A Reference Standard RIAA Preamplifier

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LAST month (p. 81), I referred to a high quality headphone amplifier that was intended to drive a pair of high-impedance 'phones directly from a moving-magnet pickup cartridge, which was built as a Christmas present a couple of years ago for a friend who sells gramophone records for a living. I thought it appropriate to build the best unit within my competence, in the hope that his use of it would enable him to advise his customers more knowledgeably about the records which came within his reach. In the event, although it was fun to make, and interesting to play with, I had reckoned without the fact that those who earn their daily bread selling records (and cassettes) have little enough time during the pandemonium of the day to listen seriously to their wares, and when the blessed relief of closing-time arrives they like nothing better than to sit, perhaps with a meditative glass of whisky, in complete and absolute silence. However, the design exercise was instructive and thought-provoking, and since I still think that the design intention was sound, I am appending this for those who would like to add an RIAA-equalised input to their headphone amplifiers.

Design philosophy

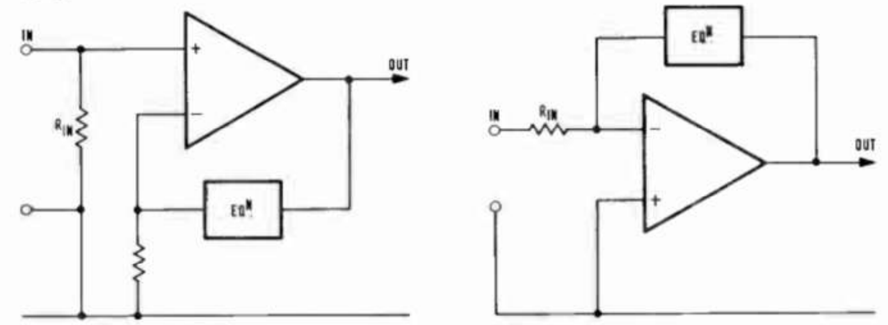
The conventional (contemporary) RIAA stage employs a series-feedback equalising network. This has the obvious advantages that the input impedance presented to the amplifier is merely that of the windings of the pickup coil which will, in general, present a noise voltage to the amplifier low enough to be ignored in comparison with the input noise of the amplifier, and that a two-transistor input amplifier stage can be used, which is the most cost-effective means of obtaining a reasonably high level of input amplification. This type of circuit is shown in schematic form in fig. 1A and in its normal realisation in fig. 2A.

The less obvious disadvantages with this type of circuit are that it does not, except with considerable and costly elaboration, present a constant load impedance to the pickup cartridge. While its mid-range impedance may be close to the nominal 47 k stipulated by cartridge manufacturers, its load impedance may differ widely from this at the upper and lower ends of the audio spectrum, and this has a detrimental effect on cartridge performance, particularly with some of the more carefully and expensively-designed units. Also, the fairly low impedance capacitative load presented to the output transistor in this configuration gives rise to a particularly unwelcome form of slew-rate limitation, in which not only the output signal but also the input impedance is modified.

To complicate matters even further, the series-feedback configuration has a gain which tends to unity at high frequencies, whereas the RIAA equalisation implicity assumes a curve which tends towards zero gain at the HF limit. With high gain systems, the approximation can be made quite good, but this also requires additional complication. Finally, for both practical and theoretical reasons it is a good idea in low level stages, if one can do it, to get one's gain entirely in the input device, and this is difficult in a non-inverting amplifier of the type required in series-feedback systems.

In the shunt-feedback system shown in schematic form in fig. 1B and in a practical embodiment in fig. 2B, none of these problems arise, since the input stage presents a constant impedance to the pickup cartridge at all frequencies and the impedance of the feedback network is sufficiently high to avoid slew rate limiting. All the gain is obtained in the first stage, and the gain tends to zero at the upper frequency limit, which facilitates an accurate equivalence to the RIAA specification. The snag is, however, that the circuit sees an input noise impedance determined by the input load resistor (or, more accurately, by this in parallel with the feedback circuit), and this, inevitably, leads to a higher input noise

