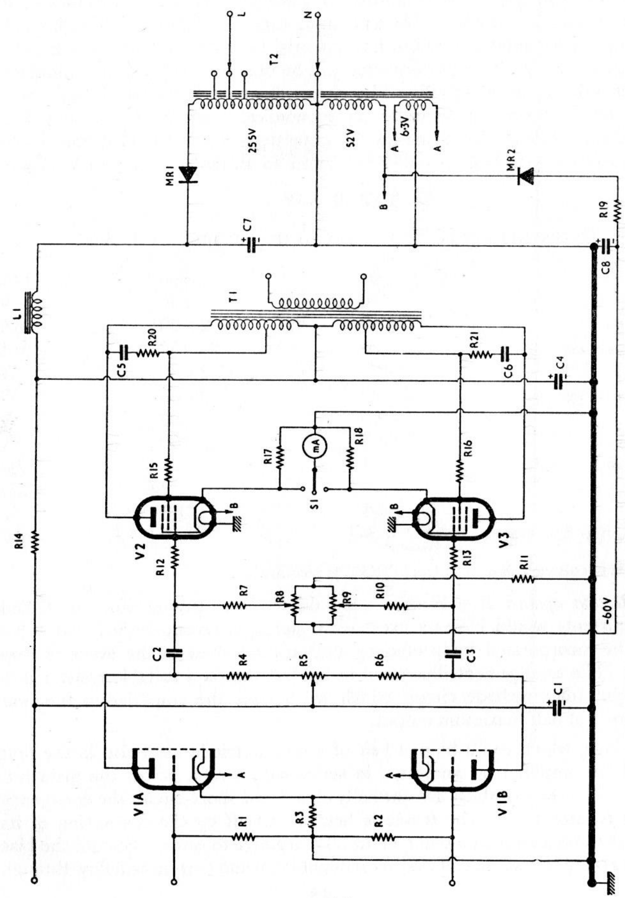


Fig. 5-5. Circuit of the KT55 50W amplifier. The earth line or chassis is connected to one side of the mains supply. If this amplifier is preceded by a triode phase-splitting stage such as that in fig. 5-8, the required signal input will be 2·5V.

5-4. KT88 100W Fixed Bias Ultra-linear Amplifier.—The circuit of this amplifier is given in fig. 5-8 and it follows, in general, the main details of the KT55 50W amplifier described in the previous section. Points of difference include the addition of a phase-splitter at the input and, of course, the h.t. power and grid bias supplies.

This amplifier provides 100W output at 5% distortion with an h.t. potential of 560V

With fixed bias, the large change in anode current necessitates a low impedance power supply and, with normal rectifier circuits, an inductance-input smoothing filter is essential. The smoothing capacitor should be of high value to prevent an instantaneous fall in h.t. potential upon the occurrence of a transient signal. Satisfactory performance will be obtained with a single inductor and a capacitance of 50-150μF. The circuit diagram shows two 160μF 450V electrolytic capacitors in series as an economical method of obtaining the required capacitance. Reducing the h.t. potential to 460V, the load impedance to 4kΩ and the grid bias to -65V results in an amplifier giving 65W output.

TABLE 5-IV

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 5-8

$V_{a\ (b)}$	460	560	V
$V_{a,\ g2}$	450	550	V
$I_{a+g2\ (o)}$	2×50	2×50	mA
$I_{a+g2\ (max\ sig)}$	2×120	2×150	mA
$p_{a+g2\ (o)}$	2×25	2×30	W
$p_{a+g2\ (max\ sig)}$	2×20	2×33	W
*V_{g1}	−65	−80	V
P_{out}	65	100	W
$R_{L\ (a-a)}$	4	4·5	kΩ
Z_{out}	6·5	6·5	kΩ
†D	3-6	3-6	%
$V_{in\ (rms)}$ (to first stage)	1 to 1·5	1 to 1·5	mV

*A bias voltage range of at least ±25% is recommended.

Protection against Bias Failure.—Should the bias supply fail, the KT88 anode currents would increase excessively and it is recommended that some device be incorporated for protecting the output valves in the event of bias failure. The arrangement illustrated in fig. 5-9 inserts a suitable resistor into the output stage cathode circuit which will enable the amplifier to function temporarily at half maximum output.

A triode, which could be one half of a double triode used also in the first stage of the amplifier, is connected in series with a relay across the main h.t. supply. The relay contacts are normally closed and short-circuit the emergency supply. The relay contacts are normally closed and short-circuit the emergency cathode resistor R1. The triode is held at cut-off by the connection of its grid to the bias supply at a point about 50V negative to earth. Should the bias fail, the grid of the triode will rise to earth potential and current will flow through

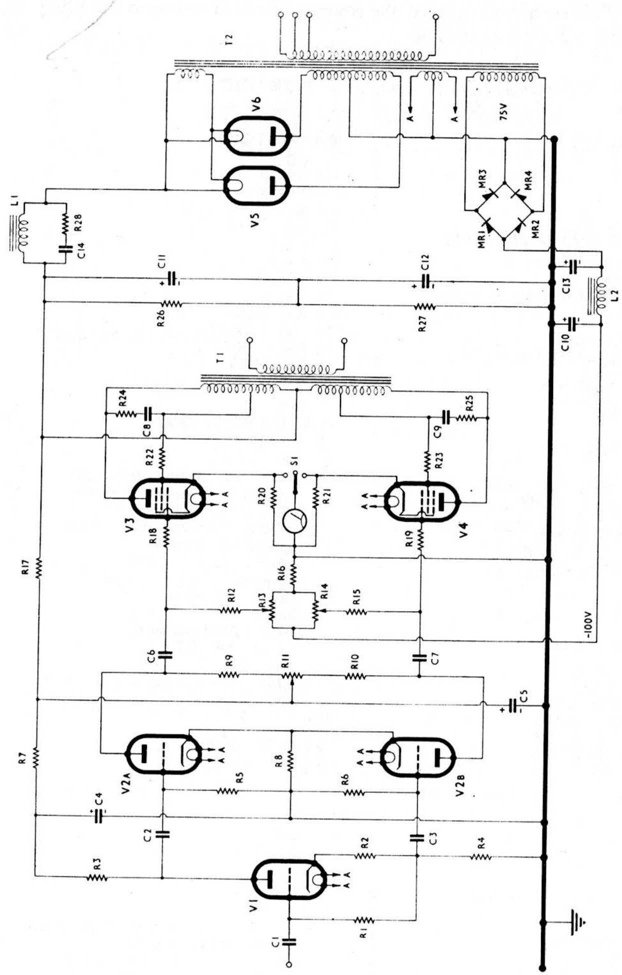


Fig. 5-8. Circuit of the KT88 fixed bias amplifier. Table 5-IV gives the operating conditions for obtaining either 65W or 100W maximum output from this circuit. R26 and R27 equalise the voltages across C11 and C12, the series-connected smoothing capacitors. C14 and R28 prevent the build-up of high voltage transients across L1.

the triode. This energises the relay, the contacts of which will open and bring into circuit the cathode bias resistor.

COMPONENT VALUES FOR FIG. 5-8

KT88 100W FIXED BIAS AMPLIFIER

G.E.C. VALVES

V1 L63/6J5
V2 B65/6SN7
V3 KT88
V4 KT88
V5, V6 } U19 (or GXU50 with delay)

RESISTORS

(20%, *0·25W unless otherwise shown)*

R1 1MΩ
R2 1·5kΩ
R3 33kΩ 1W } matched to 5%
R4 33kΩ 1W } matched to 5%
R5 470kΩ 10%
R6 470kΩ 10%
R7 33kΩ 1W
R8 1kΩ
R9 33kΩ 10% 1W
R10 10kΩ
R11 33kΩ 10% 1W
R12 100kΩ 10% 0·5W
R13 20kΩ w.w. preset
R14 20kΩ w.w. preset
R15 100kΩ 10% 0·5W
R16 10kΩ 10% 1W
R17 4·7kΩ 1W
R18 5·6kΩ
R19 5·6kΩ
R20, R21 } Meter shunts
R22 270Ω 0·5W
R23 270Ω 0·5W
R24 470-1500Ω 0·5W
R25 470-1500Ω 0·5W
R26 100kΩ 10% 1W
R27 100kΩ 10% 1W
R28 10kΩ 0·5W

CAPACITORS

C1 0·01μF
C2 0·05μF
C3 0·05μF
C4 8μF 350V
C5 8μF 450V
C6 0·1μF
C7 0·1μF
C8 1000pF
C9 1000pF
C10 8μF 250V
C11 160μF 450V } each 100+60 dual
C12 160μF 450V } each 100+60 dual
C13 8μF 250V
C14 0·01μF 750V

MISCELLANEOUS

L1 5H 325mA
L2 20H 10mA
T1 100W Ultra-linear transformer
4·5kΩ anode-anode (100W)
4kΩ anode-anode (65W)
Primary inductance : ≮ 40H
Leakage inductances :
Prim.—sec. : ≯ 6mH
½ prim.—UL tap : ≯ 6mH
T2 Mains transformer
Secondaries:
700-0-700V 325mA
6·3V 5A CT
5V 7A
75V 10mA (bias)
S1 1-pole 3-way
MR1, MR2, MR3, MR4 } 75V 10mA

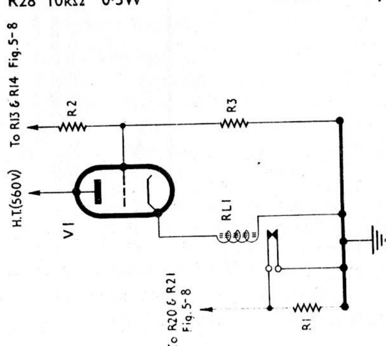


Fig. 5-9. Bias failure protection for the KT88 fixed bias amplifier. R2 is connected to the positive (earthy) junction of R13 and R14 in fig. 5-8. R2 and R3 take the place of R16 in the amplifier. The earth connection to the meter in fig. 5-8 is broken and the meter (and R20, R21) taken to R1. Component values : V1 : G.E.C. L63/6J5GT ; R1 : 330Ω 10% 5W ; R2 : 6·8kΩ 10% 0·5W ; R3 : 15kΩ 10% 0·5W ; RL1 : 20kΩ.

CHAPTER 6

CLASS B AMPLIFIERS OF 175 TO 200 WATTS

The Class B amplifier is recognised as an economical method of obtaining a high audio output because of its high efficiency and the simplicity of the circuit arrangement. The absence of a separate grid bias supply renders the valve self-protective and eliminates the need for an interlock for delaying the application of h.t. until bias is established. It is necessary to qualify this statement in view of the small degree of bias used in the output stage of the amplifier of fig. 6-5. In this case, however, the bias serves only to limit the quiescent dissipation and, unlike normal fixed bias arrangements, no harm results from running the output stage for short periods of a few minutes without this bias being fully established.

Much of the distortion associated with the Class B amplifier in the past was due to the leakage inductance of the driver stage coupling transformer but this has now been largely overcome by using the cathode-coupled bridged transformer. This arrangement was discussed in more detail in Section 1-3.

This chapter describes 175W and 200W amplifier circuits containing the DA42, an indirectly heated triode designed specifically for Class B operation.

6-1. DA42 175W Amplifier.—The complete amplifier circuit is shown in fig. 6-1 and comprises—apart from the output stage—a KT66 driver stage, which supplies the necessary 75+75V for full output, and a push-pull intermediate stage consisting of two N709 pentodes. The resistance-coupled N709 valves are required to deliver 80+80V, and this output is obtained at low distortion because of the high h.t. voltage available, provided the screen and cathode resistors are of optimum value. The input stage may utilise any small triode such as the DH77 shown, which draws an anode current of 2mA.

The combined anode and screen current of each N709 is 6mA, the anode and screen potentials being 150 and 60, respectively. An input not exceeding 1+1V to this stage gives an output of 80+80V at less than 2% distortion.

With 500V h.t. supply for the KT66 valves a bias of 45V is required, giving an anode current of 70-80mA for the pair. A common cathode resistance of 600Ω is necessary to give this bias and this is partly supplied by the resistance of the half-primaries of the coupling transformer. An additional resistor (R19), common to both valves, is usually necessary. In the transformer used for the prototype, the resistance of each half-primary was about 300Ω and R19 was found to require a value of 440Ω.

The coupling transformer bridging capacitors are not critical in value within the limits of 2μF to 16μF.

The output transformer should have low leakage inductance between the two half-primaries and between the primary and secondary. For a 100V output line (60Ω in this case), the primary/secondary ratio is 13:1.

The amplifier of fig. 6-1 gives 175W output at 6% distortion for an input of less than 1V r.m.s. and the performance of the output stage is displayed by the curves of fig. 6-3. Table 6-I gives the operating conditions.

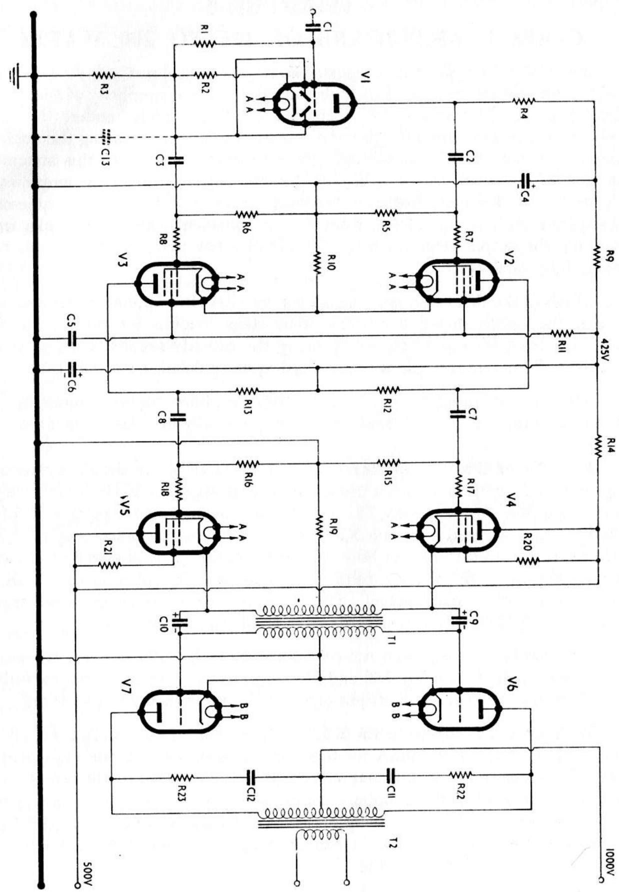


Fig. 6-1. Circuit of the DA42 175W Class B amplifier. The slight unbalance in signal in the output of V1, caused by phase shift, may be eliminated by shunting R3 with a capacitor C13, shown dotted.

COMPONENT VALUES FOR FIG. 6-1

DA42 175W AMPLIFIER

G.E.C. VALVES

V1 DH77/6AT6
V2 N709
V3 N709
V4 KT66
V5 KT66
V6 DA42
V7 DA42

RESISTORS

(20%, 0·25W *unless otherwise shown*)
R1 470kΩ
R2 1·5kΩ
R3 22kΩ 0·5W } matched to 5%
R4 22kΩ 0·5W } matched to 5%
R5 470kΩ 10%
R6 470kΩ 10%
R7 10kΩ
R8 10kΩ
R9 47kΩ 0·5W
R10 220Ω 0·5W
R11 470kΩ 10% 0·5W
R12 47kΩ 5% 5W w.w.
R13 47kΩ 5% 5W w.w.
R14 4·7kΩ 10% 1W
R15 470kΩ
R16 470kΩ
R17 10kΩ
R18 10kΩ
R19 (See text above)
R20 100Ω
R21 100Ω
R22 5kΩ 10% 10W
R23 5kΩ 10% 10W

CAPACITORS

C1 0·01μF
C2 0·01μF
C3 0·01μF
C4 4μF 450V
C5 0·5μF
C6 16μF 450V
C7 0·05μF
C8 0·05μF
C9 16μF 100V
C10 16μF 100V
C11 2000pF 1kV
C12 2000pF 1kV
C13 150 pF

TRANSFORMERS

T1 Bridged transformer (see fig. 6-7)
T2 Output transformer (see fig. 6-8)

TABLE 6-I

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. 6-1

$V_{a\ (b)}$	1 (approx)	kV
V_a	1	kV
$I_{a\ (o)}$	2×25	mA
$I_{a\ (max\ sig)}$	2×140	mA
$p_{a\ (o)}$	2×25	W
$p_{a\ (max\ sig)}$	2×50	W
V_g	0	V
$I_{g\ (max\ sig)}$	2×25	mA
P_{out}	175	W
$R_{L\ (a-a)}$	10	kΩ
z_{out}	15	kΩ
D	6	%
p_{dr}	5	W
$v_{in\ (g-g)\ (pk)}$	175	V
$V_{in\ (rms)}$ (to first stage)	≯1	V

These conditions should be used only for normal speech and music. A continuous tone at maximum output will exceed the permitted dissipation and for such applications $R_{L\ (a-a)}$ must be increased by at least 20%.

The addition of negative feedback, as described below, results in the following conditions :

P_{out}	175	W
Z_{out}	3 to 4	kΩ
D	≯2	%
$V_{in\ (rms)}$ (to first stage)	≯5	V

Negative Feedback.—An improved performance is obtained at the expense of a reduction in sensitivity by adding negative feedback. The circuit modifications are given in fig. 6-2 and this arrangement applies 12-14 db feedback, which provides a damping factor of about 4·5 and reduces the distortion to 2% at 175W.

The coupling transformer T1 introduces negligible phase shift so that feedback may be taken from the output transformer secondary to the N709 cathode circuit. The single cathode bias resistor is replaced by R1 and R2, and resistors R3 and R4 provide the feedback paths from each end of the transformer secondary.

The values of R3 and R4 are determined by the load impedance but, with a 100V line (60Ω), they are each 3·3kΩ.

Power Supply.—The power supply circuit of fig. 6-4 is unconventional in that one set of rectifiers and one transformer provide two h.t. voltages, one of which is twice the value of the other. Compared with conventional circuits, this arrangement, which is described more fully in Section 1-9, would appear to offer a saving in both initial cost and space.

The rectifiers are not equally loaded, two of them having to supply the low voltage current of 100mA in addition to the 275mA at 1000V required by the DA42 stage at maximum output. However, this is well within the capability of the four U19 rectifiers specified since their rated maximum output in this type of circuit is 500mA at 4kV. The GXU50 xenon rectifier may be used as an alternative to the U19 ; a 10-second switching delay being then required.

As the power supply impedance causes some interaction between the low and high voltage supplies, a sudden demand for maximum current would cause a reaction in the 500V supply if a conventional 4-8μF smoothing capacitor is used in the high voltage supply. Connecting three 160μF, 450V electrolytic capacitors in series achieves an effective capacitance of 50-60μF, with a working voltage of 1350. Resistors R1, R2 and R3 equalise the voltage across each capacitor. The same principle is applied to the capacitors in the low voltage supply.

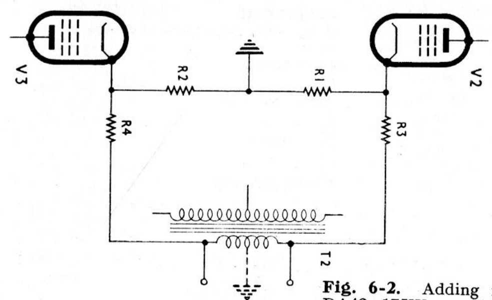


Fig. 6-2. Adding negative feedback to the DA42 175W amplifier. Component values : R1 and R2 : 330Ω 10% 0·5W ; R3 and R4 : 3·3kΩ 10% 1W (for 100V line output). Other designations are as shown on fig. 6-1.

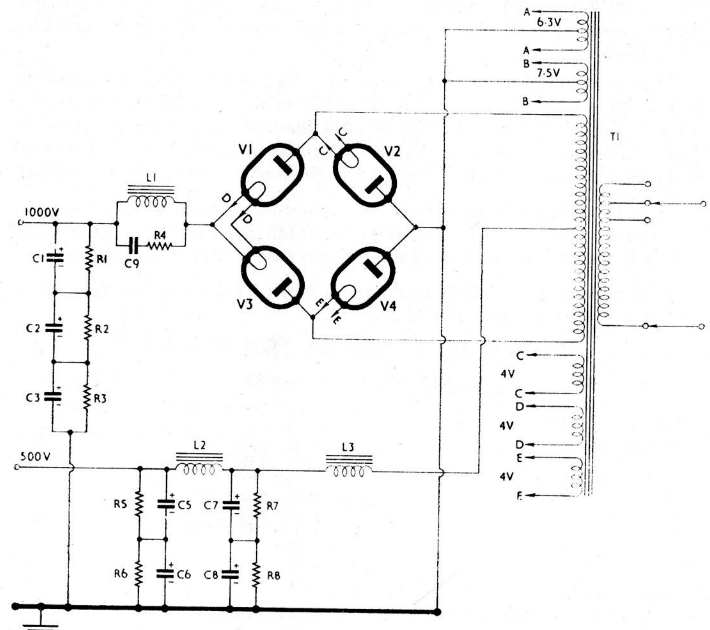


Fig. 6-4. Power supply circuit for the 175W and 200W amplifiers of figs. 6-1 and 6-5. Although three secondary windings can be used for the rectifier filaments, as shown, transformer winding is simplified if four identical windings are used. Component values are given below.

G.E.C. VALVES

V1, V2, V3, V4: U19 or GXU50

MISCELLANEOUS

L1 10H 275mA
L2 10H 100mA
L3 10H 100mA
T1 Mains transformer
Secondaries
1200V CT 275mA
6·3V 5A CT
7·5V 2·5A CT
4V 3·5A
4V 3·5A
4V 7·0A (or two 3·5A)

RESISTORS

(*10%, 1W unless otherwise shown*)
R1, R2, R3: 100kΩ
R4 10kΩ 2W
R5, R6, R7, R8: 100kΩ

CAPACITORS

C1, C2, C3, C5, C6, C7, C8: 160μF 450V (each 100 60 dual)
C9 0·01μF 1·5kV

6-2. DA42 200W Amplifier.—The amplifier of fig. 6-1 may be modified to give 200W output by increasing the output stage anode potential to 1250V and by applying about 4V negative grid bias in order to limit the quiescent dissipation.

The modifications are shown in fig. 6-5. The bias potential is obtained by inserting a resistor (R25) in the negative line of the h.t. supply for the driver stage. The value of R25 will depend upon the total h.t. current, but it will be about 47Ω in the arrangement shown. The bias is applied to the output stage grids by connecting it to the centre tap of the coupling transformer secondary.

The bias is set to 4V negative in the quiescent condition and at full output it will increase to about 5V as a result of the DA42 grid current. This has very little effect upon the performance and is in any case partially suppressed by C15.

The performance of the output stage is shown by the curves of fig. 6-6 and Table 6-II gives the operating conditions.

COMPONENT VALUES FOR FIG. 6-5

DA42 200W AMPLIFIER

G.E.C. VALVES

V1 DH77/6AT6
V2 N709
V3 N709
V4 KT66
V5 KT66
V6 DA42
V7 DA42
V8 U54

RESISTORS

(*20%, 0·25W unless otherwise shown*)
R1 470kΩ
R2 1·5kΩ
R3 22kΩ 0·5W, R4 22kΩ 0·5W: matched to 5%
R5 470kΩ 10%
R6 470kΩ 10%
R7 10kΩ
R8 10kΩ
R9 47kΩ 0·5W
R10 220Ω 0·5W
R11 470kΩ 10% 0·5W
R12 47kΩ 5% 5W
R13 47kΩ 5% 5W
R14 4·7kΩ 10% 1W
R15 470kΩ
R16 470kΩ
R17 10kΩ
R18 10kΩ
R19 (See page 66)
R20 100Ω
R21 100Ω
R22 7·5kΩ 10W
R23 7·5kΩ 10W
R24 47kΩ 5W
R25 47Ω 1W

CAPACITORS

C1 0·01μF
C2 0·01μF
C3 0·01μF
C4 4μF 450V
C5 0·5μF
C6 16μF 450V
C7 0·05μF
C8 0·05μF
C9 16μF 100V
C10 16μF 100V
C11 2000pF 1kV
C12 2000pF 1kV
C13 8μF 500V
C14 8μF 600V
C15 100μF 12V
C16 150pF

MISCELLANEOUS

L1 10H 120mA
T1 Bridged transformer (see fig. 6-7)
T2 Output transformer (see fig. 6-8)
T3 Mains transformer
Secondaries
425-0-425V 120mA
6·3V 5A CT
7·5V 2·5A CT
5V 3A

TABLE 6-II

Operating Conditions of the Output Stage of Fig. 6-5

$V_{a\,(b)}$	1·25 (approx)	kV
V_a	1·25	kV
$I_{a\,(o)}$	2×20	mA
$I_{a\,(max\ sig)}$	2×120	mA
$p_{a\,(o)}$	2×12·5	W
$p_{a\,(max\ sig)}$	2×25	W
$V_{g\,(o)}$	−4	V
$I_{g\,(max\ sig)}$	2×20	mA
P_{out}	200	W
$R_{L\,(a-a)}$	13	kΩ
Z_{out}	15	kΩ
D	6	%
p_{di}	4·5	W
$V_{in\,(g-g)\,(pk)}$	200	V
$V_{in\,(rms)}$ (to first stage)	$\ngtr 1$	V

These conditions should be used only for normal speech and music. A continuous tone at maximum output will exceed the permitted dissipation and for such applications $R_{L\,(a-a)}$ must be increased by at least 20%.

Power Supply.—The combined power supply in fig. 6-4 cannot be used because of the common negative connection. Two separate supplies are required : one inductance-input supply giving 240mA at 1250V for the output stage and one, which may be capacitance-input, of 400-500V for the earlier stages. A suggested arrangement for the latter is incorporated in fig. 6-5.

