a new amplifier

The Radford MA15 Mark II is offered to readers as a constructional item by its co-designers,

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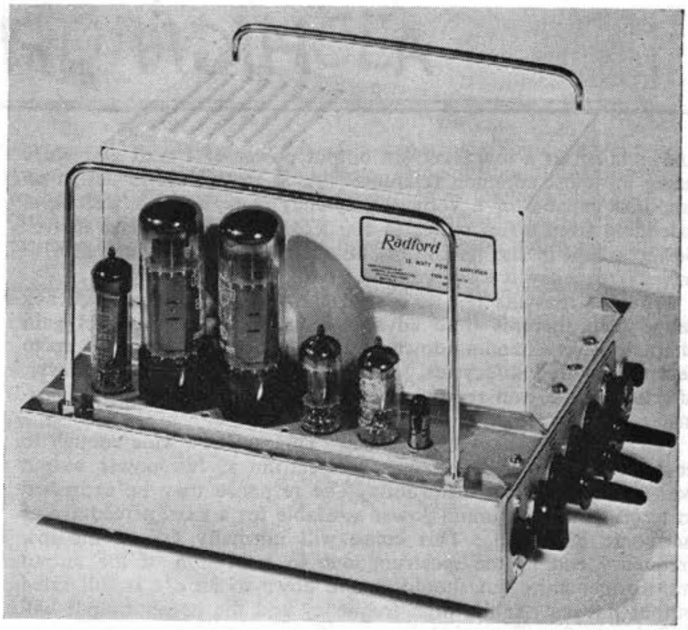
★Radford equipment has, justly, aroused a great deal of interest, and the MA15 which is discused and described by its designers in this series of articles will appeal to many of our "Do-it-yourself" readers. Full constructional details will be covered in later instalments.



THIS series of articles describes an audio frequency power amplifier of new design, having performance characteristics previously unobtained from simple circuitry. Emphasis in design is on very low phase shift in the voltage amplifying stages, and controlled phase shift in the output transformer to obtain full power with low distortion over a very wide frequency range, and yet have exceptional stability under all conditions of input waveform and output loading.

The amplifier is ideally suitable as a laboratory standard, for professional monitoring, or for sound reproduction of the highest standard. The amplifier design is a joint one, as the heading suggests, and is a development of a commercially available amplifier. Development of the new circuit was initiated when it was found that existing circuits were inadequate for use with output transformers of recent design, and that sufficient improvement could not be effected by simple modifications.

A fundamental investigation was carried out by Arthur Bailey on low phase shift circuitry and the behaviour of transformers in circuits with large overall negative feedback.

Historical survey

In the development of the high quality power amplifier over the last 25 years, important changes have taken place in addition to subtle refinements not apparent at first sight. Pre-war amplifiers with any pretence to high quality reproduction were massive and used triodes of the PX4 or PX25 class. Mains transformers were operated at low temperature rises (+10° to +20°) and were of the large window-core ratio type. Core flux densities were of the order of 8/10 K gauss, as compared with 11/13 K gauss at present, and wire densities from 1000/1200 amps per square inch as compared with 1800/2500 amps per square inch at present. Very frequently a double choke filter system was used with paper capacitors which often necessitated separating the power supply from the main amplifier.

The first major circuit development was the use of a large amount of negative feedback from the secondary winding of output transformer to the cathode of the input stage. This technique was typified in the design of the "Williamson" amplifier; (1) an amplifier with wide voltage frequency response and low distortion, using high efficiency tetrodes triode connected, in the output stage. In order to obtain a low phase shift at low frequencies, the first stage was directly coupled, and an output transformer of immense proportions was used. To reduce the phase shift at high frequencies, the primary and secondary windings of the transformer were highly sectionalised.

If comment is made in retrospect on this outstanding design, it is that the output transformer could not meet the requirements of the circuit design. With the main high frequency shunt resonance in the transformer at about 40 Kc/s, it was impossible to provide a gradual roll-off in the forward gain of the amplifier before phase shift in the transformer became serious.

High frequency stability

A modification was introduced a few years later (2) to improve the high frequency stability by the introduction of a step network across the anode load of the first stage. The bandwidth available in the circuitry and transformer was, however, insufficient for any appreciable advantage to be obtained. A large number of these amplifiers were on the verge of h.f. oscillation (if not in fact oscillatory) and were easily triggered by a sharp wave-form into producing damped wave trains. At low frequencies a switching transient would excite the amplifier to a 1–3 c/s oscillation. It was therefore, necessary to provide a low frequency roll-off in the pre-amplifier and to feed the H.T. from a separate supply.

The basic design of this amplifier was so popular however, particularly in the U.S.A., that correspondence and articles appeared in the technical press right up to 1957, (3) on how to stabilise it.

The next stage of development was probably encouraged by the introduction of the ultra-linear or distributed load output stage. This enabled the low inherent distortion level of the triode to be obtained with the power efficiency of pentodes.

New transformer materials

A design originating from the Mullard application laboratories and known as the "5–20", incorporating this type of output stage was described in 1955. (4) This was probably one of the first designs of a simple nature using a large amount of negative feedback (30dB), which had a margin of stability. The performance realised with this design (as with all other feedback designs) depended particularly on the phase characteristics of the actual output transformer used, in relation to the phase characteristics of the amplifier itself. Transformers having essentially identical characteristics in respect of shunt and leakage inductance, shunt capacitance, etc., can exhibit wide variations in performance in respect of stability and transient ringing, when connected in a negative feedback amplifier.

It should be noted that the success of the "5–20" depended on an improved output transformer, using new materials which were not available at the time of the "Williamson" design. In order to obtain optimum results from any amplifier however, it is essential that the phase and attenuation characteristics of the circuit be designed with the actual transformer used.

Specification requirements for high quality amplifiers

There are a surprising number of factors which must be considered when attempting to satisfy present day requirements in the design of audio power amplifiers. These are listed and discussed below. It must be appreciated, however, that, as many of them are conflicting, the final specification can only be arrived at by compromise. The order in which the factors appear was chosen for convenience and does not attempt to indicate their relative importance.

(1) *Voltage Frequency Response:* This is the voltage gain characteristics of the amplifier with respect to the frequency used

(continued on page 30)

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and is taken at a low level; an output power of 1 watt at 1 Kc/s being the most common reference. As it is possible to design an amplifier capable of a voltage frequency response at 1 watt level extending from a few cycles to 100 Kc/s, yet be unable to deliver power except in the middle of the audio band, some reference to power/frequency is also required.

Provided that the response is level within the audio frequency range then there is little advantage to be gained in level gain characteristics extending down to a few cycles per second or up to several hundred kilocycles. The phase response **(5)** is however, affected by the gain response, and the two must therefore be normally taken together.

(2) *Power Frequency Response:* This must be wide enough to cover the whole audio frequency spectrum at full power output with reasonably low distortion. The response may be expressed in terms of the maximum power available for a fixed percentage of harmonic distortion. This curve will normally fall at the low frequency end of the spectrum, due to saturation of the output transformer core, but should extend down to 30 c/s at full rated output power. At the high frequency end the power output will fall, due to the effect of stray capacitances, particularly the capacitive load presented to the anodes of the output valves. As the energy in some instruments is predominantly at the top end of the spectrum (cymbals for example) it is clearly necessary that the h.f. power response of the amplifier should extend to the highest possible limit, say 20 Kc/s.

It should be noted that a power frequency response which does not specify the distortion level is meaningless. It has been common, if not universal practice to ignore distortion when specifying power frequency response. It is possible for an amplifier to give its full rated power output from 20 c/s to 20,000 c/s within 1 dB of the mid-frequency range with a constant input source, but the output waveform may bear little resemblance to the input at say 30 c/s and 10,000 c/s.

(3) *Harmonic Distortion:* This should be as low as possible over the complete frequency band but not to the detriment of the stability of the complete amplifier. The requirements for the very low distortion levels of 0·1% and below at 1 Kc/s is questionable. Taken by itself, exceedingly low distortion at 1 Kc/s is no criterion whatsoever as to the performance of an amplifier for the reproduction of speech and music. However, as 0·1% distortion at mid frequency has become a standard owing to ease of attainment, it may as well be retained. The distortion level rises at low and high frequencies, and values of 1% at 30 c/s and 15,000 c/s may be regarded as being of about the same high standard as 0·1% at 1 Kc/s.

It should be noted that A.S.A. suggest a distortion level of 3% at 1 Kc/s for 0 dB level recording on a standard VU meter for professional tape recorders. In the interests of signal/noise ratio 0 dB is invariably used for peak levels.

(4) *Intermodulation Distortion:* This should be as low as possible and is more serious than harmonic distortion. Unfortunately the subjective effect is not easily equated to measurable results, but it is convenient to use one of the accepted methods of evaluation. What value of intermodulation should be regarded as a maximum is rather vague, but there is no doubt that it should be less than

EXCEPTIONAL PERFORMANCE

1%. Measurement is generally at 40 c/s and 6,000 c/s in the ratio of 4·1 amplitude ratio.

(5) *Phase Response:* This is a plot of the phase-shift of the amplifier against frequency. This response is more important than is generally assumed.

The ultimate performance of the amplifier is dependent upon this function; indeed, it is this parameter which is the key to a successful design. As will be shown, it is necessary to integrate the design of the circuitry and transformer in respect of phase response for optimum performance.

(6) *Stability:* This should be as high as possible under all conditions of load and input waveform. There should be no tendency for either h.f. or l.f. instability when either quiescent or under drive. In particular the gain and phase margins of the amplifier should be such that there is no possibility of transient distortion whatever type of load is applied to the amplifier. This will mean a phase margin of approaching 90 degrees, and a gain margin of approximately 20 dB. The subject of stability and its measurement is dealt with later.

(7) *Transient Distortion (ringing):* This has been mentioned under the heading " stability " and is due to instability under transient drive conditions produced by rapid phase shift in the system. Spurious frequencies generated by transient shock excitation will beat with incoming frequencies, distortion frequencies and intermodulation products to produce further frequencies and a consequently degraded output.

If ringing occurs at audible frequencies (such as by using very poor output transformers with too much feedback, or pre-amplifier sharp cut filters) colouration is also apparent. Some amplifiers show more than one ringing frequency. Ringing should not be less than 100 Kc/s frequency, and the amplitude as low as possible.

(8) *Transient Overload Recovery:* Amplifier advertisements sometimes quote a power output figure obtainable from constant

MA15 MK.11. Components List

Resistors

1	... ...	100k	½w.	H.S.
2	... ...	100Ω	,,	,,
3	... ...	2·2k	,,	,,
4	... ...	330k	,,	,,
5	... ...	100k	,,	,,
6	... ...	2·7k	,,	,,
7	... ...	220k	,,	,,
8	... ...	1M	,,	,,
9	... ...	150k	,,	,,
10	... ...	8·2k	,,	,,
11	... ...	33k	,,	,,
12	... ...	39k	,,	,,
13	... ...	1M	,,	,,
14	... ...	1M	,,	,,
15	... ...	470k	,,	,,
16	... ...	470k	,,	,,
17	... ...	2·2k	,,	,,
18	... ...	2·2k	,,	,,
19	... ...	390Ω	4w.	W.W.
20	... ...	390Ω	4w.	W.W.
21	... ...	1k	½w.	H.S.
22	... ...	1k	,,	,,
23	... ...	560Ω	,,	,,
24	... ...	3·9k	,,	,,
25	... ...	390Ω	,,	,,
26	... ...	2·7k	,,	,,
27	... ...	270Ω	,,	,,
28	... ...	1·8k	,,	,,
29	... ...	8·2k	,,	,,
30	... ...	Brimistor CZ3		
31	... ...	6·8k	1w.	H.S.

Condensers

1	... ...	8μF		450V.W.
2	... ...	0·1μF		400V.W.
3	... ...	500pF	5%	150V.W.
4	... ...	50μF		12V.W.
5	... ...	2μF		350V.W.
6	... ...	2μF		125V.W.
7	... ...	0·0047μF		400V.W.
8	... ...	0·1μF		400V.W.
9	... ...	0·1μF		400V.W.
10	... ...	16μF		450V.W.
11	... ...	50μF		50V.W.
12	... ...	50μF		50V.W.
13	... ...	20μF		450V.W.
14	... ...	20μF		450V.W.
15	... ...	270pF	5%	150V.W.
16	... ...	1000pF	,,	150V.W.
17	... ...	340pF	,,	150V.W.
18	... ...	8μF		450V.W.

Valves

1	... ...	EF86
2	... ...	6U8
3	... ...	EL34
4	... ...	EL34
5	... ...	EZ81

Miscellaneous

T1	... ...	Part No. T2063
T2	... ...	Part No. T2061
L1	... ...	Part No. 2004CV
Fuse	... ...	2·5 amp.

voltage power supplies. The suggestion is that although automatic bias is used, the conditions as in fixed bias apply on transients in respect of power output.