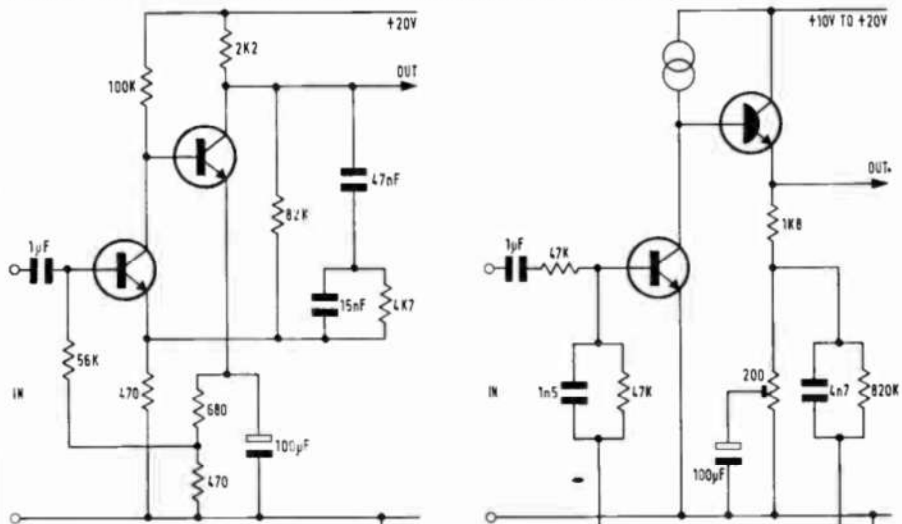
FIG. 1



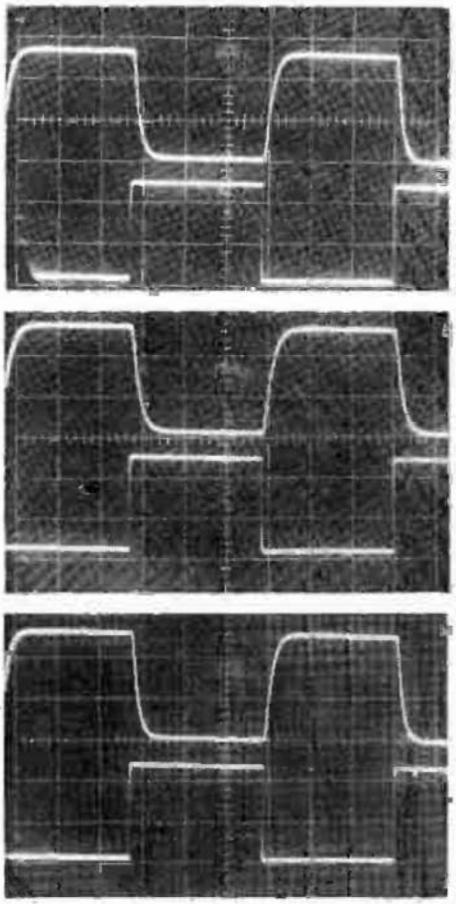
(a) 'SERIES' FEEDBACK EQUALISATION CIRCUIT

(b) 'SHUNT' FEEDBACK EQUALISATION CIRCUIT

FIG. 2 PRACTICAL 'SERIES' AND 'SHUNT' FEEDBACK RIAA EQUALISATION CIRCUITS



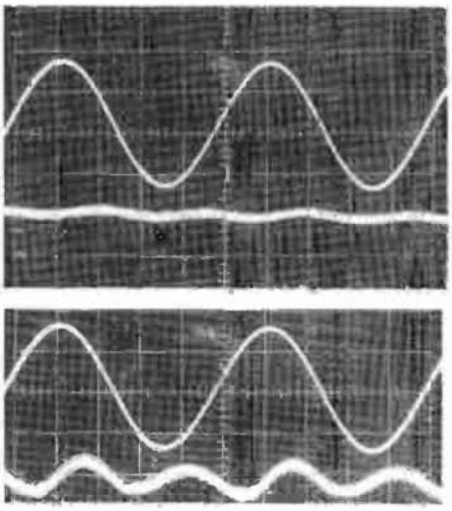
Performance of Class-A Headphone Amp. (See p 81, Jan. '79)



Transient Response of headphone amp. 10 kHz square-wave 100Ω load

Transient response of headphone amp. 10 kHz square-wave 100Ω/0·22μF load

Transient response of headphone amp. 1·0 kHz square-wave headphone load (all lower waveform input, upper waveform output)



Headphone amp. THD at 1 kHz

Headphone amp. THD at 10 kHz (note absence of crossover glitches and harmonics other than 2nd below background noise level).