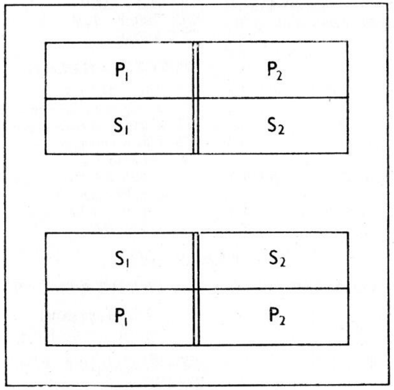


Fig. 6-7. The coupling transformer T1 in figs. 6-1 and 6-5. Winding details are as follows :

Core : Stalloy No. 4 ; square section ; no gap.
Primary : 2000+2000 turns, 28 s.w.g.
Secondary : 2000+2000 turns, 30 s.w.g.

The primary and secondary half-windings P1 and S1 are wound in the opposite direction to P2 and S2 so that, when the inner ends of each half-winding are joined, they are series-aiding. The number of turns in each of the four sections must be identical.

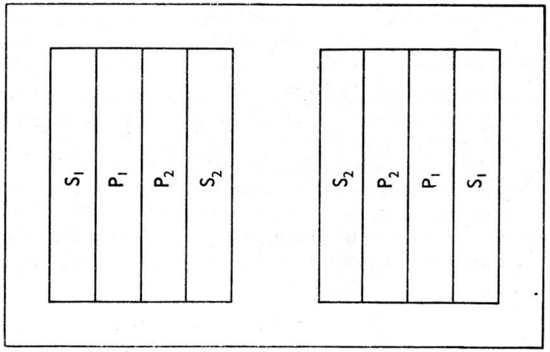


Fig. 6-8. The 100V line (60Ω) output transformer T2 in figs. 6-1 and 6-5. Winding details are as follows :

Core : Stalloy No. 66 ; square section ; no gap.
Primary : 1250+1250 turns, 28 s.w.g.
Secondary : 100+100 turns, 15 s.w.g.

All windings are wound in the same direction. The secondary is connected in series-aiding.

CHAPTER 7

CLASS AB AMPLIFIERS OF 100 TO 300 WATTS

The DA100 triode is designed for use in Class AB circuits and this chapter gives design information for Class AB1 and AB2 amplifiers which provide output powers of 175W and 270W respectively. The data includes the conditions for giving reduced outputs of 115W and 200W respectively thereby providing a choice of four amplifiers covering the range 115W to 270W.

7-1. DA100 115-175W Class AB1 Amplifier.—The circuit diagram of fig. 7-1 illustrates the output and driver stages of this amplifier and Table 7-I gives the operating data. The performance is shown by the curves of figs. 7-2, 7-3 and 7-4.

The DA100 grids are not driven positive and the preceding stage is of orthodox design and consists of a pair of triode-connected N709 or KT61 valves in a transformer-coupled push-pull Class A amplifier circuit. The output stage requires an input of 150+150V r.m.s. and this is obtained from a coupling transformer having a 1:2 ratio.

A diode is shunted across each DA100 grid circuit in order to prevent "triggering." If these diodes are absent, the output valves may be seriously damaged by excessive anode current in the event of "trigger effect" should their grids be driven positive.

Meters in the DA100 anode lines enable the grid bias to be adjusted to give the correct anode currents. These meters or other anode current measuring facilities as well as the anode circuit fuse (fig. 7-5) should always be incorporated.

A suitable power supply for the output stage is given in fig. 7-5 in which two GXU50 xenon rectifiers or U19 vacuum rectifiers are used in a bi-phase half-wave circuit having a source impedance of 400Ω. The grid bias is provided by the separate circuit also shown in fig. 7-5. The power supply for the driver and earlier stages may be of conventional design.

Table 7-II gives the data necessary for operating this amplifier at a reduced anode potential to give 115W output.

Fig. 7-1.

G.E.C. VALVES

V1 N709
V2 N709
V3 DA100
V4 DA100
V5 U709 or U78/6X4

RESISTORS

(20%, 0·25W unless otherwise shown)

R1 270kΩ
R2 270kΩ
R3 10kΩ
R4 330Ω 10% 0·5W
R5 330Ω 10% 0·5W
R6 10kΩ
R7 100Ω
R8 100Ω
R9 5kΩ
R10 5kΩ
R11 1kΩ
R12 1kΩ
R13 10kΩ 10% 20W
R14 10kΩ 10% 20W

CAPACITORS

C1 0·02μF
C2 0·02μF
C3 50μF 50V
C4 50μF 50V
C5 2000pF 1·5kV
C6 2000pF 1·5kV

TRANSFORMERS

T1 Intervalve transformer, 1 : 2
T2 Output transformer
7kΩ anode-anode (175W)
6kΩ anode-anode (115W)

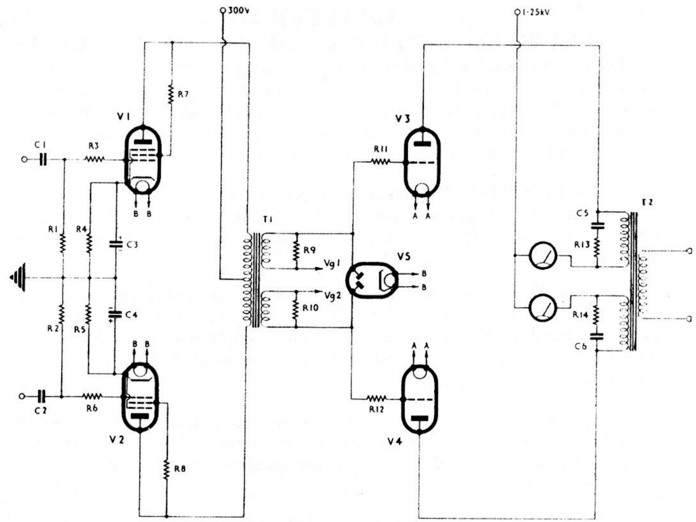


Fig. 7-1. Circuit of the driver and output stages of the DA100 Class AB1 115-175W amplifier.

TABLE 7-I

OPERATING CONDITIONS OF THE CLASS AB1 175W AMPLIFIER OF FIG. 7-1

$V_{a\ (b)}$	1·25 (approx)	kV
V_a	1·25	kV
*$I_{a\ (o)}$	2×50	mA
$I_{a\ (max\ sig)}$	2×150	mA
$p_{a\ (o)}$	2×70	W
$p_{a\ (max\ sig)}$	2×100	W
$V_{g\ (o)}$	225	V
P_{out}	175	W
$R_{L\ (a-a)}$	7	kΩ
D	3	%
$V_{in\ (pk)\ (g-g)}$	2×225	V
$V_{in\ (rms)}$ (to driver stage)	2×10	V

*Precise value unimportant and may be between 40 and 60mA.

COMPONENT VALUES FOR FIG. 7-5

POWER SUPPLY FOR DA100 AB1 AND AB2 AMPLIFIERS

G.E.C. VALVES

V1 U19 or GXU50
V2 U19 or GXU50
V3 U52/5U4G

RESISTORS

R1 10kΩ 20% 0·5W
R2 2kΩ 5W w.w.
R3 2kΩ 5W w.w.
R4 2kΩ 10% 20W w.w.

CAPACITORS

C1 4μF 1·5kV
C2 4μF 1·5kV
C3 0·01μF 1·5kV
C4 16μF 350V

MISCELLANEOUS

*L1 5H 300mA
*L2 5-25H 300mA swinging choke
L3 20H 150mA
T1 4V 7A
T2 As required
T3 Bias transformer
Secondaries
250-0-250V 100mA (1kV operation)
325-0-325V 100mA (1·25kV operation)
6V 4A CT
6·3V CT as required
5V 3A
F1 500mA
⊁50Ω for AB2

TABLE 7-II

OPERATING CONDITIONS OF THE CLASS AB1 AMPLIFIER OF FIG. 7-1 FOR REDUCED OUTPUT OF 115W

$V_{a\ (b)}$	1 (approx)	kV
V_a	1	kV
*$I_{a\ (o)}$	2×50	mA
$I_{a\ (max\ sig)}$	2×135	mA
$p_{a\ (o)}$	2×55	W
$p_{a\ (max\ sig)}$	2×90	W
$V_{g\ (o)}$	−175	V
P_{out}	115	W
$R_{L\ (a-a)}$	6	kΩ
D	3	%
$V_{in\ (pk)\ (g-g)}$	2×165	V
$V_{in\ (rms)}$ (to driver stage)	2×8	V

*Precise value unimportant and may be between 40 and 60mA.

7-2. DA100 200-270W Class AB2 Amplifier.—This amplifier, shown in fig. 7-6, requires a driver stage capable of providing the 3W driving power for the DA100 grid circuit and two KT66 in a cathode-coupled bridged transformer arrangement are recommended.

The amplifier preceding the KT66 stage is a push-pull resistance-capacitance coupled arrangement using two N709 valves as described in Chapter 1, page 14. This circuit will provide the 200+200V required at the cathodes of the KT66 driver stage.

Meters are placed in the DA100 anode lines to enable the grid bias to be adjusted for correct anode current. It is inadvisable to omit these meters or the fuse in the anode circuit (fig. 7-5). There is no danger of " trigger effect " with the Class AB2 circuit because of the low d.c. grid circuit resistance, consequently, diodes are not required in this output stage.

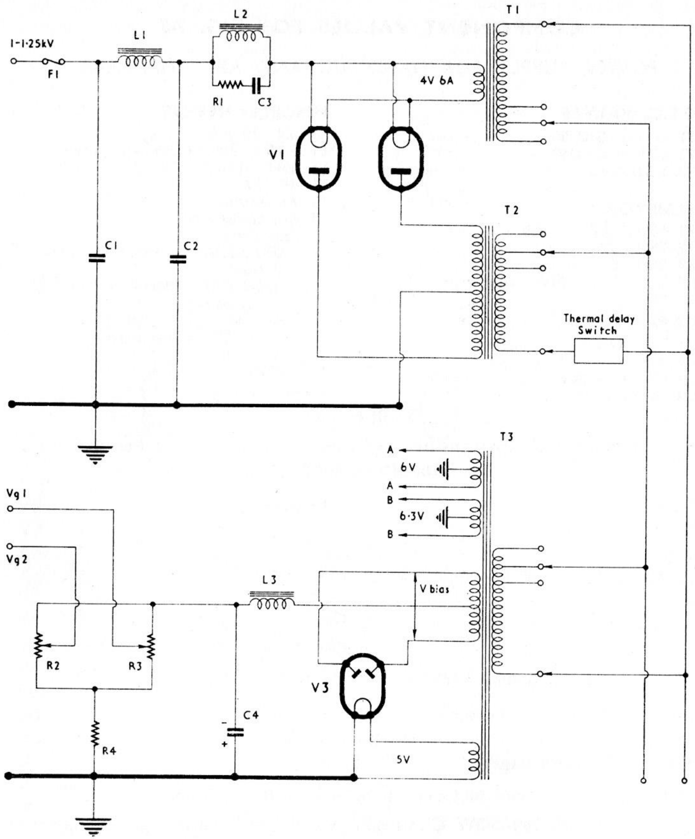


Fig. 7-5. Power supply circuit for the AB1 and AB2 amplifiers of figs. 7-1 and 7-6. The thermal delay switch at T2 is not necessary with U19 vacuum rectifiers.

The power supply of fig. 7-5 is also suitable for this amplifier and the grid bias network should provide a range of 175-250V negative to cater for differences in valve characteristics. The d.c. resistance of the inductors in the high voltage smoothing filter should not exceed 50Ω. If quality of reproduction is not important and some hum is tolerated the second inductor (L1) and capacitor (C1) may be omitted. In this event, the signal will be modulated at 100c/s ripple frequency at outputs exceeding 75% of the maximum.

The power supply for the driver and earlier stages may be of conventional design.

Table 7-III gives the output stage operating data for this amplifier and Table 7-IV the output stage characteristics for a reduced output of 200W at 1kV h.t.

TABLE 7-III

OPERATING CONDITIONS OF THE CLASS AB2 270W AMPLIFIER OF FIG. 7-6

V_a (b)	1·25 (approx)	kV
V_a	1·25	kV
I_a (o)	2×50	mA
I_a (max sig)	2×200	mA
p_a (o)	2×70	W
p_a (max sig)	2×90	W
V_g (o)	225	V
I_g (max sig)	5	mA
i_g (pk) (max sig)	20	mA
P_{out}	270	W
R_L (a−a)	7	kΩ
Z_{in}	25	kΩ
D	5	%
P_{dr}	3	W
V_{in} (pk) (g−g)	2×300 (approx)	V
V_{in} (rms) (to driver stage)	2×15	V

TABLE 7-IV

OPERATING CONDITIONS OF THE CLASS AB2 AMPLIFIER OF FIG. 7-6 FOR REDUCED OUTPUT OF 200W

V_a (b)	1 (approx)	kV
V_a	1	kV
I_a (o)	2×50	mA
I_a (max sig)	2×150	mA
p_a (o)	2×55	W
p_a (max sig)	2×85	W
V_g (o)	−175	V
I_g (max sig)	15	mA
i_g (pk) (max sig)	60	mA
P_{out}	200	W
R_L (a−a)	5	kΩ
Z_{in}	7	kΩ
D	5	%
P_{dr}	7	W
V_{in} (pk) (g−g)	2×250	V
V_{in} (rms) (to driver stage)	2×10	V

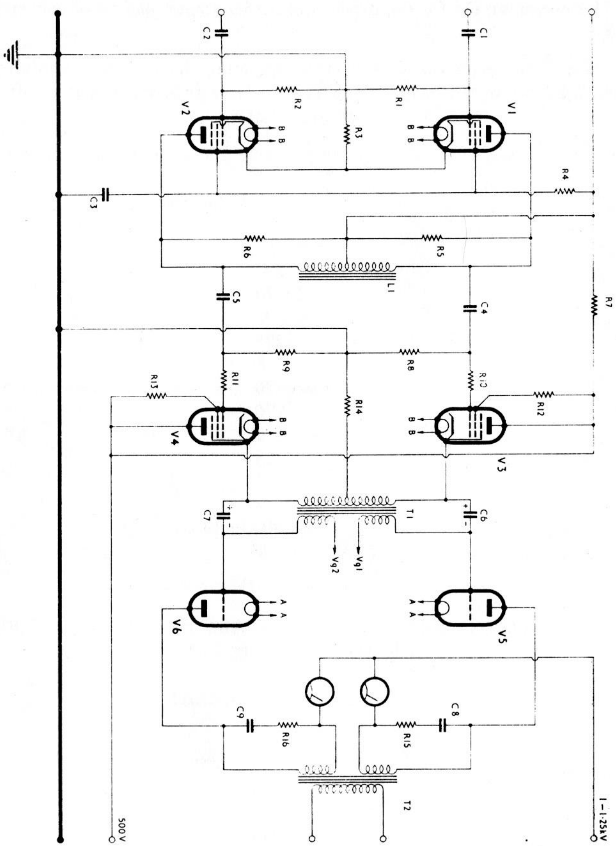


Fig. 7-6. Circuit of the DA100 Class AB2 200-270W amplifier. The cathode-coupled driver stage is preceded by two N709 beam pentodes in a push-pull voltage amplifier stage as described on page 14.

COMPONENT VALUES FOR FIG. 7-6

DA100 200-270W AB2 AMPLIFIER

G.E.C. VALVES

V1 N709
V2 N709
V3 KT66
V4 KT66
V5 DA100
V6 DA100

RESISTORS

(20%, 0·25W unless otherwise shown)

R1 220kΩ
R2 220kΩ
R3 220Ω 10% 1W
R4 220kΩ 10% 1W
R5 47kΩ 3W } matched to 5%
R6 47kΩ 3W }
R7 4·7kΩ 10% 1W
R8 470kΩ
R9 470kΩ
R10 10kΩ
R11 10kΩ
R12 100Ω
R13 100Ω
R14 (See Sect. 1-3)
R15 5kΩ 10% 50W
R16 5kΩ 10% 50W

CAPACITORS

C1 0·01μF
C2 0·01μF
C3 0·5μF
C4 0·05μF
C5 0·05μF
C6 4μF 350V
C7 4μF 350V
C8 2000pF 1·5kV
C9 2000pF 1·5kV

MISCELLANEOUS

L1 Centre-tapped inductor 100-200H
T1 Bridged transformer (See fig. 6-7)
T2 Output transformer
7kΩ anode-anode (270W)
5kΩ anode-anode (200W)

CHAPTER 8

CLASS AB AMPLIFIERS OF 300 TO 1100 WATTS

The largest amplifiers dealt with in this book are designed round the V1505, a Class AB triode of 275W dissipation, and this chapter describes the design of both Class AB1 and AB2 circuits. The AB1 amplifier will provide up to 450W output and the AB2 amplifier up to 1100W. An output exceeding 900W requires considerable driving power and the data should be closely followed in order to obtain minimum distortion.

8-1. V1505 300-450W Class AB1 Amplifier.—Fig. 8-1 is the circuit diagram of the output and driver stages of this amplifier. Little explanation is required ; the driver stage contains a pair of triode-connected KT66 valves which are coupled to the output stage by a transformer of 1:2 ratio. The spark gap across the output transformer primary is optional but, if it is used, the series resistor (R13) must not be omitted. A load impedance of 10kΩ is specified and this should not be reduced as a lower load will result in excessive anode dissipation at maximum output.

Fig. 8-5 contains the power supply circuit recommended for providing h.t. for the output stage. The bias voltage is provided by a normal full-wave rectifier circuit which is also shown in fig. 8-5. The KT66 driver stage power supply may be a conventional capacitance-input circuit since there is no variation in h.t. current. The KT66 valves require about 350V at 100mA.

Table 8-II gives the data necessary for operating this amplifier at a reduced anode potential to give 300W output and fig. 8-2 illustrates the corresponding performance. At this reduced level, a load impedance lower than the 10kΩ specified is permissible and this would provide a somewhat higher output

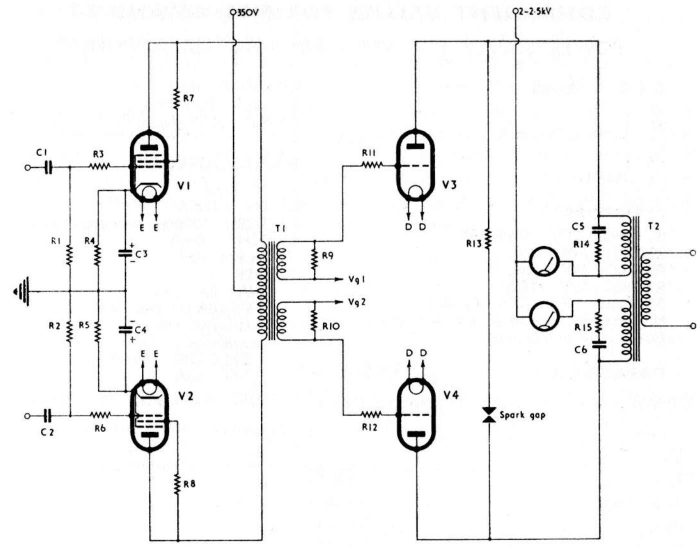


Fig. 8-1. Circuit of the output and driver stages of the V1505 Class AB1 300-450W amplifier. Component values are given below.

G.E.C. VALVES
V1 KT66
V2 KT66
V3 V1505
V4 V1505

RESISTORS
(20%, 0·25W unless otherwise shown)
R1 470kΩ
R2 470kΩ
R3 10kΩ
R4 600Ω 5% 3W
R5 600Ω 5% 3W
R6 10kΩ
R7 100Ω 0·5W
R8 100Ω 0·5W
R9 33kΩ 10% 0·5W
R10 33kΩ 10% 0·5W
R11 1kΩ
R12 1kΩ
R13 10kΩ 10% 100W (see p. 86)
R14 10kΩ 10% 50W
R15 10kΩ 10% 50W

CAPACITORS
C1 0·02μF
C2 0·02μF
C3 25μF 50V
C4 25μF 50V
C5 2000pF 1·5kV
C6 2000pF 1·5kV

TRANSFORMERS
T1 Intervalve transformer 1:2
T2 Output transformer
10kΩ anode-anode

COMPONENT VALUES FOR FIGS. 8-5 AND 8-7

POWER SUPPLY FOR V1505 AB1 AND AB2 AMPLIFIERS

G.E.C. VALVES

V1, V2, V3, V4 } U19 or GXU1
V5 U52/5U4

RESISTORS

R1 10kΩ 20% 0·5W
*R2 2kΩ 5W w.w.
*R3 2kΩ 5W w.w.
*R4 2kΩ 10% 10W
R5 700Ω 10% 25W (fig. 8-7)
In Class AB1, may be increased to 5kΩ to reduce current consumption.

CAPACITORS

C3 0·01μF 3kV
C4 16μF 350V
C5 4μF 350V (fig. 8-7)

MISCELLANEOUS

F1 HV type
L1 5H 600mA
L2 5-20H 600mA swinging choke
L3 20H 100mA
T1 As required
T2 2·5V 5A
2·5V 5A
2·5V 10A (or two 5A)
T3 Bias transformer
Secondaries :
220-0-220V 85mA
14V 13A

TABLE 8-I

OPERATING CONDITIONS OF THE CLASS AB1 450W AMPLIFIER OF FIG. 8-1

$V_{a\,(b)}$	2·5 (approx)	kV
V_a	2·5	kV
*$I_{a\,(o)}$	2×50	mA
$I_{a\,(max\ sig)}$	2×200	mA
$p_{a\,(o)}$	2×130	W
$p_{a\,(max\ sig)}$	2×275	W
$V_{g\,(o)}$	−155	V
P_{out}	450	W
$R_{L\,(a-a)}$	10	kΩ
D	3	%
$V_{in\,(pk)\,(g-g)}$	2×150	V
$V_{in\,(rms)}$ (to driver stage)	2×10	V

*See section 8-3,

TABLE 8-II

OPERATING CONDITIONS OF THE CLASS AB1 AMPLIFIER OF FIG. 8-1 FOR REDUCED OUTPUT OF 300W

$V_{a\,(b)}$	2 (approx)	kV
V_a	2	kV
*$I_{a\,(o)}$	2×65	mA
$I_{a\,(max\ sig)}$	2×180	mA
$p_{a\,(o)}$	2×135	W
$p_{a\,(max\ sig)}$	2×200	W
$V_{g\,(o)}$	−125	V
P_{out}	300	W
$R_{L\,(a-a)}$	10	kΩ
D	3	%
$V_{in\,(pk)\,(g-g)}$	2×120	V
$V_{in\,(rms)}$ (to driver stage)	2×8	V

*See section 8-3,

8-2. V1505 600-1100W Class AB2 Amplifier.—The recommended circuit for Class AB2 operation of the V1505 is shown in fig. 8-6 which also contains the KT66 driver stage with its preceding amplifier. Bridged transformer coupling is used, as described in Chapter 1, page 5, between the KT66 and V1505 stages. Up to 900W will be obtained with a driver anode supply of 475V at a total current of 130mA. The bias required for the KT66 valves for this operating point is 40V and this is provided by an effective cathode resistance of 600Ω per valve. Part of this resistance is provided in each cathode line by the half-primaries of the coupling transformer, the remainder (R13) being inserted in the common earth line to the transformer centre-tap.

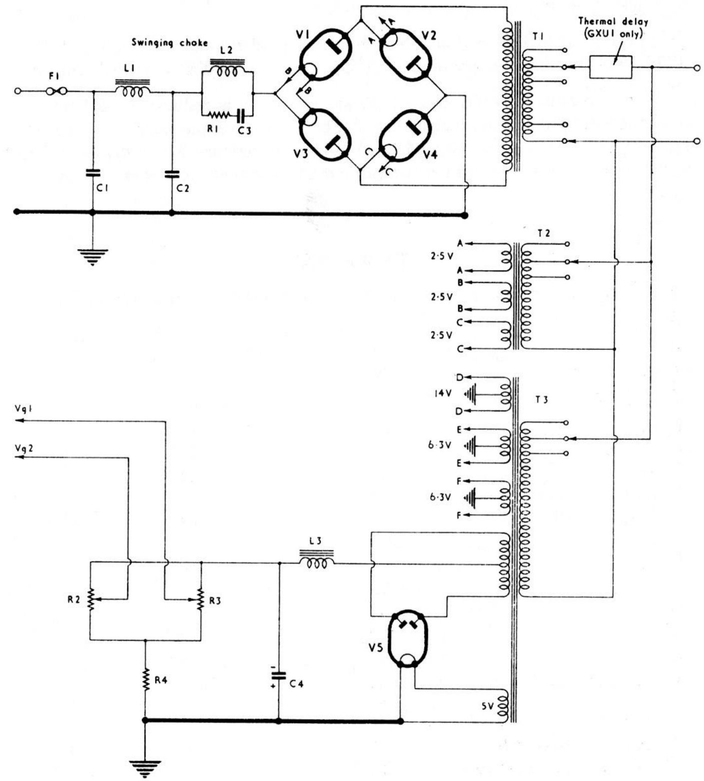


Fig. 8-5. Power supply circuit for the V1505 Class AB1 and AB2 amplifiers in figs. 8-1 and 8-6.

To obtain 1100W output, the V1505 valves will require a driving power approximately double that needed for 900W and a considerable signal voltage is necessary. The KT66 valves (the KT88 is also suitable) must operate at an anode supply of 575V and they will each consume 55mA. The required bias of about 55V is produced by an effective cathode resistance of 1000Ω per valve made up as above.

A load impedance of 10kΩ is specified and this should not be reduced as a lower load will result in excessive anode dissipation at the half-power point—where maximum dissipation occurs in Class AB2.

The power supply circuit is the same as that recommended for the Class AB1 arrangement and is shown in fig. 8-5.

The curves in figs. 8-3 and 8-4 show the performance in detail and Table 8-III gives the output stage operating data for either 900W or 1100W output.

For an output of 600W the V1505 anode supply is reduced to 2kV and the resulting performance is illustrated in fig. 8-2. The output stage data are given in Table 8-IV. At this reduced level, a load impedance lower than the 10kΩ specified is permissible and this would provide a somewhat higher output.

