Series 5000 stereo control preamplifier

Part 1.

Designed as the 'perfect partner' to our Series 5000 MOSFET stereo amplifier, this preamp offers many unique features.

David Tilbrook

IN DESIGNING the Series 5000 Stereo Control Preamplifier we have had to look closely at the facilities and options available on existing preamplifiers. The concept that evolved was to design a preamplifier with the accent on versatility and control. We judged that most Series 5000 components will be built by those interested in high quality sound, who will therefore spend a considerable time taping, either from source material or dubbing from one deck to another. The preamp should therefore offer good facilities for taping and tape dubbing.

Tape facilities

Two tape inputs have been provided, either of which can be selected as a source input or used through a conventional tape monitor circuit. Dubbing from one deck to another is simply a matter of selecting the playback deck as an input and selecting the record deck with the tape monitor facility. Now if a tape deck were put into record mode and then selected as a source, its output would be connected via the preamplifier back to its own record input. This forms a feedback loop that would lead to oscillation, probably within the audio spectrum and at full power. To prevent this, this preamplifier automatically mutes the record output to any tape deck that is selected as a source input. If any other source is selected the muting is removed. Thus, both decks can be used simultaneously to record source material.

Metering

Another facility that is *essential* for high quality taping is a really good set of level meters. Unfortunately, those supplied on most cassette and reel to reel decks are of the −20 to +3 VU type and as such are poor indicators of the true audio level. To overcome this problem the Series 5000 Preamp has been provided with two wide dynamic range LED level meters, each with 3 dB resolution and covering the range −48 to +9 dB, so a dynamic range of 57 dB can be displayed. The LED level meter displays *both the peak and the average* of the audio signal at any instant, with the peak indicator having a rapid-attack/slow-decay characteristic that ensures any transient can be easily seen on the display. The prototype has been tested with single pulses and even with pulses as short as 50 µs the transient was accurately recorded and easily spotted.

A 400 Hz sinewave oscillator is also provided and is selected by one of the positions of the tape input selector. This enables calibration of 0 dB on the LED level meters to match the 0 dB reading on the tape deck meters.

This LED level meter will also be valuable in many other projects where accurate monitoring of an audio signal is required. For this reason we have given it a separate project number (ETI-458). Full details were given in the June issue, so many constructors may already have built the LED level meter.

The preamplifier is provided with both *master* and *monitor* volume controls. The master control varies the output to the LED level meters and to the line and tape outputs. In this way the record level can be adjusted. The monitor volume control varies the signal level on the monitor output. Normally the power amplifier would be connected to this output so that the volume can be adjusted without affecting the record level. If the monitor level control is turned fully up, however, and the master control used as the main volume control, the LED level meters indicate the output level below 1.2 V, which corresponds to full power from the power amp. So, in this configuration, the LED level meters function as output level indicators, with full power before clipping occurring at 0 dB.