Under short transient conditions the peak h.t. current is drawn from the smoothing capacitor across the h.t. supply. A long transient burst will partially discharge the capacitor and the dynamic levels will change through the amplifier. This can cause severe distortion and the amplifier may, in fact, cut off. This type of amplifier uses quiescent anode currents lower than for normal Class " A " operation and a lower load impedance, and the operating condition is referred to as "Low Loading ".

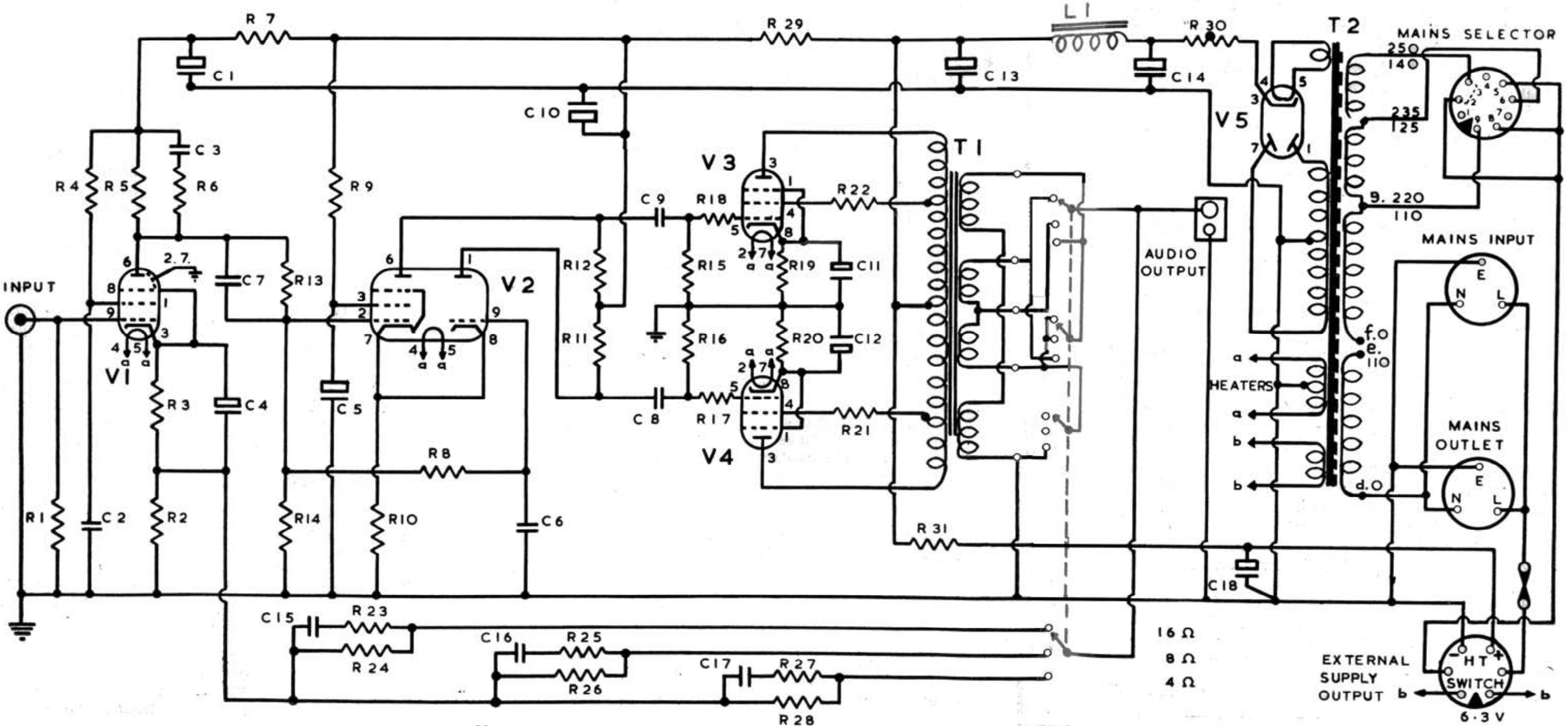
All amplifiers suffer from this defect to a certain degree but it is improved by using optimum dynamic working conditions. An amplifier should show no signs of limiting under sustained transients up to its full rated power output and should show rapid recovery of dynamic operating level after a power output overload from a sustained transient. Tone burst testing can be used to investigate these conditions.

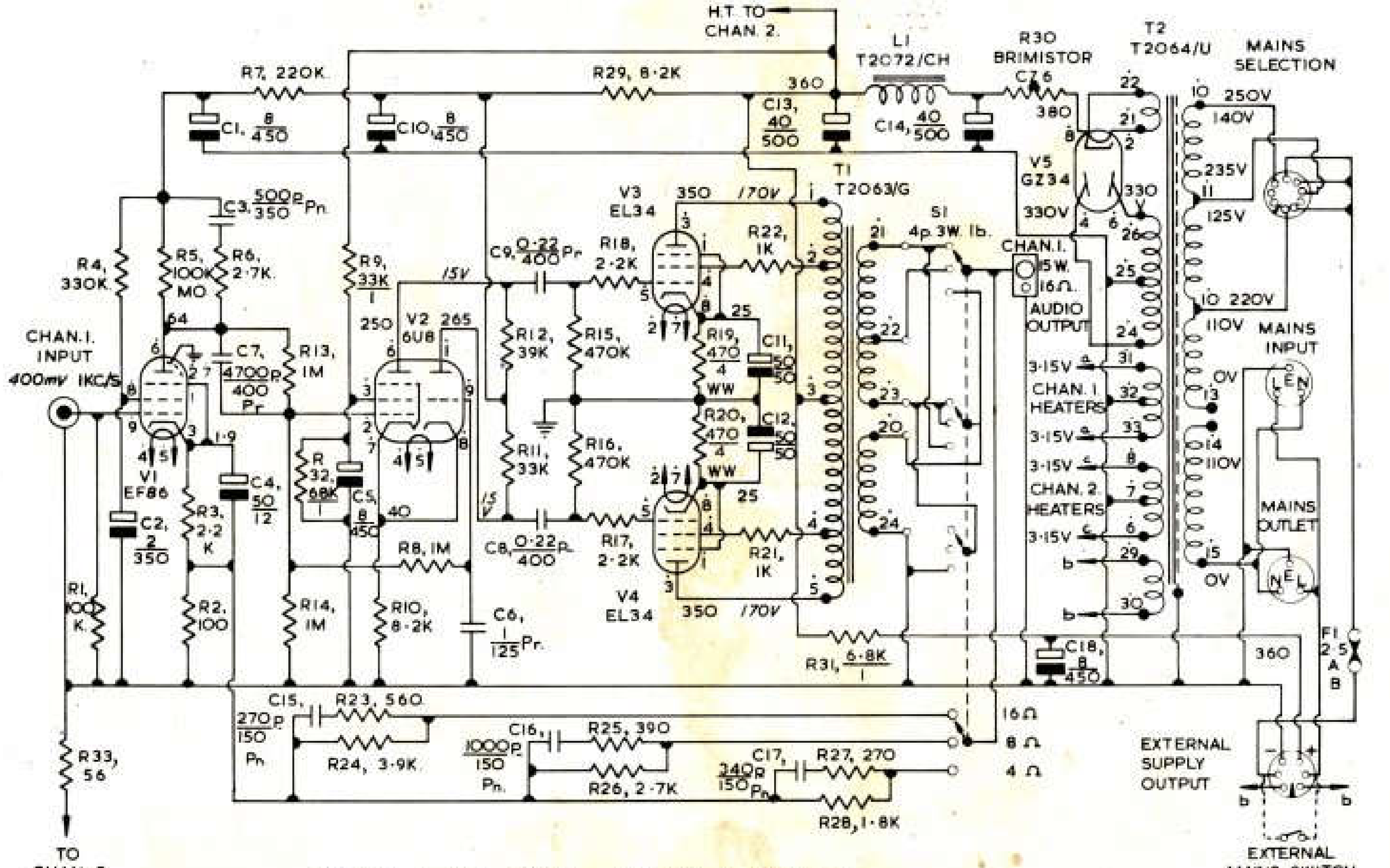
(9) *Transfer Linearity:* The linearity of the transfer curve gives a general indication as to basic soundness of the amplifier design. More specific information can be obtained by precise measurement of the effects of non-linearity such as harmonic and intermodulation distortion. However, it is a basic requirement in the design of a high quality amplifier which must be satisfied before more subtle and refined design features are considered.

(10) *Damping Factor:* This is the ratio of the rated load impedance of the amplifier and is therefore a measure of the regulation of the amplifier output. The output impedance of the amplifier should be as low as possible so that the maximum electrical damping can be applied to the loudspeaker. Also the loudspeaker distortion will be reduced somewhat due to this negative feedback effect. It appears that if the amplifier output impedance is less than about one tenth of the optimum load (i.e. a damping factor of more than 10) then any subsequent reduction of output impedance has no noticeable effect on the loudspeaker output. Providing that the damping factor is therefore, greater than 10–15, the effect of the output impedance is completely swamped by the loudspeaker impedance and the actual value of damping factor above this value is immaterial.

Necessary requirements

This list of requirements may seem rather long, but it is necessary if the design of an amplifier is to be adequate. Some of the factors can only be evaluated after the prototype has been built and the initial tests have been made. Nevertheless, it is wise to keep them in mind at the design stage to avoid too many trials (and errors!).


R 7
R 29
L 1
R 30
T 2
MAINS SELECTOR
250
140
235
125
220
110
C 1
C 13
C 14
C 10
V 5
C 3
R 4
R 5
R 6
R 9
V 3
T 1
C 9
R 18
R 22
R 12
R 15
R 19
C 11
AUDIO
OUTPUT
MAINS INPUT
INPUT
2.7.
C 7
R 13
V 2
V 1
R 11
R 16
R 20
C 12
HEATERS
C 4
C 5
C 8
R 17
R 21
V 4
MAINS
OUTLET
R 8
R 1
C 2
R 2
R 3
R 14
R 10
C 6
R 31
C 18
C 15
R 23
R 24
C 16
R 25
R 26
C 17
R 27
R 28
16 Ω
8 Ω
4 Ω
EXTERNAL
SUPPLY
OUTPUT
HT
SWITCH
6·3 V



CHANNEL 1 ONLY SHOWN — CHANNEL 2 IDENTICAL

+ − ELECTROLYTIC CAPACITOR

POWER AMPLIFIER MA 15. MK 2.

Later Circuit included for comparison

A NEW HIGH FIDELITY

PART 2 — DESIGN CONSIDERATIONS —

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THE amplifier described here was designed for commercial purposes, so the performance cost ratio is therefore of major importance. The circuit developed is only slightly more involved than the *Mullard* "5-20." Considering first the input and driver portion of the amplifier, the essential requirements are: (a) Ability to drive the output valves to full power output with low distortion over the audio frequency range; (b) One effective low frequency phase shift; (c) Very low phase shift at high frequencies.

Many amplifiers use the direct coupled, long-tailed pair phase inverter as is used in the *Mullard* (**fig. 1**) circuit. This circut fulfils the low frequency requirements, but due to Miller Effect in the first triode (V2A) the high frequency response is poor. The anode to grid capacitance of V2A produces an effective capacitance of about 75–100pf at the anode of V1. As the output impedance of V1 is effectively 100K ohms, the frequency for −3dB response (and what is more important 45° lag in phase) occurs at between 15 and 20 Kc/s. This is clearly not good enough as the phase shift will profoundly affect the performance of the amplifier. For this analysis the effect of correction networks will be ignored as they can be applied to any amplifier; it is the inherent weaknesses in the circuits that are being considered.

Phase Splitting Circuits

The other factor in the circuit that will cause phase-shift at high frequencies is the high impedance existing at the anodes of V2a and V2b. This is high, due to the values of anode loads used, and to the high anode slope resistance of the valves when operated with anode currents in the order of ½ mA. If these anodes are fed into the grid circuits of tetrodes or pentodes there would not be a very serious effect, but with ultra-linear stages there is a quite pronounced Miller Effect because of the variation in screen grid potential in the output valves. The cure for this appears to be to reduce the Miller capacitance or, alternatively, the impedance of the drive to the grids of the output valves.

Apart from the long-tailed pair just dealt with, there are several other phase splitting circuits that can be used. These have been discarded for a variety of reasons. The circuits will now be briefly analysed so that the reasons for discarding them will be seen.

The split load (or concertina) phase splitter, **fig. 2** has many good points, but also one or two bad ones. The circuit is self balancing but the two outputs are mutually dependent to a large degree (*Ref. 5*). Further, the circuit can be DC coupled to a previous stage, but as the gain of the phase splitter is less than unity, an additional stage would still be required. This would give rise to low frequency instability, unless the stage is DC coupled to the stage feeding the phase inverter, introducing drift troubles quite apart from the difficulty in designing such a stage.

The next related circuit is that of *Jeffrey's* (*Ref. 6*). This circuit can be arranged to overcome the difficulties of AC coupling, but the problem of high frequency response remains. In this circuit it is difficult to raise the −3dB point above 10 Kc/s due to the impedance multiplying properties of the triode valve (*Ref. 7*). This circuit is shown in **fig. 3**.