TABLE 8-III

OPERATING CONDITIONS OF THE CLASS AB2 900-1100W AMPLIFIER OF FIG. 8-6

$V_{a\ (b)}$	2·5 (approx)	2·5 (approx)	kV
V_a	2·5	2·5	kV
*$I_{a\ (o)}$	2×50	2×50	mA
$I_{a\ (max\ sig)}$	2×275	2×300	mA
$p_{a\ (o)}$	2×130	2×130	W
†$p_{a\ (max)}$	2×275	2×275	W
$V_{g\ (o)}$	−155	−155	V
$I_{g\ (max\ sig)}$	30	60	mA
$i_{g\ (pk)\ (max\ sig)}$	75	140	mA
P_{out}	900	1100	W
$R_{L\ (a-a)}$	10	10	kΩ
Z_{in}	8-12	8-12	kΩ
D	6	7	%
p_{dr}	10‡	20§	W
$V_{in\ (pk)\ (g-g)}$	2×230	2×280	V
$V_{in\ (rms)}$ (to first stage)	2×5	2×6	V

*See section 8-3, page 96.

†$p_{a\ (max)}$ occurs at half output.

‡Driver stage operating at 475V, 130mA.

§Driver stage operating at 575V, 110mA.

TABLE 8-IV

OPERATING CONDITIONS OF THE CLASS AB2 AMPLIFIER OF FIG. 8-6 FOR REDUCED OUTPUT OF 600W

$V_{a\ (b)}$	2 (approx)	kV
V_a	2	kV
*$I_{a\ (o)}$	2×60	mA
$I_{a\ (max\ sig)}$	2×230	mA
$p_{a\ (o)}$	2×120	W
$p_{a\ (max\ sig)}$	2×170	W
$V_{g\ (o)}$	−125	V
$I_{g\ (max\ sig)}$	40	mA
$i_{g\ (pk)\ (max\ sig)}$	95	mA
P_{out}	600	W
$R_{L\ (a-a)}$	10	kΩ
Z_{in}	8-12	kΩ
D	6	%
p_{dr}	10†	W
$V_{in\ (pk)\ (g-g)}$	2×200	V
$V_{in\ (rms)}$ (to first stage)	2×4	V

*See section 8-3, page 96.

†Driver stage operating at 475V, 130mA.

COMPONENT VALUES FOR FIG. 8-6

V1505 600-1100W AB2 AMPLIFIER

G.E.C. VALVES

V1 N709
V2 N709
V3 KT66
V4 KT66
V5 V1505
V6 V1505

RESISTORS

(20%, 0·25W unless otherwise shown)

R1 470kΩ
R2 470kΩ
R3 220Ω
R4 100Ω
R5 100Ω
R6 1·5-2·5kΩ 5W
R7 100kΩ 10%
R8 100kΩ 10%
R9 10kΩ
R10 10kΩ
R11 100Ω
R12 100Ω
R13 See Sect. 8-2, page 92
R14 10kΩ 10% 50W
R15 10kΩ 10% 50W

CAPACITORS

C1 0·02μF
C2 0·02μF
C3 4μF 600V
C4 4-16μF 350V
C5 4-16μF 350V
C6 2000pF 1·5kV
C7 2000pF 1·5kV

TRANSFORMERS

T1 Intervalve transformer 1:2
T2 Bridged transformer (see fig. 6-7)
T3 Output transformer
10kΩ anode-anode

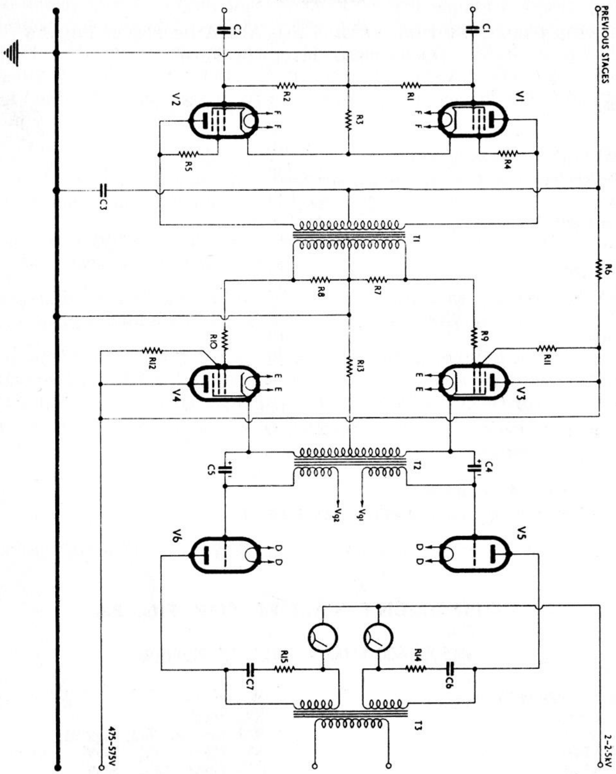


Fig. 8-6. Circuit of the V1505 Class AB2 600-1100W amplifier.

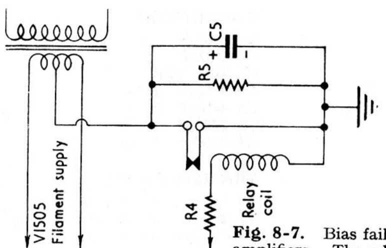


Fig. 8-7. Bias failure protection for the V1505 amplifiers. The relay coil is connected in series with R4 (fig. 8-5) and the normally-open contacts are held closed by the bias supply current. Bias failure opens the contacts and the V1505 filaments are then auto-biased by R5.

8-3. Quiescent Current in AB1 and AB2.—The quiescent currents selected (65mA and 50mA for anode potentials of 2kV and 2·5kV respectively) are not critical and any current of this order may be used. In both Class AB1 and AB2, the quiescent dissipation is low, at 150W, which is a desirable feature. However, too low a current will result in increased distortion whereas currents in excess of those recommended will not reduce distortion. These two values of current are chosen to give similar orders of the crossover distortion which occurs at full output as a result of the regulation of the grid bias supply circuit. Although a stabilised bias supply would remove the effect at the cost of some circuit elaboration it should be noted that the effect is virtually absent up to 90% of full output.

8-4. Protection against Bias Failure.—In the simple arrangement shown in fig. 8-7 a relay, with its operating coil connected to the bias supply, inserts a resistor in the V1505 filament circuit when its normally-open contacts are released by bias failure. The value of the resistor is chosen so that the V1505 valves operate with auto-bias at 100mA anode current per valve and give a reduced output of 200-300W.

CHAPTER 9
MULTIPLE-PAIR PUSH-PULL AMPLIFIERS

When an output exceeding 100W is needed, the KT88 valve may be used in multiple pairs in parallel push-pull instead of a single pair of larger valves.* One of the advantages of this method is the low cost of the power supply which is required to give an h.t. of only 550V. Another advantage is that a valve failure in the output stage merely reduces the available output power with a probable increase in distortion.

The circuit diagram of fig. A-1 illustrates the use of ten valves in a fixed bias ultra-linear output stage and this arrangement gives 400W output. More or less than five pairs of valves can be used, depending upon the power required.

A single control is used for grid bias adjustment and this simplifies the amplifier at the expense of somewhat higher distortion and lower output. For this to be successful the power supply must have good regulation and a suitable circuit is included in fig. A-3.

It is not essential to use accurately matched valves but the total current in each half of the push-pull stage should be equalised as closely as possible. This becomes easier with an increasing number of pairs and is facilitated by the cathode current meter built into the amplifier. The individual cathode currents will vary from about 35mA to about 60mA and each valve should be measured in turn and the valves sorted into two groups of approximately similar total current.

When the valves have been sorted and the two groups plugged into their respective halves of the amplifier, the optimum operating condition is obtained by adjusting the grid bias with R5 (fig. A-3) so that the current drawn by any single valve does not exceed 60mA. In this way an output of 400W will be obtained at a distortion of about 5%.

When using six or more valves in this type of amplifier, the value of the grid return resistors R14 and R15 is of importance and a low value is desirable. In order to facilitate the production of the necessary distortion-free input signal of 55+55V r.m.s., the output stage is driven by a pair of cathode followers V2B and V3B. With four output valves the grid resistors could be increased to 100kΩ and the cathode followers dispensed with. However, as they form parts of double triodes, the saving in cost is insignificant and, on the whole, it is preferable to retain them.

The method of measuring the cathode current of each KT88 valve is shown in fig. A-1. A resistor of 10Ω is inserted in the cathode lead of each valve (i.e. R38, R39, etc.) and a meter M1 is connected across this resistor through switch S1 and a series resistor (R48, R49, etc.).

It was found convenient to use a meter with a full-scale deflection of 200μA, the value of the series resistor being such that the meter indicated 0-200mA. At full output each cathode current is about 100mA to 125mA. It may be preferred to substitute the individual cathode series resistors with a single resistor inserted between the meter and switch but the possibility of instability should be borne in mind as the individual resistors act as cathode circuit isolators.

The ultra-linear output transformer must have low leakage inductance between : primary and secondary ; half-primary and half-primary ; and each half-primary and its tap. The absolute values of leakage inductance will depend upon the number of valves used but a 400W transformer used in the prototype had the following characteristics :

Primary Inductance :	4H
Leakage Inductances :	
Primary to secondary :	0·75mH
Half-primary to half-primary :	0·75mH
Each half-primary to tap :	1·5mH

To prevent ultrasonic oscillation, resistor/capacitor combinations are connected between each tap and the anode terminal of each half-primary. In the prototype, C10 and R60 were also found to be desirable, the values used being 3500pF and 1kΩ respectively.

Negative feedback may be added to this amplifier in the normal way, from the secondary of the output transformer to the cathode circuit of the input valve. The values of the resistors R2 and R3 in fig. A-1 are determined by the amount of feedback and the ratio of the transformer.

Fig. A-2 illustrates a recommended layout for the output valves and is self-explanatory. Adequate ventilation should be provided ; if in doubt, a temperature sensitive paint should be used (see footnote on page 9).

Power Supply.—The design of the power supply is an important factor in the satisfactory operation of an amplifier of this type. The regulation should be good ; better than 10% with a current variation of 400-1200mA was obtained in the prototype. This order of regulation was achieved by using the xenon filled rectifier GXU1 which, in this application, is considerably under-run at a PIV of 1kV as against the rated PIV of 10kV. For up to six output valves, the smaller xenon rectifier GXU50 is suitable.

The circuit of a complete power supply is given in fig. A-3 and this provides the lower h.t. voltage required by the earlier stages as well as the grid bias supply.

A single inductance-input filter is shown and, with a smoothing capacitance of 150-200μF (obtained by series-connecting two larger capacitors), this should be satisfactory for most purposes. A further filter section may be inserted if desired. The smoothing inductor(s) should have a value of 2-3H and a d.c. resistance of about 25Ω.

TABLE A-I

OPERATING CONDITIONS OF THE OUTPUT STAGE OF FIG. A-1

$V_{a\ (b)\ (o)}$	570	V
$V_{a\ (b)\ (max\ sig)}$	530	V
$V_{a,\ g2\ (o)}$	565	V
$V_{a,\ g2\ (max\ sig)}$	525	V
$I_{a+g2\ (o)}$ (per valve)	35 to 60	mA
$I_{a+g2\ (max\ sig)}$ (per valve)	100 to 125	mA
$I_{a+g2\ (o)}$ (total)	450	mA
$I_{a+g2\ (max\ sig)}$ (total)	1200	mA
$p_{a+g2(o)}$ (per valve)	35	W
$p_{a+g2\ (max\ sig)}$ (per valve)	25	W
V_{g1} (approx)	—75	V
P_{out}	400	W
$R_{L\ (a\text{-}a)}$	1	kΩ
Z_{out}	1·2	kΩ
D	5 to 7	%
$V_{in\ (rms)}$ to first stage	250	mV

If 10db of negative feedback is applied, the last three values become 400Ω, 2% and 750mV respectively.

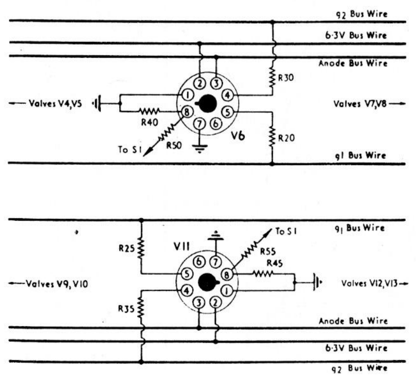


Fig. A-2. Recommended layout for the output valves of the 400W amplifier.

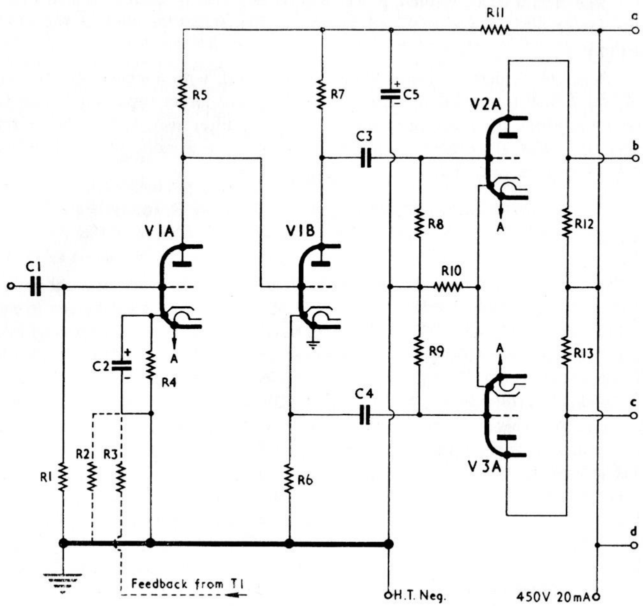
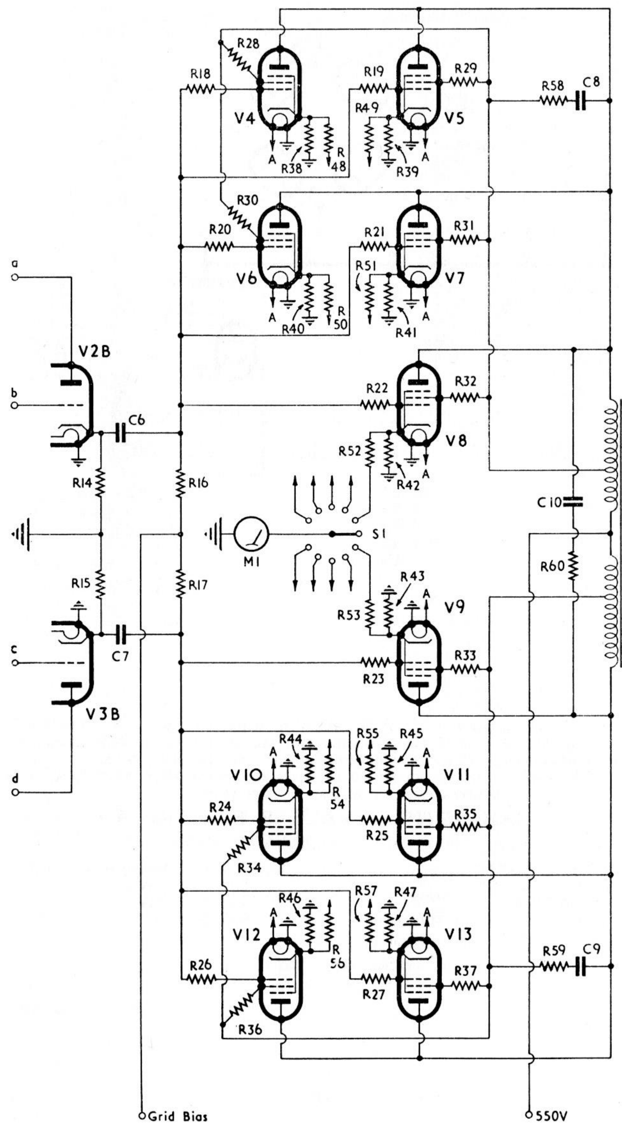


Fig. A-1. Circuit of the 400W multiple-pair amplifier, Component values are given



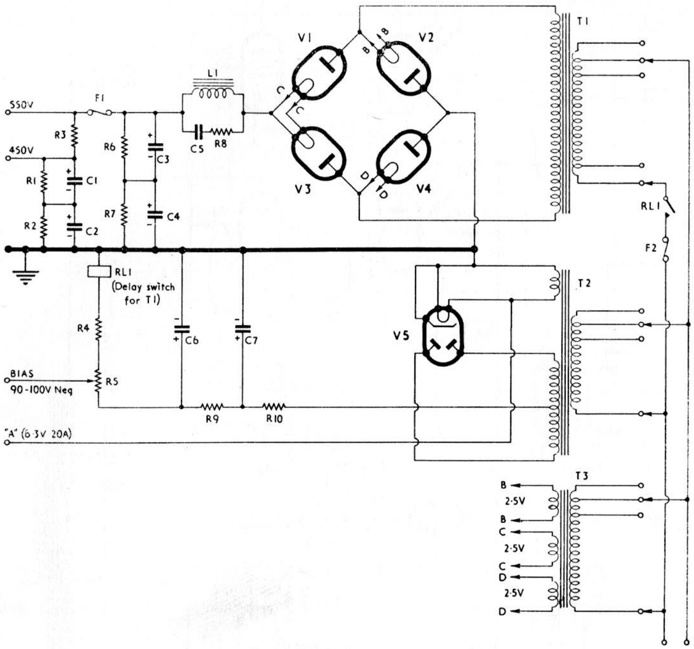


Fig. A-3. Power supply circuit for the 400W multiple-pair amplifier. The winding of T3 is simplified if four identical secondaries are used. To prevent mains-borne interference with nearby receivers, generated by the xenon rectifiers, a pair of capacitors (0·05μF, 700V a.c.) should be connected in series across the 700V winding and their junction connected to the negative (earth) line.

COMPONENT VALUES FOR FIG. A-1

KT88 400W AMPLIFIER

G.E.C. VALVES

V1-V3 B65/6SN7
V4-V13 KT88

R48-R57 Meter series resistors,.
R58 1kΩ 2W
R59 1kΩ 2W
R60 See p. 113

RESISTORS

(*20%, 0·25W unless otherwise shown*)

R1 1MΩ
R2, R3 } See page 113
R4 1kΩ
R5 100kΩ 0·5W
R6 15kΩ 0·5W } matched to 5%
R7 15kΩ 0·5W }
R8 1MΩ
R9 1MΩ
R10 4·7kΩ 10% 0·5W
R11 33kΩ 0·5W
R12 220kΩ 1W } matched to 5%
R13 220kΩ 1W }
R14 47kΩ 1W } matched to 5%
R15 47kΩ 1W }
R16 47kΩ 10% 0·5W
R17 47kΩ 10% 0·5W
R18-R27 15kΩ
R28-R37 220Ω
R38-R47 Meter shunts, see p. 113

CAPACITORS

C1 0·01μF
C2 50μF 12V
C3 0·01μF
C4 0·01μF
C5 8μF 450V
C6 0·25μF
C7 0·25μF
C8 1000pF
C9 1000pF
C10 See p. 113

MISCELLANEOUS

M1 200μA meter (see p. 113)
S1 1-pole 11-way switch b.b.m.
T1 400W Ultra-linear transformer
1 kΩ anode-anode
Primary inductance : ≮ 4H
Leakage inductances :
Prim.—sec. : ≯ 750μH
½ prim.—UL tap : ≯ 750μH

COMPONENT VALUES FOR FIG. A-3

POWER SUPPLY FOR KT88 400W AMPLIFIER

G.E.C. VALVES

V1-V4 GXU1
V5 U709 or U78/6X4

RESISTORS

R1 100kΩ 10% 0·5W
R2 100kΩ 10% 0·5W
R3 4·7kΩ 20% 2W
R4 22kΩ 20% 1W
R5 10kΩ wirewound
R6 47kΩ 10% 1W
R7 47kΩ 10% 1W
R8 10kΩ 20% 1W
R9 22kΩ 20% 1W
R10 1·5kΩ 20% 0·25W

CAPACITORS

C1 32μF 350V
C2 32μF 350V
C3 300μF 350V } each 100+200 dual
C4 300μF 350V }
C5 0·01μF 500V
C6 2μF 250V
C7 2μF 250V

MISCELLANEOUS

L1 2H 1200mA 25Ω
T1 700V 1200mA*
T2 Bias transformer
Secondaries
150-0-150V 10mA
6·3V 20A
T3 2·5V 5A
2·5V 5A
2·5V 10A (or 2×5A)
F1 2A
F2 5A

**10V taps on primary*

OUTPUT TRANSFORMERS AND STABILISATION

The output transformers used in the prototypes of the ultra-linear amplifiers described in this book were components of good quality made by various manufacturers. The leakage inductance was found to vary considerably, not so much between units by different makers, as between the dates of manufacture. The most recently made transformers incorporated improved methods of winding which gave very low values of leakage inductance. As a guide to this improvement, the values are tabulated below :

Relative Age of Transformer	Leakage Inductance (mH)			Comments
	P-S	½P-½P	½P-UL tap	
Early	20-25	20	50	a-g2 capacitors required for stability
Medium	10-15	10	25	Stable without capacitors
Late	5-10	5	5	Stable with 30db of negative feedback

The capacitors mentioned in the last column are those shown connected between the anode and screen of each output valve in the ultra-linear amplifier circuit diagrams in previous chapters. They have a value of 1000-2000pF and are connected in series with resistors of 470-1500Ω.

All amplifiers and transformers were examined with a square wave generator and oscilloscope at a frequency of 5-10kc/s and it was found that the overshoot produced was roughly proportional to the leakage inductance. The earlier transformers produced an overshoot of about 25% and a ringing frequency below 40kc/s. In the later types the overshoot is reduced to less than 10% at a frequency above 100kc/s.

When negative feedback was applied it was found essential with the two earlier groups and desirable with the latest transformers to attenuate the high frequency response of the amplifier. The method of doing this is shown by the circuit of fig. B-1, in which the attenuation is achieved by a capacitor and resistor in series between the anode of the first triode amplifier stage and earth. A slightly different arrangement is necessary when the first stage consists of a balanced phase inverter such as the one used in the KT88 50W amplifier of fig. 5-2 (page 55). In this, the attenuating components (C6, R10 and C7, R11) are incorporated in each half of the phase inverter in order to maintain the balance of the stage. The principle is the same, however, and the following comments apply to both arrangements.

Referring now to fig. B-1, the optimum values of C1 and R1 will depend upon the degree of attenuation required but, with the latest transformers, values of 500pF and 10kΩ gave an attenuation of 3db at 25kc/s and 10db at 40kc/s. With these values, complete stability was obtained with feedback applied. With earlier transformers the capacitor and resistor were required to have values up to 2000pF and down to 1kΩ. All these values assume a valve impedance of about 10kΩ but if V1 in fig. B-1 is a pentode-connected Z729 with an anode load of 100-330kΩ, suitable values for C1 and R1 would be 50-200pF and 100-10kΩ respectively.

To ensure stability at very low frequencies, below the cut-off frequency of the output transformer, it is advisable to insert low frequency attenuation

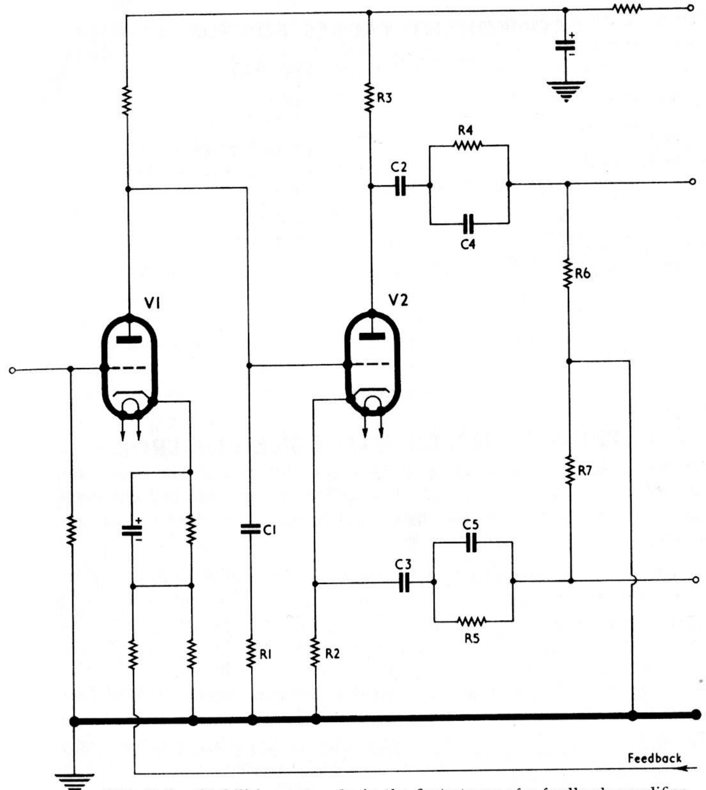


Fig. B-1. Stabilising networks in the first stages of a feedback amplifier.

between the first and second stages within the feedback loop. The recommended arrangement is shown in fig. B-1 and consists of a capacitor and shunt resistor in each grid line between the two stages, designated C4, R4 and C5, R5. Assuming the following grid resistors R6 and R7 to be each 1MΩ, values of 5000pF and 3·3MΩ will attenuate frequencies below 25c/s and will be satisfactory with an output transformer of high primary inductance.

Low values of transformer primary inductance will need correspondingly lower and higher values, respectively, of capacitors and resistors. Values of 2000pF and 10MΩ will be found adequate for transformers of poor quality in terms of primary inductance. Low frequency instability ("motor-boating") in amplifiers containing poor-quality output transformers may be eliminated by incorporating these networks.

COMPONENT VALUES FOR FIG. B-1

STABILISING NETWORKS

VALVES

V1, V2 } B65/6SN7

RESISTORS

(20%, 0·25W unless otherwise shown)

R1 1-10kΩ

R2 15kΩ 0·5W } matched to 5%
R3 15kΩ 0·5W }

R4 3·3MΩ } see text above
R5 3·3MΩ }

R6 1MΩ

R7 1MΩ

CAPACITORS

C1 500-5000pF (see text above)

C2 0·05μF

C3 0·05μF

C4 5000pF

C5 5000pF

CHAPTER 10

PRE-AMPLIFIERS AND TONE CONTROL

This chapter discusses some of the requirements of pre-amplifiers for use with the power amplifiers described in earlier chapters and includes details of two complete pre-amplifier circuits containing record and tape equalisation and tone controls.

