Transistor driver for valve amplifiers

Design for Williamson and other output stages

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There are certainly a lot of valve audio amplifiers still in use, and many of them have an inherent quality of performance that makes it reasonable to give them a thorough repair, with or without an accompanying modernization. The work needed for repair may tend to grow, however, since it may not be sufficient to replace valves and a few electrolytic capacitors. A general degradation of components may have taken place, and in nearly all instances of modernization it should be advantageous to replace the rectifying valve by silicon or maybe selenium rectifiers. So there may be some doubt as to what is really needed and what is worthwhile.

For those who are interested in giving their valve amplifier a positive modernization that will result in obvious improvements, a description is here given of a transistor driving amplifier that can replace the voltage amplifying stages of many existing power amplifiers. The Williamson amplifier[1] has been chosen as a typical example for the discussion that follows, because it is a well-known design. Other amplifier designs that have been used for comparison are those designs by Mullard[2] and by GEC[3].

The original idea was to design an amplifier with a bandwidth sufficiently in excess of the output transformer bandwidth, so that the only phase shift to take account of should be that of the transformer. A d.c. amplifier with a bandwidth of about 1MHz was thought to be sufficient. Direct coupling from the input stage to the signal grids of the output valves leaves the output transformer as the only cause of phase shift at the low frequency end, and the shift tends to only 90°. So there are no problems of instability from negative feedback at the low frequency end, provided that the usual precautions as to supply line filtering are taken.

At the high frequency end of the transformer passband there is usually one main resonant frequency, often at about 100kHz, around which the phase shift passes 90° by a considerable amount but does not reach 180°. It was thought therefore that with a bandwidth of at least 1MHz for the driver, the normal amount of 20dB frequency-independent feedback should be
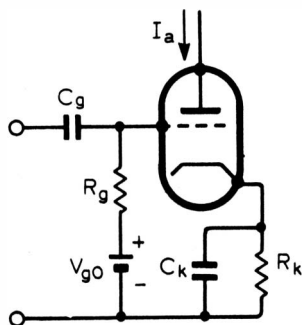

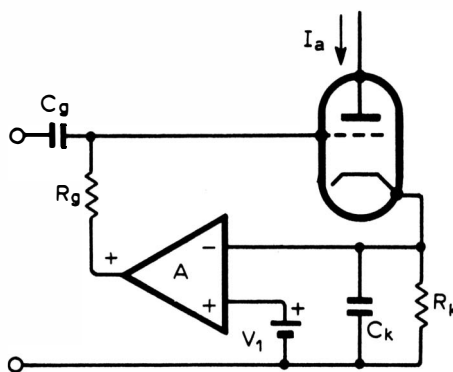Fig. 1. A constant voltage V_{go} is in this circuit added to a normal cathode bias.

Fig. 2. Further development of the circuit in Fig. 1 by means of a gain function A.

allowable without instability. This was found to be the case for several output transformers, at least with a resistive load. With the Williamson transformer and output stage according to the original design, an essentially flat amplitude-frequency response was obtained up to 1MHz, and there was some stability margin.

If a loudspeaker or some load of a complex nature is inserted, the phase shift tends to become too large, and the only way to stability is then to reduce the closed-loop gain. So extended bandwidth is no radical solution for stability at the high frequency end in the same way as direct coupling is for low frequencies. And after all, the aim should not be amplification up to radio frequencies, but an l.f. amplifier with a defined upper frequency limit. This does not mean that it is a wasted effort to start with a large bandwidth for the

driver. On the contrary, by starting with a bandwidth of 1MHz, the high-frequency response can be exactly formed up to this frequency, using simple operational amplifier techniques, and so it can easily be changed to suit different output transformers. The output voltage of the driver is sufficient even for large output tubes such as the KT88; they are assumed to work in class A or AB in the design that follows.

Output valve biasing

When direct coupling to the output valves is used, the grids can still be kept at zero potential for the quiescent point, with a normal cathode bias for class A or AB operation. But this is not necessary and in my opinion not at all the best way. Let us therefore look at other ways of biasing. For the sake of simplicity, single valve biasing is discussed first, and the valves shown as triodes with the usual assumption of zero grid current, i.e. anode and cathode currents are identical. If thus I_a is the anode current of a triode and R_k the outer cathode resistance, the negative grid voltage with a normal cathode bias is $V_{gk} = I_a R_k$.