The floating paraphase (or anode follower) phase inverter has a good high frequency response if it is phase corrected as shown in **fig. 4**. Difficulty is again experienced with unwanted low frequency time constants due to C_1 and, more important still, the coupling capacitors between other stages, which because of voltage considerations, do not lend themselves to DC coupling. By the use of DC bias supplies, it is possible to overcome the low frequency difficulties, but the stability and cost problems then far outweigh the advantages gained.

After investigating the above circuits, it was realised that the best approach would be to modify the long-tailed pair type of phase splitter. Ignoring for the moment the problem of driving the output valves, the main drawback of the circuit is the Miller feedback from anode to grid of the first triode V2a. The obvious way of overcoming this is to use a pentode, or perhaps to apply neutralisation to the circuit. Neutralisation was in fact tried by feeding back from the anode of the second triode to the grid of the first one by means of a small capacitance. This was effective in reducing the phase shift up to about 50 Kc/s, but after this the phase shift became worse. This was traced to the voltage at the anode of the second triode not remaining in phase with the voltage at the grid of the first one. This may be expected as the capacitive loading of the output valves will begin to be noticeable at these frequencies. This loading has the usual effect of retarding the phase at the output and hence the neutralisation will be ineffective at these frequencies due to difference in phase and magnitude.

Using a pentode valve instead of a triode valve for V2a produced a marked effect on the performance of the amplifier (**fig. 5**). The phase shift was much reduced at the high frequency end of the

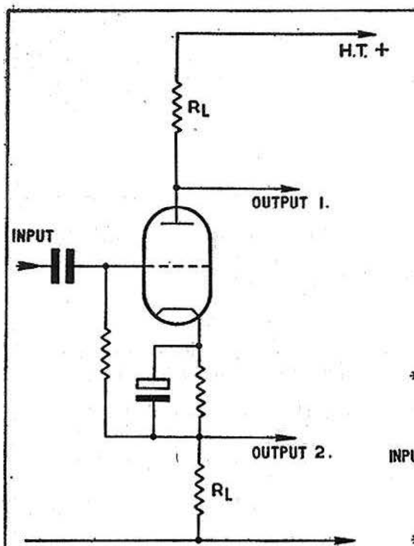

FIG.2.

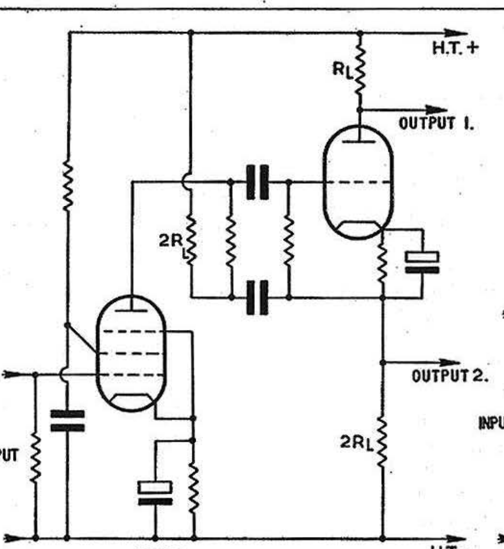

FIG.3.

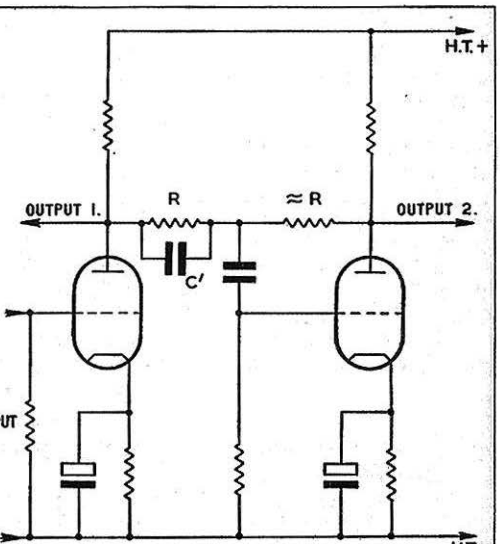

FIG.4.

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AMPLIFIER

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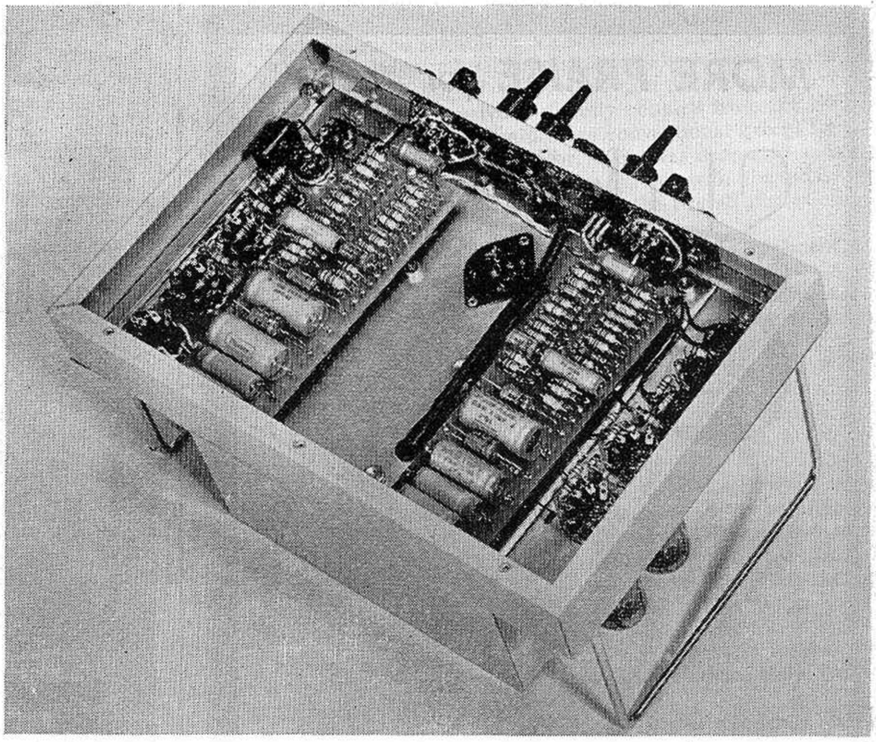

spectrum, as can be seen from the comparative curves of magnitude and phase in **fig. 6.** These curves are for the input pentode V1, and the phase inverter feeding an open curcuit. This illustrates the improvement in the performance of V1, due to the reduced capacitive loading of V2a. These results are very satisfactory and suggest that a double pentode as a long-tailed pair may be advantageous. This was not pursued any further as no suitable valve appears to be available. It should be noted, however, that the second valve in the phase inverter is driven in a grounded grid condition. Miller effect is thus very small, and in practice there may be no advantage in the use of a pentode in this position.

Running the EL 34 as a pentode, the input capacitance is about 15pF. and the anode to grid capacitance is 1·0pF. The gain of the valve when operating into its optimum load is approximately 10. This gives an input capacitance under running conditions of 15+11=26pF. The total capacitive load on the driver valve will therefore be in the region of 30pF.

Under ultra-linear conditions the gain of the valve falls to approximately 8 times, thus the reflected capacitance at the grid is 8pF. There will also be feedback from the screen grid to control grid. The C_{g2}-g_1 capacitance measured 5·2pF, and as the screen grid on the EL 34 runs as 43% of AC anode potential for minimum distortion, the capacitance reflected back due to this amounts to $0{\cdot}43 \times 9 \times 5{\cdot}2$ which equals 20 pF approximately. The total input capacitance is thus 20+15+8=43pF. Allowing 5pF for the previous valve, the capacitance load is approximately 50pF. As will be seen this figure is only a factor of about 1·7 times greater than the pentode connection. In practice it may therefore be as well to use a lower value of driver anode load resistor in the ultra-linear connection.

From a distortion point of view there is little to choose between either connection if the valves are driven to full power. At low signals, however, the ultra-linear circuit gives less distortion and its performance on reactive loads is also better (*Ref. 8*). The final deciding factor was a consideration of the internal regulation of the output stage. The anode slope resistance of a pentode is much greater than that of the same stage run in the ultra-linear mode. The difference is commonly a factor of 10 or thereabouts. When operating into a fixed load of the correct value the difference does not matter, except as "damping factor". The importance arises in the performance with high impedance loads; e.g. a moving coil loudspeaker at high frequencies. The rise in amplification of the output stage so produced can lead to instability and even if the stability margin of the amplifier is such that this does not happen, the amplifier will be that much nearer instability with a corresponding degradation in the transient response. Ultra-linear connection of the output stage was therefore chosen.

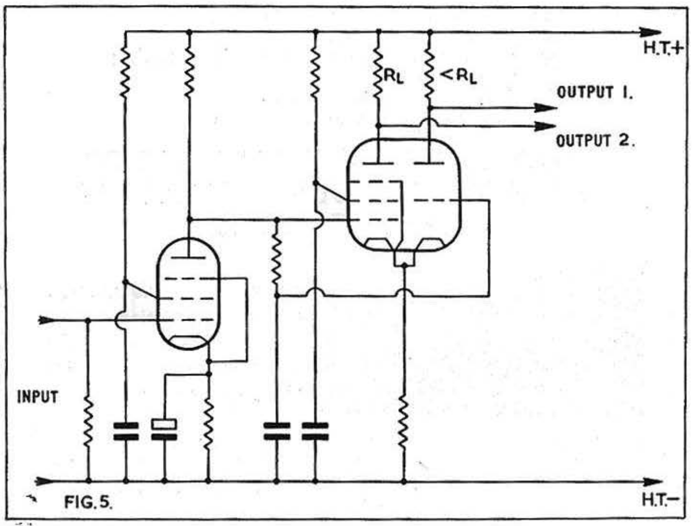

FIG.5.

Reverting to the phase splitter, it will be seen that it has to operate into a capacitive load of about 50pF in the ultra-linear connection. For a −3dB point at 300 Kc/s, this means that the input impedance of the output valves should be about ten kilohms.

Miller Effect

Experimentally, it was found that this value was much lower than necessary; values of 50 kilohms giving better performance. Here again Miller effect is the cause. The input capacitance of a valve due to Miller effect is (1+M) Ca-g, where "M" is the gain of the valve and Ca-g is the anode/control grid capacitance. This is for a resistive load and at high frequencies the output transformer is certainly not purely resistive. Measurements indicated that at the anodes of a pair of push-pull ultra-linear EL 34's driving a 15-watt transformer (the type specified in the final circuit) the effective capacitance load was approximately 1400pF each side of the centre tap.

Exact measurement is difficult owing to the circuit being push-pull. For this reason the capacitance was measured from one anode to earth, and this was assumed to be four times the anode to anode capacitance. The error in assuming this should not be so serious as to invalidate the conclusions drawn from the results. In any case the results of this assumption bear out the measured results, so there cannot be serious discrepancies occurring. The capacitance effective at the anodes of the valves (anode to anode) measured 700pF within close limits up to frequencies of 70Kc/s, after which it varied as shown in **fig. 7**. These results are of course for one particular transformer and will vary with different transformers, but the general shape is typical for multi-section windings. The part of the characteristic when the reactance becomes inductive will be considered as negative capacitance for ease of analysis.

The "equivalent" circuit for the output stage is shown in **fig. 8.**, values being lumped to form a single ended circuit. The capaci-

(*Continued on page 93*)

tance presented at the anode of the output valve is not of course constant, but varies in the manner previously described. However, up to 95 Kc/s it can be assumed to be constant, when the following explanation will be correct.

At the lower frequencies, before the reactance of C1 affects the phase angle of the output voltage to an appreciable extent, the effect of the Miller feedback of C2 will be to increase the apparent input capacitance of the output stage. This will give a high frequency roll-off as expected. As the frequency rises the phase of the output voltage will retard due to the falling reactance of C1 until when $X_{c1}=R_L$ the vector diagram will appear as shown in **fig. 9**. It is now apparent that the feedback is now at about 45° compared

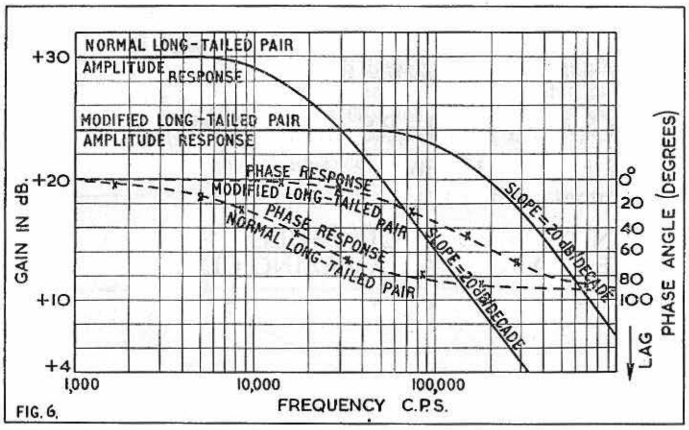

FIG. 6.

with the input voltage and as the frequency rises still further it will tend towards the in-phase (negative feedback) condition. The negative feedback will reduce the apparent input resistance of the output valves and driving circuit and so will reduce the phase shift at these high frequencies. It is this effect that enables use to be made of a higher value of driver output impedance than would otherwise be thought necessary.