9-1. The Purpose of a Pre-Amplifier.—This can be summed up, within the context of this book, in three requirements : (*a*) the amplification of small signals, of the order of millivolts, to a level suitable for operating a power amplifier without adding hum and other noise ; (*b*) frequency " equalisation " of the output from gramophone records, tape, etc. ; (*c*) the provision of variable tone controls, which, to satisfy present-day demands requires at least bass and treble cut and bass lift.

These three requirements are dealt with in more detail in the following paragraphs.

Amplification.—A comprehensive pre-amplifier of the type described later in this chapter must be suitable for use with microphones, tape playback heads and low-sensitivity pickups.

The output voltage of a typical crystal microphone of the diaphragm type connected directly to the grid of the first valve is about 2mV for a sound pressure of 1 dyne/cm^2. This is roughly equivalent to loud speech a foot or so from the microphone and normal conversation will give about 0·5mV. Signal magnitudes of this order will also be obtained at the input valve grid from a moving-coil or ribbon microphone connected via an appropriate step-up transformer. With such low level input voltages the hum and noise level at the grid must not exceed about 1·5μV for a signal-to-noise ratio approaching 50db (300:1), which is an acceptable standard for high-fidelity equipment.

The output from tape playback heads varies from about 1·5mV, at 1000c/s, for the cheapest kinds up to about 10mV for the more expensive heads. Obviously, the greater the output, the better will be the signal-to-noise ratio obtained with any given tape. Magnetic tape is recorded at constant peak flux density so that the head output falls with decreasing frequency at the rate of 6 db per octave. Consequently, the output at 50c/s will be about 24db less than that at 1000c/s, that is, about 0·1mV for a low-output head. It follows that, with this type of head, a signal-to-noise ratio of 40db (100:1) will necessitate a noise level at the grid no higher than about 1μV.

An input stage which meets the requirements of high signal-to-noise ratio with adequate gain can be built by using the Z729, a valve especially designed for this type of application. The pre-amplifier circuits of figs. 9-2 and 9-3 both contain a tape playback amplifier of this kind, the valve being V1 in each case. Certain precautions must be observed, particularly with the wiring of grid, heater and cathode circuits and in the use of suitable components in order to take full advantage of the performance of the valve itself.

The grid and cathode wiring must be compact and direct in order to minimise the electrostatic pick-up of hum and the magnetic loop formed by the input terminal, grid, cathode and earth must be as small as possible. A single earth connection is essential and this should be close to the input socket, which should be screened. The part of the chassis containing the input stage may, with advantage, be totally enclosed with metal screens. The heater wiring should consist of a twisted pair and should be kept well clear of the grid circuit. The heaters should be wired up from the supply end starting with the output valve and finishing with the input stage. In this way, the heater supply wires running to the first stage will carry only the current drawn by the input valve and the electro-magnetic field about these wires will be at a minimum.

Care in the choice of components used in the input stage is necessary and all capacitors should be of high quality if they are not to introduce noise due to variable insulation resistance etc. Resistors should be of the high-stability type in order to minimise noise produced by thermal agitation (Johnson noise) and other effects. The valve socket should be of unloaded plastics or ceramic material and should preferably contain a central screening boss, this being earthed together with contacts 2 and 7 which connect to the internal screening system of the Z729 valve.

Equalisation. As explained above, the frequency characteristic of the output from a tape playback head or gramophone pickup is not level and a suitable equalising network must be incorporated in the pre-amplifier in order to reproduce these signals with a flat frequency characteristic.

Such a network introduces a loss in amplification which must be made good and it is convenient to combine it with a valve amplifier, the amplification of which makes up for the insertion loss of the correcting circuit. By arranging the equalising network in the form of a negative feedback loop in this amplifier stage, valve distortion is reduced to a minimum.

The equalisation basically comprises bass boost of 10-15db at 50c/s for record reproduction and 24db or more at 50c/s for tape playback. This is achieved by suitably attenuating the higher frequencies, full gain being obtained only at the lowest frequencies. Fig. 9-1 is a simplified diagram of this circuit,

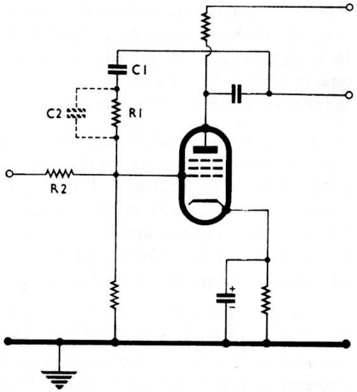


Fig. 9-1. Simplified circuit showing how the response of an amplifier may be modified by placing a suitable net work in the feedback loop.

the corresponding valves in the full circuits of figs. 9-2 and 9-3 being V2 in each case.

The feedback loop can be made to modify the response of this amplifier stage in any desired way by introducing into it a suitable frequency selective network. To obtain bass boost, this network consists basically of C1 and R1 in fig. 9-1 and the values of these components are chosen so that the impedance of C1 becomes greater than the resistance of R1 at a particular low frequency. The amount of feedback thus becomes smaller with decreasing frequency and the amplification increases, giving bass boost. For record reproduction, high frequency de-emphasis is introduced by the addition of C2.

In the complete pre-amplifiers, a suitable equalising network is automatically switched into circuit for each type of input source. For radio reproduction the gain of this stage is reduced uniformly at all frequencies, by introducing pure resistance into the feedback loop, to suit the larger input signal from a radio tuner.

As stated above, when the signal is from a tape playback head considerable attenuation of all but the lower frequencies is necessary for true equalisation and an additional stage must precede the equalising amplifier. The requirements for this input stage were discussed under " Amplification " on page 97.

An additional amplifier is not necessary for record equalisation because the required bass boost is considerably less and, normally, the input signal is larger.

The gain control follows the equalising stage as the signal level is such that any noise generated by a potentiometer may be neglected.

Tone Control.—Having achieved a signal of uniform frequency characteristics from any of the input sources by the action of the equalising stage, tone controls can be introduced for adjusting the balance and frequency range to suit listening conditions, programme quality and personal taste.

Both pre-amplifier designs incorporate tone controls, those in the circuit of fig. 9-3 being somewhat more comprehensive. The details of the tone control circuits are given in the descriptions of the pre-amplifiers below.

9-2. Pre-Amplifier Circuit No. 1.—The circuit diagram of this pre-amplifier is given in fig. 9-2. This design, the simpler of the two described in this chapter, utilises three Z729 valves and is suitable for feeding an amplifier requiring about 0·5V input. V1 is the tape playback amplifier and the sensitivity at the input socket of this stage is 1mV. The output of this stage and the input signals from the pickup and radio sockets are connected to the grid of V2 via the input selector switch S1A.

COMPONENT VALUES FOR FIG. 9-2

PRE-AMPLIFIER CIRCUIT NO. 1

G.E.C. VALVES

V1, V2, V3 } Z729

RESISTORS

(10%, 0·25W unless otherwise shown)

*R1 2·2MΩ 5%
*R2 5%, *R3 5% } See page 102
*R4 3·3kΩ 5%
*R5 47kΩ 5%
*R6 220kΩ 5%
*R7 470kΩ 5%
*R8 220kΩ 5%
*R9 1MΩ 5%
*R10 470kΩ 5%
R11 220kΩ
R12 220kΩ
R13 47kΩ
R14 470kΩ
R15 1MΩ
R16 10MΩ
R17 1MΩ
R18 10MΩ
R19 220kΩ
R20 22MΩ
*R21 3·3kΩ 5%
*R22 220kΩ 5%
*R23 1MΩ 5%
R24 500kΩ Log.
R25 2·2MΩ
R26 47kΩ
R27 1·5kΩ 5%
R28 100kΩ
R29 470kΩ
R30 220kΩ 5%
R31 1MΩ Log.
R32 22kΩ 5%
R33 22kΩ

*High Stability

CAPACITORS

C1 16μF 350V
C2 25μF 25V
C3 0·1μF
C4 0·05μF
C5 33pF 10%
C6 330pF 10%
C7 68pF 10%
C8 220pF 10%
C9 470pF 10%
C10 25μF 25V
C11 0·1μF
C12 0·05μF
C13 8μF 350V
C14 0·005μF
C15 25μF 25V
C16 0·1μF
C17 0·05μF
C18 2000pF 10%
C19 0·02μF
C20 16μF 450V
C21 1000pF 10%
C22 2000pF 10%
C23 0·05μF

SWITCHES

S1 2-pole 4-way m.b.b.
S2 1-pole 4-way m.b.b.

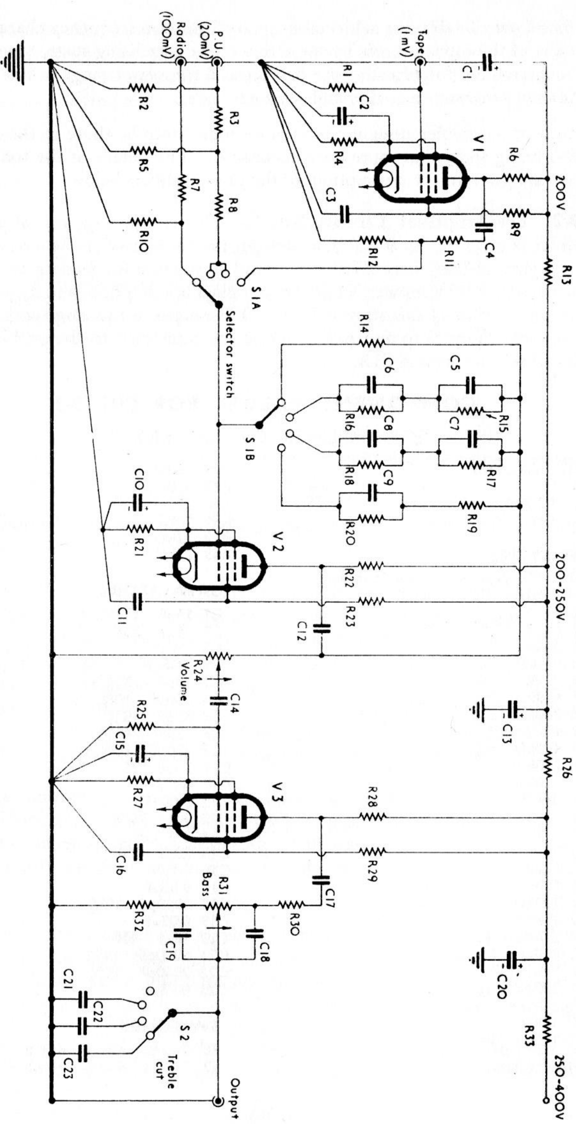


Fig. 9-2. Circuit of Pre-Amplifier No. 1. All switches are shown fully anticlockwise and the arrows at the potentiometers indicate clockwise rotation.

Switch S1 has four positions and in clockwise order these are : radio (100mV), pickup 78 r.p.m. (20mV), pickup 33 r.p.m. (20mV) and tape playback (1mV). S1A selects the required input signal and S1B automatically inserts into the feedback loop of V2 the appropriate equalising network.

The values of resistors R2 and R3, in the pickup input circuit, will depend upon the type of pickup used. Variation of the value of R2 will alter the load impedance presented to the pickup and variation of R3 will alter the input sensitivity to suit the output of the pickup. With many moving iron (variable reluctance) pickups R2 should be open-circuited (R2=∞Ω) and R3 short-circuited (R3=0Ω) but if this results in full output with the volume control at less than half-travel, a suitable value of R3 will attenuate the pickup output and so prevent distortion in V2. A value of 47kΩ for R2 will provide a suitable load impedance for most magnetic pickups. The output of a crystal pickup can be made similar to a magnetic type by making R2 100kΩ and reducing its output by making R3 1MΩ (see fig. 9-8).

The tone control stage consists of V3 and its associated bass and treble controls R31 and S2. Both controls utilise conventional circuits, the bass control being a potentiometer which boosts the bass as the slider travels away from the centre in one direction and cuts as it is moved in the opposite direction, with the centre position giving a level response. The treble control consists of a four-position switch giving three degrees of cut and one level position.

The input sockets and the first valve should be grouped together with a single earth connection to the chassis. The earth connection from the equalising stage (V2) should be returned to the same point. The two selector switches may be of the bakelite wafer type, with make-before-break contacts.

The grid coupling capacitors are chosen so that their combined effect produces attenuation below 30c/s in order to reduce rumble.

As the output impedance of the pre-amplifier is about 22kΩ, up to four feet of ordinary screened wire or up to twenty feet of coaxial cable may be connected without high frequency loss.

This pre-amplifier may be used as a substitute for the built-in tone controls of the 912-PLUS amplifier when the flexibility of a separate control unit is preferred. The h.t. feed for the pre-amplifier is taken from the junction of C11 and R28 in fig. 3-3 (page 28). The output of the pre-amplifier, fed into the first grid of V2 in fig. 3-3, provides the correct input signal for the 912-PLUS.

9-3. Pre-Amplifier Circuit No. 2.—This pre-amplifier was designed for use with the KT88 50W amplifier described in Chapter 5* but it may be used to feed any power amplifier requiring an input signal of 0·5V. It is very suitable for use with the G.E.C. 912-PLUS amplifier as a substitute for the built-in tone control stage. Details of its use with the 912-PLUS are given in Chapter 3.

The pre-amplifier circuit is illustrated in fig. 9-3 and it is designed to operate from a gramophone pickup, a tape playback head (for which a separate input stage, V1, is used), a high-impedance microphone and a radio tuner. Choice of any of these inputs is provided by a selector switch which also automatically modifies the pre-amplifier sensitivity and equalisation to suit the selected input source.

The six positions of the input selector switch are given in clockwise order by the table below which also shows the input sensitivity, input impedance and an indication of the equalisation obtained at each position.

Switch Position	Sensi-tivity*	Input Imped-ance	Frequency Correction	
			Bass Boost at	Treble Cut at
Radio	0·2V	470kΩ	—	—
78 r.p.m. Records	14mV	50kΩ	400c/s	6kc/s
British Microgroove Records ...	10mV	50kΩ	500c/s	3kc/s
American Standard Records ...	12mV	50kΩ	600c/s	2kc/s
Tape (7½in/s)	4mV	220kΩ	1·5kc/s	—
Microphone (high impedance) ...	1mV	1·1MΩ	—	—

*Input for 0·5V r.m.s. output from pre-amplifier.

Equalisation Networks.—The equalising stage is of the negative feedback type as described in Sect. 9-1 above and it utilises another G.E.C. Z729 low noise pentode. The various equalisation characteristics are illustrated in figs. 9-4 to 9-7. The tape playback equalisation is based on the recommendation that tape recorded at 7½in/s should be played back with a time constant of 100μs and that treble loss in recording should be offset by pre-emphasis.† This is equivalent to a bass boost of 6db per octave from 1500c/s and is obtained by C10 and R19 in the feedback loop. At other tape speeds, these components should be as follows :

Speed	*Time Constant*	*C10*	*R19*
3¾in/s	200μs	1000pF	220kΩ
15in/s	35μs	150-220pF	220kΩ

The selector switch may be provided with more positions if these additional equalisation characteristics are required to be built into the pre-amplifier. It should be noted that R17 is connected across C10 merely to eliminate switch clicks and does not form part of the equalisation circuit.

In general, it is undesirable to compensate for tape losses by treble boosting on playback as this will accentuate tape hiss. However, a degree of treble boost

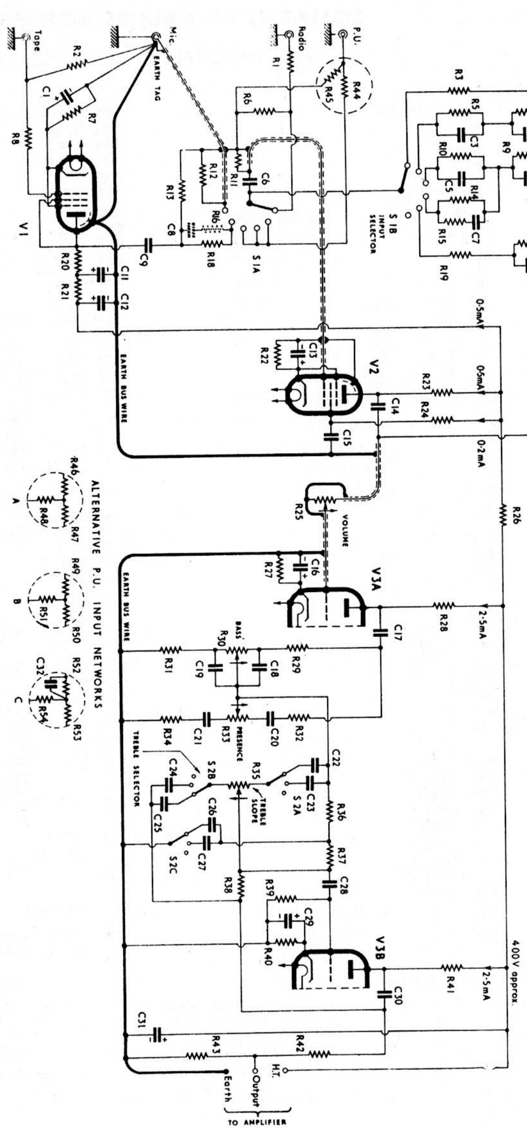


Fig. 9-3. Circuit of Pre-Amplifier No. 2. All switches are shown fully anticlockwise and the arrows at the potentiometers indicate clockwise rotation. The circuit is arranged to show the method of connecting all earth returns to a bus which is earthed to the chassis only at the microphone input socket. Details of the alternative pickup input networks are given

COMPONENT VALUES FOR FIG. 9-3

PRE-AMPLIFIER CIRCUIT NO. 2

G.E.C. VALVES

V1 Z729
V2 Z729
V3 B329/12AU7

RESISTORS

(10%, 0·25W unless otherwise shown)

R1 470kΩ
*R2 220kΩ 5%
R3 470kΩ
R4 10MΩ
R5 820kΩ 5%
R6 470kΩ
*R7 3·3kΩ 5%
*R8 10kΩ 5%
R9 22MΩ
R10 1·5MΩ 5%
R11 2·2MΩ
R12 2·2MΩ
R13 1MΩ
R14 1MΩ 5%
R15 10kΩ
R16 100kΩ (see page 107)
R17 22MΩ
R18 220kΩ
R19 220kΩ
*R20 100kΩ 5%
R21 22kΩ
*R22 3·3kΩ 5%
*R23 220kΩ 5%
R24 1MΩ
R25 500kΩ Log.
R26 22kΩ
R27 1kΩ
R28 47kΩ
R29 100kΩ
R30 1MΩ Log.
R31 10kΩ
R32 47kΩ
R33 500kΩ Log.
R34 3·3kΩ
R35 2MΩ Log.
R36 100kΩ
R37 100kΩ
R38 680kΩ
R39 1·5MΩ
R40 1kΩ
R41 47kΩ
R42 10kΩ
R43 150kΩ
R44 100kΩ
R45 100kΩ
R46 47kΩ
R47 100kΩ
R48 10kΩ
R49 100kΩ
R50 100kΩ
R51 2·2kΩ
R52 100kΩ
R53 100kΩ
R54 2·2kΩ

CAPACITORS

C1 25μF 12V
C2 470pF 5%
C3 33pF 5%
C4 220pF 5%
C5 33pF 5%
C6 0·02μF
C7 82pF 5%
C8 100pF 5%
C9 0·05μF
C10 470pF 5%
C11 8μF 450V
C12 8μF 450V
C13 25μF 12V
C14 0·05μF
C15 0·1μF
C16 25μF 12V
C17 0·1μF
C18 5000pF 5%
C19 0·04μF
C20 1000pF 5%
C21 0·02μF
C22 680pF 5%
C23 330pF 5%
C24 100pF 5%
C25 47pF 5%
C26 470pF 5%
C27 220pF 5%
C28 0·002μF
C29 25μF 12V
C30 0·02μF
C31 8μF 500V
C32 220pF 5%

SWITCHES

S1 2-pole 6-way m.b.b.
S2 3-pole 3-way m.b.b.

High stability

on playback is an advantage when the playback head itself is responsible for high frequency loss. A suitable amount of treble boost may be introduced by shunting R18, in the output circuit of V1, by C8 as shown in broken line in fig. 9-3. C8 should not exceed 100pF for $7\frac{1}{2}$in/s and the effect of this capacitor is shown in fig. 9-4. The limiting resistor R16, in series with C8, should be 100kΩ.

Gramophone record equalisation is obtained in the same way as for tape, e.g. for American Standard records bass boost is provided by C4 and R14 with treble roll-off or de-emphasis by C7. Resistor R15, in series with C7, is inserted to limit feedback at very high frequencies which would cause instability. It was not found necessary to include this resistor in the other feedback paths. Resistors R4, R9 and R17 eliminate switch clicks and, in addition, limit the bass boost.

The record equalisation networks are designed for use with moving coil or moving iron (variable reluctance) pickups which give an output voltage proportional to recorded velocity. The pre-amplifier input impedance and sensitivity are suitable for use with most pickups of this type, which give a maximum output of 10mV to 50mV. When using a pickup of different characteristics, such as a crystal type, the input network must be appropriately modified. Normally, with the optimum load impedance of about 1MΩ, a good quality crystal pickup gives an approximately flat response from a record, in other words it provides its own bass boost to compensate the falling bass characteristic of the record. However, it has been found that a very smooth frequency response, extending up to the high frequency peak, can be obtained if a crystal pickup is connected to a load lower than the optimum. The resulting response is very similar to that of a velocity pickup and it is then corrected by the equalising network as for a magnetic type. Fig. 9-8 shows the response of a Collaro " Studio P " crystal pickup to a British microgroove test record plotted with a normal load of 1MΩ and with a load of 100kΩ. The fall in bass in the latter curve is caused by the internal capacitance of the pickup and the result closely resembles the British Microgroove recording characteristic, also shown in the graph.

As the output of a crystal pickup is considerably greater than that of a magnetic type, the input network of the pre-amplifier must be modified to attenuate the signal to the level required for correct operation of V2. Inset in fig. 9-3 are three alternative pickup input networks. Inset " A " is for use with magnetic pickups having a maximum output greater than 50mV. Maximum output is here defined as that given by a recorded velocity of 7cm/s and corresponds to high modulation on an average record. Inset " B " is for use with good-quality popular crystal pickups, such as the Collaro " Studio P." Inset " C " is for best quality crystal pickups having a less pronounced treble peak, such as the Collaro " Transcription."

Tone Control.—The gain control is inserted after the equalising stage and the various tone controls are built into a following double triode amplifier stage. V3A feeds two frequency selective potential divider circuits which provide variable control of bass and " presence." Logarithmic controls are used for R30 and R33 in order to give a level response when their sliders are at the mid position. The effect of these controls is shown in fig. 9-9.

The presence control alters the level of all frequencies above about 1500c/s. It therefore alters the balance between high and low frequencies so that an increase in presence, i.e. turning the control clockwise from mid-travel thereby lifting the high frequencies, gives a more " forward " quality. This is as though the listener were to move nearer the orchestra or speaker. Conversely, anticlockwise rotation from the mid position simulates the effect of an orchestra receding and approximates to listening from the back of the concert hall.

The signal passes from these controls to V3B, to which is applied negative feedback, the loop consisting of R38 in conjunction with isolating resistors R36 and R37. Associated with these resistors are the treble boost capacitors C22 and C23, and the treble cut capacitors C24 to C27. Appropriate combinations of these capacitors are switched into circuit by the treble selector S2 and the response curve so obtained may be modified by the treble slope potentiometer R35. This control is inoperative when S2 is at the level position but an overall cut or *boost* is obtained by rotating the potentiometer when S2 is at either of the *cut* positions. The variation in treble response with rotation of the treble slope control at both " cut " positions of S2 is shown in figs. 9-10 and 9-11.

If a normal logarithmic potentiometer is used for treble slope, it will attenuate the treble as it is rotated clockwise. An inverse log potentiometer must be used to obtain increasing treble with clockwise rotation if this is preferred.

Rumble Filter.—Within the feedback loop of V3B are two coupling networks, C28 R39 and C30 R42 R43, each having a time constant of 3ms. As a result, low frequencies are attenuated and positive feedback is introduced below 50c/s, the combined effect giving a flat response down to 30c/s below which the response falls steeply down to about 15c/s, when the coupling capacitors elsewhere in the pre-amplifier provide additional attenuation.

The effect of the rumble filter is shown in fig. 9-9. If it is desired to attenuate below 40c/s instead of 30c/s, R39 and R43 should be made 1MΩ and 100kΩ respectively.

Output Impedance.—The negative feedback on V3B maintains a low output impedance over the greater part of the frequency range of the pre-amplifier. The actual value is approximately equal to R42, i.e. about 10kΩ and up to five feet of ordinary screened wire may be connected to the output terminal with no deterioration in high frequency performance.

Construction.—The circuit diagram, fig. 9-3, is drawn in a way that illustrates the method of earthing all the components to a continuous earth wire which is connected to the chassis only at a tag mounted on the microphone input socket.

The two switches are bakelite wafer types, S1 having each pole on a separate wafer whilst S2 is a single wafer 3-pole 3-way switch.

The pre-amplifier earth bus wire should be connected to the main amplifier chassis only by a single wire which may conveniently be the screening braid on the signal output cable. This cable itself may be one of the elements of a multi-core cable between pre-amplifier and main amplifier.

The precautions to be taken in constructing a pre-amplifier of this type are described on page 11 and if these are borne in mind and the circuit diagram intelligently interpreted, the construction should be quite straightforward. The original article, mentioned on page 102, in which constructional information and illustrations were given, should be consulted by those with little experience in building this type of equipment.