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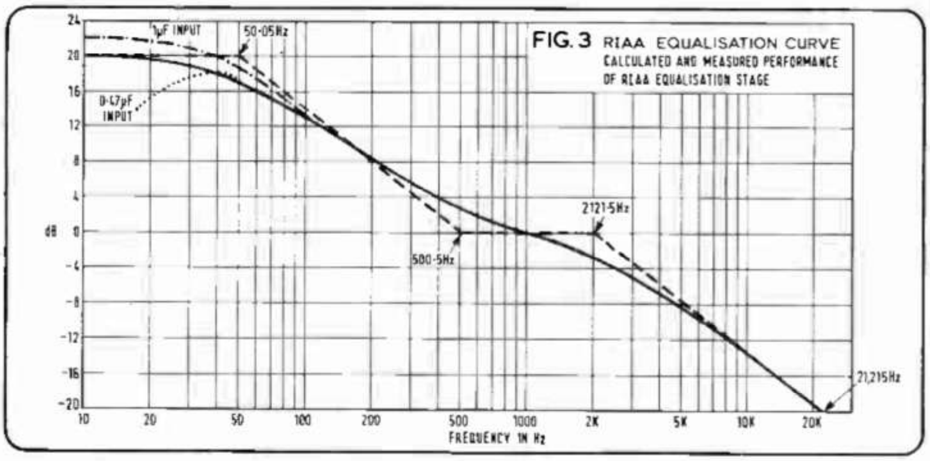
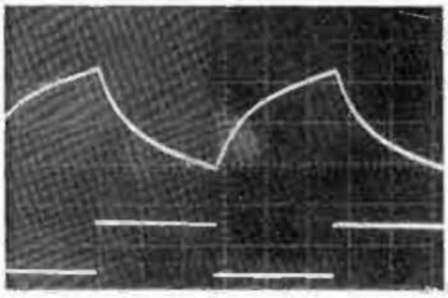
than could be obtained by the use of a circuit *of equivalent quality* using series-feedback. This drawback, and the unthinking calculations based on an assumed 20 kHz bandwidth, has effectively ensured that all commercial designs employ series-feedback, with a happy acceptance of all the problems that this generates—some of which have only recently become known at all widely.

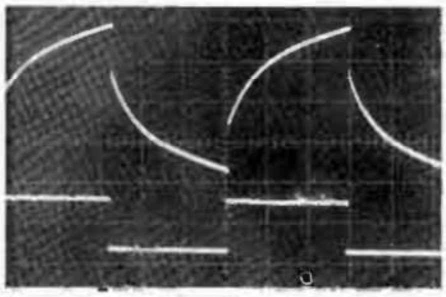
In defence of shunt-feedback systems, it should be said that these have lower distortion, other things being equal, that they generally allow the input transistors to be used under circumstances in which device noise (as distinct from circuit noise) is at its optimum, and that the RIAA stage has only a bandwidth of some 500 Hz, because of the falling frequency response of this—shown in fig. 3. This gives a calculated S/N ratio, referred to a 3 mV input signal (typical for some of the better cartridges) of some 74 dB, which is more than adequate. Also, the kind of noise produced by this type of stage is more of a rustle than a hiss, and this is subjectively less objectionable—as the CCIR weighting curve implies.

It is for these reasons then, rather than natural pig-headedness, that I prefer, and have always used shunt-feedback RIAA input stages, and my preference has been confirmed by experiments made with the *ad-hoc* listening panels of audiences at lectures. If an input equalising stage is made which can be switched from series to shunt-feedback and back again; and if the components chosen are of optimum values for each configuration to give the best possible noise performance; and it can be demonstrated with signal generator and millivoltmeter that the two arrangements have an identical frequency response over the range 30 Hz–20 kHz; if the two unidentified forms are merely labelled 'A' and 'B', in my experience audiences can generally distinguish between one and the other, and, if asked to give a preference, invariably choose the shunt form. Subsequently trying the same experiment on a critical friend, who actually makes gramophone records, I was properly rebuked on the grounds that it was irrelevant which one was preferred, what *was* important was which one gave a reproduced sound closest to the original tape. So, since it was at his house, a





Response of Shunt-feedback type of RIAA-equalised input to square-wave signal (fig. 6a)



Response of conventional (Series-feedback) RIAA-equalised input stage to square-wave signal
Note: starting transients only zero when closed-loop gain infinitely high (fig. 6b)

Headphone amplifier errata: in the power supply (fig. 3), the 'B' and 'case' connections of the negative voltage regulator (MC 7912CK) should have been transposed; the 0·038μF capacitor in the parts list should have been 0·033μF. Finally, 5% tolerance resistors should be suitable and where not specified, the capacitor type is not critical