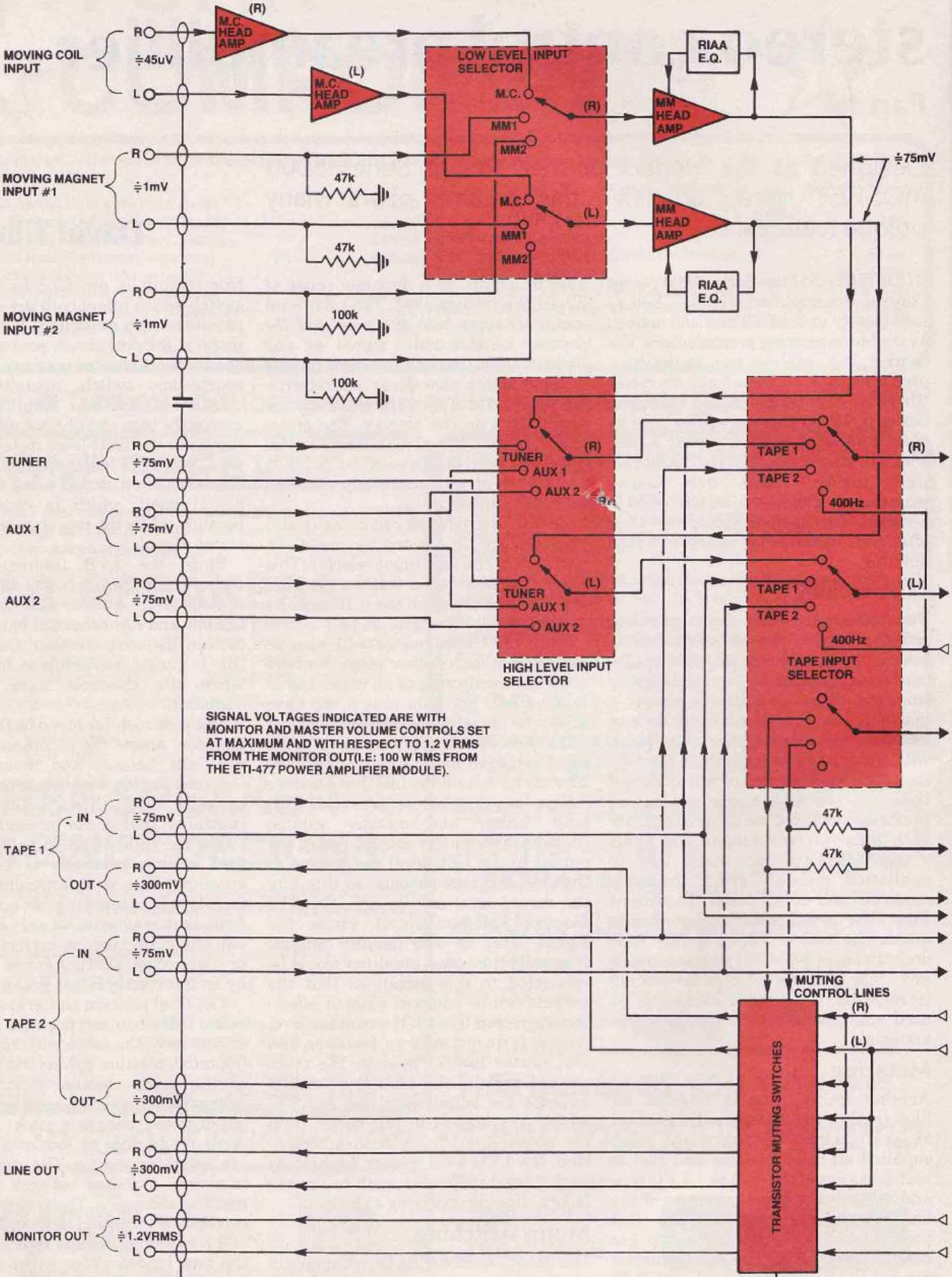
Mono switching

The ability to mono the two channels is another feature essential in a preamplifier. In the Series 5000 Preamp this feature is provided by the *mode* switch which offers both pre- and post-monitor mono switching. If the source mono is selected (L+R, source) the two channels are shorted together before the source-tape switch, actually at the outputs of the line amplifiers. This monos the tape and line outputs, and the inputs to the LED level meters. In this way the record levels on the tape decks can be easily matched using the source signal itself, which is virtually impossible when the two channels have a stereo signal.