Power Supplies.—The pre-amplifier was designed to draw its power supplies from the KT88 50W amplifier (Chapter 5) and the h.t. line smoothing is the minimum necessary for adequate ripple attenuation in order to maintain the h.t. as high as possible. When using the KT88 amplifier, no added h.t. decoupling is necessary because the stabilising circuits in that amplifier and its good push-pull balance contribute greatly to its freedom from instability at low frequencies. It is possible that further smoothing and/or decoupling may be required when using the pre-amplifier with a different type of power amplifier. This point is covered in Chapter 3 when using the pre-amplifier as part of the G.E.C. 912-PLUS (see page 33).

Performance.—The KT88 50W amplifier requires 0·5V r.m.s. input for maximum output and the pre-amplifier will supply this signal at a harmonic distortion comparable with that of the KT88 amplifier. The distortion does not increase with variation of the controls but to avoid the possibility of distortion in the equalising stage (V2) all input levels must be adjusted so that the full output of 50W from the KT88 amplifier (corresponding to 0·5V pre-amplifier output) is not obtained until the pre-amplifier gain control is beyond the centre of its travel. Given this condition, an increase in input signal level of nearly 20db can be accepted with no increase in distortion.

The signal-to-noise ratio of the complete equipment comprising the pre-amplifier and the KT88 amplifier is —76db relative to 50W with the gain control at minimum and the tone controls set for level response. With the gain control at maximum and the input sockets short-circuited to earth, the following signal-to-noise ratios were obtained at each position of the input selector on a prototype : Radio : —69db ; A.S. Records : —63db ; Tape : —52db ; Microphone : —55db.

CHAPTER 11

HIGH-FIDELITY AUDIO AMPLIFIER

Class AB_1; Power Output, 15 Watts

Circuit Description

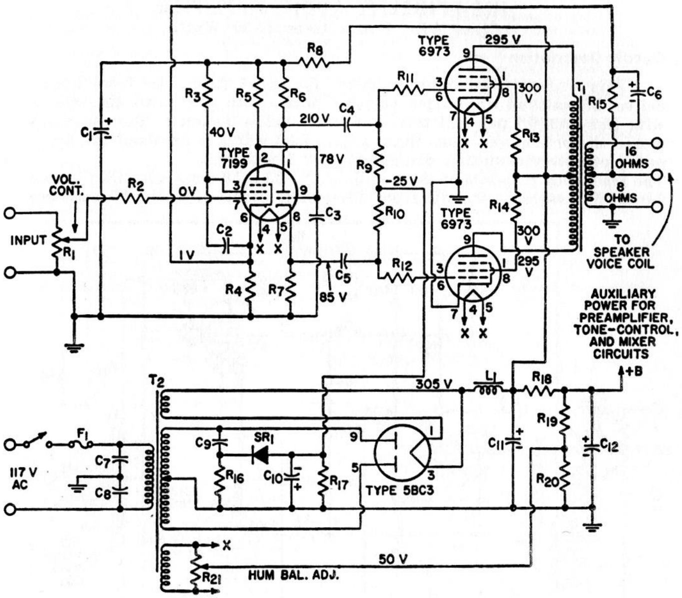
This high-fidelity audio power amplifier can deliver 15 watts of rms output power with less than 0.4 per cent total harmonic distortion and less than 1.5 per cent intermodulation distortion. The frequency response of the amplifier is flat within ± 0.5 dB from 20 Hz to 60 kHz, and the sensitivity is such that the rated output of 15 watts is obtained for an input of 1.2 volts rms. The total hum and noise, with the input shorted, is 84 dB below 15 watts. The circuit operates from a 117-volt ac power line. The transformer-coupled ac input power is converted to dc operating power for the amplifier stages by the 5BC3 full-wave rectifier. Heater power for the amplifier tubes and the rectifier are obtained from the 6.3-volt and 5-volt secondary windings, respectively, on the rectifier power transformer (T_2).

A high-gain pentode voltage amplifier is used as the input stage for the audio power amplifier. The output of this stage is direct-coupled to the control grid of a triode split-load type of phase inverter. The use of direct coupling between these stages minimizes phase shift and, consequently, increases the amount of inverse feedback that may be used without danger of low-frequency instability. A low-noise 7199 tube, which contains a high-gain pentode section and a medium-mu triode section in one envelope, fulfills the active-component requirement for both the pentode input stage and the triode phase inverter. Potentiometer R_1 in the input circuit of the 7199 pentode section is the volume control for the amplifier.

The plate and cathode outputs of the phase inverter, which are equal in amplitude and opposite in phase, are used to drive a pair of pentode-connected 6973 beam-power tubes used in a class AB_1 push-pull output stage. The 6973 output tubes are biased for class AB_1 operation by the fixed negative voltage applied to the control-grid circuit from the rectifier circuit. Fixed bias is used because a class AB amplifier provides highest efficiency and least distortion for this bias method.

Transformer T_1 couples the audio-amplifier output to the speaker. The taps on the secondary of this transformer match the plate-to-plate impedance of the output stage to the voice-coil impedance of an 8- or 16-ohm speaker. Negative feedback of 19.5 dB is coupled from the secondary of the output transformer (speaker voice coil) to the cathode of the input stage to reduce distortion and to improve circuit stability.

Fixed-bias operation of the output stage requires that the power supply provide very good voltage regulation because the plate current of the 6973 tubes varies considerably with the signal level. The conventional choke-input type of power supply used provides the required regulation. The fixed bias for the output stage is obtained from one-half the high-voltage secondary winding of power transformer T_2 through a capacitance-resistance voltage divider and the 20-milliampere, 135-volt selenium rectifier. Potentiometer R_{21} connected across the 6.3-volt secondary winding of transformer T_2 provides a hum balance adjustment for the audio power amplifier. The wiper arm of this potentiometer is connected to the junction of a resistive voltage divider across the output of the power supply. The resulting positive bias voltage applied to the tube heaters minimizes heater-to-cathode leakage and substantially reduces hum.



Parts List

C_1=40 μF, electrolytic, 450 V.
C_2, C_4, C_5=0.25 μF, paper, 400 V.
C_3=3.3 pF, ceramic or mica, 600 V.
C_6=150 pF, ceramic or mica, 400 V.
C_7, C_8=0.05 μF, paper, 400 V.
C_9=0.02 μF, paper 600 V.
C_{10}=100 μF, electrolytic, 50 V.
C_{11}=80 μF, electrolytic, 450 V.
C_{12}=40 μF, electrolytic, 450 V.
F_1=Fuse, 3 amperes
L_1=Choke, 3 H, 160 mA, dc resistance 75 ohms or less, Triad C-13X or equiv.
R_1=Volume control, potentiometer, 1 megohm
R_2=10000 ohms, 0.5 watt
R_3=0.82 megohm, 0.5 watt
R_4=820 ohms, 0.5 watt
R_5=0.22 megohm, 0.5 watt
R_6, R_7=15000 ohm ±5 per cent, 2 watts
R_8=3900 ohms, 2 watts
R_9, R_{10}=0.1 megohm, 0.5 watt
R_{11}, R_{12}=1000 ohms, 0.5 watt
R_{13}, R_{14}=100 ohms, 0.5 watt
R_{15}=8200 ohms, 0.5 watt
R_{16}=15000 ohms, 1 watt
R_{17}=68000 ohms, 0.5 watt
R_{18}=4700 ohms, 2 watts
R_{19}=0.27 megohm, 1 watt
R_{20}=47000 ohms, 0.5 watt
R_{21}=Hum balance adjustment, potentiometer, 100 ohms, 0.5 watt
SR_1=Selenium rectifier, 20 mA, 135 volts rms
T_1=Output transformer, (having 8-ohm tap for feedback connection) for matching impedance of voice coil to 6600-ohm plate-to-plate tube load; 50 watts; frequency response, 10 to 50000 Hz; Stancor A-8056 or equiv.
T_2=Power transformer, 360-0-360 volts rms, 120 mA; 6.3 V., 3.5 A; 5 V., 3 A; Stancor 8410 or equiv. (see Note 1)

Notes: 1. For stereo operation from a single power supply, the power transformer T_2 must be replaced by one that has a higher current rating. A Stancor Type 6315 or equivalent (370-0-370 volt rms, 275 mA) is recommended.
2. If the amplifier oscillates or "motorboats," reverse ground and feedback connections in secondary of output transformer T_1.

CHAPTER 12

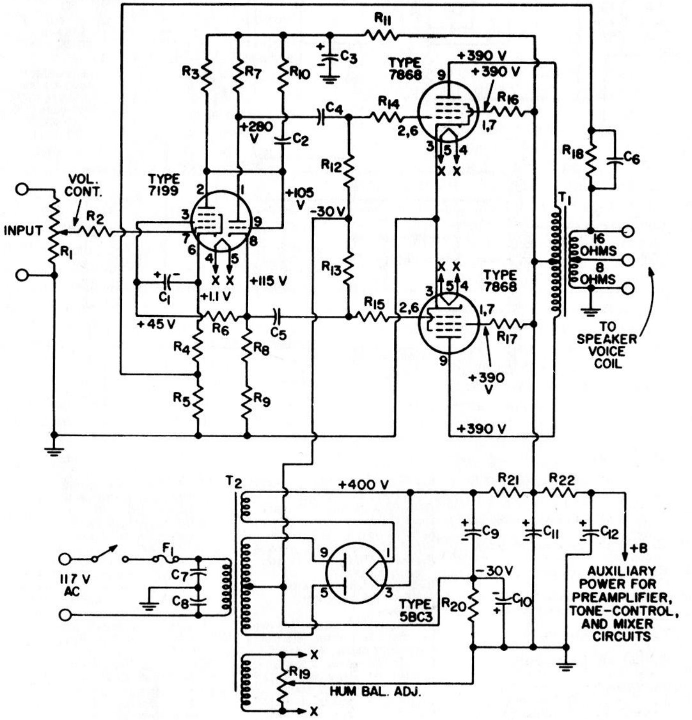
HIGH-FIDELITY AUDIO AMPLIFIER

Class AB_1; Power Output, 30 Watts

Circuit Description

This audio power amplifier can deliver 30 watts of rms output power with less than 0.7 per cent total harmonic distortion and less than 1.5 per cent intermodulation distortion. The frequency response of the amplifier is flat within ±0.5 dB from 15 Hz to 40 kHz. The total hum and noise, with the input shorted, is 85 dB below 30 watts. The rated output of 30 watts is obtained for an input of 1 volt rms.

The 30-watt amplifier is essentially identical to the 15-watt amplifier (circuit 26-9) except that it uses 7868 beam power tubes in the output stage to develop the higher audio power output and uses a resistive network in the negative leg of the power supply, rather than a separate rectifier, to supply the fixed-bias voltage for the output stage. A potentiometer (R_{19}) connected across the 6.3-volt heater winding also provides the hum balance adjustment for the 30-watt amplifier.



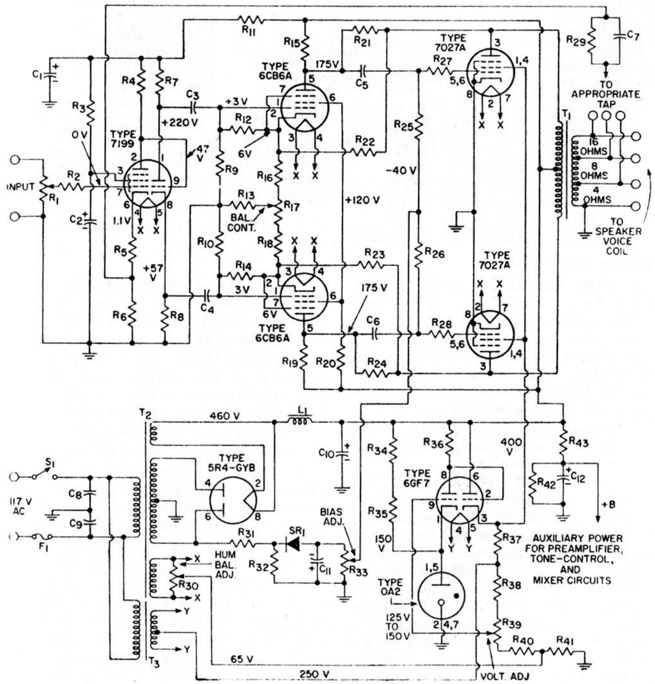
Parts List

C_1=25 μF, electrolytic, 50 V
C_2=22 pF, ceramic or mica, 600 V
C_3=80 μF, electrolytic, 450 V
C_4, C_5=0.25 μF, paper, 600 V
C_6=0.01 μF, paper, 600 V
C_7, C_8=0.05 μF, paper, 600 V
C_9, C_{11}=40 μF, electrolytic, 500 V
C_{10}=100 μF, electrolytic, 50 V
C_{12}=20 μF, electrolytic, 450 V
F_1=Fuse, 3 amperes, 150 V
R_1=Volume control, potentiometer, 1 megohm
R_2=10000 ohms, 0.5 watt
R_3=0.22 megohm, 0.5 watt
R_4=820 ohms, 0.5 watt
R_5=10 ohms, 0.5 watt
R_6=0.18 megohm, 0.5 watt
R_7, R_8=15000 ohms ±5 per cent, 2 watts
R_9=1000 ohms, 0.5 watt
R_{10}=22000 ohms, 0.5 watt
R_{11}=2000 ohms, 2 watts
R_{12}, R_{13}=0.1 megohm, 0.5 watt
R_{14}, R_{15}=1000 ohms, 0.5 watt
R_{16}, R_{17}=56 ohms, 0.5 watt
R_{18}=270 ohms, 0.5 watt
R_{19}=Hum balance adjustment, potentiometer, 100 ohms, 0.5 watt
R_{20}=220 ohms, 100 watts
R_{21}=50 ohms, 10 watts
R_{22}=10000 ohms, 2 watts
T_1=Output transformer (having 16-ohm tap for feedback connection) for matching impedance of voice coil to 6600-ohm plate-to-plate tube load; 50 watts; frequency response, 10 to 50000 Hz; Stancor A-8056 or equiv.
T_2=Power transformer, 375-0-375 volts rms, 160 mA; 6.3 V., 5 A; 5 V., 3 A; Thordarson type T22R33 or equivalent (see Note 1).

Notes: 1. For stereo operation from a single power supply, the power transformer T_2 must be replaced by one that has a higher current rating. A Stancor Type 6315 or equivalent (370-0-370 volts rms, 275 mA) is recommended.
2. If amplifier oscillates or "motorboats," reverse ground and feedback connections in secondary of output transformer T_1.

HIGH-FIDELITY AUDIO AMPLIFIER

Class AB_1; Power Output, 50 Watts



Preliminary Adjustments

The following adjustments should be made before operation:

(1) With rectifier out of socket, adjust Bias Adj. R_{33} for −40 volts between the wiper arm and ground bus.

(2) With speaker connected, adjust Screen-Grid Voltage Adj. R_{39} for 400 volts between pin 3 of 6GF7 and ground bus.

(3) With input shorted, adjust Hum Bal. Adj. R_{30} for minimum hum from speaker.

(4) With input open and Vol. Cont. set for maximum volume, adjust Bal. Cont. R_{17} for minimum hum from speaker.

Parts List

C_1, C_2=40 μF, electrolytic, 450 V
C_3, C_4=0.02 μF, paper, 400 V
C_5, C_6=1 μF, paper, 400 V
C_7=0.002 μF to 4-ohm tap; 0.0015 μF to 8-ohm tap; or, 0.001 μF to 16-ohm tap; paper, 400 V
C_8, C_9=0.05 μF, paper, 600 V
C_{10}=20 μF, electrolytic, 450 V
F_1=Fuse, 5 amperes
L_1=Choke, 8 H, 250 mA, dc resistance 60 ohms, or less
R_1=Volume control, potentiometer, 0.5 megohm
R_2=4700 ohms, 0.5 watt
R_3=0.82 megohm, 0.5 watt
R_4=0.22 megohm, 0.5 watt
R_5=820 ohms, 0.5 watt
R_6=10 ohms, 0.5 watt
R_7 R_8= 15000 ohms, 2 watts
R_9, R_{10}=1.5 megohms, 0.5 watt
R_{11}=33000 ohms, 2 watts
R_{12}, R_{14}=1.3 megohms, 0.5 watt
R_{13}=47 ohms, 0.5 watt
R_{15}, R_{19}=0.15 megohm, 0.5 watt
R_{16}, R_{18}=390 ohms, 0.5 watt
R_{17}=AC balance control, potentiometer, 500 ohms
R_{20}=0.15 megohm, 1 watt
R_{21}, R_{24}=0.33 megohm, 1 watt
R_{22}. R_{23}=0.12 megohm, 2 watts
R_{25}, R_{26}=0.1 megohm, 0.5 watt
R_{27}, R_{28}=47000 ohms, 0.5 watt
R_{29}=600 ohms to 4-ohm tap; 820 ohms to 8-ohm tap; or, 1200 ohms to 16-ohm tap; 0.5 watt
R_{30}=Hum balance adjustment, potentiometer, 100 ohms
R_{31}=0.12 megohm, 5 watts
R_{32}, R_{34}, R_{35}, R_{37}=33000 ohms, 2 watts
R_{33}=Bias adjustment, potentiometer 5000 ohms,
R_{38}=10000 ohms, 1 watt
R_{39}=Screen-grid voltage adjustment, potentiometer, 25000 ohms, 2 watts
R_{40}=15000 ohms, 2 watts
R_{41}=12000 ohms, 2 watts
R_{42}=0.22 megohm, 2 watts
R_{43}=22000 ohms, 2 watts
SR_1=Selenium rectifier, 20 mA, 135 volts rms
T_1=Output transformer for matching impedance of voice coil to 5000-ohm plate-to-plate tube load; 50 watts; frequency response, 10 to 50000 Hz; United Transformer Corp. LS55 or equiv.
T_2=Power transformer, 600-0-600 volts rms, 200 mA, 6.3 V., 5 A; 5 V., 3 A; Thordarson 22R36 or equiv. (see Note 1)
T_3=Filament transformer, 6.3 volts, center tapped, 1 ampere; Thordarson 21F08 or equiv.

Notes: 1. For stereo operation from a single power supply, the following changes are required: (a) The power transformer T_2 must be replaced by one that has a higher current rating; a United Transformer Corporation Type H-93 or equivalent (600-0-600 volts rms, 300 mA) is recommended. (b) The 5000-ohm Bias Adj. potentiometer R_{33} should be replaced by two 10000-ohm potentiometers (one for each channel) connected in parallel. (c) A second 5R4-GYB rectifier tube should be connected in parallel with the one used for monaural operation. (Connect the 5R4-GYB tubes so that the two sections of each tube are in parallel with the corresponding sections of the other tube; do not use separate tubes for each section of the rectifier circuit.)

2. If the amplifier oscillates or "motorboats," reverse ground and feedback connections in secondary of output transformer T_1.

Circuit Description

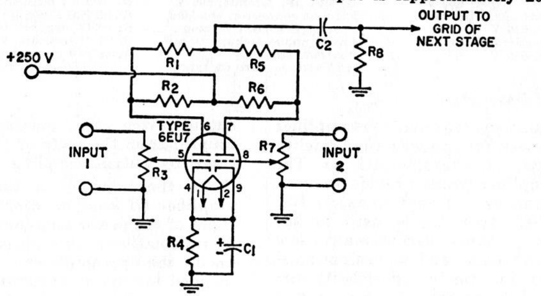
This four-stage audio power amplifier can deliver 50 watts of rms power output with less than 0.1 per cent total harmonic distortion and less than 1 per cent intermodulation distortion. The frequency response of the amplifier is flat within ±0.5 dB from 10 Hz to 50 kHz. Sensitivity is 0.4 volt rms input for 50 watts output. The total hum and noise is 70 dB below 50 watts.

The 50-watt amplifier, like the 15-watt and 30-watt high-fidelity amplifiers (circuits 26-9 and 26-10), uses a 7199 low-noise triode-pentode as an input amplifier and phase-splitter, but has a push-pull driver stage, which uses 6CB6 sharp-cutoff pentodes. The superior performance of this amplifier can also be attributed, in part, to the use of a 450-volt plate supply and a 400-volt electronically regulated grid-No. 2 supply for the 7027A beam power tubes in the output stage and to the use of inverse-feedback loops from the plates to the grids of the output tubes, from the plates of the output tubes to the cathodes of the driver tubes, and from the voice-coil winding of the output transformer to the cathode of the input amplifier. Additional features are the operation of all heaters at a positive voltage with respect to ground and use of a balancing adjustment (R_{30}) in the heater-supply circuit to minimize hum, a grid-No. 2 voltage adjustment (R_{39}), a grid-No. 1 bias adjustment (R_{33}) for the 7027A output tubes, and an ac-balance adjustment (R_{13}) which may be used to balance the outputs of the push-pull stages. Operation of the 50-watt amplifier is essentially the same as that of the 15- and 30-watt amplifiers.

CHAPTER 13

TWO-CHANNEL AUDIO MIXER

Voltage Gain from Each Grid of 6EU7 to Output is Approximately 20



Parts List

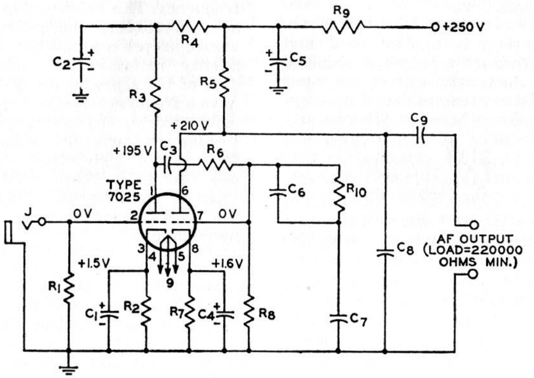
C_1=10 μF, electrolytic, 25 V
C_2=0.05 μF, paper, 400 V
R_1, R_5, R_8=1 megohm, 0.5 watt
R_2 R_6=0.1 megohm, 0.5 watt
R_3, R_7=Potentiometers, 0.1 megohm, audio taper
R_4=1200 ohms, 0.5 watt

Circuit Description

This high-fidelity mixer circuit can be used to combine audio-frequency program material from two sources. Each signal channel consists of a one-stage voltage amplifier using one section of a 6EU7 low-noise twin-triode. Each section of the mixer can provide a voltage gain of about 20, and can handle an input signal of about 0.2 volt rms without overloading. The dc plate supply of +250 volts (nominal value) for the mixer stages can usually be obtained from an auxiliary tap on the power supply for the audio power amplifiers.

PREAMPLIFIER FOR MAGNETIC PHONOGRAPH PICKUP

With RIAA Equalization



Sensitivity=3 millivolts rms input for output of 0.55 volt at frequency of 1000 Hz.

Parts List

C_1, C_4=25 μF, electrolytic, 25 V
C_2, C_5=20 μF, electrolytic, 450 V
C_3=0.1 μF, paper, 600 V
C_6=0.0033 μF±5 per cent, paper, 600 V
C_7=0.01 μF±5 per cent, paper, 600 V
C_8=180 pF±5 per cent, ceramic or mica, 500 V (includes capacitance of output cable)
C_9=0.22 μF, ceramic, 500 V
J=Input connector, shielded, for high-impedance magnetic phono pickup (10 mV output, approx.)
R_1=Value depends on type of magnetic pickup used. Follow pickup manufacturer's recommendations
R_2, R_7=2700 ohms, 0.5 watt
R_3, R_5=0.1 megohm, 0.5 watt
R_4=39000 ohms, 0.5 watt
R_6=0.47 megohm, 0.5 watt
R_8=0.68 megohm, 0.5 watt
R_9=15000 ohms, 1 watt
R_{10}=22000 ohms, 0.5 watt

Circuit Description

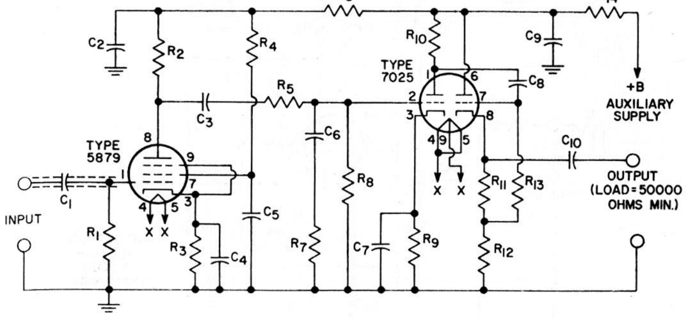
This two-stage audio preamplifier is intended for use with high-fidelity magnetic phonograph pickups. The two amplifier stages provide an over-all circuit gain of approximately 150. The 7025 twin triode used in the circuit features exceptionally low hum and noise and is designed especially for use in high-fidelity circuits that operate at low signal levels. The preamplifier is ideally suited for use as the low-level input stage for audio power amplifiers such as the 50-watt unit, circuit 27-11. For use with audio power amplifiers such as the 15- and 30-watt units, circuits 27-9 and 27-10, which require higher input signals, another low-level amplifier (e.g., the tone-control amplifier, circuit 27-20) must be inserted between the preamplifier and the power amplifier to obtain the full rated output. The heater and dc operating power required for the preamplifier can usually be obtained from the power-supply circuit for the power amplifier.

The audio signal from the phonograph pickup is applied to J and coupled through a length of shielded cable to the control grid of the input stage of the preamplifier. The interstage coupling between the two amplifier sections of the preamplifier includes an RIAA equalization network (R_{10} and C_6). This network compensates for the Orthophonic recording characteristic* introduced into a record disc by the manufacturer. The output from the preamplifier is coupled from the plate of the second stage by output coupling capacitor C_9 to the input of a tone-control amplifier (if used) or directly to the input of the power amplifier. Because of its relatively high output impedance, the preamplifier is recommended for use in systems in which the preamplifier is mounted on the same chassis as the power amplifier and/or tone-control amplifier. The preamplifier may be used at distances up to 6 feet from the following amplifier provided that the capacitance of capacitor C_8 is reduced approximately 30 picofarads for each foot of shielded cable used for the audio-frequency connection between the preamplifier and the following amplifier.

* To achieve wide frequency and dynamic ranges, manufacturers of commercial recordings use equipment which introduces a non-uniform relationship between amplitude and frequency. This relationship is known as a "recording characteristic." To assure proper reproduction of a high-fidelity recording, therefore, some part of the reproducing system must have a frequency-response characteristic which is the inverse of the recording characteristic. Most manufacturers of high-fidelity recordings use the RCA "New Orthophonic" (RIAA) characteristic for discs and the NARTB characteristic for magnetic tape.