It is possible, although not often used in practise, to modify the influence of the anode current on this bias voltage by the addition of a constant voltage to the circuit, either in series with the cathode or, normally with less effort, in the grid circuit, shown as the voltage V_{go} in Fig. 1. The grid bias voltage is now $V_{gh} = V_{go} - I_a R_k$.

It is important to note that V_{go} may be positive as indicated in the figure, or negative. In the first case a larger resistance R_k is required than for simple cathode bias, which makes the grid voltage more dependent on the anode current, i.e. there is a better stabilization of the quiescent point. In the case of negative polarity for V_{go}, the grid voltage becomes less dependent on the anode current, as R_k must be diminished. For growing negative values of V_{go}, it becomes in the limit equal to the desired grid voltage. Then R_k must go to zero and the result is a constant grid voltage.

A grid bias that has exactly the same dependence on the combination of a constant voltage and the anode current

as that of Fig. 1, but with improved means for the choice, can be obtained by a circuit as shown in Fig. 2. With the notations according to this figure, and provided that the operational amplifier of voltage gain A has zero offset, the constant part of the grid bias is $V_{go} = AV_1$, and the total grid bias becomes

$$V_{gk} = AV_1 - (1+A)I_aR_k$$

In this circuit R_k can be a small resistance, which is an advantage for large output tubes where the power dissipated in R_k for a normal bias may be considerable. Most important is, however, the ease of adjustment to a desired bias.

The bypass capacitor C_k has retained its function, and the time constant R_kC_k is chosen as for normal cathode bias. However, if R_k is small, so that it causes only negligable feedback by itself, the bias time-constant may be introduced by a separate RC-link, either before or after the amplifier.

In the foregoing figures the bias and signal voltages have been mixed in the usual way by a grid resistor R_g and a coupling capacitor C_g. If a full signal feedback from R_k is wanted, corresponding to an unbypassed cathode resistance in Fig.1, some other type of mixing circuit is needed. This also holds, if direct coupling of the signal to the valve grid is used.

Arrangement of amplifier

To explain the main features of the complete push-pull amplifier, its layout is first shown by the simplified circuit of Fig. 3. The circuit comprises three differential stages, namely a signal input stage, a biasing stage for the sensing of the currents in the output valves, and between these a mixing and amplifying stage that drives the valves. It is a symmetrical circuit throughout for the input signals, and the necessary d.c. balance is obtained at the emitter side of the input stage, in the figure by means of the potentiometer R_1. Another important feature is that the differential stages are all supplied by a current source at the emitter side, instead of just by a common emitter resistor. A high common-mode rejection ratio is thereby obtained, which means that the input signal and the negative feedback around the amplifier can be fed differentially to the input stage without danger of adverse secondary effects.

The current source for the mixing stage, a single transistor in Fig.3, acts with the differential pair as a common-mode amplifier for the signals from the preceding biasing stage, so that the two stages together give a common-mode voltage gain from cathodes to grids that corresponds to the gain function denoted by A in Fig. 2. The gain to a sufficiently good approximation is