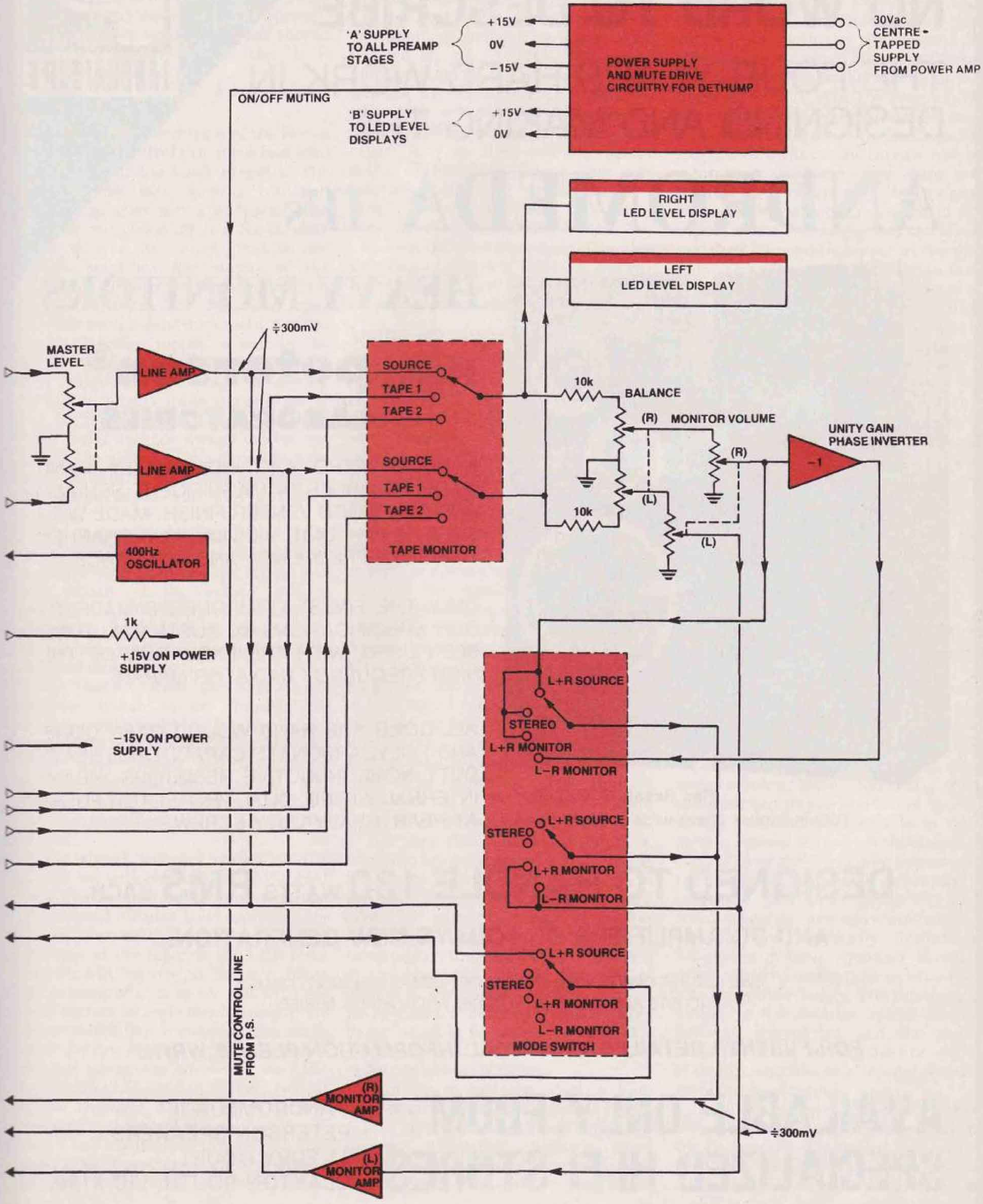
Since the L+R (source) position monos the channels before the balance control it can also be used to optimise the left and right channel balance for a certain listening position. Once again, this is almost impossible to do properly when the channels have different signals.

The post-monitor mono facility (L+R, monitor) shorts the channels together *after* the balance and monitor level controls. In this way the stereo content is kept valid through the balance control so that it can be used to select either the right or the left channel and feed it simultaneously to both loudspeakers. This is a particularly useful facility when listening for a suspected fault that may exist on only one of the two channels, such as cartridge faults, or more often, skating errors produced by an incorrectly set-up tone arm.

The final position on the mode switch is the L-R (monitor) position. This acts in precisely the same way as the L+R (monitor) position but inverts the phase of the right channel first. This is primarily to facilitate easy detection of out-of-phase channels such as would occur if cartridge or loudspeaker leads are wired incorrectly. The mode switch is simply switched between the L+R and L−R positions. The system is set up correctly if the sound has a full bass end and a stationary image in the centre of the two loudspeakers when the mode switch is in the L+R position. If the bass is better on the L−R position, however, ▶



SIGNAL VOLTAGES INDICATED ARE WITH MONITOR AND MASTER VOLUME CONTROLS SET AT MAXIMUM AND WITH RESPECT TO 1.2 V RMS FROM THE MONITOR OUT (I.E: 100 W RMS FROM THE ETI-477 POWER AMPLIFIER MODULE).