HIGH-FIDELITY PREAMPLIFIER FOR TAPE-HEAD PICKUP

With NARTB Equalization



Sensitivity=3 millivolts rms input for output of 0.55 volt at frequency of 1000 Hz.

Parts List

C_1=0.047 μF, ceramic, 400 V
C_2=40 μF, electrolytic, 450 V
C_3=0.1 μF, ceramic, 400 V
C_4=25 μF, electrolytic, 25 V
C_5=0.22 μF, ceramic, 400 V
C_6=0.015 μF, ceramic, 400 V
C_7=25 μF, electrolytic, 25 V
C_8=0.22 μF, ceramic, 400 V
C_9=40 μF, electrolytic, 450 V
C_{10}=0.47 μF, ceramic, 400 V
R_1=1 megohm, 0.5 watt
R_2=0.1 megohm, 0.5 watt
R_3=1000 ohms, 0.5 watt
R_4=0.47 megohm, 0.5 watt
R_5=0.22 megohm, 0.5 watt
R_6=22000 ohms, 0.5 watt
R_7=3300 ohms, 0.5 watt
R_8=3.3 megohms, 0.5 watt
R_9=1500 ohms, 0.5 watt
R_{10}=0.1 megohm, 0.5 watt
R_{11}=1500 ohms, 0.5 watt
R_{12}=15000 ohms, 0.5 watt
R_{13}=0.47 megohm, 0.5 watt
R_{14}=4700 ohms, 0.5 watt

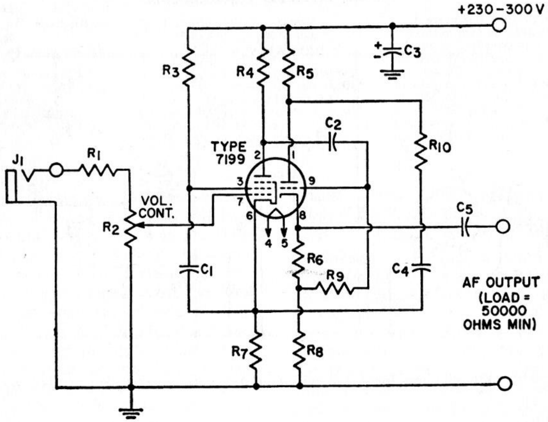
Circuit Description

This three-stage preamplifier provides the amplification necessary to increase the output from a tape-head pickup to the level required to drive an audio power amplifier. The circuit uses a 5879 low-noise sharp-cutoff pentode in a high-gain input voltage amplifier, one section of a 7025 twin triode in a second voltage amplifier, and the other section of the 7025 in a cathode-follower output stage. Because of the low-impedance cathode-follower output circuit, the preamplifier may be installed at distances up to 50 feet from the following stage (tone-control or power amplifier) without adverse effect upon its frequency-response characteristics. The preamplifier is intended for use as the low-level input stages for an audio power amplifier, such as the 50-watt unit (circuit 27-11) or, when followed by another low-level amplifier (e.g., the tone-control amplifier, circuit 27-20), the 15- or 30-watt unit (circuit 27-9 or 27-10). The heater and dc operating power for the preamplifier can usually be obtained from the power supply for the power amplifier.

The preamplifier provides an over-all circuit gain of 180. An input of 3 millivolts rms at the input terminals, is amplified by the pentode and triode voltage amplifiers to develop an output of approximately 0.55 volt rms at the cathode of the cathode-follower output stage. The interstage coupling between the pentode and triode voltage amplifiers equalizes the playback frequency response of the preamplifier to compensate for the NARTB recording characteristic introduced into the magnetic tape by the manufacturer. (See footnote for circuit 27-16.) The output of the preamplifier is coupled by capacitor C_{10} to the input of the audio power amplifier or to the input of an intermediate tone-control amplifier.

PREAMPLIFIER FOR CERAMIC PHONOGRAPH PICKUP

Cathode Follower (Low-Impedance) Output



Parts List

C_1=0.1 μF, paper, 400 V
C_2=0.01 μF, paper, 400 V
C_3=20 μF, electrolytic, 400 V
C_4=0.25 μF, paper, 400 V
C_5=0.22 μF, paper, 600 V
J_1=Input connector, shielded, for high-impedance ceramic phono pickup (0.5-volt output)
R_1=1.8 megohms, 0.5 watt
R_2=Volume control, potentiometer, 0.5 megohm, audio taper
R_3=0.82 megohm, 0.5 watt
R_4=0.22 megohm, 0.5 watt
R_5, R_8=4,000 ohms, 0.5 watt
R_6=4700 ohms, 0.5 watt
R_7=1000 ohms, 0.5 watt
R_9=1 megohm, 0.5 watt
R_{10}=1800 ohms, 0.5 watt

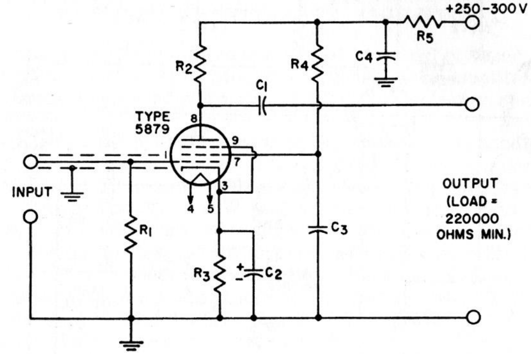
Circuit Description

This two-stage preamplifier is intended for use with a high-impedance ceramic phonograph pickup. The circuit features a cathode-follower (low-impedance) output which makes it possible to install the preamplifier at distances up to 50 feet from the succeeding stage (tone-control or power amplifier). The preamplifier operates from a dc supply of 230 to 300 volts and a heater supply of 6.3 volts. These voltages can usually be obtained from the power supply for the power amplifier in the audio system.

The preamplifier uses a 7199 triode-pentode in a high-gain pentode input stage and a triode cathode-follower output stage. These stages provide the amplification necessary to increase the output from a crystal phonograph pickup, applied at J_1, to the level required to drive an audio power amplifier. The output of the preamplifier, coupled from the cathode of the 7199 triode section, may be applied directly to the power amplifier, or to an intermediate tone-control amplifier.

LOW-DISTORTION PREAMPLIFIER

For Low-Output, High-Impedance Microphones



Sensitivity=3 millivolts rms input for output of 220 millivolts.

Parts List

C_1=0.047 μF, paper, 400 V
C_2=25 μF, electrolytic, 25 V
C_3=0.22 μF, paper, 400 V
C_4=40 μF, electrolytic, 450 V
R_1=2.2 megohms, 0.5 watt
R_2=0.1 megohm, 0.5 watt
R_3=1000 ohms, 0.5 watt
R_4=0.47 megohm, 0.5 watt
R_5=22000 ohms, 0.5 watt

Circuit Description

This single-stage preamplifier is intended for use with a high-fidelity, high-impedance crystal or dynamic microphone. The circuit uses a 5879 low-noise sharp-cutoff pentode in a conventional amplifier circuit that has a high-impedance output, a voltage gain of approximately 70, and a flat frequency response over the audio range. Because of its high output impedance, the preamplifier should be mounted on the same chassis as the power amplifier and tone-control amplifier (if used). Heater and dc power for the circuit can be obtained from the power supply for the audio power amplifier.

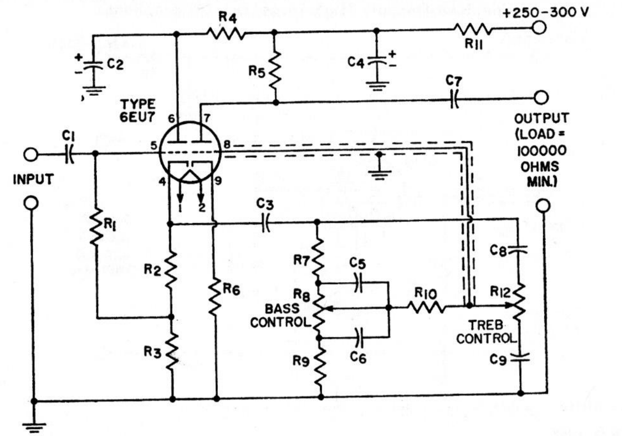
BASS AND TREBLE TONE-CONTROL AMPLIFIER

Parts List

C_1=0.047 μF, paper, 400 V
C_2, C_4=20 μF, electrolytic, 450 V
C_3=0.1 μF, paper, 400 V
C_5=0.0022 μF, paper, 400 V
C_6=0.022 μF, paper, 400 V
C_7=0.22 μF, paper, 400 V
C_8=220 pF, ceramic or mica, 500 V
C_9=0.0022 μF, paper, 400 V
R_1=0.47 megohm, 0.5 watt
R_2=1500 ohms, 0.5 watt
R_3, R_{11}=15000 ohms, 0.5 watt
R_4=22000 ohms, 0.5 watt
R_5, R_7, R_{10}=0.1 megohm, 0.5 watt
R_6=1000 ohms, 0.5 watt
R_8=Bass control, potentiometer, 1 megohm, audio taper
R_9=10000 ohms, 0.5 watt
R_{12}=Treble control, potentiometer, 1 megohm, audio taper

Circuit Description

This high-fidelity tone-control amplifier uses a 6EU7 low-noise twin triode in a two-stage amplifier cascade that consists of an input cathode follower connected to a triode voltage amplifier through a frequency-sensitive (tone-control) interstage coupling network. The bass and treble controls in the coupling network can be adjusted to provide up to 16 dB of boost or attenuation (cut) at 30 Hz and at 15 kHz. With the bass and treble controls set at the mid-range positions, the amplifier provides an



Sensitivity=0.5 volt rms input for output of 1.25 volts with controls set for flat response.

over-all voltage gain of approximately 2.5, and its frequency response is flat within ±1dB from 30 Hz to 15 kHz.

The tone-control amplifier is designed for use immediately ahead of an audio power amplifier, such as the 15-, 30-, or 50-watt unit (circuit 27-9, 27-10, or 27-11, respectively). Operating power for the tone-control circuit can usually be obtained from the power supply for the power amplifier. For operating convenience, the volume control on the power amplifier may be physically located on the tone-control chassis. In this case, it is advisable to insert a 1-megohm resistor in place of the volume control on the power amplifier. If partial compensation for the reduced high- and low-frequency sensitivity of the ear at low volume levels is desired, the volume-control potentiometer may be replaced by a loudness-control potentiometer.

CHAPTER 14

MULLARD TWENTY-WATT AMPLIFIER

The circuit is designed to give the highest standard of sound reproduction when used in association with a suitable pre-amplifier, a high-grade pick-up head and a good-quality loud-speaker system.

Two Mullard output pentodes, type EL34, rated at 25W anode dissipation, form the output stage of the circuit. These are connected in a push-pull arrangement with distributed loading, and give a reserve of output power of 20W with a level of harmonic distortion less than 0.05%. The intermediate stage consists of a cathode-coupled, phase-splitting amplifier using the Mullard double triode, type ECC83. This stage is preceded by a high-gain voltage amplifier incorporating the Mullard low-noise pentode, type EF86. Direct coupling is used between the voltage amplifier and phase splitter to minimise low-frequency phase shifts.

The main feedback loop includes the whole circuit, the feedback voltage being derived from the secondary winding of the output transformer and being injected in the cathode circuit of the EF86. The amount of feedback applied around the circuit is 30dB, but in spite of this high level, the stability of the circuit is good and the sensitivity is 220mV for the rated output power. The level of hum and noise is 89dB below the rated 20W.

The rectifier used in the power-supply stage is the Mullard full-wave rectifier, type GZ34. This provides sufficient current for the amplifier (about 145mA) and also for a pre-amplifier and radio tuner if (about 40mA) required.

CIRCUIT DESCRIPTION

Input Stage

The EF86 input stage of the circuit of Fig. 1 provides high-gain voltage amplification, the stage gain being approximately 120 times. High-stability, cracked-carbon resistors are used in the anode, screen-grid and cathode circuits, and they give an appreciable improvement in the measured level of background noise compared with ordinary carbon resistors.

The stage is coupled directly to the input of the phase splitter. The purpose of this is to minimise low-frequency phase shift in the amplifier and to improve the low-frequency stability when negative feedback is applied. A CR network (C1, R3) connected across the anode load produces an advance in phase and thus improves the high-frequency stability of the amplifier.

Intermediate Stage

The second stage of the circuit uses a Mullard double triode, type ECC83, and fulfils the combined function of phase splitter and driver

20W AMPLIFIER

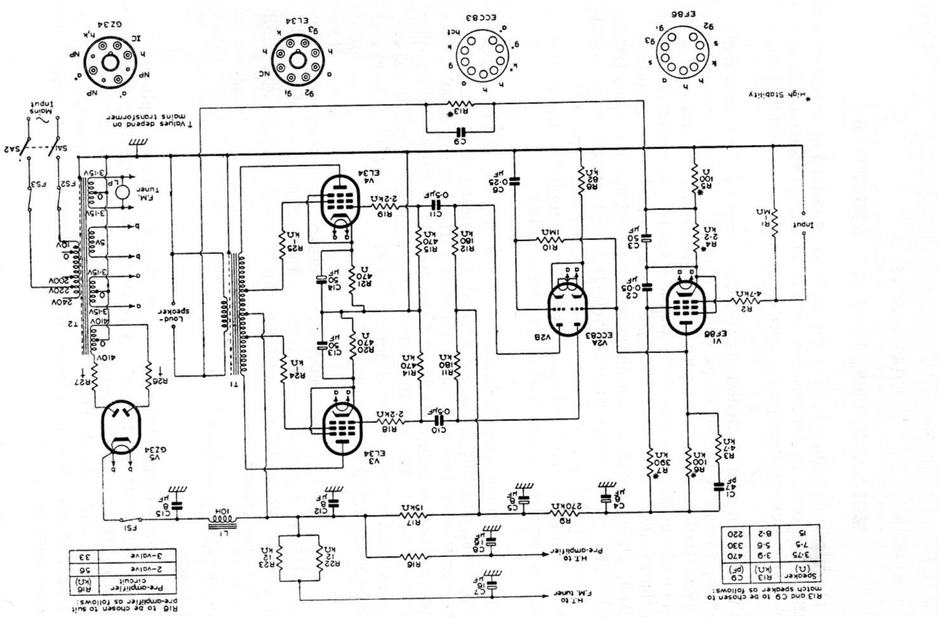


Fig. 1—Circuit diagram of 20W amplifier

LIST OF COMPONENTS *20W AMPLIFIER*

Resistors

Circuit ref.	Value	Tolerance (±%)	Rating (W)
R1	1 MΩ	20	¼
R2	4·7kΩ	20	¼
R3	4·7kΩ	10	¼
[1]R4	2·2kΩ	10	½
[1]R5	100 Ω	5	½
[1]R6	100 kΩ	10	½
[1]R7	390 kΩ	10	½
R8	82 kΩ	10	¼
R9	270 kΩ	10	½
R10	1 MΩ	20	¼
[2]R11	180 kΩ	10	½
[2]R12	180 kΩ	10	½
[1]R13 for 3·75Ω speaker	3·9kΩ	5	½
for 7·5Ω speaker	5·6kΩ	5	½
for 15Ω speaker	8·2kΩ	5	½
[3]R14	470 kΩ	10	¼
[3]R15	470 kΩ	10	¼
R16 for 2-valve pre-amp.	56 kΩ	10	1
for 3-valve pre-amp.	33 kΩ	10	1
R17	15 kΩ	20	½
R18	2·2kΩ	20	¼
R19	2·2kΩ	20	¼
R20	470 Ω	5	3
R21	470 Ω	5	3
R22	12 kΩ	20	6
R23	12 kΩ	20	6
R24	1 kΩ	10	½
R25	1 kΩ	10	½
R26 and R27	Values depend on mains transformer		6

1. High stability, cracked carbon
2. Matched to within 5%
3. Preferably matched to within 5%
4. Wire wound

Capacitors

Circuit ref.	Value	Description	Rating (V)
C1	47 pF	silvered mica[5]	
C2	0·05μF	paper	350
C3	50 μF	electrolytic	12
C4	8 μF	electrolytic	450
C5	8 μF	electrolytic	450
C6	0·25μF	paper	350
C7	16 μF	electrolytic	450
C8	8 μF	electrolytic	500
C9 for 3·75Ω speaker	470 pF	silvered mica[6]	
for 7·5Ω speaker	330 pF	silvered mica[6]	
for 15Ω speaker	220 pF	silvered mica[6]	
C10	0·5 μF	paper	350
C11	0·5 μF	paper	350
C12	8 μF	paper	500
C13	50 μF	electrolytic	50
C14	50 μF	electrolytic	50
C15	8 μF	paper	500

5. Tolerance, ±10%
6. Tolerance, ±5%

Valves

Mullard EF86, ECC83, EL34 (two), GZ34

Valveholders

B9A (noval) nylon-loaded, with screening skirt (for EF86) McMurdo, XM9/AU, Skirt 95
B9A (noval) nylon-loaded (for ECC83), McMurdo XM9/AU
B8-O (International octal) (three, for EL34s and GZ34), McMurdo B8/U

20W AMPLIFIER

Miscellaneous

Mains input plug, 3-way. Bulgin, P340
Mains switch, 2-pole. Bulgin, S300
Mains selector. Clix, CTSP/2
H.T. supply socket (f.m. tuner), 4-way. Elcom, S.04
H.T. supply socket (pre-amplifier), 6-way. Elcom, S.06
Fuseholders (three). Belling Lee, L356

Fuses, 2A (two): 250mA (one)
Lampholder. Bulgin, D180/Red
Pilot lamp. 6.3V, 40mA
Input socket, coaxial. Belling Lee, L.734/S
Output plug, 2-pin. Bulgin, P350
Tagboard (10-way) (two). Bulgin, C114

Output Transformer T1

Primary Impedance:
7kΩ for 20% screen-grid taps
6·6kΩ for 43% screen-grid taps

Mains Transformer T2

Primary: 10–0–200–220–240V
Secondaries: H.T., 410–0–410V, 180mA
3·15–0–3·15V, 4A
3·15–0–3·15V, 2·5A
0–5V, 3A

Smoothing Choke L1

Inductance: 10H at 180mA
Resistance: 200Ω

Commercial Components

Manufacturer	Output Transformer Type No. 20% Taps	Output Transformer Type No. 43% Taps	Mains Transformer Type No.	Choke Type No.
Colne	03070	03069	03068	03071
Elden	486A	486	477	478
Gardners	AS.7034	AS.7034	RS.3175	CS.5142
Gilson	W.O.1342	W.O.866	W.O.775	W.O.1340
			W.O.917	W.O.1341
Hinchley	1532	1377	1441	1528
Parmeko	P2913	P2647	P2646	P463
Partridge	—	P3878	P3877	C10/180
	—	P6878	P6877	—
Savage	—	4B14	4B32-1	—
Wynall	W1900C	W1552C	W1584	W1585

amplifier. The phase splitter is a cathode-coupled circuit which enables a high degree of balance to be obtained in the push-pull drive signal applied to the output stage.

With the high line voltage available, the required drive voltage for an output power of 20W is obtained with a low level (0.4%) of distortion. The anode load resistors R11 and R12 should be matched to within 5%, R12 having the higher value for optimum operation. Optimum balance is obtained when the effective anode loads differ by 3%. The grid resistors R14 and R15 of the output stage should also be close-tolerance components because they also form part of the anode load of the driver stage. High-frequency balance will be determined largely by the wiring layout because equality of shunt capacitances is required. Low-frequency balance is controlled by the time constant C6R10 in the grid circuits of the triode sections, and the time constant chosen in Fig. 1 will give adequate balance down to very low frequencies.

A disadvantage of the cathode-coupled form of phase splitter is that the effective voltage gain is about half that attainable with one section of the valve used as a normal voltage amplifier. However, as the mutual conductance of the ECC83 is high (100), the effective gain of the cathode-coupled circuit is still about 25 times.

Output Stage

The main feature of interest in the output stage is the use of two EL34s with partial screen-grid (or distributed) loading, the screen grids being fed from tappings on the primary winding of the output transformer. The best practical operating conditions are achieved with this type of output stage when about 20% of the primary winding is common to the anode and screen-grid circuit.

The anode-to-anode loading of the output stage is 7k ohms and, with a line voltage of 440V at the centre-tap of the primary winding of the output transformer, the combined anode and screen-grid dissipation of the output valves is 28W per valve. With the particular screen-grid-to-anode load ratio used, it has been found that improved linearity is obtained at power levels above 15W when resistors of the order of 1k ohm are inserted in the screen-grid supply circuits. The slight reduction in peak power-handling capacity which results is not significant in practice.

Separate cathode-biasing resistors are used in the output stage to limit the out-of-balance direct current in the primary winding of the output transformer. The use of other balancing arrangements has not been thought necessary although it is likely that some improvement in performance, particularly at low frequencies, would result from the use of d.c. balancing. It is necessary in this type of output stage for the cathodes to be bypassed to earth even if a shared cathode resistor is used. Consequently, a low-frequency time constant in the cathode circuit cannot be eliminated when automatic biasing is used.

NEGATIVE FEEDBACK

Negative feedback is taken from the secondary winding of the output transformer to the cathode circuit of the input stage. In spite of the high level of feedback used (30dB), the circuit is completely stable under open-circuit conditions. At least 10dB more feedback (obtainable by reducing the value of R13) would be required to cause high-frequency instability. The most probable form of instability would be oscillation with capacitive loads, but this is most unlikely to occur even with very long loudspeaker leads.

POWER SUPPLY

The power supply is conventional and uses a Mullard indirectly-heated, full-wave rectifier, type GZ34, in conjunction with a capacitive input filter. The values of the limiting resistors R26 and R27 will depend on the winding resistances of the mains transformer used.

Their purpose, when required, is normally one of voltage control only. Where a transformer with a very low winding resistance is used, a secondary voltage rated at 400-0-400V may be found adequate. The rating of the mains transformer is such that about 40mA may be drawn from the h.t. supply to feed a pre-amplifier circuit and tuner in addition to the normal current required (about 145mA) for the amplifier.

Extra decoupling will be needed for these ancillary supplies. The smoothing components R22, R23 and C7 can only be chosen when the type of tuner to be used is known. The values given in Fig. 1 would be suitable for typical current and voltage requirements of approximately 40mA and 200V. The components R16 and C8 depend on the pre-amplifier to be used. The values given in Fig. 1 refer to the 2- and 3-valve pre-amplifiers described in the next chapter.

PERFORMANCE

At the full rated output, the distortion without feedback is well below 1% and with feedback is below 0.05%. The distortion rises to 0.1% for an output power of 27W. The loop-gain characteristic is such that a level of at least 20dB of feedback is maintained from 15c/s to 25c/s, and of at least 26dB down to 30c/s.

Measurements of intermodulation products were made in the prototype amplifier using a carrier frequency of 10kc/s and a modulating frequency of 40c/s. The ratio of modulating amplitude to carrier amplitude was 4 : 1. With the combined peak amplitudes of the mixed output at a level equivalent to the peak sine-wave amplitude at an r.m.s. power of 20W, the intermodulation products, expressed in r.m.s. terms, totalled 0.7% of the carrier amplitude. At an equivalent power of 27W, the products totalled 1% of the carrier amplitude. The beat-note distortion between equal-amplitude signals at frequencies of 14 and 15kc/s is 0.25 and 0.3% at equivalent powers of 20 and 27W respectively, and between frequencies of 9 and 10kc/s it is 0.2 and 0.25% at the same equivalent powers.

FREQUENCY RESPONSE

The frequency- and power-response characteristics show that, at an output of 1W, the characteristic is flat (±1dB) compared with the level at 1kc/s from 2c/s to 100kc/s and the power response characteristic is flat (±0.5dB) from 30c/s to 20kc/s.

It is important that adequate power-handling capacity is available at the low-frequency end of the audible range and this is determined chiefly by the characteristics of the output transformer.

SENSITIVITY

The sensitivity of the amplifier measured at 1kc/s is 6.5mV for an output of 20W when no feedback is applied, and approximately 220mV with feedback, the loop gain being 30dB. The sensitivity, with feedback, at the overload point (27W) is approximately 300mV.

The level of background noise and hum is 89dB below 20W, measured with a source resistance of 10k ohms. This is equivalent to a signal of about 5.5µV at the input terminals. It is possible to increase the overall sensitivity of the amplifier by 6dB while still maintaining a low background level, a high loop gain and a good margin of stability, but various design requirements of associated pre-amplifier circuits (the need for a high signal-to-noise ratio, for example) render a higher sensitivity a doubtful advantage.

PHASE SHIFT AND TRANSIENT RESPONSE

Emphasis has been laid in the amplifier on a good margin of stability and, consequently, the phase shift is held to a comparatively low level. The shift is only 20° at 20kc/s. Excellent response to transient signals is obtained, the rise time of the amplifier being of the order of 5µ sec.

OUTPUT IMPEDANCE

The output stage has a low inherent output impedance, and this is further lowered by the use of negative feedback. It is approximately 0.3 ohms with a 15 ohms load for an output of 20W at frequencies of 40c/s, 1kc/s and 20kc/s. This gives a damping factor of about 50.

CHAPTER 15

MULLARD TWO- AND THREE-VALVE PRE-AMPLIFIERS

The circuits described in this chapter have been designed principally for use with the 20 watt amplifier previously described. They can, however, be used with any power amplifier which does not require an input signal greater than 250mV for full output. Input facilities are provided for magnetic and crystal pick-up heads, tape-recorder playback heads, radio tuner units and (in the 2 valve circuit) a microphone. An auxiliary input position is provided for ancillary equipment.

An additional output position, which is independent of the tone controls, is also provided with both circuits, enabling programmes to be taken to a tape amplifier while they are being fed from the normal output position to a power amplifier. Both the auxiliary input and the additional output positions have jack sockets which are situated at the front of the chassis.