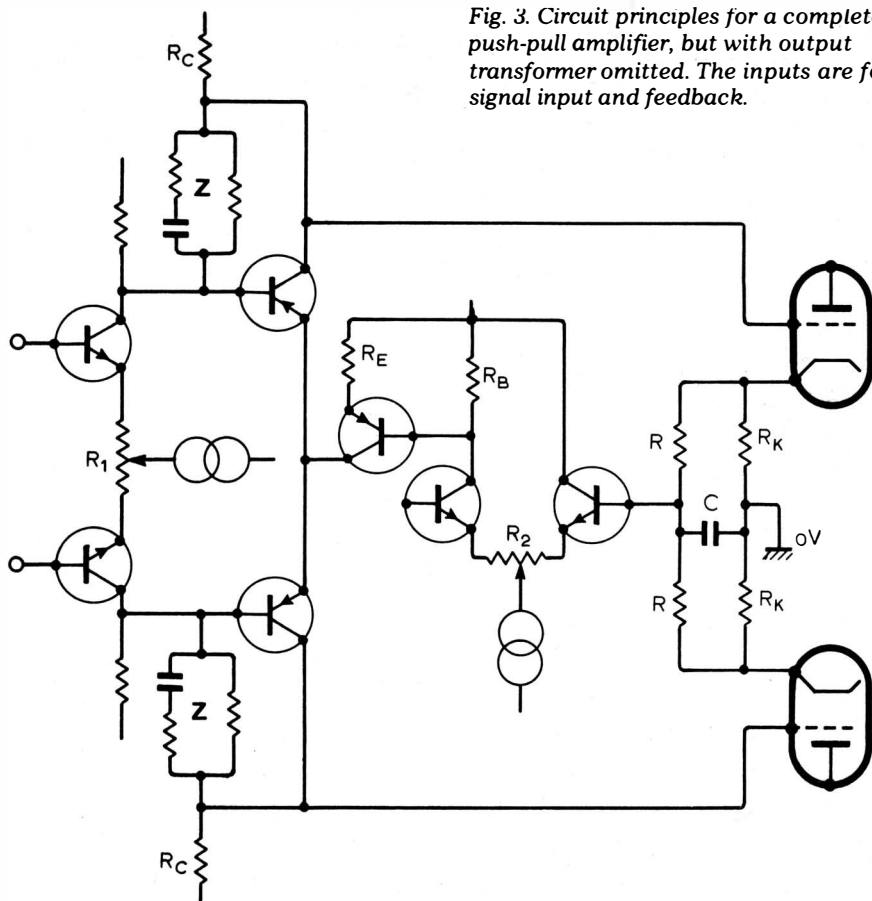$$|A| = \frac{R_B}{R_2} \cdot \frac{R_C}{2R_E}.$$

Fig. 3. Circuit principles for a complete push-pull amplifier, but with output transformer omitted. The inputs are for signal input and feedback.

Resistance R_2 is selectable for choice of voltage gain.

As the amplified part of the bias is a common-mode one, it corresponds to a common cathode resistance with the value $R_kA/2$, and the time constant of the RC-link is $RC/2$. The constant part of the grid bias is simply an offset voltage, effected by an adjustment of the potentiometer R_2, which is therefore found to have the double function of determining the gain by its resistance value and the constant voltage by its adjustment.

The resistances of R_k may be so small that their direct influence on the valve bias becomes negligible. They cause a small lowering of the effective valve transconductances.

Because the collector resistances of the mixing stage become fairly large, there is ample signal amplification available in this stage for local feedback to be applied. This is used in the amplifier for determining the response by means of the impedances Z.

Amplifier design

The complete amplifier is shown in the circuit diagram of Fig. 4. Although the number of components has grown, the fundamental simplicity as evidenced by Fig.3 is retained, and there are not any hidden difficulties such as the need for tricky adjustments or special demands on the power supply voltages, which may vary within large limits. The demands on filtering are not very large either, since the current sources for the differential stages reduce hum. Only the negative high tension voltage needs a certain stabilization.

The input common-emitter long-tailed pair of Fig. 3 is a dual n-p-n transistor Tr_1, and it is completed by an n-channel dual f.e.t. Tr_2, the two transistors of which are used as input source followers. This makes the amplifier compatible with valve amplifiers with regard to input impedance as well as to independency of the characteristics of the driving source. So all that is required of the preamplifier is that it shall give sufficient voltage.

The d.c. balancing potentiometer R_{41}, a 15-turn trimming potentiometer, has been moved away from the main signal path into the f.e.t. source circuit, where it gives a smooth adjustment of the differential balance. By this change the two resistors R_7 and R_8 also become more freely selectable for their function to determine the local feedback of the stage and the gain of the amplifier. They should be matched, so as not to cause additional asymmetry to be balanced out. It is the combination of f.e.t. and bipolar transistor pairs that gives the good input property, together with an easily variable amplification and a large bandwidth. Dual transistors must be used to reduce temperature drift, see later.