'A' SUPPLY TO ALL PREAMP STAGES
+15V
0V
−15V
ON/OFF MUTING
'B' SUPPLY TO LED LEVEL DISPLAYS
+15V
0V
POWER SUPPLY AND MUTE DRIVE CIRCUITRY FOR DETHUMP
30Vac CENTRE-TAPPED SUPPLY FROM POWER AMP
RIGHT LED LEVEL DISPLAY
LEFT LED LEVEL DISPLAY
≑300mV
MASTER LEVEL
LINE AMP
LINE AMP
SOURCE
TAPE 1
TAPE 2
SOURCE
TAPE 1
TAPE 2
TAPE MONITOR
10k
10k
BALANCE
(R)
(L)
MONITOR VOLUME
(R)
(L)
UNITY GAIN PHASE INVERTER
−1
400Hz OSCILLATOR
1k
+15V ON POWER SUPPLY
−15V ON POWER SUPPLY
L+R SOURCE
STEREO
L+R MONITOR
L−R MONITOR
L+R SOURCE
STEREO
L+R MONITOR
L−R MONITOR
L+R SOURCE
STEREO
L+R MONITOR
L−R MONITOR
MODE SWITCH
MUTE CONTROL LINE FROM P.S.
(R) MONITOR AMP
(L) MONITOR AMP
≑300mV

then the system has an out-of-phase channel. The surest way to perform this test is to position the loudspeakers directly facing one another and approximately 50 mm apart. The bass should drop *dramatically* when the L–R position is selected.

Block diagram

The overall configuration of the Series 5000 Stereo Control Preamp is best seen by looking at the block diagram. The preamp has both moving coil and moving magnet cartridge inputs. The moving coil phono input is fed directly to a moving coil input preamplifier which amplifies the output of the cartridge up to typical moving magnet signal voltages. The *low level input selector* switch determines which of the three cartridge inputs is sent to the RIAA-equalised phono preamp stage. The output of this stage is around 75 mV for a 1 mV input signal to one of the moving magnet inputs. The *high level input* selector switch selects between the output of the phono preamp and one of three other line level inputs (the tape input and two auxiliary inputs).

The output of the high level input selector is fed to the *tape input* selector together with the two tape inputs and the output of the inbuilt 400 Hz oscillator. A third set of contacts on the switch is used to drive the *muting* transistors whenever a tape input is selected as a source.

The output from the tape input selector switch is fed through the master level control and line amplifiers to the tape and line outputs, and to the LED level meters and tape-source switch, through the balance and monitor volume controls and via the monitor amplifiers to the main monitor output.

The signal voltages shown on the circuit are with respect to a 1.2 V signal from the monitor output and with the master and monitor level controls fully up. In this condition the average signal voltages at the input to the LED level meters will be around 300 mV. Since the preamplifier is never run with the level meters off scale it is impossible for the preamplifier to overload any stage after the line amplifiers; this is another distinct advantage offered by the LED level meters. As stated earlier, typical signal voltages at the input to the master level control are around 75 mV for a 1 mV input signal to one of the MM inputs. Most moving magnet cartridges produce maximum output signal voltages of the order of 60 to 80 mV. Some higher output cartridges may produce signals as high as 120 to 140 mV but these are generally the exception. Allowing for a signal input voltage of, say, 140 mV, the output voltages from the main phono amplifier will be approximately 10 V RMS. So this stage must be provided with adequate supply voltage and slew rate capabilities so that overload cannot occur. Fortunately this is not particularly difficult, necessitating a supply voltage of around 15 V and a slew rate of approximately 1 V/µs.

A more detailed discussion of the slew rate requirements of phono stages will be included in part two of this series of articles on the Series 5000 Preamp.

Using op-amps?!