To find this ultimate value of input resistance it is merely necessary to assume that the effect of R_L can be neglected, when the gain of the output stage becomes $g_m \times jx_{c1}$. The feedback current will therefore be approximately $V_g \times g_m\, jx_{c1} / -jX_{c2}$ which equals $C1/C2 \times g_m$ as the frequencies are the same for both capacitors. For the ultra-linear EL 34 this makes a value of $1400 \times 1000/3{\cdot}5 \times 11{\cdot}0$ as the g_m is 11 mA/V. This is approximately 40,000 ohms and is obviously the reason for the phase shift being lower than was expected. If the driving impedance is 50 kilohms this will have the effect of a step network with a ratio of 2·2:1 and will therefore assist in maintaining stability to a certain extent. The ultimate effect, however, will be capacitive as the gain of the output stage progressively falls below unity.

Above 95 Kc/s this explanation will not hold until the frequency is above 160 Kc/s when the reactance again becomes capacitive. Between these frequencies the reactance is inductive and the

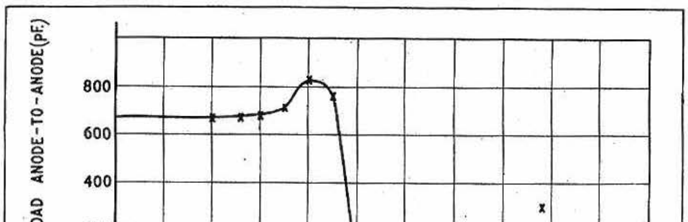

FIG.7.

feedback due to this reactive component will cause a reflected negative resistance (along with a capacitive component). This is due to the effect of the first transformer resonance. This will of course make the phase response worse than expected and the lagging phase angle tends towards 90°. To prevent this having too serious an effect it is necessary that all other phase shifts at this frequency are as small as possible.

The valve found to be suitable as V2 is the 6U8 (ECF 82) triode-pentode. This has a high mutual conductance in both sections, and a triode amplification factor of 40. In addition all the important capacitances are low. This valve is directly coupled from the EF 86 pentode first amplifier. The values are conventional except for the reduced value of screen grid resistor. This was decreased as it reduces the anode potential of the valve (along with the phase inverter cathode) and also somewhat reduces the overall distortion at full power output.

The final circuit evolved was shown in the first instalment. Negative feedback is returned to the cathode of V1 as usual. A step network is included in the anode load of V_1 to assist in maintaining h.f. stability. This step starts fairly high up in the treble response of the amplifier and the small increase in distortion at high audio frequences due to the reduction in feedback is justified by the gain in stability. The step is complete (i.e. attenuation is maximum and phase has returned to about zero) before the critical frequency for instability is reached, so the step network will have no degrading effect on the stability of the amplifier.

It is important to note that the design of the network must be based on a knowledge of the phase characteristics of the amplifier and the actual transformer used. If a different transformer is used then the step network must be redesigned for minimum distortion at high frequencies and maximum attenuation with low phase shift at the critical frequency. The network specified in this design was obtained by reference to the measured Nyquist diagram and was checked by transient tests. A discussion on networks in feedback circuits is beyond the scope of this article and reference should be made to Crowhurst (*Ref. 9*).

The input valve V1 is coupled to V_{2a} via a low frequency step network. This network is included to maintain an adequate low frequency stability margin. The feedback is taken from the second-

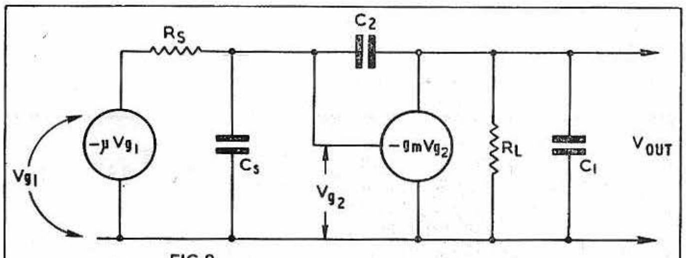

FIG.8.

ary of the output transformer and is fed back via the feedback resistors and associated phase correcting parallel networks. The R/C combination in parallel with the feedback resistors gives phase advance at high frequencies but ceases to do so before such phase advance would cause the stability margin of the amplifier to be degraded. It is perhaps interesting to note that using the conventional parallel R/C network instead of the step network specified, the stability margin is only 8dB at 2·7 Mc/s.

The final performance figures of the amplifier are rather exceptional. This is due to the fact that full advantage has been taken of recent advance in output transformer design by using improved circuitry. The transformer has its first parallel resonance at 150 Kc/s and using a conventional phase splitter the performance was limited by the circuitry rather than the output transformer. From this it is obvious that for the ultimate in amplifier performance, the amplifier must be carefully designed as a whole and not made up of several known circuits lumped together.

Regarding the transformer, designers of feedback amplifiers have always stressed the importance of the output transformer in the performance of their amplifiers. The performance obtainable from the amplifier described depends even more on the accuracy of the transformer, and for this reason no specification is given. This transformer is a precision component and cannot be manufactured under average conditions. Material quality and manufacturing methods must be under rigid control and inspection, or very wide variations will be exhibited in the finished product, thus making nonsense of the circuit technology. This problem was apparent in the manufacture of prototypes when it was found that no two transformers were the same. Variations were found to be due to

(Continued on page 95)

differing spaces taken up by different lead out methods, and wire tensioning.

Protoype transformers were wound on single coil automatic winders and each operator has his own techniques which can amount to a considerable variation on such a sectionalised coil. The amplifier described is in commercial production and the output transformer coils are wound on heavy-duty precision multi-winding machines. Two machines are used, one for primary sections and the other for secondary sections. Each mandrel holds twelve coils and with a number of mandrels available a shuttle service continues between the machines with mandrels in reserve for each machine. Continuous economic production is thereby assured and variation between individual transformers is neglibly small.

The design of an output transformer is still largely empirical and a fantastic amount of time can be spent on ultra-linear transformers varying resonances in frequency and amplitude by winding section size, and phase shift by interconnection. A badly sectionalised transformer may not be much better than one wound with lumped windings. In the transformer specified the first parallel resonance occurs at 150 Kc/s and this has the effect of a peaking phase advance circuit. This first resonance is put to use in maintaining stability, the phase advance occurring where the phase

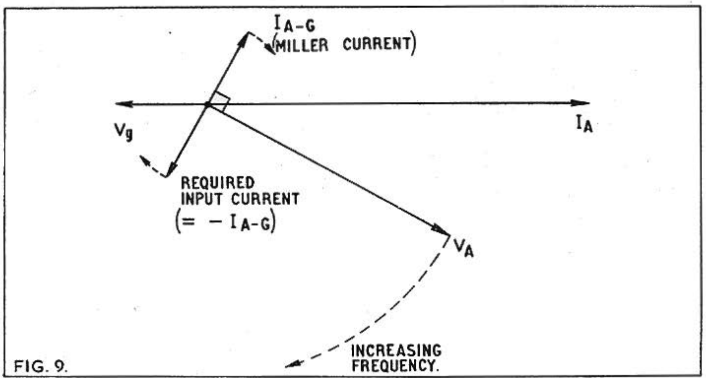

FIG. 9.

margin is in its most critical state. The effect of the peak is to improve the phase margin by an estimated 10–15°. The next transformer resonance that causes a noticeable effect on the gain and phase margins occurs at about 2·7 Mc/s and this is completely swamped in the final design as the loop gain has then fallen to about −26 dB.

Regarding the measurement of the stability of the amplifier, the simple method widely used in obtaining stability margins is unfortunately not a correct measure of the true stability of the circuit. The usual method is to reduce the feedback resistor until instability results, at the same time increasing the feedback capacitor across it so as to maintain the same time constant. At low frequencies this is quite correct as the feedback factor is increased as the ratio between the feedback resistors is changed. At high frequencies, however, the effect is not the same as the feedback capacitor acts as a short circuit if the frequency is high enough. It is therefore clear that there is already 100% feedback at very high frequencies and no adjustment of the feedback resistor will alter this. The only way of measuring gain and phase margin for analysis and design is to break the feedback loop and take plots of gain and phase against frequency around the complete loop. This was done with the complete amplifier, and it is these results that are quoted in the test figures.

The test figures, which will be given in the next instalment, were taken on the amplifier described; the tests being designed, so far as possible, to give a complete picture of the performance of the amplifier. In particular the stability tests were designed to provoke the instability so prevalent in many current designs using high factors of negative feedback.

[1] *Wireless World—May* 1947.
[2] *Wireless World—August* 1949.
[3] *"The care and treatment of audio feedback amplifiers", W. B. Bernard—Audio—January* 1957.
[4] *Wireless World—May/June* 1955.
[5] *High Fidelity Sound Engineering (Newnes), N. H. Crowhurst, p.* 129.
[6] *E. Jeffrey. Wireless World—August* 1947.
[7] *"Economical High-Gain Amplification", A. R. Bailey. Wireless World—January* 1960.
[8] *High Fidelity Sound Engineering (Newnes), N. H. Crowhurst, p.* 120.
[9] *High Fidelity Sound Engineering (Newnes), N. H. Crowhurst, chap.* 6.
[10] *"Why do amplifiers sound different?", N. H. Crowhurst. Radio & T.V. News—March* 1957.

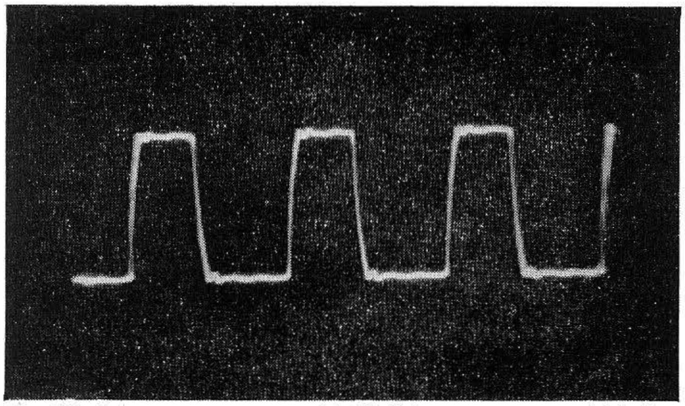

Fig. 16. *Square-wave response with 10 Kc/s square wave and 16 ohms resistive load.*

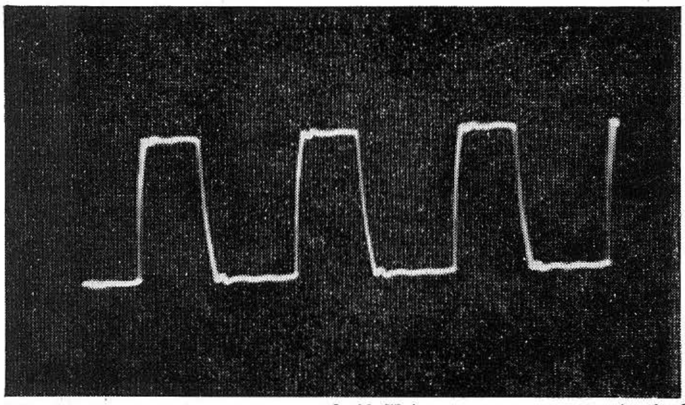

Fig. 17. *Square-wave response with 10 Kc/s square wave and a load of 16 ohms resistive with 0·01 microfarad in parallel.*

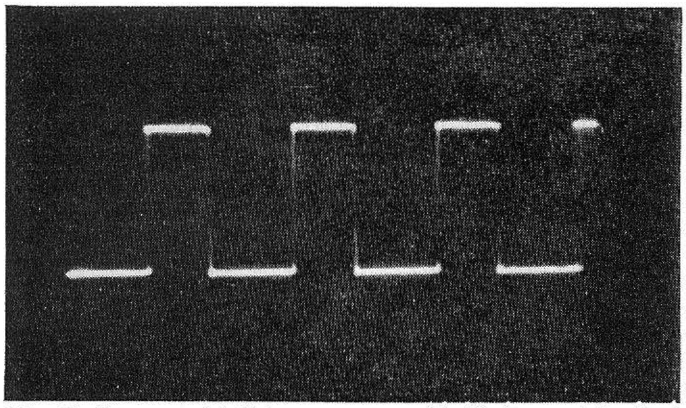

Fig. 18. *Response of 1 Kc/s square wave with 16 ohms resistive load.*

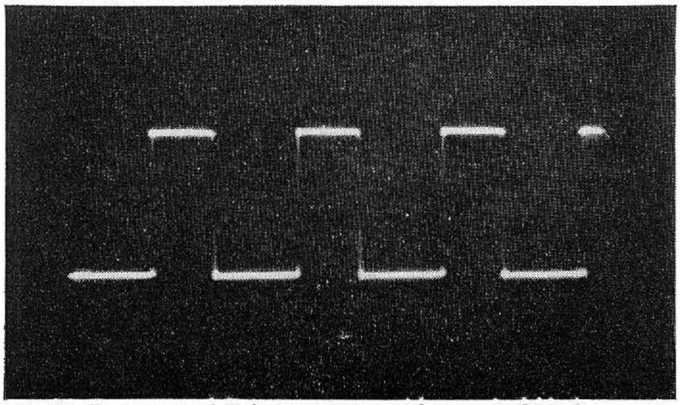

Fig. 19. *Response to 1 Kc/s square wave with a nominal 15 ohms moving coil loudspeaker as the amplifier load.*

A NEW HIGH

THE previous two articles in this series have dealt with (*part one*) the historical survey of the scene and the specification requirements for high grade amplifiers and (*part two*) the design considerations. This part deals with the performance of the completed amplifier, as tested, and concludes the main report presented by the authors. However, since it is probable that many of our readers will wish to follow this further, we have arranged for the preparation of constructional details for publication, and these will be introduced in our next number. These details will follow the lines of the kit version of the amplifier which is available with metalwork.