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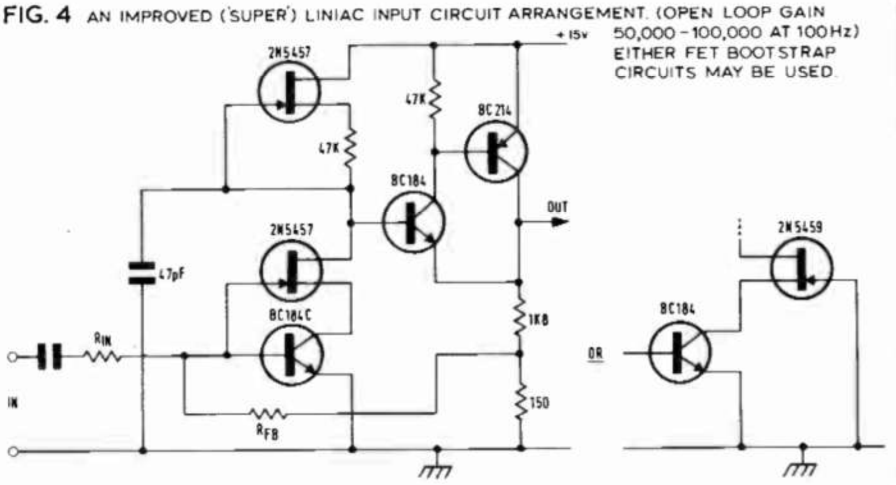
master tape and equivalent record were duly produced for a two-way equivalence trial against the original. Happily for my theories, my friend's choice was also for the shunt-type circuit as being the closest to the master.

I know that listening panels, under carefully monitored blind-comparison trials, have been shown recently not to be able to distinguish, with certainty, between several differing power amplifiers fed with identical signals. On that basis, my experiments were not adequately rigorous, in that the 'A' and 'B' positions could be seen, and a click indicated the changeover, so that although the audience did not know which was which, and I tried to remain silent and neutral, nevertheless I might have influenced the result. However, in this case the differences are not so small that they cannot be seen on an oscilloscope—in spite of the identity of the response curves in the audio band—when a step-function input is applied to the system. This is presumably due to the differing gain asymptotes of the two forms, and is shown in fig. 6.

FIG. 4 AN IMPROVED ('SUPER') LINIAC INPUT CIRCUIT ARRANGEMENT. (OPEN LOOP GAIN 50,000 - 100,000 AT 100Hz) EITHER FET BOOTSTRAP CIRCUITS MAY BE USED.