The basis of the Series 5000 Preamplifier is a new audio op-amp by Signetics, the NE5534N and its low noise complement, the NE5534AN. These devices are capable of truly *superb* performance. There is a popular misconception among audiophiles that op-amp stages cannot 'sound' as good as discrete designs. This undoubtedly has come about due to the introduction of some earlier amplifiers that used 741s, or similar devices, which are simply not suitable for audio applications. The noise performance of these earlier op-amps was mediocre and their slew rate figures were typically less than 1 V/µs. If we consider the output stage of a preamplifier, for example, the maximum output signal voltage from the stage is required to be around 1 V and this dictates a slew rate substantially higher than is possible from a 741 op-amp. The result is slew limiting distortion or TIM (transient intermodulation distortion).

The slew rate figures for the 5534, however, are *in excess* of 10 V/µs, giving the device a large-signal bandwidth of around 10 MHz and maintaining the open-loop gain at 10 kHz at greater than 6000. This is important since it ensures that the negative feedback loop of the stage will not be degraded at higher frequencies. The noise performance of the 5534 is equally good, being equal to or better than the best discrete input stages for input impedances above 1000 ohms.

In order to test the 5534, a high quality moving magnet input stage and a line amplifier stage were built with discrete transistors. Actually these stages were originally designed for the Series 5000 Preamplifier before the 5534s were available. These stages were compared with similar stages designed around the 5534s and both objective and subjective testing was carried out. Both the discrete and IC stages gave measured distortion figures less than 0.001% at all operating frequencies up to 20 kHz. The noise figures were similar and none of the stages gave any indication of slew rate limiting. The subjective testing was carried out by comparing the two phono amps and then the two line amps. The input signal was sent via a selector switch to the inputs of the two stages and the levels matched by potentiometers at the outputs. The wipers of the pots were then connected to a second set of contacts on the selector switch. This configuration allows both the input and the output of the unused stage to be disconnected from the circuit so that no possibility of interaction exists.

After many hours of experimenting with this circuit we were *still unable to determine which stage was operating at any one time.* The op-amp circuit, however, contains approximately half the number of components of the discrete design.

A complete data sheet for the NE5534N will be included with part two of this series.

Tone controls?

As stated earlier, the Series 5000 Preamplifier is intended as a high quality linear preamp, and as such, the conventional tone control circuits have been omitted. This is becoming the accepted practice on high quality preamplifers, since the usual tone control facilities are very seldom used in good sound systems. More importantly, their presence can impair the sound quality unless provision is available to completely remove them from the circuit. If control over the frequency response of the system is required it is usually to overcome problems associated with room acoustics, and conventional tone controls are practically useless. A 1/3-octave graphic equaliser is much more effective and can be incorporated into the system easily. The equaliser is placed in the monitor output circuit, between the output and the power amplifier input, leaving the line output of the preamp free as a general purpose non-equalised output.

In following articles we will describe the specifications and construction of the Series 5000 Preamp as well as the problems that must be overcome to ensure the best possible performance.●

Series 5000 stereo control preamplifier

Designed as the perfect partner to our Series 5000 MOSFET stereo amplifier, the stereo control preamplifier has been designed with equal attention to quality and detail. This month David Tilbrook explains the principles behind the design of the moving coil and moving magnet stages and gives details on constructing them.

David Tilbrook

JUST AS a loudspeaker represents a non-linear load to the output stage of a power amplifier, a moving magnet or moving coil cartridge represents a non-linear source impedance to the input stage of a preamplifier. This is the cause of many of the problems associated with any preamp.

Both moving coil and moving magnet cartridges generate electrical signals through the interaction of a coil of wire and a magnetic field. The signal voltage produced is therefore proportional to the relative velocity between the coil and the magnet assemblies. This relationship is predicted by Faraday's law of induction, expressed mathematically as:

$$\varepsilon = -\frac{d\varnothing}{dt}$$

where ε is the signal voltage at any instant and $\varnothing$ is the magnetic flux.