Voltage frequency response of the amplifier

The measured frequency response of the amplifier at an output power of 1 watt is shown in **fig. 10.** From this it will be seen that the response is virtually flat over the full audio frequency range. The h.f. response is 2 dB's down at 70 Kc/s, and this is due to the phase-advance capacitor in the feedback loop and the falling internal response due to the step-network. This result is the kind to be expected with an exceptional margin of stability. If the amplifier were marginally stable, then the resultant peaking would give an apparently greater bandwidth, but at the expense of a degraded phase response.

Power frequency response and harmonic distortion

This is plotted in **fig. 11,** the distortion level being fixed at 1% for the power measurement. This may seem somewhat high but was chosen as it is just before the amplifier starts to limit, and is therefore a fairly accurate measure of the peak power that the amplifier will handle without serious distortion. As will be seen from the graph, the rated power output is obtainable over the frequency range of 27 c/s to 28 Kc/s with approximately 20 watts available in the middle of the frequency range.

In order to give some idea of the distortion at full power output, a graph is plotted of distortion at 15 watts against frequency. This is shown in **fig. 12.** Here it will be seen that from below 30 c/s to 15 Kc/s the distortion level is less than 0·5%, and over the majority of the range, less than 0·1%. The distortion at the high frequency end can be reduced by decreasing the size of the step in the h.f. response. As already explained, the step design is a compromise of the h.f. power and stability margin. A signal of 400 mV r.m.s. is required for a power output of 15 watts r.m.s.

Intermodulation product distortion

This was measured using the S.M.P.E. method. The frequencies used were 40 c/s and 6 Kc/s in the ratio of 4:1 respectively. The peak waveform amplitude was set to that for 15 watts sine wave output, and the

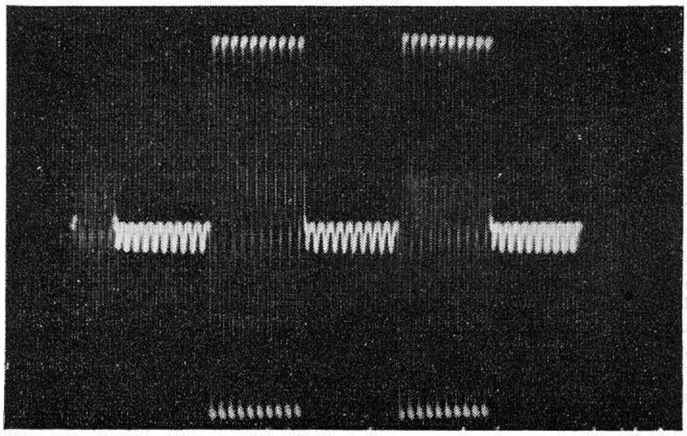

Fig. 24. *As* **Fig. 21** *but with 100% voltage overload compared with the rated 15 watts output.*

FIDELITY AMPLIFIER

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—PART 3 SUMMARY AND TEST DETAILS

amplitude of the intermodulation products measured. These were 0·092% of second order and 0·074% of third order, making a total intermodulation level of 0·12%. This is considered to be very satisfactory.

Phase/frequency response

This is plotted in **fig. 10.** It will be seen that the phase response is steady at about 180 degrees over the full audio frequency range. As there are no rapid phase-shifts it could be expected that the transient response may be correspondingly good. This is borne out in fact by the test results shown later.

Stability

The stability of the amplifier was determined by taking the gain and phase characteristics over the frequency range of 0·1 c/s to 3·0 Mc/s. This very wide range is necessary if the gain and phase margins are to be obtained. The open loop gain and phase plots are shown in **fig. 13,** where it will be seen that the h.f. phase margin is 90 degrees (at several frequencies) and the gain margin is 25 dB at 2·6 Mc/s. These values are plotted in the form of a Nyquist diagram in **figs. 14** and **15.** These values were taken with a resistive load of 16 ohms in the 16 ohm output position which is the worst condition. The stability factor is improved in the 4 ohm output condition.

These results mean that the amplifier is unconditionally stable for any type of output load, as a pure capacitive load can only add a maximum of 90 degrees of phase shift compared with a resistive load, and in practice somewhat less than this due to the output resistance of the transformer. This has been demonstrated to be true; no combination of capacitive and resistive loads has been found to make the amplifier unstable. Indeed, there has always been a phase margin of more than 20 degrees irrespective of the load type. The gain margin drops on capacitive load as would be imagined, but has not been measured under any condition at less than 6 dB.

To try to give a complete picture of what happens to the amplifier response under all conditions of loading would be an impossible task, but the maintenance of adequate stability is shown by the square wave tests referred to under " Transient Distortion ". It is worth mentioning that one early prototype of the amplifier had a gain margin of over 30 dB when measured by the method of reducing the feedback resistor as mentioned before. When checked by the open loop Nyquist diagram, the actual gain margin proved to be only 8 dB; it is suggested that reviewers of amplifiers bear this in mind for future stability tests.

Referring to the low frequency end of the diagram little comment is necessary except that as the gain and phase margins are adequate (—20 dB and 82° respectively) there will be no tendency for switching transients to shock excite low frequency oscillations in the system.

Transient distortion

The completed amplifier was tested with square waves having a rise-time of better than 0·5 microsecond. The results are shown in **figs. 16-20** inclusive. The tests at 10 Kc/s show that ringing is nearly entirely absent, the overshoot being negligible. In fact, the total deviation from a flat top is only about 2%. The rise-time is also very good and is in the region of 2·5 microseconds. The effect of adding a capacitance of 0·01 microfarad across the 16 ohm load cannot be distinguished and so the length of loudspeaker leads will not affect the the stability or transient response to any noticeable degree.

The 1 Kc/s square waves are very good, as would be expected, and it is noticeable that using a moving-coil loudspeaker load does not affect their shape. This was also expected but it was nice to see the theory verified. The square wave at 50 c/s shows some droop as would be expected and corresponds to a low-frequency cut-off of about 5 c/s. As the amplifier is not intended to operate at sub-sonic frequencies this result was also considered to be quite satisfactory. The overall transient response is therefore very good under all normal load conditions and should produce no colouration.

Transient overload recovery

The amplifier was tested on resistive load with tone-burst waveforms from 10 Kc/s down to 50 c/s. Other types of load had no observable

(*continued on page 152*)

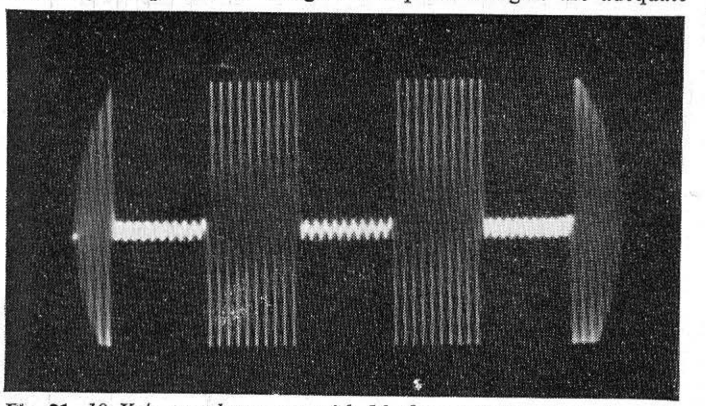


Fig. 21. *10 Kc/s tone-burst test with 16 ohms resistive load. 15 watts r.m.s. output during the period of maximum output.*

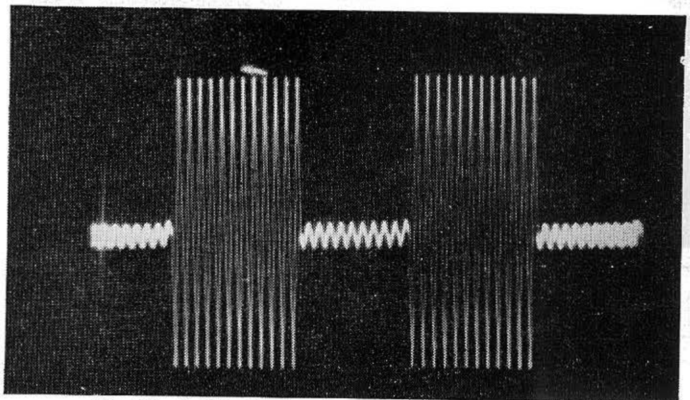


Fig. 22. *As* **Fig. 21** *but the tone frequency is 1 Kc/s.*

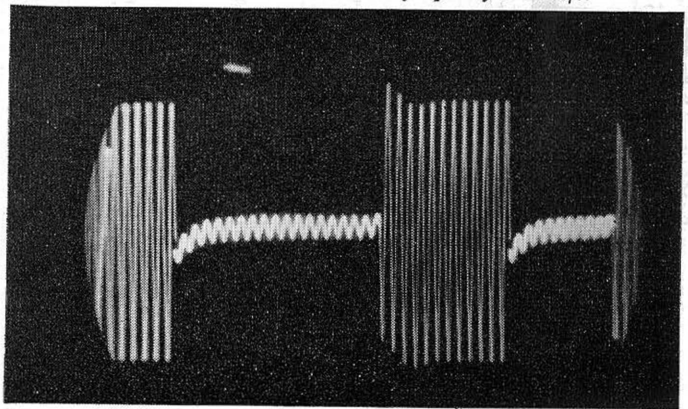


Fig. 23. *As* **Fig. 21** *but the tone frequency is 50 c/s.*

* *Senior lecturer in electrical engineering at Bradford Institute of Technology.*
† *Radford Electronics Ltd. Bristol.*

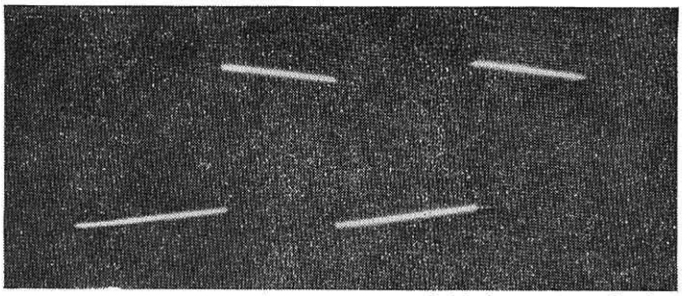


Fig. 20. *Response to 50 c/s square wave with 16 ohms resistive load.*

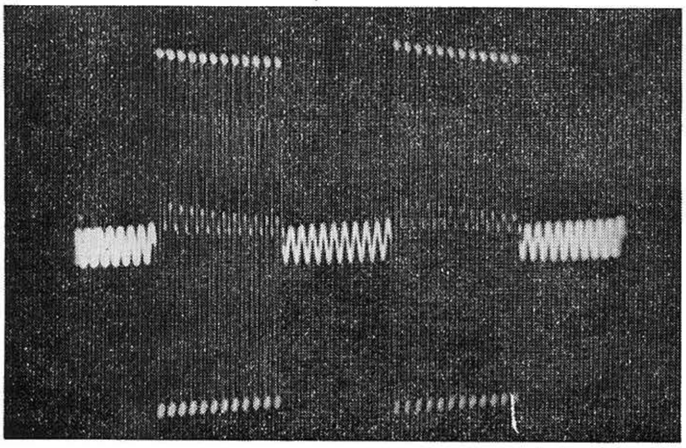
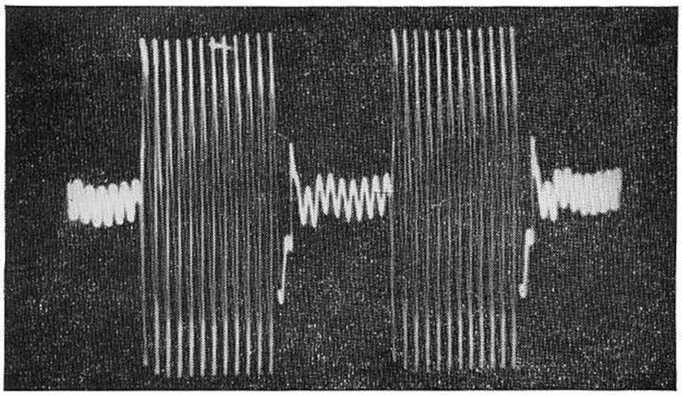
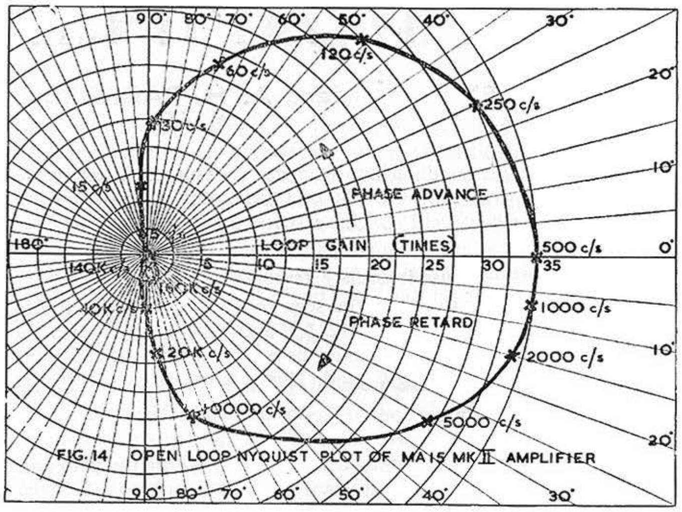