All the input sockets in each circuit are connected to one switch which selects one input at a time. In both circuits this switch also short-circuits the unused sockets to earth, an arrangement which considerably reduces the amount of 'break-through' between channels. The positions of the switch, from left to right, are: Auxiliary, Radio, Tape, Microphone (in the 2 valve circuit only) Microgroove and 78 r.p.m.

The equalisation for disc recordings conforms to the present R.I.A.A. characteristics which has been adopted by most of the major recording companies. The tape-playback characteristic is intended for use with high-impedance heads when replaying pre-recorded tapes at a speed of $7\frac{1}{2}$ inches per second.

The tone controls used in both circuits cover a wide range of frequencies and provide boost and cut for high and low frequencies. Switched high- and low-pass filter networks are included in the 3-valve circuit so that unwanted signals such as rumble and record scratch can be eliminated.

CIRCUIT DESCRIPTIONS

Two Mullard high-gain pentodes, type EF86, are used in the 2-valve circuit of Fig. 2. In the 3-valve circuit shown in Fig. 3, two EF86s and a Mullard double triode, type ECC83, are used.

First Stage of 2- and 3-Valve Circuits

All the equalisation takes place in the first stage of both circuits, and is achieved by means of frequency-selective feedback between the anode and grid of the first EF86. This arrangement has been chosen because the grid-circuit impedance of the first stage should be low. A low

impedance at this grid lessens hum pick-up and reduces the effect of plugging-in external low-impedance circuits. Furthermore, the arrangement also results in low gain in the first stage. Hence, Miller effect between the anode and grid of the first EF86, which can be troublesome when high values of resistance are used in series with the grid, is reduced.

Series resistors are used in the input circuits so that the sensitivity and impedance of any channel can be adjusted accurately. The component values given in Figs. 2 and 3 are intended for sources encountered most frequently, but the sensitivity and impedance (the impedance of the input channels includes the grid impedance of the EF86 modified by the feedback components as well as the impedance of the input network) of each channel can be altered by changing the value of the appropriate series resistor.

Second Stage of 2-Valve Circuit

No feedback is used in the second stage of the 2-valve pre-amplifier. The full output from the anode of the second EF86 is taken to the passive tone-control network and an auxiliary output is taken from the junction of R20 and R21 to the programme-recording jack socket.

Second Stage of 3-Valve Circuit

The second stage of the 3-valve pre-amplifier circuit of Fig. 3 has a linear frequency-response characteristic and its gain is reduced by a small amount of negative feedback applied at the control grid by way of the resistor R23. The output of this stage is taken from the anode to the tone-control network, and also to the programme-recording output socket. Because of the negative feedback, the tone controls have very little effect on the frequency response at this anode.

*Fig. 1—Output-attenuating networks for use with **both** pre-amplifiers*

Third Stage of 3-Valve Circuit

The output stage of the 3-valve pre-amplifier is made up of the filter circuits. The low-pass filter is situated between the two sections of the ECC83, and is arranged on the switch SB. It incorporates a Mullard wound pot-core inductor, type WF1428, in a pi-type network. The high-pass filter is arranged on the switch SC and consists of two RC networks in a feedback loop around the second triode section of the valve.

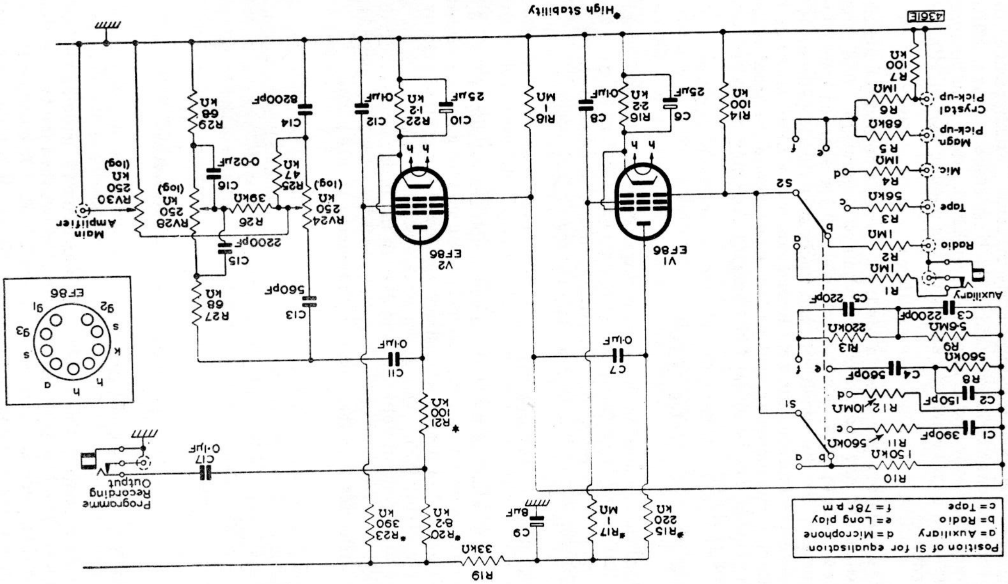


Fig. 2—*Circuit diagram of* ***two-valve*** *pre-amplifier*

LIST OF COMPONENTS FOR TWO-VALVE PRE-AMPLIFIER

Resistors

Circuit ref.	Value	Tolerance (±%)	Rating (W)
R1	1 MΩ	10	¼
R2	1 MΩ	10	¼
R3	56 kΩ	10	¼
R4	1 MΩ	10	¼
R5	68 kΩ	10	¼
R6	1 MΩ	10	¼
R7	100 kΩ	10	¼
R8	560 kΩ	10	¼
R9	5·6MΩ	10	¼
R10	150 kΩ	10	¼
R11	560 kΩ	10	¼
R12	10 MΩ	10	¼
R13	220 kΩ	10	¼
R14	100 kΩ	10	¼
[1]R15	220 kΩ	10	½
R16	2·2kΩ	10	½
[1]R17	1 MΩ	10	½
R18	1 MΩ	10	½
R19	33 kΩ	10	½
[1]R20	8·2kΩ	10	½
[1]R21	100 kΩ	10	½
R22	1·2kΩ	10	½
[1]R23	390 kΩ	10	½
RV24	250 kΩ logarithmic potentiometer		
R25	47 kΩ	10	¼
R26	39 kΩ	10	¼
R27	68 kΩ	10	¼
RV28	250 kΩ logarithmic potentiometer		
R29	6·8kΩ	10	¼
RV30	250 kΩ logarithmic potentiometer		
R48 (Fig. 1a)	680 kΩ	10	¼
R48 (Fig. 1b)	270 kΩ	10	¼
R49 (Fig. 1a)	180 kΩ	10	¼
R49 (Fig. 1b)	150 kΩ	10	¼

1. High stability, cracked carbon

Valves

Mullard EF86 (two)

Capacitors

Circuit ref.	Value	Description	Rating (V)
C1	390 pF	silvered mica	
C2	150 pF	silvered mica	
C3	2200 pF	silvered mica	
C4	560 pF	silvered mica	
C5	220 pF	silvered mica	
C6	25 μF	electrolytic	12
C7	0·1 μF	paper	350
C8	0·1 μF	paper	350
C9	8 μF	electrolytic	350
C10	25 μF	electrolytic	12
C11	0·1 μF	paper	350
C12	0·1 μF	paper	350
C13	560 pF	silvered mica	
C14	8200 pF	silvered mica	
C15	2200 pF	silvered mica	
C16	0·02μF	paper	
C17	0·1 μF	paper	350

Tolerance of silvered mica capacitors is 10%

Valveholders

B9A (noval) nylon-loaded, with screening skirt.
McMurdo XM9/AU, skirt 95 (two).

Miscellaneous

Supply input socket, 6-pin. Bulgin, P194
Input socket (six), coaxial. Belling Lee, L.604/S
Input jack. Igranic, P71
Output socket, coaxial. Belling Lee, L.604/S
Output jack. Igranic, P71
Selector switch, 6-way rotary.
Shirley Laboratories, SBL M/S/190
Specialist Switches, SS/592
(Note: Details of proprietary switches may not be identical with those shown in the diagram.)
Tagboard, 10-way (two). Bulgin, C125; Denco

LIST OF COMPONENTS FOR THREE-VALVE PRE-AMPLIFIER

Resistors

Circuit ref.	Value	Tolerance (±%)	Rating (W)
R1	1 MΩ	10	¼
R2	1 MΩ	10	¼
R3	56 kΩ	10	¼
R4	100 kΩ	10	¼
R5	2·2MΩ	10	¼
R6	100 kΩ	10	¼
R7	8·2MΩ	10	¼
R8	2·2MΩ	10	¼
R9	100 kΩ	10	¼
R10	390 kΩ	10	¼
R11	270 kΩ	10	¼
R12	180 kΩ	10	¼
R13	100 kΩ	10	¼
[1]R14	270 kΩ	10	½
R15	3·9kΩ	10	½
[1]R16	1·5MΩ	10	½
R17	33 kΩ	10	½
R18	220 kΩ	10	¼
R19	1 MΩ	10	¼
[1]R20	100 kΩ	10	½
R21	1·2kΩ	10	½
[1]R22	470 kΩ	10	½
R23	3·9MΩ	10	¼
RV24	250 kΩ logarithmic potentiometer		
R25	47 kΩ	10	¼
R26	39 kΩ	10	¼
R27	68 kΩ	10	¼
RV28	250 kΩ logarithmic potentiometer		
R29	6·8kΩ	10	¼
R30	270 kΩ	10	¼
R31	33 kΩ	10	¼
R32	22 kΩ	10	¼
R33	1·2kΩ	10	½
R34	10 kΩ	10	¼
R35	22 kΩ	10	¼
R36	12 kΩ	10	½
R37	270 kΩ	10	¼
R38	33 kΩ	10	¼
R39	47 kΩ	10	¼
R40	56 kΩ	10	¼
R41	56 kΩ	10	¼
R42	1·5MΩ	10	¼
R43	220 kΩ	10	¼
R44	47 kΩ	10	½
R45	1·5kΩ	10	½
R46	220 kΩ	10	¼
RV47	50 kΩ logarithmic potentiometer		
R48 (Fig. 1a)	680 kΩ	10	¼
R48 (Fig. 1b)	270 kΩ	10	¼
R49 (Fig. 1a)	180 kΩ	10	¼
R49 (Fig. 1b)	150 kΩ	10	¼

1. High stability, cracked carbon

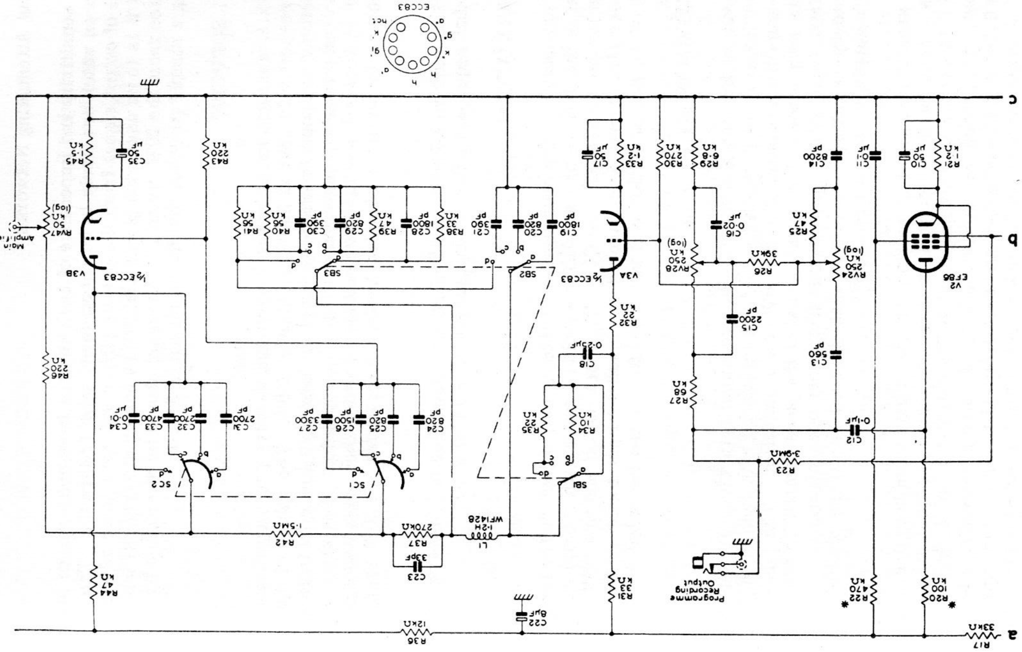
LIST OF COMPONENTS FOR THREE-VALVE PRE-AMPLIFIER

Capacitors

Circuit ref.	Value	Description	Rating (V)	Circuit ref.	Value	Description	Rating (V)
C1	330 pF	silvered mica		C18	0·25 μF	paper	350
C2	820 pF	silvered mica		C19	1800 pF	silvered mica	
C3	2700 pF	silvered mica		C20	820 pF	silvered mica	
C4	330 pF	silvered mica		C21	390 pF	silvered mica	
C5	330 pF	silvered mica		C22	8 μF	electrolytic	350
C6	50 μF	electrolytic	12	C23	33 pF	silvered mica	
C7	0·1 μF	paper	350	C24	820 pF	silvered mica	
C8	8 μF	electrolytic	350	C25	820 pF	silvered mica	
C9	0·1 μF	paper	350	C26	1500 pF	silvered mica	
C10	50 μF	electrolytic	12	C27	3300 pF	silvered mica	
C11	0·1 μF	paper	350	C28	1800 pF	silvered mica	
C12	0·1 μF	paper	350	C29	820 pF	silvered mica	
C13	560 pF	silvered mica		C30	390 pF	silvered mica	
C14	8200 pF	silvered mica		C31	2700 pF	silvered mica	
C15	2200 pF	silvered mica		C32	2700 pF	silvered mica	
C16	0·02 μF	paper		C33	4700 pF	silvered mica	
C17	50 μF	electrolytic	12	C34	0·01 μF	silvered mica	
				C35	50 μF	electrolytic	12

Tolerance of silvered mica capacitors is 10%

Fig. 3 - Circuit Diagram of Three-Valve Pre-Amplifier



LIST OF COMPONENTS FOR THREE-VALVE PRE-AMPLIFIER

Inductor, L1

Wound Ferroxcube pot core, Mullard type WF1428

Valves

Mullard EF86 (two), ECC83

Valveholders

B9A (noval) nylon-loaded, with screening skirt
McMurdo, XM9/AU, Skirt 95 (three)

Miscellaneous

Supply input socket, 6-way. Bulgin, P194
Lampholder. Bulgin, D/675/1/Red
Pilot lamp. 6·3V, 150mA
Input socket (four), coaxial. Belling Lee, S.604/S
Input jack. Igranic, P71
Output socket, coaxial. Belling Lee, S.604/S
Output jack. Igranic, P71
Selector switch SA, 5-way rotary
Specialist Switches, SS/593/A
Low-pass filter switch SB, 4-way rotary
Specialist Switches, SS/593/B
High-pass filter switch SC, 4-way rotary
Specialist Switches, SS/593/C
Set of 3 switches. Shirley Laboratories, SBL F/5/62
(Note: Details of proprietary switches may not be identical with those given in the diagrams.)
Tagboard, 5-way (one). Bulgin, C120
Tagboard, 10-way (three). Bulgin, C125

Position of SA for equalisation	Position of SB:	Position of SC:
a = Auxiliary	a = 5kc/s	a = 160c/s
b = Radio	b = 7kc/s	b = 80c/s
c = Tape	c = 9kc/s	c = 40c/s
d = Long Play	d = Flat	d = 20c/s
e = 78 r.p.m.		

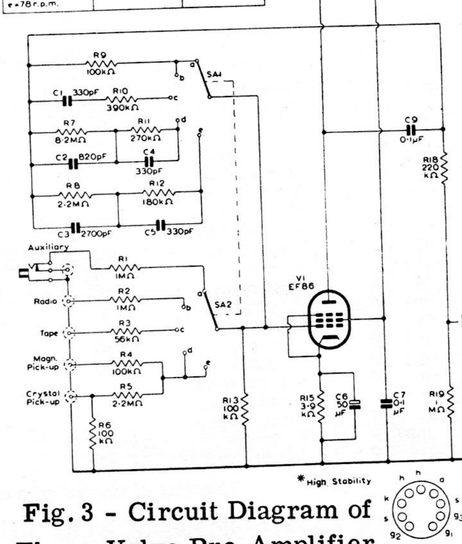


Fig. 3 - Circuit Diagram of Three-Valve Pre-Amplifier

Output Attenuating Networks

The sensitivity and impedance of each channel have been chosen for sources which are likely to be encountered most frequently. The level of output signal from the pre-amplifier will have to be adjusted if it is to be used with power amplifiers having different sensitivities from the 20W circuit. The signal should be attenuated by using a simple potential divider as shown in Fig. 1.

H.T. SUPPLIES

The h.t. currents taken by the 2-valve and 3-valve pre-amplifiers are respectively 3mA at 230V and 6mA at 250V. The values of the smoothing components for these supplies depend on the combination of pre-amplifier and power amplifier. Appropriate values are given in Tables 1 and 2. The points from which these supplies are taken are shown in the complete circuit diagram for the 20W amplifier.

The l.t. current taken by the 2-valve pre-amplifier is 0.4A at 6.3V and that required by the 3-valve circuit is 0.7A at 6.3V.

CONSTRUCTION

The connection to the programme-recording output jack on the front panel of the 3-valve pre-amplifier should be made with coaxial cable from the junction of R27 and C13. This jack socket and that of the 2-valve circuit may be connected to a coaxial output socket on the back of the pre-amplifier if it is needed for a fixed recorder.

PERFORMANCE

The values for hum and noise in the pre-amplifiers which are quoted below for each input channel have been measured with each of the pre-amplifiers connected to the 20W power amplifier. The measurements were made at the output socket of the power amplifier when the input terminals of the pre-amplifier were open-circuited. The frequency-response curves were also obtained with these combinations of pre-amplifier and power amplifier.

The sensitivity figures given provide full outputs from both pre-amplifiers. Total harmonic distortion in the two-valve pre-amplifier is less than 0.15% at the rated output level and is only 0.24% at ten times this output. Total harmonic distortion in the 3-valve pre-amplifier is less than 0.1% at the rated output and only 0.65% at ten times the rated output. A rapid increase in distortion does not occur until the pre-amplifiers are considerably overloaded.

MAGNETIC PICK-UP POSITION

	2-valve pre-amplifier	3-valve pre-amplifier
Input Impedance	100k ohms	100k ohms
Sensitivity at 1 kc/s		
(a) microgroove	4.8mV	7mV
(b) 78 r.p.m.	13mV	12mV
Hum and Noise (below 20W)		
(a) microgroove	55dB	53dB
(b) 78 r.p.m.	57dB	58dB

This channel is most suitable for pick-up heads of the variable-reluctance type, but moving-coil types which have higher outputs can be used if a larger value of series resistance is included.

CRYSTAL PICK-UP POSITION

	2-valve pre-amplifier	3-valve pre-amplifier
Input Impedance	100k ohms	100k ohms
Sensitivity at 1kc/s		
(a) microgroove	70mV	150mV
(b) 78 r.p.m.	210mV	270mV
Hum and Noise (below 20W)		
(a) microgroove	55dB	53dB
(b) 78 r.p.m.	57dB	58dB

Low- and medium-output crystal pick-up heads can be used for this channel. The input is loaded with the 100k ohms resistor (causing bass loss) in order that it characteristic shall approximate to that of a magnetic cartridge, and to allow the same feedback network to be used. This produces the best compromise with most types of pick-up head.

However, if the head is not suitable for this form of loading, or if its output is too high, then it can be connected to the auxiliary input socket, which is discussed fully below. With this channel, the pick-up output is fed into a 1M ohm load which compensates automatically for the recording characteristic.

TAPE PLAYBACK INPUT CHANNEL

	2-valve pre-amplifier	3-valve pre-amplifier
Input Impedance	80k ohms (approx)	80k ohms(approx)
Sensitivity at 5kc/s	4mV	2.5mV
Hum and Noise (below 20W)	52dB	47dB

Each channel is intended for replaying pre-recorded tapes using high-impedance heads, and the characteristics adopted result in good performance with these heads. If a greater sensitivity is required, the value of the series input resistor can be decreased until the desired sensitivity is obtained.

MICROPHONE INPUT CHANNEL
(two-valve circuit only)

Input Impedance	1M ohm
Sensitivity	7.5mV
Hum and Noise (velow 20W)	44dB

The microphone input channel is intended for use with high-impedance systems such as crystal microphones or magnetic microphones with transformers

RADIO INPUT CHANNEL

	2-valve pre-amplifier	3-valve pre-amplifier
Input Impedance	1M ohm	1M ohm
Sensitivity	330mV	250mV
Hum and Noise (Below 20W)	63dB	63dB

With the values of impedance and sensitivity quoted above, this channel should meet most requirements. Other values can easily be obtained, however, by altering the feedback resistor and the series input resistor. If the input impedance of the channel is too high, it can be reduced by connecting a resistor of the appropriate value between the input end of the series resistor and the chassis.

AUXILIARY INPUT CHANNEL

It can be seen from the circuits that the auxiliary channel is identical with the radio input channel in both pre-amplifiers. The channels can therefore be used for high-output crystal pick-ups, or for a tape pre-amplifier. If it is desired to use a microphone with the 3-valve circuit, a separate microphone pre-amplifier or an input-mixing pre-amplifier can be connected to the auxiliary socket. The auxiliary input is taken to a jack socket at the front of the chassis. This makes it easier to connect a portable tape recorder. The jack-socket termination of the auxiliary input position is such that insertion of the jack disconnects the coaxial socket on the rear panel.

TONE CONTROLS

Low-impedance controls have been adopted in the 2-valve circuit so that any capacitance resulting from the use of long coaxial leads between the pre-amplifier and main amplifier will have a minimum effect on the output impedance of the pre-amplifier.

AUXILIARY OUTPUT POSITION

The output from the second **EF86** of both circuits is available at this auxiliary position enabling a record of programme material to be made with tape equipment. This additional output is taken to a jack socket at the front of the chassis. Excellent recordings can be made even when the input signal is derived from a low-output magnetic pick-up.

An output of about 250mV is available at this socket, and the impedance is low. Recording equipment plugged into this socket should not have an input impedance less than 500k ohms. The tone controls and the filter networks are inoperative when this output is used.

FILTER NETWORKS (three-valve circuit only)

The characteristic of the low-pass filter has a slope of approximately 20dB per octave. The components of the network are arranged for operation at 5, 7 and 9kc/s in positions (a), (b) and (c) respectively of the switch SB in Fig. 3. Position (d) of the switch gives a flat characteristic.

The characteristic of the high-pass filter has a slope of approximately 12dB per octave. Operation is at 160, 80 and 40c/s respectively at positions (a), (b) and (c) of the switch SC. Position (d) of the switch custs at 20c/s: this is considered preferable to allowing the response to continue to lower frequencies because of the possibility of the amplifier or loudspeaker being overloaded.

CHAPTER 16

BRIMAR VP10 MONO PRE-AMPLIFIER DESIGNED BY THORN RADIO VALVES AND TUBES LTD

The Brimar VP10 mono high fidelity pre-amplifier circuit is suitable for driving any power amplifier which requires an input of 130mV for full power output such as the Brimar VA12 20-Watt amplifier.

The VP10 pre-amplifier incorporates full equalisation in accordance with the British Standard recommendation, and controls are provided for adequate bass and treble cut or boost, input selection and output level. Low hum and noise level is obtained without stringent layout specifications.

Tone Controls: Continuously variable

Bass: ± 17.5dB at 30c/s referred to 1kc/s.
Treble: ± 18.5dB at 15kc/s referred to 1kc/s.

Harmonic Distortion: Estimated to be better than 0.03% for an output of 130mV at 1kc/s.

Power Supply: H.T. supply 250/260V at 2.85mA.
Heater supplies. 6.3V, 0.6A
Panel Bulb 6.3V, 0.14A

INTRODUCTION

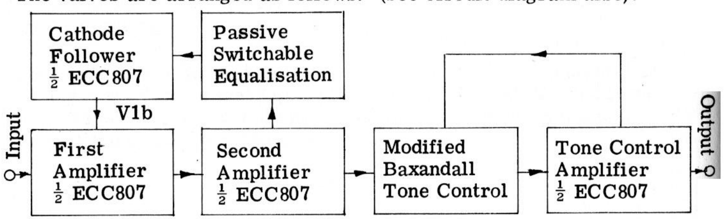
Input facilities are selected as the input selector switch is rotated clockwise for magnetic pick-up, microphone, radio tuner, and tape playback. The inputs are selected one at a time, leaving the other input sockets open circuited.

The output voltage level is controlled by a 1M ohm potentiometer which performs the function of a volume control.

Two Brimar ECC807 double triodes are used in this circuit. These valves are specially designed for use in high gain pre-amplifier stages operating from low level inputs.

Circuit Description

The valves are arranged as follows: (see circuit diagram also).



SPECIFICATION SUMMARY

	Magnetic Pick-up	Microphone	Radio Tuner	Tape Replay
Sensitivity For an output of 120mV into 1M . Bass and treble controls flat.	4mV at 1kc/s	9mV at 1kc/s	32mV at 1kc/s	4mV at 5kc/s
Hum and Noise Below 20W Using VP10 and VA12 amplifiers	-53dB Input s/c	-56dB Input o/c	-58dB Input o/c	-54dB Input s/c
Equalisation	R.I.A.A. u Groove recommendation.	Flat ± 0.5dB 10c/s to 15kc/s	Flat ± 0.5dB 10c/s to 20kc/s	+ 21dB lift at 30c/s (B.S. Recommendation)
Feedback Factor (approx) For an output of 130mV Bass and treble controls flat output level control maximum.	37dB at 1kc/s	37dB at 1kc/s	56dB at 1kc/s	36dB at 5kc/s
Maximum Input Without amplitude distortion of the output	420mV	500mV	500mV	370mV

First Amplifier

This stage operates at full gain which is of the order of 70 times. Although the cathode of this stage is connected to R3 (680 ohms) which is not bypass, the hum pick-up is negligible because R3 is shunted by the output impedance of the cathode follower V1b which is further lowered due to the additional gain of V1a and V2a.