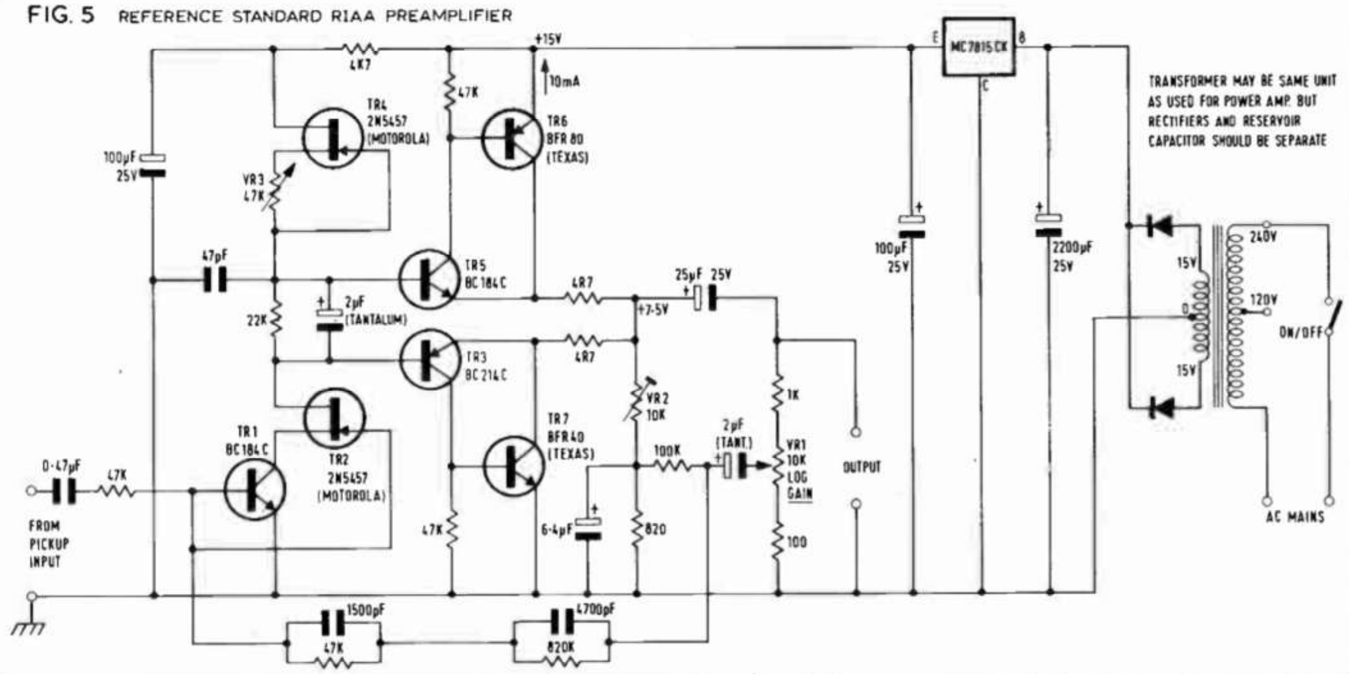
Circuit design

In my 75 watt amplifier design (*HFN/RR* '72/'73), I used a high-gain inverting amplifier stage based on a single transistor operating into an FET constant-current source, with a Darlington transistor used as an output emitter-follower impedance converter. This configuration gave a stage gain from a single transistor which was in the range 3000–4000, and may have inspired the integrated circuit version of this circuit (with a current-mirror added at the input) which appeared the following year as the Motorola MC3401 and its equivalents.

However, consideration of the gain characteristics of the Liniac circuit suggested that the gain was mainly limited by the output admittance of the amplifying device, and by its collector-base internal capacitance. If, therefore, this input transistor were 'bootstrapped' by an FET, the output impedance would be that of the FET, some 10 or 20 times higher, while, since the FET was operating in a 'grounded-grid' (or more accurately 'grounded-gate') configuration, the output to input capacitance could be ignored. This arrangement is shown in fig. 4, and might be styled a 'super Liniac'. In practice, it slightly simplifies the HF phase-shift characteristics if the gate of the FET and the base of the input transistor are connected together. A typical, open-loop low frequency gain for such an arrangement, with FET load and Darlington transistor emitter-follower output, is in the range 50,000–100,000 with a unity-gain bandwidth of about 20 MHz.

An improvement in the performance can be obtained if the Darlington transistor is replaced by an *n-p-n/p-n-p* compound pair emitter follower, and an improved output

FIG. 5 REFERENCE STANDARD RIAA PREAMPLIFIER



driving capability is gained by making this a push-pull Class-A arrangement, as used in the circuit shown in fig. 5. During experiments in the generation of very low distortion sine-waves, a parallel 'T' oscillator built around a circuit of this type gave a measured THD at 1 kHz of about 0·0008% at 1 volt RMS output, which suggests that the circuit is capable of a high degree of linearity. Again, the harmonic distortion residues, so far as they are measurable, are second harmonic—which is the type which is least objectionable to the ear.

The circuit shown in fig. 5 has adequate output capability to drive high-impedance headphones (such as the Sennheiser HD414) directly (since it was intended for this use), in which case the present gain control potentiometer VR_1 can be used as a volume control. If a pair of slider pots is used for this function, this also allows easy channel balance. These pots should be of 'log' form, with the 'log' end connected to the 25 μF capacitor, since this is the minimum output position. In the power amplifier (*HFN/RR* Jan. '79, p. 81), it was suggested that the gain controls should either be a similar pair of slider pots or a twin concentric-spindle dual gang.

Setting up the circuit involves adjusting VR_2 so that the output DC voltage at the junction of the two 4·7 ohm resistors is half the available HT (*ie*, 7·5 V) and then adjusting VR_3 so that the current through each channel from the HT supply is 10 mA (*ie*, 20 mA total). The calculated and measured agreement between the output of this stage and the RIAA equalisation curve is better than ±0·5 dB from 30 Hz to 21 kHz. A suitable printed circuit layout is shown in fig. 7.