Fig. 25. *As* **Fig. 22** *but with 100% voltage overload compared with 15 watts output.* **Fig. 26.** *As* **Fig. 23** *but with 100% voltage overload compared with 15 watts output.*



TEST PERFORM



effect and the results are therefore not included. The waveforms corresponding to full power output are very good, except for the appearance of the one taken at 50 c/s. The transient shift in the zero axis was unexpected but was traced to the subharmonics of the tone-burst extending down to zero frequency. The tone-burst can be considered as a square-wave modulated sine-wave and will therefore have a strong retinue of harmonics extending both sides of the carrier frequency. Feeding the tone-burst into an R/C circuit with a low frequency cut-off at 5 c/s produced a nearly identical result, and so the effect is due to the amplifier response not extending down to d.c. Feeding in the tone-burst through a high-pass filter cutting off at 20 c/s gave identical amplifier input and output waveforms. This is another reason for incorporating a rumble filter in the pre-amplifier.

The transient overload of the amplifier was then tested by doubling the voltage drive of the tone-burst. This would correspond with a power output of 60 watts if the amplifier could develop such a power. As was expected, the amplifier limited the peak output, but the recovery time in all instances was less than half-a-cycle of the input signal. There is therefore no tendency towards blocking, and consequently no distressing results if the amplifier is accidentally overloaded. This result is very satisfactory and it is difficult to see how it could be improved.

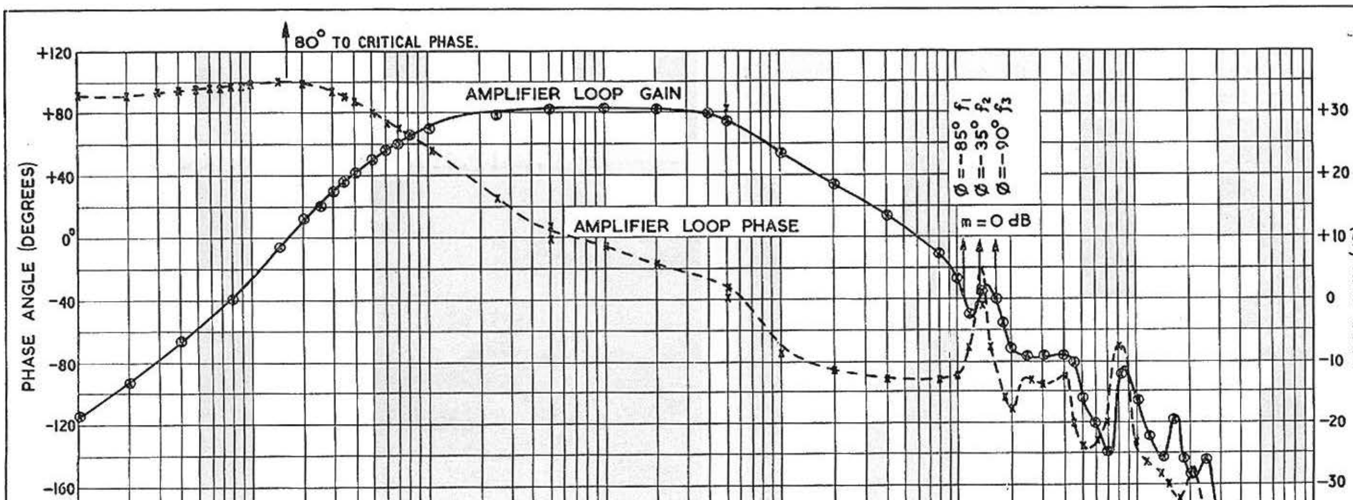


Fig. 13. *Open-loop gain and phase response of the complete amplifier. The minimum phase margin is 90° at a frequency of 162 Kc/s and the gain margin is =26 dB, at a frequency of 2.4 Mc/s.*

NCE OF THE AMPLIFIER

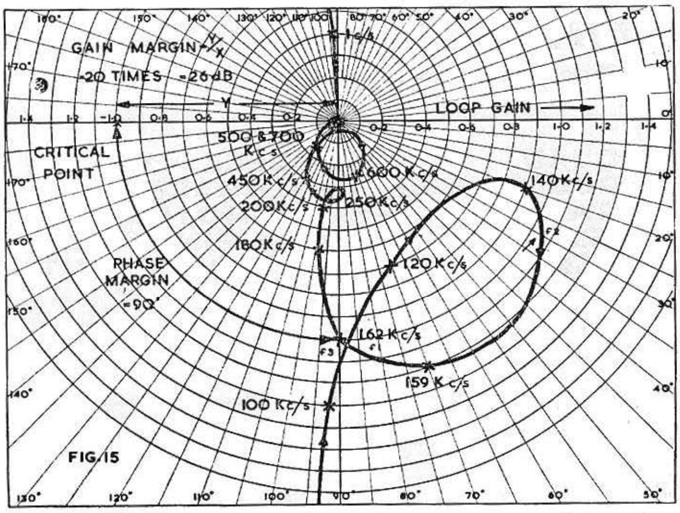


Fig. 15. *Expanded open-loop Nyquist diagram showing h.f. gain phase margins.*

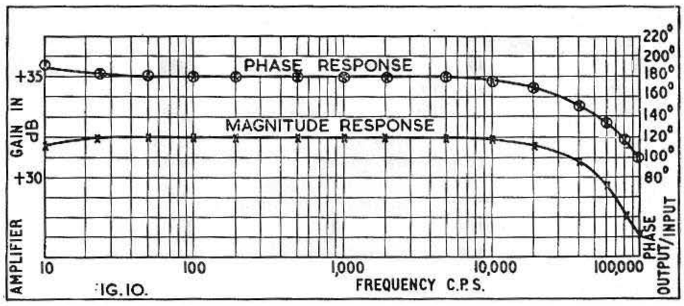


Fig. 10. *Gain and phase response of the complete amplifier, output power 1 watt (constant).*

Transfer linearity

As would be expected from the very low distortion figures the transfer linearity is very good, and limiting of the peaks is the first factor restricting the output power. There is therefore little point in giving the transfer characteristics, as on an oscillograph it appears as a perfectly straight slanting line that limits abruptly at each end. As the distortion is so low the distortion measurement gives a much better indication of linearity than the transfer characteristic could do.

Damping factor

This was measured as 60 at 1000 c/s and is considered adequately high for all applications.

Summary

The fact that so-called high fidelity feedback amplifiers *do* sound different from each other has been bothering engineers for some time, and independent investigations have been carried out to find the reason. There appears to be no doubt that the stability margin is the factor responsible, although the actual ringing frequencies generated by transient inputs must be important. The aim must be for the highest stability in respect of gain and phase with ringing frequencies as high as possible with low amplitudes. It has been reported (*Ref. 10*) that amplifiers having a stability margin greater than 18 dB cannot be detected as being different from each other.

In this summary it would not be correct to omit reference to amplifiers using transistors. In fact, it may appear anachronistic to be designing valve a.f. power amplifiers at this time. One of the authors (A.R.B.) has been intimately concerned with transistor technology in the last few years, and is particularly interested in the design of transistor power amplifiers. It is considered that the manufacture of a transistor amplifier having the specification of the amplifier described is economically impracticable at the present time.

The description of this amplifier has been somewhat lengthy but necessary to convey the rather comprehensive requirements in respect of a modern a.f. power amplifier. Many amplifiers are used in the chain from the studio microphone to the loudspeaker, and cascaded distortions are normally additive. It is hoped this article will provide some guide to the specification of future amplifiers for use in this chain.

In conclusion, it is desired to acknowledge the use of the facilities at the Bradford Institute of Technology where much of the academic work and testing has been carried out.

CONSTRUCTIONAL DETAILS

For the benefit of those readers who prefer to follow their own methods of layout and construction it should be noted that all the components of the amplifier, including the transformers, are available commercially. The construction is not critical in respect of layout although a few points, if obvious, may be worth mentioning. For example, if a possible hum level of —80 dB below 15 watts is anticipated, then care must be taken in respect of hum loops. Even if a low flux density grain oriented mains transformer is used the output transformer should be kept at least 3 inches away from it to avoid injected hum.

The signal earth should be connected to the chassis in one place only (the input socket generally). The filter capacitors must be insulated from the chassis and earthed independently to prevent ripple currents from flowing in signal circuits.

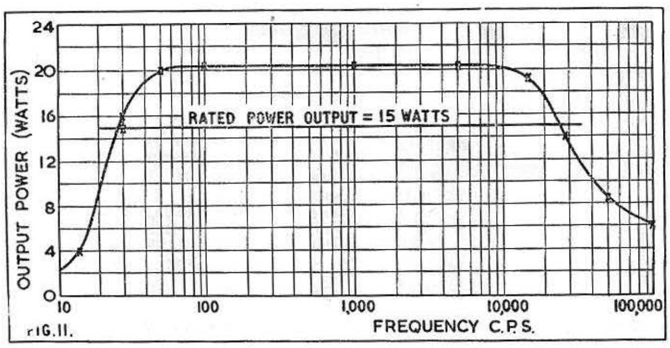


Fig. 11. *Variation in the available power output for a fixed distortion of 1%.*

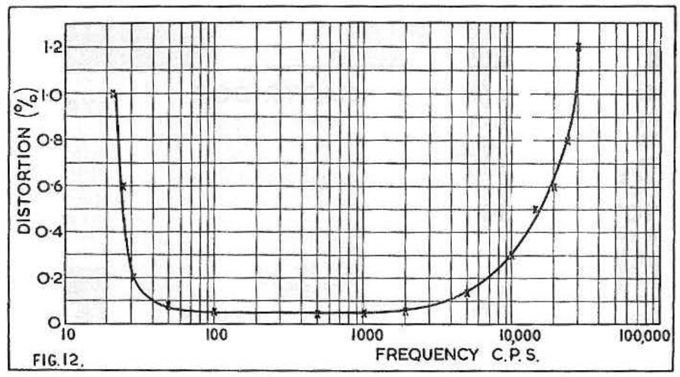


Fig. 12. *Variation in the amplifier distortion with frequency. Output power = 15 watts r.m.s. (constant).*

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A NEW HIGH FIDELITY

PART FOUR—GENERAL CONSTRUCTION

FOR the more technical and experienced readers, the practical information given in this issue will be unnecessary, although probably of interest. The more practical constructor will no doubt find it not only of considerable help, but instructive. The method of construction of the amplifier described, and the component parts listed, are relevant to the factory made model. It is appreciated that some constructors will wish to make the amplifier in a different form from that described. This is quite in order providing certain basic requirements are satisfied which will be described later.

Output Transformer

Sufficient information is given for the constructor to build the amplifiers from any parts available, with any form of construction desired. It was stressed in an earlier issue that the output transformer is a very important component and is absolutely necessary if it is desired to obtain the performance specified. This does not mean that no other output transformer will work in this circuit. To obtain the optimum results (however good or bad this might be) with any other transformer, it will be necessary to adjust the feedback values both in regard to amount of feedback and phase correction, and l.f. and h.f. step networks.

Chassis layouts are also shown in respect of a dual version (**fig. 5**) and one of 25 watts nominal power output (**fig. 4**). The relevant component changes are also detailed (**figs. 13, 14**). Further extensions are also practicable such as a dual 25 watt and a dual 35 watt. The 35 watt amplifier is obtained merely by using fixed bias instead of the conventional cathode bias. Tests on a prototype developed for foreign markets showed an output of 35 watts per channel with less than 0·1% distortion mid-band, and an output of 50 watts per channel at less than 1% distortion using the standard 25 watt output transformer.