The mixing stage has been developed to a cascode configuration, which is very important with regard to harmonic distortion because the output voltage swing is large. It is also important that

the Miller feedback capacitance is kept very low so that the loading on the preceding stage can be controlled as desired, and the amplifier as a whole be given sufficient bandwidth. The main local feedback is by means of the emitter resistors R_{13} and R_{14}, but they need not be matched as their counterparts R_7 and R_8, as the balancing action of R_{41} is amplified by the input stage.

Local feedback by the two impedances Z starts at a value of about 12dB for low frequencies, but increases within the frequency range 20 to 200kHz to about 26 dB. It forms the amplitude response as shown in Fig.5, curve A. The impedances Z do not cause any common-mode feedback but act together for the differential feedback, so they do not need matching for their action. However, matching is needed for the collector loads of transistors Tr_5 and Tr_6 for symmetry in driving the output valves. The two collector resistors R_{15} and R_{16} should be matched, and also the impedances as they also load the collectors.

As to the valve common-mode biasing, there are only two alterations from the simplified circuit of Fig.3. One is that the potentiometer for adjustment of the constant voltage part has been split up in two fixed resistors, R_{19} and R_{20}, and a 15-turn trimming potentiometer, R_{42}. This makes the selection of resistances for a desired value of the amplification fairly easy, and provides for a smooth adjustment of the constant voltage. The other change, mainly for temperature

drift is that the current source for the mixing stage, Tr_9 and Tr_{10}, is a complementary pair amplifier.

The gain as defined by Fig.2 is nearly 70, which means that the bias circuit corresponds to a common-cathode resistor of 350 ohms. A common-mode constant grid voltage of about +5V is added by adjustment of R_{42}. The quiescent grid-cathode voltage is about −45V and the valves work in class A.

A negative feedback that senses the differential direct voltages across the cathode resistors has also been added to the circuit. It consists of the matched resistor pairs R_{31}, R_{32} and R_5, R_6 together with the capacitors C_3, C_4. This feedback is coupled to the amplifier inputs and has an upper frequency limit of about 1Hz. It has the same stabilizing effect on the balance between the tube currents as two separate cathode resistors of 200 ohms, connected together in a long-tailed pair configuration but without influence on the common-mode bias.

All the above values are easily changed for desired bias conditions, but a general discussion of valve biasing is outside the scope of this article.

A capacitance of 22 μF was originally used for C_5, but is omitted in the circuit of Fig. 4. However, output triodes in

class A with a high load impedance is the only case where the capacitance may be omitted to some advantage.

Response and distortion

The amplitude-frequency response of the complete amplifier is shown in Fig. 5: without feedback by curve B, and with 20dB overall negative feedback by curve C. The low-frequency response for small signals is flat down to 10Hz both with and without feedback. Exact curves showing the fall below 10Hz are not interesting, but it is possible to select a value for C_2 that gives an optimum response to square waves at low frequencies.

There is a dip in transformer response at about 50kHz, which cannot be eliminated by simple feedback circuits. It causes some ringing in square-wave tests, which of course has nothing to do with instability. The capacitance of C_6 in the feedback loop has, however, been chosen so large that it has a damping influence on the ringing. The series resistance of R_{36} has been chosen as a compromise to give about the same frequency response when loaded by a certain broadband loadspeaker as with a resistive load. A capacitance inserted as C_6 in the feedback loop without a series resistance often gives a good frequency response with a resistive load, but oscillations when a loadspeaker is connected. Its influence on the feedback must therefore always be carefully checked.