The signal voltage produced at any instant is proportional to the rate of change of flux with respect to time, i.e: $\frac{d\varnothing}{dt}$

The design of the cartridge must ensure that a linear relationship exists between the position of the stylus cantilever assembly and the magnetic flux. In this way changes in the position of the stylus give rise to changes in the magnetic field intensity. So the rate of change of stylus position with respect to time will be proportional to the signal voltage, i.e:

$$\varepsilon \,\alpha \frac{dx}{dt}$$

where ε is the signal voltage and x is the stylus displacement from its equilibrium position.

This means that the waveform actually 'on' the grooves is not proportional to the signal voltage itself. Instead it is proportional to the integral of the signal waveform. If a square wave, for example, is to be produced from a record, the waveform as seen in the groove with a microscope will be a triangle wave.

Since the signal voltage is proportional to the velocity of the stylus, the signal slope is proportional to the *acceleration* of the stylus. In order for high signal slopes to be reproduced accurately by the cartridge it is important that the mass of the stylus cantilever assembly be kept as small as possible. At the same time, however, it is important to realise that the cartridge cantilever assembly and its associated suspension and magnet/coil system form a resonant mass-spring system analogous to a complex electrical series resonant circuit.

At one particular frequency, called the resonant frequency, the impedance of the cartridge will no longer be related linearly to the driving force on the stylus, and distortion results. To overcome this problem two techniques are used simultaneously.

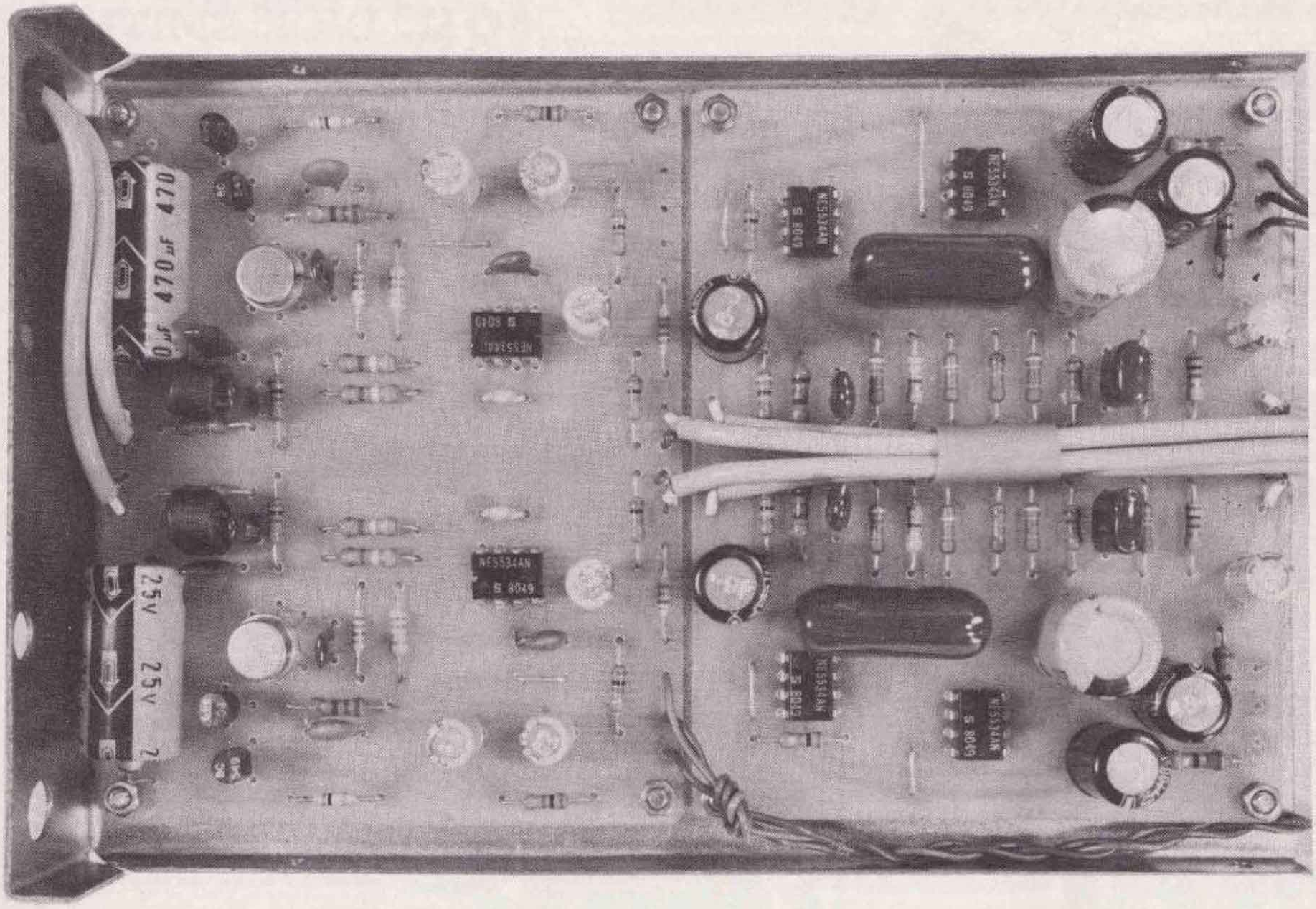
Firstly the resonant frequency of the cartridge is moved to a frequency below the audio spectrum. Using the damped mass-spring model of a magnetic cartridge we can predict that the resonant frequency will depend on the mass of the stylus cantilever assembly and on the 'springiness' of the cantilever's suspension. This springiness is characterised by a number, often given the symbol k, called the spring constant. Spring constant is defined in terms of the force needed to bring about a certain compression or extension of the spring. Stiffer springs have a higher value for k. The spring constants associated with magnet cartridges, however, are so small that the numbers are hard to interpret. For this reason cartridge manufacturers usually specify this quantity by quoting the reciprocal of the spring constant, $\frac{1}{k}$, called *compliance*. Stiffer suspension systems have lower compliance figures.