Constructional details

1. *Chassis.* The amplifier contains five major component parts in the chassis assembly as follows:—Chassis surround, valve chassis, power chassis, base and power chassis cover. The valve and power supply chassis are simple plates with a small returned edge running lengthwise for rigidity and strength (drawings **figs. 6, 7** and **8** refer). This method of construction is used commercially because of its ease of manufacture in quantity by press tools, and the ease of sub-assembly production.

Those who wish to make their own chassis may find it easier to

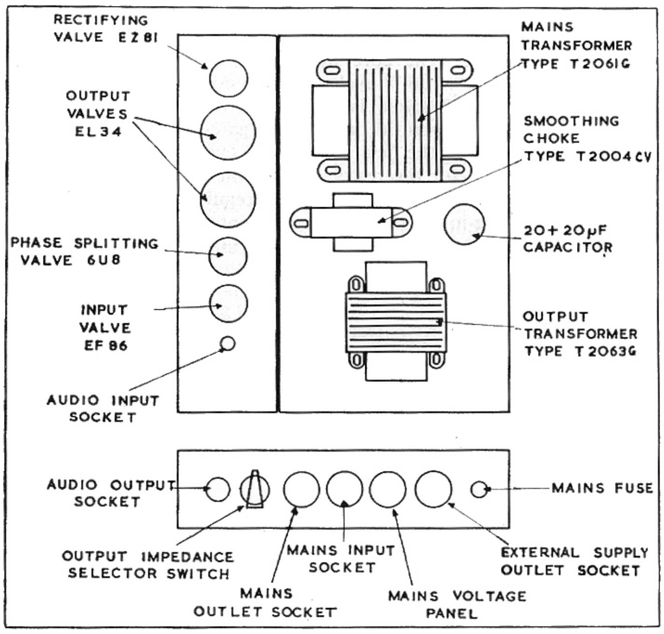

Fig. 3. (*above*) *MA* 15 *Mk. II Pictorial Layout* **Fig. 9.** (*below*) *Parts List MA* 15 *Mk. II*

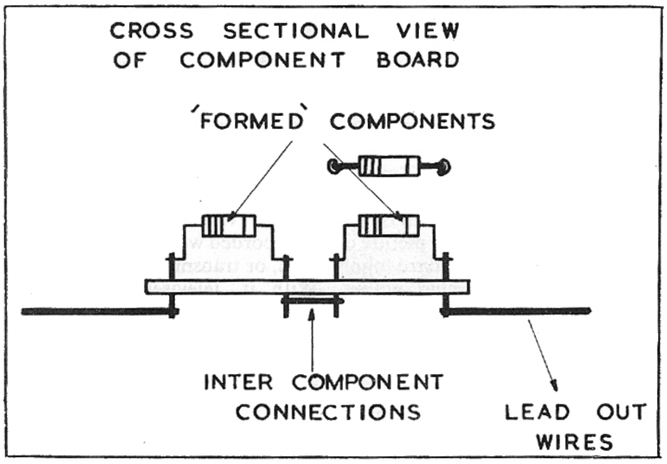

Fig. 2. *Component forming for fitting to pin type board*

MA 15 Mk. II Power Amplifier Parts List **Fig.9.**

Resistors

R.1	100K	5%	½w	R.17	2·2K	5%	½w
R.2	100 ohm	5%	½w	R.18	2·2K	5%	½w
R.3	2·2K	5%	½w	R.19*	390 ohm		4w
R.4	330K	5%	½w	R.20*	390 ohm		4w
R.5*	100K	5%	½w	R.21	1K	5%	½w
R.6	2·7K	5%	½w	R.22	1K	5%	½w
R.7	220K	5%	½w	R.23	560 ohm	5%	½w
R.8	1M	5%	½w	R.24	3·9K	5%	½w
R.9	33K	5%	1w	R.25	390 ohm	5%	½w
R.10	8·2K	5%	½w	R.26	2·7K	5%	½w
R.11	33K	5%	½w	R.27	270 ohm	5%	½w
R.12	39K	5%	½w	R.28	1·8K	5%	½w
R.13	1M	5%	½w	R.29	8·2K	5%	½w
R.14	1M	5%	½w	R.30	CZ3		
R.15	470K	5%	½w	R.31	6·8K	5%	1w
R.16	470K	5%	½w	R.32	68K	5%	1w

*All resistors are High Stability Carbon, except R.5 (Metal Oxide) R.19 and R.20 (Wirewound), R.30 (Brimistor).

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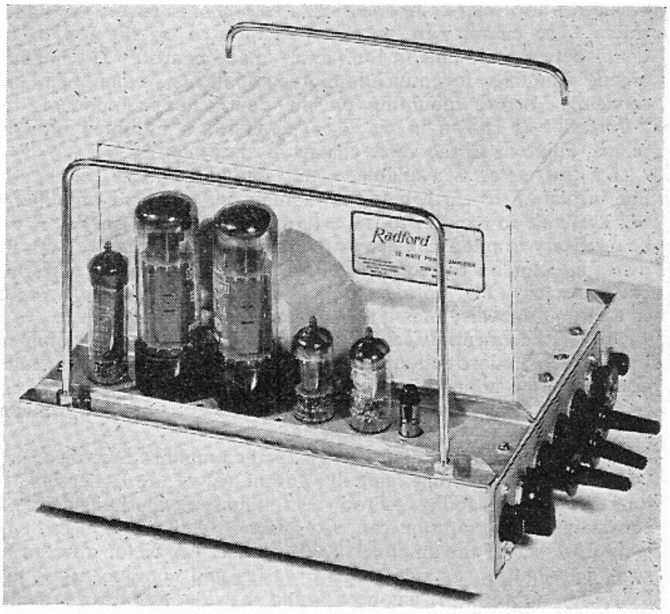
AMPLIFIER

The following text, drawings and components list cover the basic constructional details for the Radford MA 15 amplifier. However, as will be seen from the text, slight modifications will enable constructors to change the product into its MA 25 Mk 2 form. Also, by extending the construction, the stereo version, STA 15, can be produced. In order to avoid interfering with the authors' manuscript, all figure numbers have been printed, but the diagrams not required for the MA 15 have been omitted, and similarly none of the metalwork drawings has been printed. These unpublished drawings are now being lithographed, and will be available by September 1st (price 1/9d, post paid) for readers who would like them. It should be noted that Nos 2, 3, 9, 10, 11, 13, 14, 15, 16 and (front cover) 17 are used in this number. Nos 1, 4, 5, 6, 7, 8 and 12 are omitted.



use the conventional inverted tray. It is advisable to get the metal parts zinc-plated if possible. Since the bread-board disappeared, one of the great problems of the home constructor has been to obtain metal work of a suitable type. The cost of obtaining the "one-off" chassis, from the local metalworking firm, with all the holes cleanly punched is normally quite prohibitive. The amateur, therefore, often has to resort to buying a standard chassis blank, drilling the small holes and getting the larger holes out with tank or fly cutters.

It will be noticed that holes or slots are provided for cooling. The amplifier base should also have louvres or some other air inlet and be fitted with stand-off buffers.

2. *Wiring.* A description is given of the methods of wiring of the factory built amplifier. Layout is not critical, although if a hum level of −85 dB below 15 watts is anticipated, it can only be obtained if care is taken to eliminate hum loops and common impedance paths. (Average values on production amplifiers are −85 dB to −90 dB below 15 watts.)

The electrical earth should be connected to the chassis in one place only (near the input socket generally). The filter capacitor should be insulated from the chassis and earthed independently to prevent ripple currents flowing in signal circuits. Suggested earthing arrangements are shown in **fig. 1**.

The basic chassis wiring is shown in **fig. 10**. This is mainly concerned with connections from the mains and output transformers, and the smoothing choke, and must be done before the terminal board can be fitted. Each wire has been allocated a number to ease tracing of wires. These wires may be either sleeved or laced together.

The component board should be considered as a sub-assembly. All the components can be mounted and flying leads connected for subsequent fixing and wiring into the chassis after the basic chassis wiring has been done. The factory made board uses high grade SRPB suitable for tropical use, but commercial grade is suitable for less arduous conditions. The values of components mounted on the board can be obtained by reference to the parts list, **fig. 9**. Lead

Capacitors			
C.1	8µf	450v	Electrolytic
C.2	0·1µf	400v	Polyester
C.3	500pf	350v	Polystyrene 5%
C.4	50µf	12v	Electrolytic
C.5	2µf	350v	”
C.6	1µf	125v	Polyester
C.7	0·0047µf	400v	”
C.8	0·1µf	400v	”
C.9	0·1µf	400v	”
C.10	16µf	450v	Electrolytic
C.11	50µf	50v	”
C.12	50µf	50v	”
C.13	20µf +	450v	”
C.14	20µf		
C.15	270pf	150v	Polystyrene 5%
C.16	1,000pf	150v	”
C.17	340pf	150v	”
C.18	8µf	450v	Electrolytic (in same can as C.1)
Valves			
V.1	EF86		
V.2	6U8 (ECF82)		
V.3	EL34		
V.4	EL34		
V.5	EZ81		
Transformers			
T.1	Output	T.2063/G	Radford Electronics Ltd.
T.2	Mains	T.2061/G	”
L.1	Choke	T.2004/CV	”

Miscellaneous Components		
Valveholders—International Octal	2 off	McMurdo
Valveholders—B9A	3 off	McMurdo
Phono Input Socket 79/756		Carr Fastener
Phono Input Insulation Plate 79/780		Carr Fastener
Mounting Plate for 20+20µf Electrolytic		
Output Socket 2 pin 76/092		Carr Fastener
Output Plug 2 pin 76/1054		Carr Fastener
4 pole 3 way Switch		A.B. Metal
Mains Outlet Plug and Socket P.438		Bulgin
Mains Input Plug and Socket P.360		Bulgin
Voltage Selector 81/118		Carr Fastener
Fuseholder L.578		Belling & Lee
Fuse Link $\frac{5}{8} \times \frac{3}{16}$ in. 2·5 amp		
$\frac{3}{8}$ in. Grommets	2 off	
$\frac{1}{2}$ in. Grommets	3 off	
Stand-off Terminals W.6003	4 off	Harwin
6 pin Plug and Socket P.194		Bulgin
Phono Plug		Carr Fastener
Pointer Knob		
Chassis Metalwork		
Tagboard approx. $2\frac{1}{2} \times 7\frac{1}{2} \times \frac{3}{32}$ in.		
Double Taper Panel Pins H.2081	72 off	Harwin
P.V.C. insulated connecting wires 7/0076 in.	14/0076 in.	
Sleeving, Tinned Copper Wire, Screws, Nuts, Washers.		

(continued overleaf)

out wires and inter-component links are connected on the underside. (The circuit part numbers are printed on the top of the board between the pins in the factory made board.) This method greatly simplifies indentification of components, replacement and servicing. Before mounting on the board, the components are "formed" as shown in **fig. 2**. The board is mounted on ½-in. pillars, adjacent to the valve chassis as shown in **fig. 11**.

If a test meter is available, a test should be made on the HT line insulation to earth before switching on. If a pre-amplifier is not available to connect to the external supply outlet socket, it will be necessary to put a link across the mains switch circuit. With the mains selector on the correct voltage tapping the mains may then be connected. **Fig. 16** gives typical voltage readings with respect to earth on the valve pins taken with a model 8 Avometer.

One of the essential factors of a successful design for equipment for amateur construction, is flexibility. A design which relies upon accurate adjustments on test and critical voltages, etc., for its performance, is bound to present the constructor with problems sooner or later.

All the specified component parts of the amplifier described have very conservative ratings and the output valves are under-run for long life and reliability. From tests it has been established that considerable deterioration can take place in valves and electrolytic capacitors before the amplifier needs service. A factor affected by such deterioration or sub-standard valves and components is low frequency stability. It has been found that low frequency stability can be maintained more satisfactorily by feeding the screen of V.2 from a potential divider instead of a series resistance. This modification of the original circuit published is shown in **fig. 15** and is incorporated in the wiring diagrams and parts list.

The remainder of the components are mounted on or between the valve holders and insulated standoffs, and in no instance should a component be loosely connected in the wiring. Before fitting the component board the wiring should be given a final check against the wiring diagrams.