The branch R_{22} and C_7 between the

Fig. 4. Complete circuit diagram of the power amplifier. Valves work in class A as C_5 is made zero (see text).
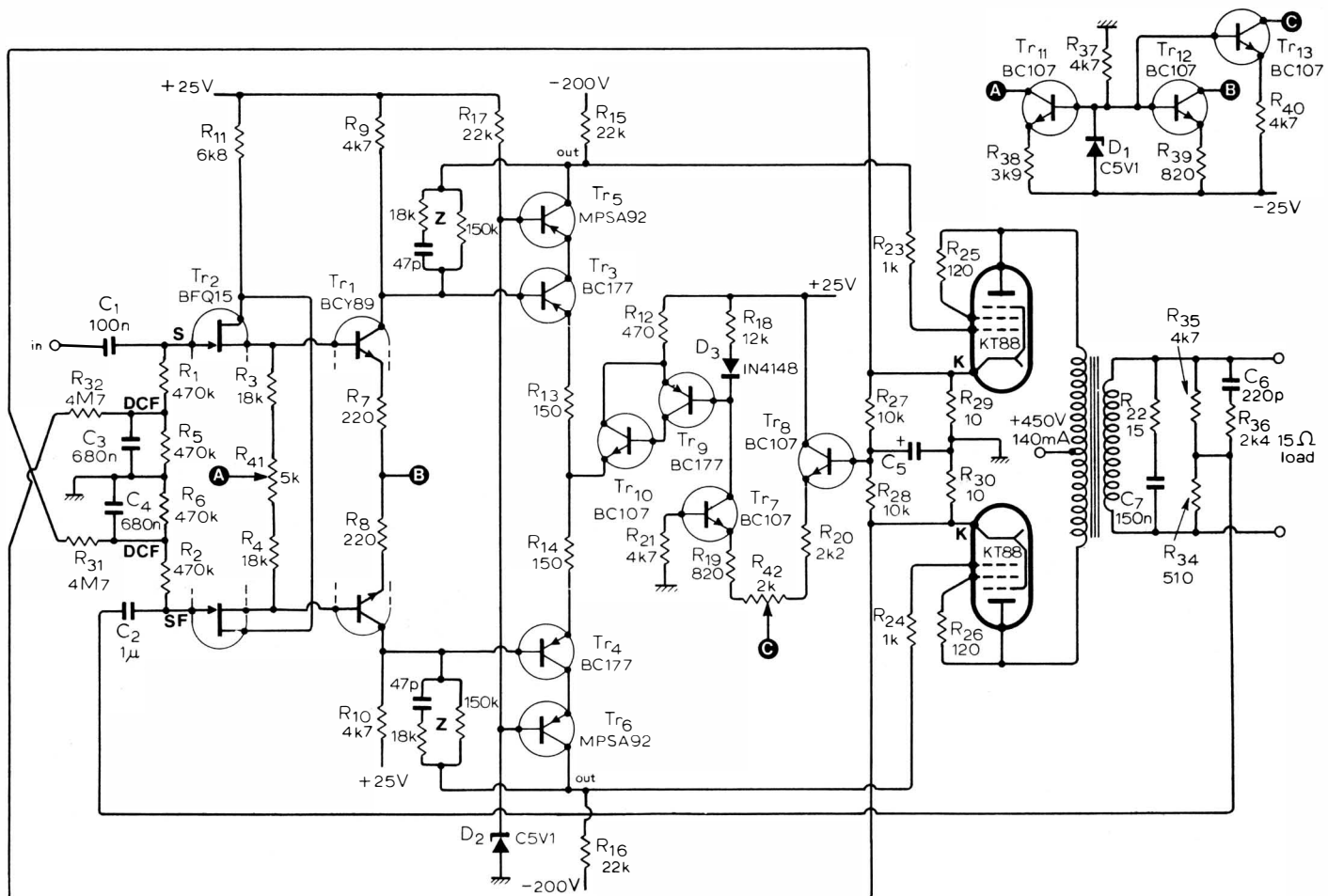
output terminals has been found valuable with several output transformers, and is therefore recommended. It has no effect on the response within the audible band, but represents a resistive load at high frequencies. Values are not critical.

It has been an aim to choose about the same high frequency limit for the response without feedback as in the improved version design by Williamson to make a comparison of the final result fairly easy. It could be an advantage, however, to choose a lower high frequency limit by a change of the impedances Z.

Total harmonic distortion of the driver is quite low. For 30V r.m.s. output on each side it is only about 0.05% at low frequencies and rises to about 0.1% at 20kHz. This leads to a low distortion for the whole amplifier even without overall feedback: at 1kHz this distortion is only 0.08% for 10W and 0.2% for 15W output power.

The overall feedback works fully within the audible band, but the maximum output power falls at the low and high frequency ends. At a distortion of less than a quarter of a percent the available output power with resistive load is 20W at middle frequencies and 15W at 20Hz and 15kHz.