Second Amplifier Stage

The first stage is capacity coupled to the second stage via C11. This stage is grid current biased by R17. R16 provides some current feedback, which helps to reduce distortion.

V1a has a gain of 70 without feedback, V2a a gain of about 60 and the total gain of the two stages is 4, 200. This is appropriately reduced by feedback to give the required sensitivity. Feedback is obtained from the anode circuit V2a and fed to V1a via the cathode follower V1b.

If, during switching from one input facility to another, a momentary open circuit occurs in the feedback loop, instability is likely to occur because of the high loop gain. To prevent this, a "make-before-break" selector switch is used.

Equalisation Network

Equalisation for magnetic pick-up and tape replay is achieved by frequency selective networks, and for microphone and radio by resistive networks.

Magnetic Pick-up

Equalisation in accordance with RIAA recommendation is provided (see magnetic pick-up response). The gain of the pre-amplifier, when switched to this channel, can be controlled by R8, which in turn controls the amount of feedback. The sensitivity of this channel is sufficient for any of the magnetic pick-ups presently available.

At the anode of V2a, the base lift continues to increase below 20c/s. C10, R19, which couple the signal into the Baxendall circuit provide a low frequency droop below 20c/s which ensures correct equalisation for the complete pre-amplifier.

At the low frequency end, when the lift is a maximum, the pre-amplifier still has about 34dB of total feedback. The maximum input voltage that can be safely used without causing amplitude distortion is 420mV.

Microphone Input Channel

This channel is designed for a high quality low output microphone. The frequency response is flat (within ± 0.5dB). Phase shift correction is provided by C6 connected across R12. The gain of this channel can be controlled by R7. This channel has 54dB of feedback. To ensure level response at very low frequencies C10, R19 are included in the feedback loop.

Radio Tuner Input

The frequency response is flat over the whole audio range. With the sensitivity figure of 31.5mV, it should be suitable for any type of tuner. Feedback in this channel is 61dB. Phase correction is provided by C5 connected across R11. Once again C10, R19 are included in the feedback loop. The gain of this channel can be controlled by R6.

Tape Replay

The required equalisation characteristic depends upon the replay head to be used. The performance of a head is affected by the quality of the front gap, its depth, its relation to the back gap, the iron losses, self capacitance loss, and the condition of the working face. In this channel nearly 21dB's of lift at 30c/s is provided.

This should be adequate for playing back recorded tape at a speed of 3.75 inch/sec. from a reasonable quality tape replay head. At maximum lift the feedback is at least 33dB. Hence the distortion is low. As for the magnetic pick-up C10, R19 are left outside the feedback loop to provide the desired low frequency droop. The gain of this channel can be controlled by R5.

Cathode Follower

The required feedback boltage is applied to the input valve cathode through the cathode follower (V1b). This cathode follower has a gain of 0.6 and provides a low impedance feedback voltage source to the input valve.

Modified Baxendall Tone Control and Amplifier

This is a "see-saw" type of circuit. When the treble and bass controls are at a mid-position, the gain of the stage is unity. The base lift and cut is 17.5dB at 30c/s and treble lift and cut is 18.5dB at 15kc/s.

The lift or cut is initially provided at the end of the frequency scale without appreciably affecting the mid frequency range. A very high degree of feedback is present even at full lift, hence the distortion at this stage is extremely low. The degree of lift and cut provided should be adequate to provide compensation for room acoustics, studio and recording deficiencies, record surface noise etc.

Bandwidth

This was measured in the microphone channel. The tone controls were set in the "flat" position and the output control was in its maximum output positon. The generator had an impedance of 600 ohms and the output voltage was developed across a 1M ohm load. The frequency response was flat at 10c/s and -1dB at 20kc/s.

Sensitivity

Measurements were made with a generator impedance of 600 ohms. Tone controls-flat, volume control maximum, output across 1M ohm.

Hum and Noise

For this measurement the pre-amplifier was connected to Brimar VA12, which gives an output of 20 watts with an input of 130mV. DB measurements were made with respect to 20 watts with input appropriately short-circuited or open-circuited. The power amplifier was terminated with its appropriate resistive load.

Tone Controls

The performance of the tone controls are plotted in the tone control response. The range is for bass control ± 18.5dB's. Generator impedance was 600 ohms and the pre-amplifier load was 1M ohm.

Harmonic Distortion

Very low harmonic distortion in the pre-amplifier made any realistic measurements impossible for 120mV output. Hence measurements were made with an output of 1V r.m.s. The estimated distortion figure for 120mV rated output was 0.03%, since distortion is approximately proportional to output at these levels.

Overload

Measurements were made to see the effect of overloading the inputs. A visual inspection of amplitude distortion was made using a Tektronix 535 series oscilloscope with 53/54C plug-in unit.

Overload factor

Magnetic Pick-up:	=	105	=	40dB
Microphone:	=	109	=	41dB
Radio:	=	16	=	25dB
Tape Replay:	=	92.5	=	39dB

Transient Response

A good quality pre-amplifier should have a good transient response. A transient pulse was obtained from a Tone Burst Generator. This Tone Burst Generator has facilities to vary the on/off ratio. The output was observed on the oscilloscope and was compared with the input. The output was completely free from any ringing or high frequency attenuation effect.

A Short Note on Layout

The specified performance of the amplifier was obtained without any special plan for layout. Due care should be taken to keep the heater leads, which should be twisted, well away from the input grid. The heater supply is balanced with respect to earth by means of resistance from each lead. Two 330 ohms resistors are wired in series across the heater leads, the centre tap is connected to earth. The power supply to the pre-amplifier should be adequately smoothed.

D.C. Conditions

Valve Type	Point of Measurement	Voltages	D.C. Range of AVO. 8*
V1a	Anode	135	1,000 Volts
V1a	Cathode	1.15	10 Volts
V1b	Anode	207	1,000 Volts
V2a	Anode	140	1,000 Volts
V2b	Anode	160	1,000 Volts
V2b	Cathode	1.4	10 Volts

* 1,000 Volts range resistance = 20M ohms.
10 Volts range resistance = 200k ohms.

Component Lists

Resistors (All ¼ W Rating)

Circuit Ref.	Value	Tol.	Circuit Ref.	Value	Tol.
RO	1 M	H.S.	R14	10M + 8.2M	H.S.
R1	220 k	H.S.	R15	1 M	± 10%
R2	1 M	H.S.	R16	470	± 10%
R3	680	H.S.	R17	8.2 M	± 10%
R4	22 k	± 10%	R18	100 k	H.S.
R5	33 k	H.S.	R19	1 M	± 10%
R6	270 k	H.S.	R20	120 k	± 10%
R7	27 k	H.S.	R21	1 M	± 10%
R8	47 k	H.S.	R22	1.8M	± 10%
R9	560 k	H.S.	R23	120 k	± 10%
R10	8.2M	H.S.	R24	100 k	± 10%
R11	470 k	H.S.	R25	3.3k	± 10%
R12	470 k	H.S.	R26	15 k	± 10%
R13	820 k	H.S.			

Potentiometers

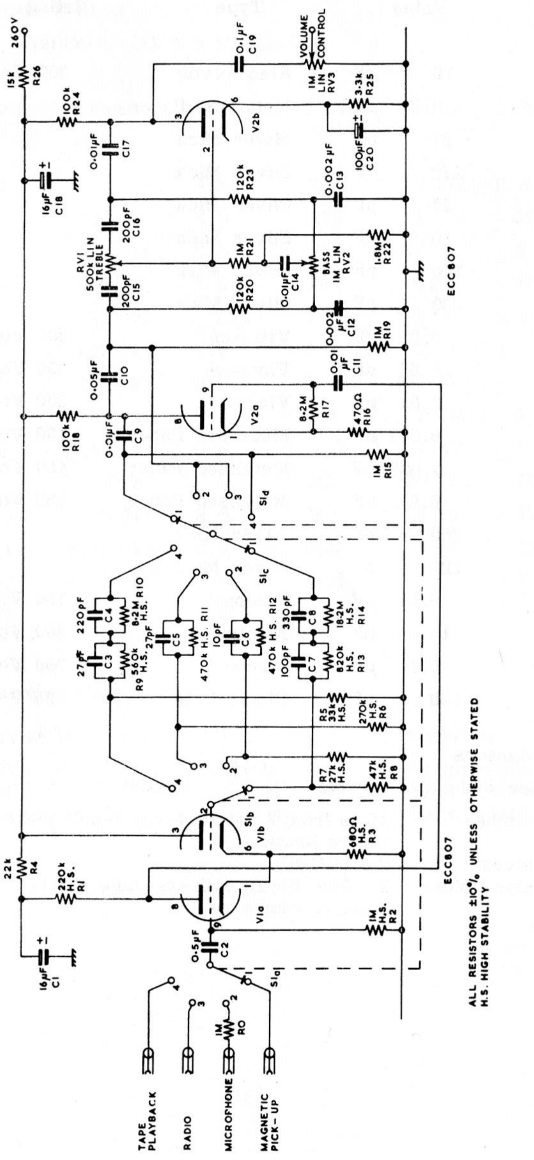
Circuit Ref	Value	Function
RV1	500 k	Treble control
RV2	1 M	Bass control
RV3	1 M	Output level control

Circuit Ref	Value		Type	Rating
C1	16	µF	Electrolytic	300 Volts
C2	0.5	µF	Metalised Paper	
C3	27	pF	Silver Mica	
C4	220	pF	Silver Mica	
C5	27	pF	Silver Mica	
C6	10	pF	Silver Mica	
C7	100	pF	Silver Mica	
C8	330	pF	Silver Mica	
C9	0.01	µF	Visconol	300 Volts
C10	0.05	µF	Visconol	300 Volts
C11	0.01	µF	Visconol	300 Volts
C12	0.002	µF	Metalised Paper	150 Volts
C13	0.002	pF	Metalised Paper	150 Volts
C14	0.01	µF	Metalised Paper	150 Volts
C15	200	pF	Silver Mica	
C16	200	pF	Silver Mica	
C17	0.01	µF	Visconol	150 Volts
C18	16	µF	Electrolytic	300 Volts
C19	0.1	µF	Visconol	300 Volts
C20	100	µF	Electrolytic	25 Volts

Miscellaneous

Input and output sockets. coaxial

Switches:	Made from 2 wafers 4-way 2-pole make-before-break.
Valves:	2 - ECC807
Valve Holders:	2 - B9A (noval) with screening skirt.
	2 - Screening cans.



VP10 AUDIO PRE-AMPLIFIER

CHAPTER 17
POWER AMPLIFIER

The Brimar VA12 high sensitivity, high fidelity, 20 watt, audio power amplifier circuit. The output stage uses two Brimar EL506 pentodes in the class AB1 push-pull ultra linear mode.

The amplifier may be driven from a suitable pre-amplifier such as the Brimar VP10.

The valve and rectifier complement is as follows:

2	-	EL506	pentode output power valves,
1	-	ECC83	double triode amplifier driver and phase splitter,
1	-	ECC807	double triode high voltage gain input valve, and
4	-	BY105	forming a silicon H.T. rectifier bridge.

PERFORMANCE

Power Output and Distortion:	20 watts at 1 kc/s with total harmonic distortion of less than 0.1%
Sensitivity:	130mV for 20 watts output.
Bandwidth:	Flat to 0.1dB between 30 c/s and 20 kc/s at 20 watts.
Feedback:	36dB overall feedback.
Hum and Noise:	80dB below 20 watts with input short circuited.

CIRCUIT DESCRIPTION

For the first two stages of the amplifier a high gain, low noise valve type ECC807 is used (Ref. V1a and V1b in the circuit diagram). The total voltage gain of a double triode ECC807, connected in cascade, is approximately 4900 times. Used as the input stages of the VA12 Amplifier, however, a total gain of only 380 times was required. Consequently R5 and R6 form an attenuator for the first stage and R10 and R16 an attenuator for the second stage.

It can be seen from the circuit diagram that the first three stages of the amplifier are d.c. coupled. This improves the low frequency stability when negative feedback is applied by minimising low frequency phase shift in the amplifier.

A silicon voltage regulator diode BZY88/C9V1 is used in the cathode circuit of V1a in order to give a stabilised d.c. reference to the amplifier. This is necessary for the amplifier which, apart from being d.c. coupled over the first three stages also employs d.c. feedback from the anode of V2a to the grid of V1a. Any drift or changes in H.T. voltage is, therefore, compensated for by the use of the voltage regulator diode.

V1a, V1b and V2a form part of a negative feedback loop with R7, R12, R13 and C8. This only becomes operative at frequencies below 100c/s, which compensates for the lack of low frequency feedback, due to the limitations of the output transformer, which forms part of the overall feedback loop (Ref. C15 and R27 in the circuit diagram). The negative feedback loop formed by R7, R12 and R13 also controls the overall d.c. gain. C8 causes the amount of negative feedback to be reduced as the signal frequency is raised.

To prevent high frequency instability within the amplifier a CR network (Ref. C3, R2 in the circuit diagram) is connected in parallel with R3 the anode load of V1a, giving an advance in phase.

Miller effect in V1b and V2a increases the attenuation and phase shift of the coupling circuits at high frequencies. These effects are reduced by the use of C5 and C7 which, thereby, further improves the high frequency stability of the amplifier.

The third stage, the driver and phase splitter of the amplifier uses the ECC83 double triode (Ref. V2a and V2b in the circuit diagram). To ensure sufficiently equal drive to the EL506 grids, the resistors R18, R19 and R20, R21 should be matched to within 5%.

At the junction of R11 and R14 the voltage is arranged to be the same as that of the grid of V2a. This voltage is applied to the grid of V2b, decoupled to audio by C12. The arrangement so made, provides a measure of d.c. compensation for any change of components etc. occuring in the circuitry of V2b.

The gain of the ECC83 in the cathode coupled form of phase splitter used is about 26 times. This is sufficient to provide a drive of 15V r.m.s. (distortion 0.3%) to each grid of the output stage, needed, for a total output power of 20 watts.

For the output stage two valves, type EL506, are connected in a push-pull, Class AB1, ultra linear mode. The anode-to-anode loading of the output stage is 6.6k ohms with an H.T. voltage of 438V at the centre-tap of the primary winding of the output transformer. Primary taps at 43% feed the screen grids. Anode and screen grid dissipations are 17 and 2.2 watts respectively.

Direct current, out-of-balance in the primary winding of the output transformer, is limited by the use of separate cathode-biassing resistors R26 and R25 for the two EL506 valves.

Negative feedback is taken from the secondary winding of the output transformer to V1a cathode circuit. With 36dB of feedback applied, as in circuit diagram, the amplifier is completely stable.

The H.T. supply is provided by a winding on the mains transformer and four BY105 silicon H.T. rectifiers connected in a bridge.

The bridge rectifier is connected across the mains transformer 0-320V secondary winding and is capable of supplying 150mA d.c. 108mA is drawn by the amplifier, leaving a further 42mA available for a pre-amplifier and f.m. tuner.

For such auxiliary equipment additional decoupling will be necessary since, in order not to alter the H.T. voltage for the amplifier input stage, the auxiliary supply much be connected to the reservoir condenser as shown in the circuit diagram. The mains transformer also contains two additional heater windings for use with auxiliary equipment.

NOTE: Due to the use of silicon rectifiers it is possible for 521 volts H.T. to appear immediately on switch-on, and remain until the valves draw current. This no load voltage can reach this value due to:

(a) High mains voltage.
(b) High secondary voltage. The rated secondary voltage, having allowed for a voltage drop in the winding, being for the loaded condition.
(c) Two heater windings of the transformer not being used.

It is important, therefore, that capacitors across which the H.T. voltage appears should have suitable surge voltage ratings.

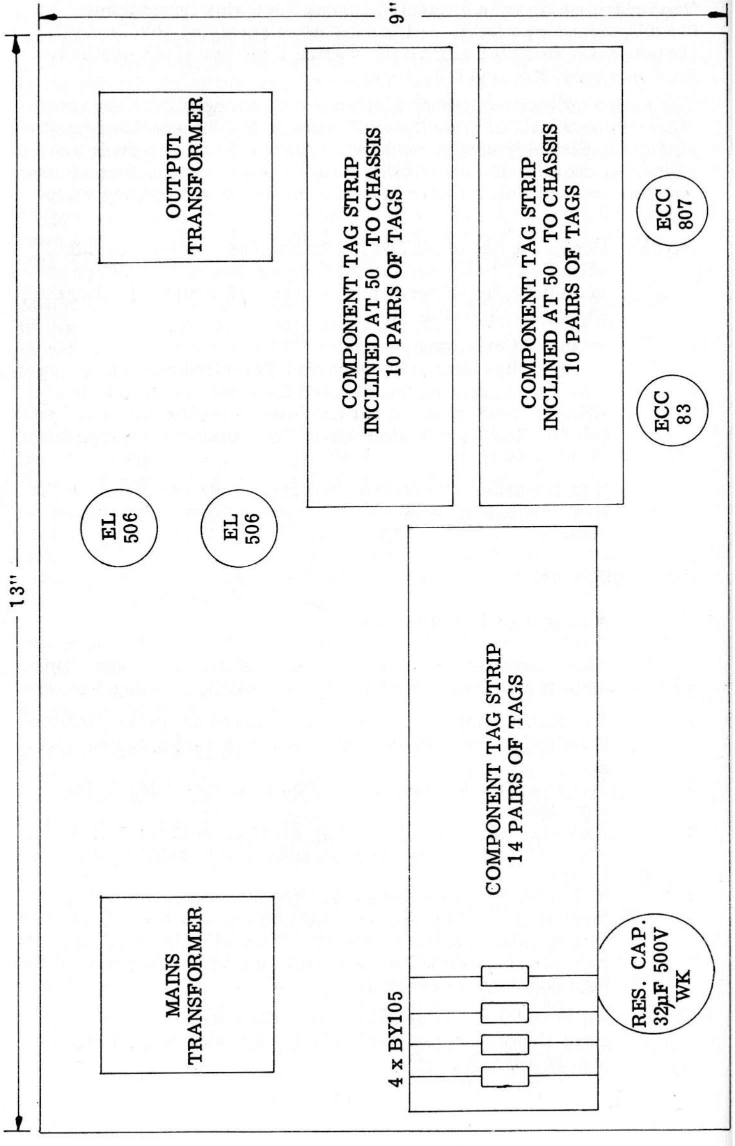
CONSTRUCTION

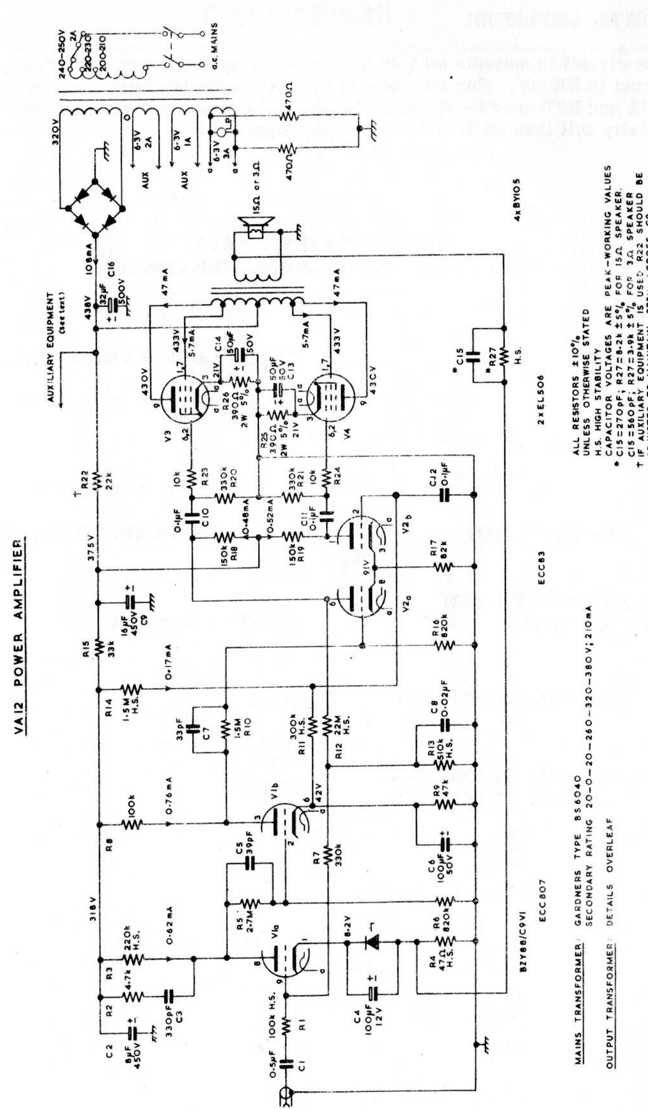
A suggested layout is shown.

The chassis is made from 16s.w.g. aluminium sheet. Important points to note when constructing the amplifier are as follows:-

1. The ECC807 input valve must be mounted as far as possible from the mains transformer in order to minimise hum pick-up.
2. The amplifier has only one earth point, this being in the input stage.
3. Valve heaters are connected in parallel, with the ECC807 input valve being connected directly to the transformer heater winding.
4. High-stability, cracked carbon resistors are used in the input stage as they give an improved level of noise compared with standard carbon resistors. They are also used as feedback components and at points throughout the amplifier where high stability is essential.

NOTE: All resistors and capacitors are either mounted on the tag strip shown in the suggested layout or between the tag strip and the valve holders.


9"
13"
OUTPUT TRANSFORMER
COMPONENT TAG STRIP INCLINED AT 50 TO CHASSIS 10 PAIRS OF TAGS
COMPONENT TAG STRIP INCLINED AT 50 TO CHASSIS 10 PAIRS OF TAGS
ECC 807
ECC 83
EL 506
EL 506
COMPONENT TAG STRIP 14 PAIRS OF TAGS
MAINS TRANSFORMER
4 x BY105
RES. CAP. 32μF 500V WK

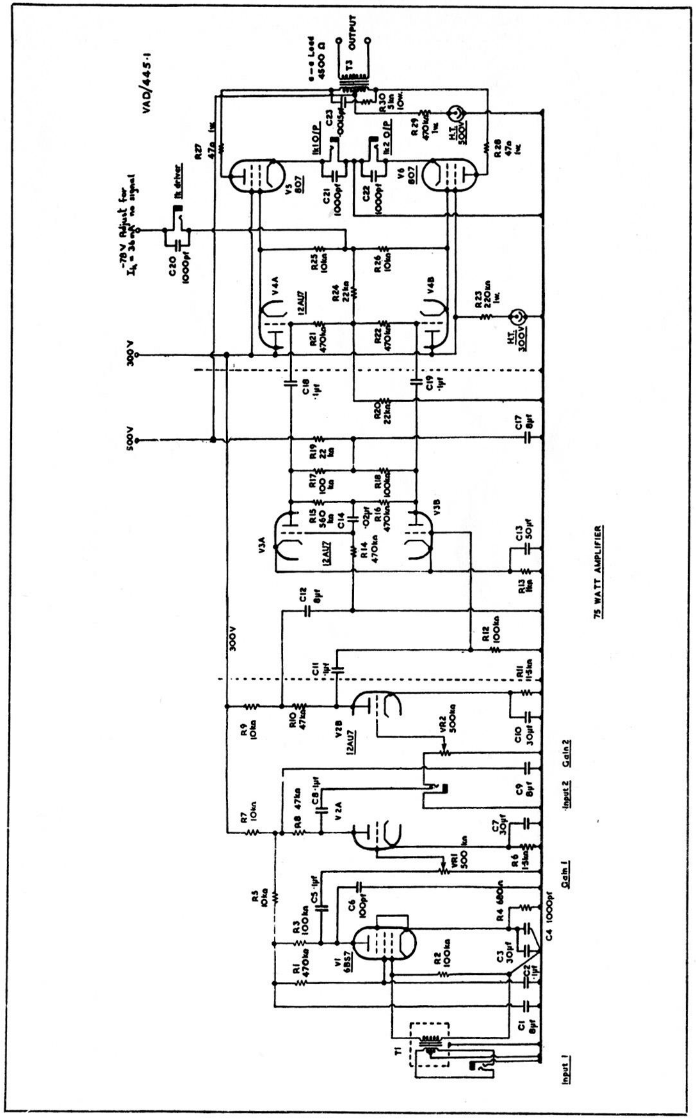

VA12 POWER AMPLIFIER
AUXILIARY EQUIPMENT (see text)
a.c. MAINS
240–250V
220–230
200–210
320V
6·3V 2A
6·3V 1A
6·3V 3A
AUX
AUX
15Ω or 3Ω
4xBY105
2xEL506
ECC83
ECC807
BZY88/C9V1
ALL RESISTORS ±10% UNLESS OTHERWISE STATED
H.S. HIGH STABILITY
CAPACITOR VOLTAGES ARE PEAK-WORKING VALUES
* C15=270pF; R27=8·2k ±5% FOR 15Ω SPEAKER.
C15=560pF; R27=3·9k ±5% FOR 3Ω SPEAKER
† IF AUXILIARY EQUIPMENT IS USED R22 SHOULD BE ADJUSTED TO MAINTAIN 375V ACROSS C9
MAINS TRANSFORMER: GARDNERS TYPE BS6040
SECONDARY RATING 20–0–20–260–320–380V; 210mA
OUTPUT TRANSFORMER: DETAILS OVERLEAF

GUITAR AMPLIFIER

The circuit is suitable as a 30 watt guitar amplifier when the input signal is 220mV. For increase in sensitivity the feedback components (C15 and R27) may be disconnected from the loudspeaker. The sensitivity will then be 7mV for 30 watts output.

OUTPUT TRANSFORMER
USED IN VA12 POWER AMPLIFIER CIRCUIT

TYPE	ULTRA LINEAR PUSH-PULL
RATING	30 W
PRIMARY LOAD	6.6k OHMS ANODE TO ANODE
PRIMARY SCREEN GRID TAPS	43% OF 6.6k OHMS
PRIMARY INDUCTANCE	200 H
PRIMARY/SECONDARY LEAKAGE INDUCTANCE	20 mH



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