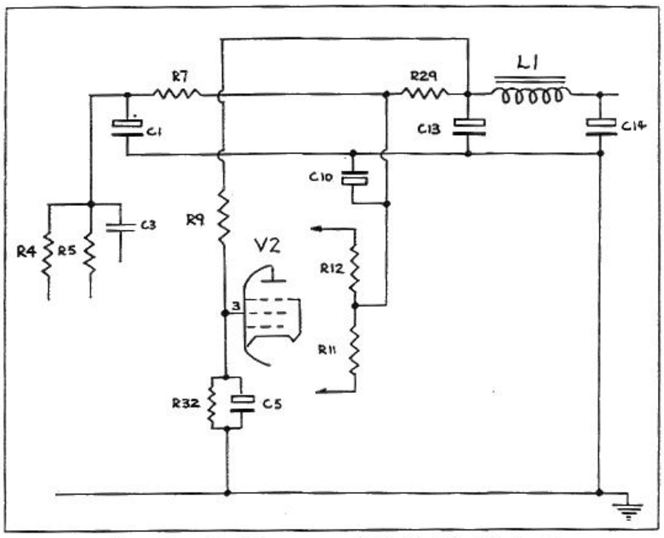


Fig. 15. *Modification to MA 15 Mk. II circuit*

SUMMARISING—(*Construction Order.*)

1. Assemble chassis surround components, and wire (**fig. 10 stage I**).
2. Wire valveholder chassis and fit to surround with power chassis (**fig. 10 stage II**).
3. Wire assembled chassis (**fig. 10 stage III**).
4. Fit transformers and choke, and wire (**fig. 12 stage IV**).
5. Make, fit and wire component board sub-assembly (**fig. 11 stage V**).
6. Inspect carefully to see that everything has been done—CORRECTLY—before testing.

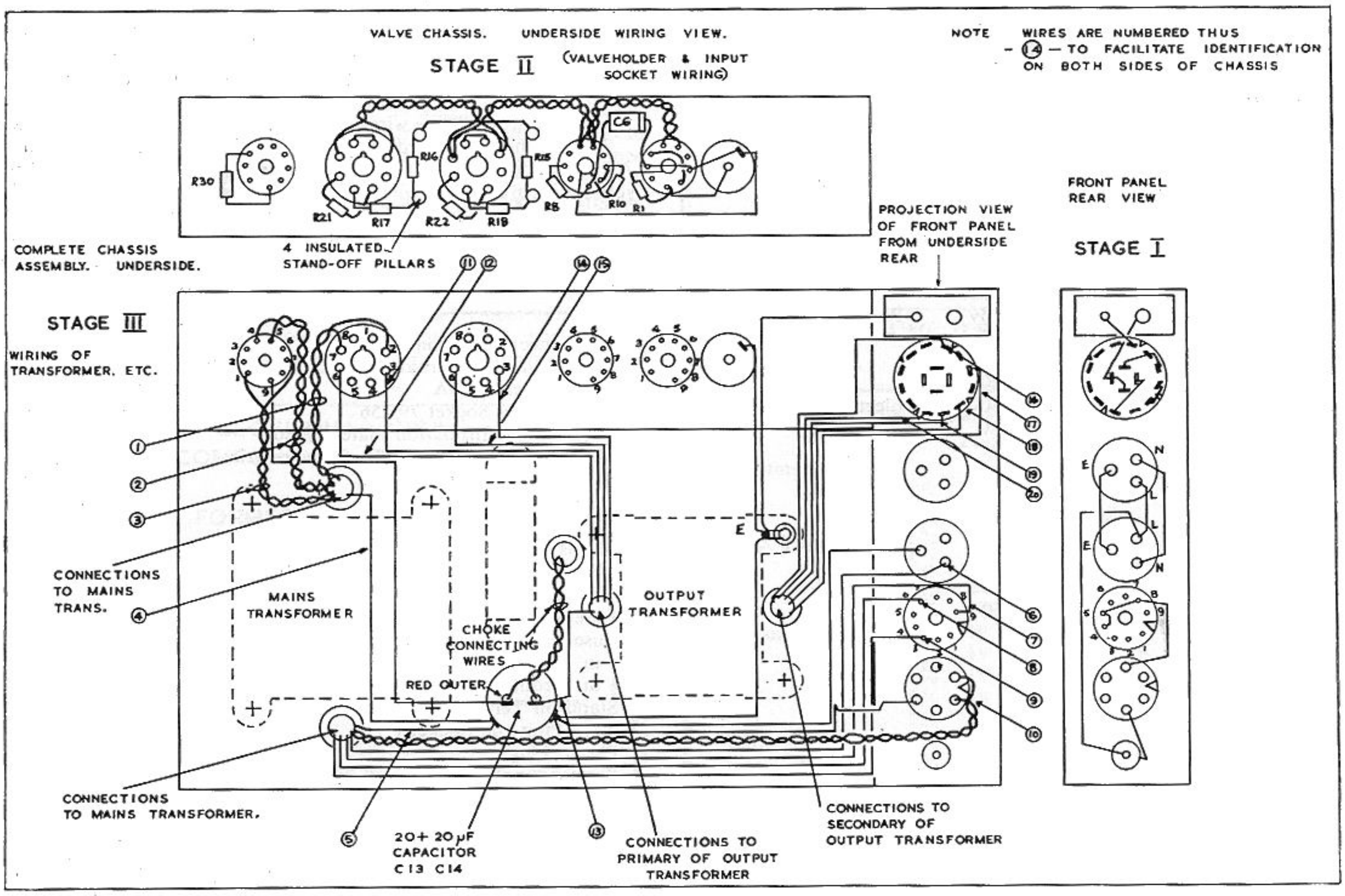


Fig. 10. *Basic chassis wiring stages I, II and III*

Resistors. R.19 470 ohm, 5%, 4w, Wirewound; R.20 470 ohm, 5%, 4w, Wirewound; R.30 CZ6, Brimistor. **Capacitors.** C.3 680pf, 150v, Polystyrene 5%; C.13 40μf and C.14 40μf, 500v, Electrolytic. **Transformers.** T.1 T.2075/U; T.2 T.2062/U; L.1 T.2008/CH, Radford Electronics Ltd. **Valves.** V.3; V.4 KT88; V.5 GZ34.

Fig. 13. *Component changes for a 25 watt amplifier (MA 25 Mk. II)*

Resistors. R.30 CZ6, Brimistor. **Capacitors.** C.13 40μf and C.14 40μf, 500v, Electrolytic. **Transformers.** T.2 T.2064/U; L.1 T.2072/CH, Radford Electronics Ltd. **Valves.** V.5 GZ34.

Fig. 14. *Component changes for a dual 15 watt amplifier (STA 15).*

Valve	Valve Pin								
	1	2	3	4	5	6	7	8	9
EF86	90v	0	1·9	3·15 a.c.	3·15 a.c.	62	0	1·9	0
6U8	260	26	230	3·15 a.c.	3·15 a.c.	255	40	40	30
EL34	24	3·15 a.c.	365	360	—	—	3·15 a.c.	24	—
EZ81	340 a.c.	—	400	—	—	—	340 a.c.	—	—

Voltage on reservoir capacitor C.14 = 400v.
Voltage on smoothing capacitor C.13 = 375v.
Voltage between pins 4 and 5 of EZ81 = 6·3v.
All readings taken with model Avo 8 meter.
In each case the voltage range which gave the greatest deflection was used.

Fig. 16. *Typical voltage readings on MA 15 Mk. II circuit*

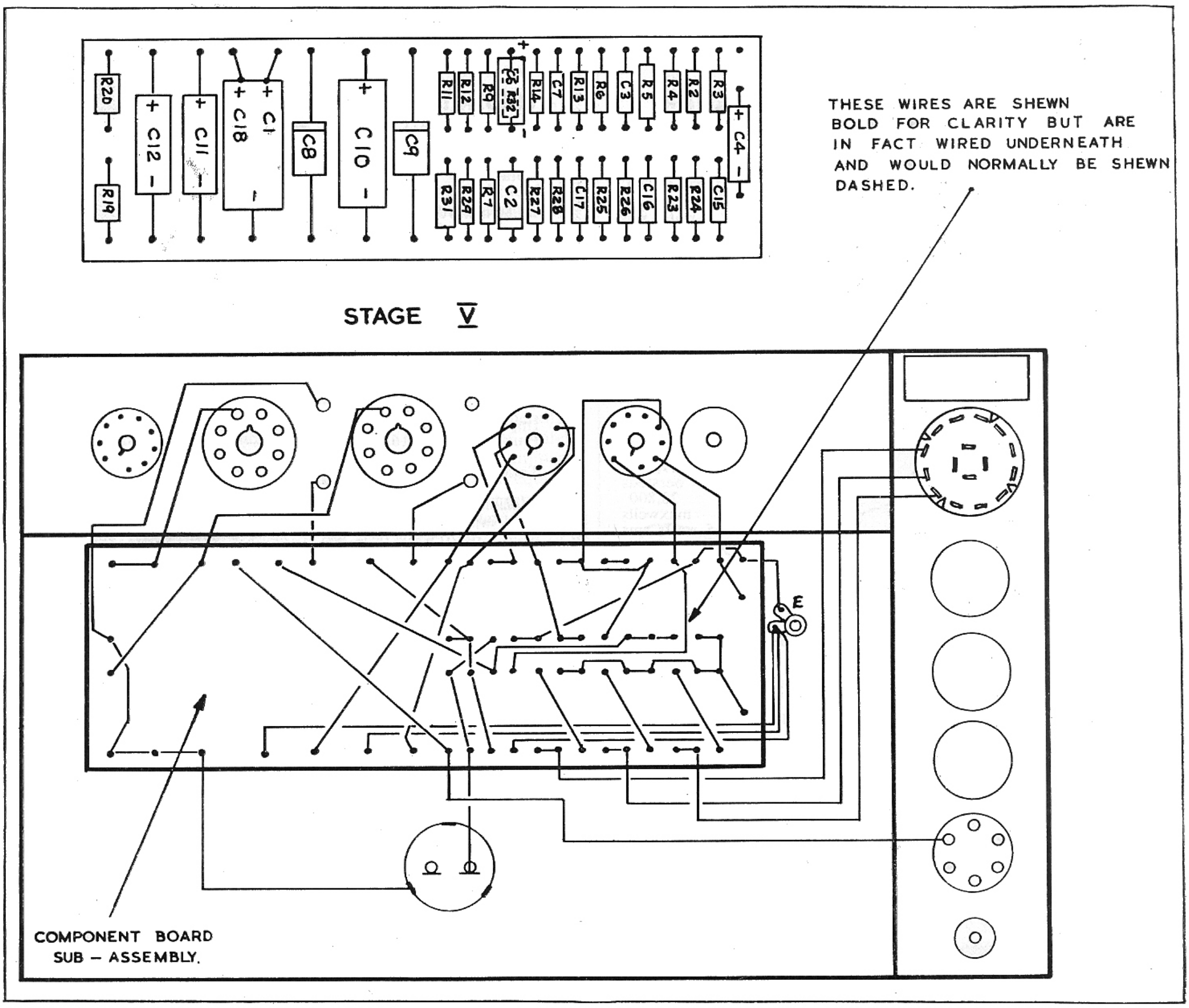


Fig. 11. *Chassis and component board wiring stage V*

Notes on Radford MA15 HiFi News 1962

Outlines the view that the output transformer needed "immense proportions" to achieve low phase shift at low frequencies, and the high frequency shunt resonance was too low to allow a gradual roll-off in forward gain before OT resonance interaction. Then associates the large number of amplifiers that had stability problems with the original Williamson design even though such amplifiers didn't use the required output transformer, which is a bit rich given the output transformer required for the MA15 Mk2.

The MA15 uses a custom output transformer made only by Radford using a specific construction process, and the design specifically needs retuning for any other model of output transformer as the higher 27dB of NFB makes stability tuning so much more custom. The use of better laminations and winding wire and optimisation of winding layout pushes the first resonance out to 150kHz, compared to about 60-70kHz for Williamson's OT.

The design includes an input stage step network to maintain HF stability, and a more complex R/C step network across the feedback resistor.

Mullard 20W circuit V2A 12AX7 triode has Miller capacitance of 75-100pF. Replacing V2A (and V2B) with a pentode significantly improves the HF response as shown in Fig.6, although the final amplifier uses a 6U8 where V2B is a triode. The article only makes a passing reference about the performance of V2B as a triode not being a disadvatage.

UL operation of EL34 imposes a significant Miller capacitance loading on V2A-V2B.
- EL34 pentode has input capacitance of 15pF and Cag of 1.0pF, causing an effective input capacitance of circa 15pf+(1+10)*1pF+4pf=30pF (gain of 10, stray of 4pF).
- EL34 UL has lower gain (x8) but also screen capacitance (5.2Pf) Miller effect of about 0.43x(8+1)x5.2=20pF, which increases effective input capacitance of circa 20+15+(8+1)*1+6=50pF (stray of 6pF).
- compared to the Mullard 20W design, the MA15 lowers the PI anode load resistances from 180k down to 33k-39k (to balance the gain difference between pentode and triode sections).

Williamson use of concertina PI with direct coupling from input stage is summarised as 'introducing drift troubles quite apart from the difficulty in designing such a stage'. The article then procedes to use direct coupling between input stage and PI stage in the MA15. Counter to this view is that the input valve would need to be quite poor to cause a lowered anode voltage from grid leakage, and similarly for the PI valve where both faulty conditions would restrict PI output swing.

Comment is made that output stage pentode mode operation not only has a lower damping factor, but at HF that can interact with high impedance speaker loading to reduce stability margin or cause unstable operation.

LF performance does not show any dramatic phase change from 1-100Hz, with a slight hump around 15Hz, likely due to pushing the coupling cap poles to below a few Hz , and adding some lead compensation with C7/R13. No primary inductance value is identified.