The total harmonic distortion, measured at 20Hz, 1kHz, and 15kHz and with an output power of 10 and 15W is summarized in the table below. The figures are given in percentage distortion, but include what there may be of hum and noise in the prototype amplifier.

Power output (W)	Total harmonic distortion (%)		
	20Hz	1kHz	15kHz
10	0.05	0.01	0.1
15	0.1	0.02	0.25

Circuit working conditions

In all d.c. amplifiers there is a temperature drift that must be taken account of. In this case there are really two, namely a common-mode drift in the biasing circuit and a differential drift for the signal path. Drift in the output valves is not considered.

An obvious cause of common-mode bias drift is the difference in change of base-emitter voltage with temperature for the transistor pair Tr_7 and Tr_8. The two transistors should be of the same current amplification class, BC107A in the prototypes, in which case the difference may be assumed to be 0.1mV/deg C at the most. The drift voltage is equal in its effect to a false reading of the direct input voltage on the base of Tr_8, and results in a corresponding shift of the anode currents of the valves.

If an ambient temperature change as large as ±20deg C is assumed, the false reading is not more than ±2mV, which is less than 0.3% of the above-
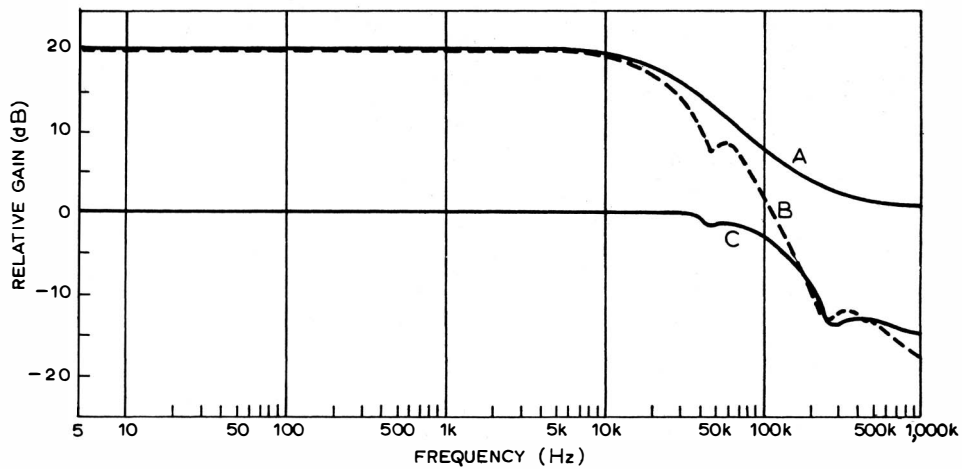

Fig. 5. Amplitude frequency response curves for the driver (A) and for the complete amplifier without (B) and with feedback (C).

mentioned d.c. input voltage, being about 700mV. However, the two transistors must be mounted close together, so that they experience the same ambient temperature change. Preferably they should be plastics transistors and clamped together, but a dual transistor is not necessary.

There is also a temperature drift from differences in the internal heating of the transistors, for instance at power supply variations. This is kept low by means of low collector currents. For the same reason the design assures a small collector current for Tr_9 in the current source, and the transistor drift is partly balanced out by D_3. The balance is not as good as for a couple of equal transistors, but here the drift is inside the feedback loop and has less influence on the valve currents, about one third of that of the preceding transistor pair.

The main cause of differential drift is the input dual f.e.t. Although its thermal drift of gate-source voltage difference for specified working conditions is less than $40\mu V$/deg C, its drift in the circuit may be larger, on account of shifts of quiescent points. There is also up to $10\mu V$/deg C drift in the dual bipolar transistor, and some additional drift from the transistor pair Tr_3, Tr_4. As a summation a temperature drift of up to $100\ \mu V$/deg C referred to the input of the amplifier will be assumed.