Rumble filter

Everybody has prejudices, I suspect, and one of mine is that I do not like using gramophone input circuits without rumble filtering, because I believe that the average amount of recorded rumble on commercial pressings is uncomfortably high. However, debating this matter with myself when this circuit was originally designed, I concluded that the aim should be to tell the truth, without the concealment of defects—if any there be. So, I didn't include a rumble filter, and then discovered to my surprise that the performance was still acceptable on all the records I particularly liked, and turning this matter over in my mind subsequently, I have decided to leave well alone. If anyone thinks that this is a pity, well, the rumble filter circuit of my 75 watt design will marry into this design without difficulty ●

Components list for two channels and psu

Resistors		**Capacitors**	
4R7	4 off	47 pF	2 off
100R	2 off	1500 pF	2 off
820R	2 off	4700 pF	2 off
1 k	2 off	0·47 μF	2 off
4 k7	2 off	2 μF Tant.	4 off
22 k	2 off	6·4 μF Tant.	2 off
47 k	8 off	25 μF 25 V elec.	2 off
100 k	2 off	100 μF 25 V elec.	3 off
820 k	2 off	2200 μF 25 V elec.	1 off
Pots		**Semiconductors**	
		BC184C	4 off
Dual 10 k log.		BC214C	2 off
10 k preset	2 off	BFR40 (Texas)	2 off
47 k preset	2 off	BFR80 (Texas)	2 off
		2N5457 (Motorola)	4 off
		MC7815 CK	1 off
		IN4001 (or equiv.)	2 off

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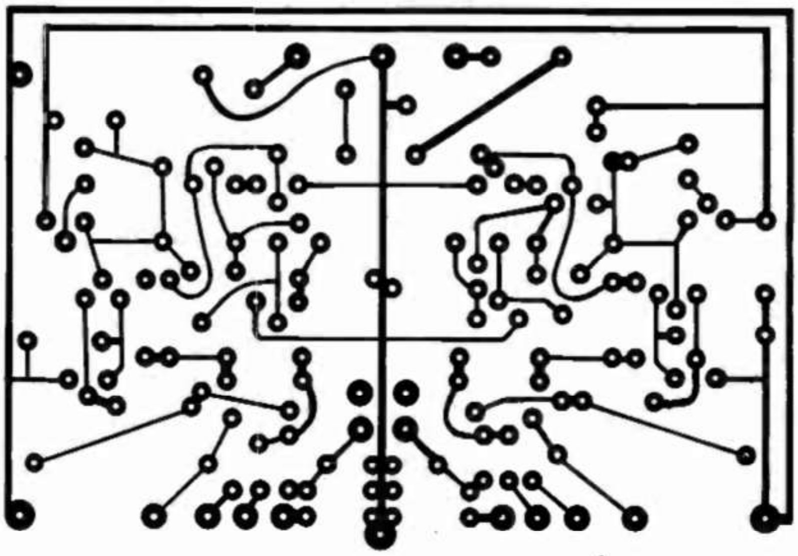
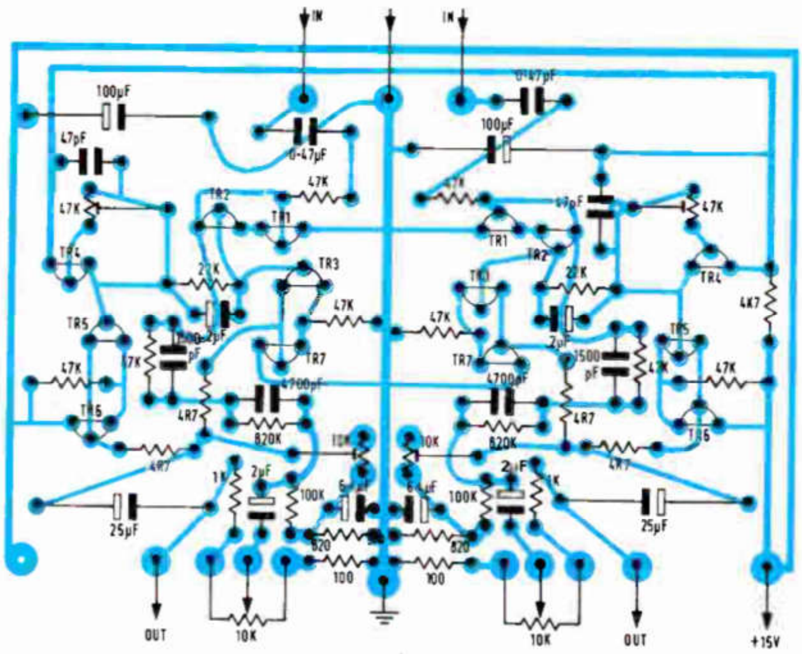
FIG. 7



(a) PRINTED CIRCUIT BOARD LAYOUT FOR RIAA PREAMPLIFIER (COPPER SIDE)



(b) COMPONENT LAYOUT FOR RIAA PREAMPLIFIER

VIEWED FROM COPPER SIDE