To find what the above drift means as a drift in quiescent current for the valves, the d.c. feedback from the cathodes to the input circuit will first be assumed inoperative. The differential voltage amplification to the grids is 450 and the transconductance is 10mA/V, which gives 0.45mA/deg C differential drift for the anode currents, or ±9mA for a change in temperature of ±20deg C. This is at the limit of what should be allowable, but, on the other hand, fairly wide limits as to the causes are assumed.

The picture of drift changes radically, however, if the d.c. differential feedback
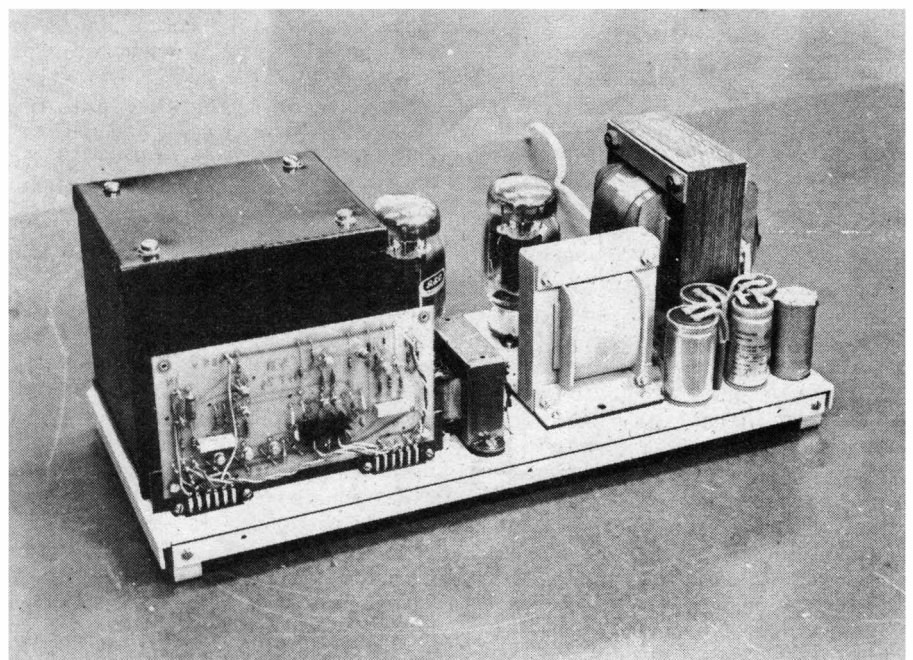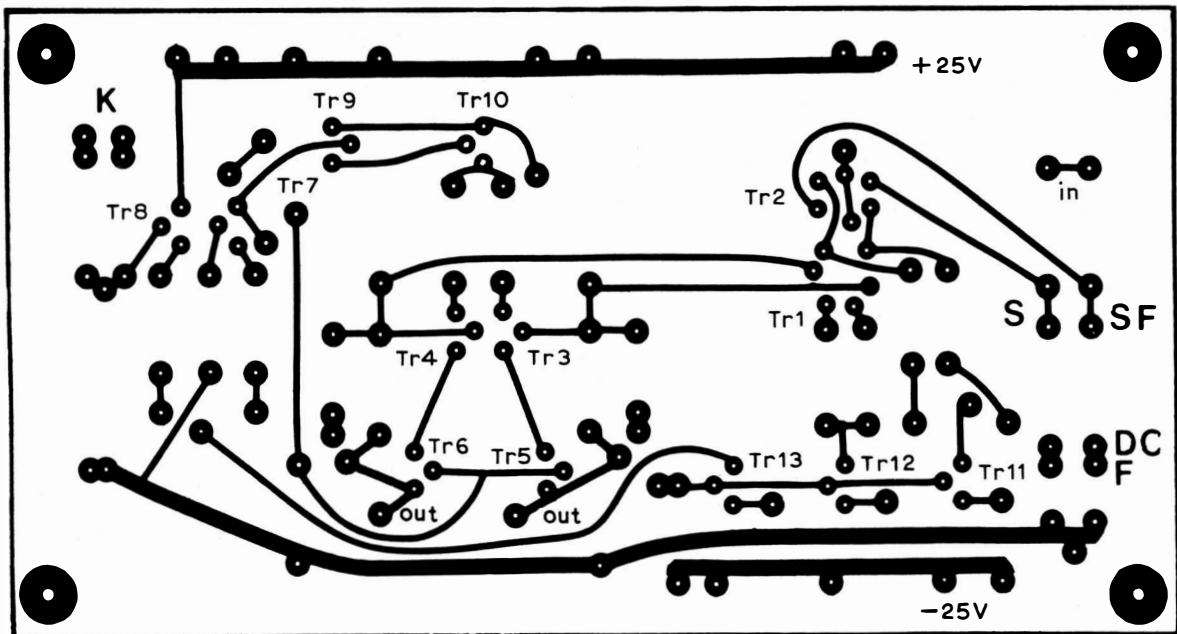
is inserted. The feedback is 14dB from
d.c. to about 1Hz, and the above anode
drift becomes less than ±2mA for a
±20deg C temperature change. The
feedback also reduces d.c. drift from
other causes, such as changes of
component values with time. Its equi-
valence to a pair of separate cathode
resistances has already been shown.

The above feedback may, on the
whole, be regarded as a possibility
rather than a necessity, and 14dB is
certainly more than necessary. The time
constant in the feedback circuit is so
large that temporary deviations from
symmetry in the signal (musical) vol-
tage should not cause appreciable d.c.
shifts.

Stabilization is needed for the nega-
tive high tension voltage, because a
±10% variation of this voltage would
cause too large variations in the valve
bias. A simple stabilization, for instance
by means of a series resistance from
a −300V supply feeding a chain of six
0.4W, 33V zener diodes is sufficient. The
voltage is of course not critical.

Constructional details

The layout of the circuit on a printed
circuit board or otherwise is not critical.
It has already been mentioned that the
two transistors of the pairs Tr_3, Tr_4 and
Tr_7, Tr_8 should be mounted for close
thermal connexion, and so should Tr_9 be
with D_3. To avoid heating effects from
the collector resistors R_{15} and R_{16},
mount them with the valves, and not on
a p.c. board. The circuit should be
mounted away from the mains trans-
former and filtering choke to avoid
induced hum from stray magnetic
fields. It should also be kept away from
any hot air stream or heat radiation
from the valves. These precautions do
not cause any problem, as the circuit
may be given fairly small dimensions.
Simple metal shields have been used in
the prototype amplifiers.

Five-percent resistors have been
use, and for matched pairs a 2%

difference is acceptable, although a
closer tolerance may be required for the
resistor pairs in the d.c. feedback, or the
value of R_1 may prove not to be
sufficient

In a first construction, the d.c.
feedback should be omitted, and put
into effect only as a finishing touch.

For the positive and negative supply
voltages of 25V in Fig. 4 the recom-
mended values are 25 to 30V, but there
is no need for symmetry. The value of
collector currents for the cascode stage
is 6 to 7mA. The currents of the other
stages are evident from the values of the
resistors R_{38}, R_{39} and R_{40}, since the
voltage across these is about 4.3V.

Any transistor in the Philips
BFQ10-16 family may be used at the
input, and there are of course also other
replacement types, for instance the
Siliconix E401. There are also a number
of replacements for the Motorola
MPS-A92, for instance MPS-U60, BFT19
(RCA), and BFW43 and BFW44 (SGS-
Ates). There are numerous replace-
ments for BC107 and BC177, and also for
the dual transistor BCY89, which is the
least expensive of the BCY87-89 family.

Concluding remarks

One reason for the choice of KT88
valves connected as triodes was that
they put high demands on the driver,
and so are suitable for presentation of
driver qualities. The same valves con-
nected as pentodes or with a distributed
load are more easily driven because the
Miller capacitance is lower. An obvious
conclusion is therefore that the driver
should suit most power amplifiers
except for very large ones that require
several output valves in parallel.

The ratio between the negative high

tension voltage for the cascode stage
and the maximum grid peak-to-peak
voltage is about two. When smaller
output tubes are used, such as EL34,
EL506 or EL84, the negative voltage
should be lowered, but the above ratio
not made smaller — a value between
two and three is preferred. The collector
currents for the cascode stage should be
maintained, and the collector resistors
chosen accordingly.

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Logic design course

Digital System Design is the name of a course
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address.