As stated earlier, the cartridge resonant frequency is a function of both the mass and the compliance of the cantilever and suspension system. The damped resonance mass-spring model of a magnetic cartridge predicts that the resonant frequency will be given by the equation:

$$f = \frac{1}{2\pi\sqrt{(mC)}}$$

where m is the mass of the cantilever/stylus system and C is the compliance of the stylus suspension system. ▶

Our adaptation of a famous scene from Irving Stone's book about Michelangelo, 'The Agony and the Ecstasy'. During the time Michelangelo was painting his masterpiece frescoes in the Sistine Chapel the Pope continually asked ". . . when will you make an end?". Likewise, Roger Harrison has kept asking David Tilbrook when the Series 5000 would finish. The answer was the same in both cases! (Thanks to the cartoonist, Brendan Akhurst, and the choirboys: Jack O'Donnell of Altronics, Dick Smith, and Gary Johnston of Jaycar. Collyn Rivers looks on from the sidelines).



The low level Series 5000 Preamp input stages — for moving magnet and moving coil cartridges — are housed in a steel enclosure inside the main preamp case. This is a view inside the enclosure; moving coil stage to the left, moving magnet stage at right.

(Notice that the equation for the resonant frequency of magnetic cartridges has exactly the same form as the equation for the resonant frequency of an electrical resonance circuit, i.e:

$$f = \frac{1}{2\pi\sqrt{(LC)}}$$

where C in this case is capacitance and L is inductance. The fact that a mechanical system like the stylus cantilever assembly of a cartridge should be described by such an equation is another striking example of the consistency of nature.)

The equation predicts that the resonant frequency of the cartridge can be decreased by increasing either the mass or the compliance. Since the mass of the moving parts in the cartridge must be kept small so the stylus can respond quickly to changes in the record groove, the compliance must be increased until a suitably low resonant frequency is obtained. Most high-quality magnetic cartridges have resonant frequencies below 10 Hz.

The second technique used to overcome problems associated with this resonance characteristic is to decrease the Q of the system by damping the resonance with a suitable combination of mechanical and electrical losses. Mechanical damping is obtained by deliberately introduced friction within the cantilever suspension system. The cantilever is often terminated into a rubber mounting block for this purpose. The electrical damping comes about as a direct consequence of the law of conservation of energy. The cartridge is acting as a generator, delivering power to the input resistance of the preamplifier. Since energy is absorbed by this load resistance the Q of the cartridge resonance is decreased.

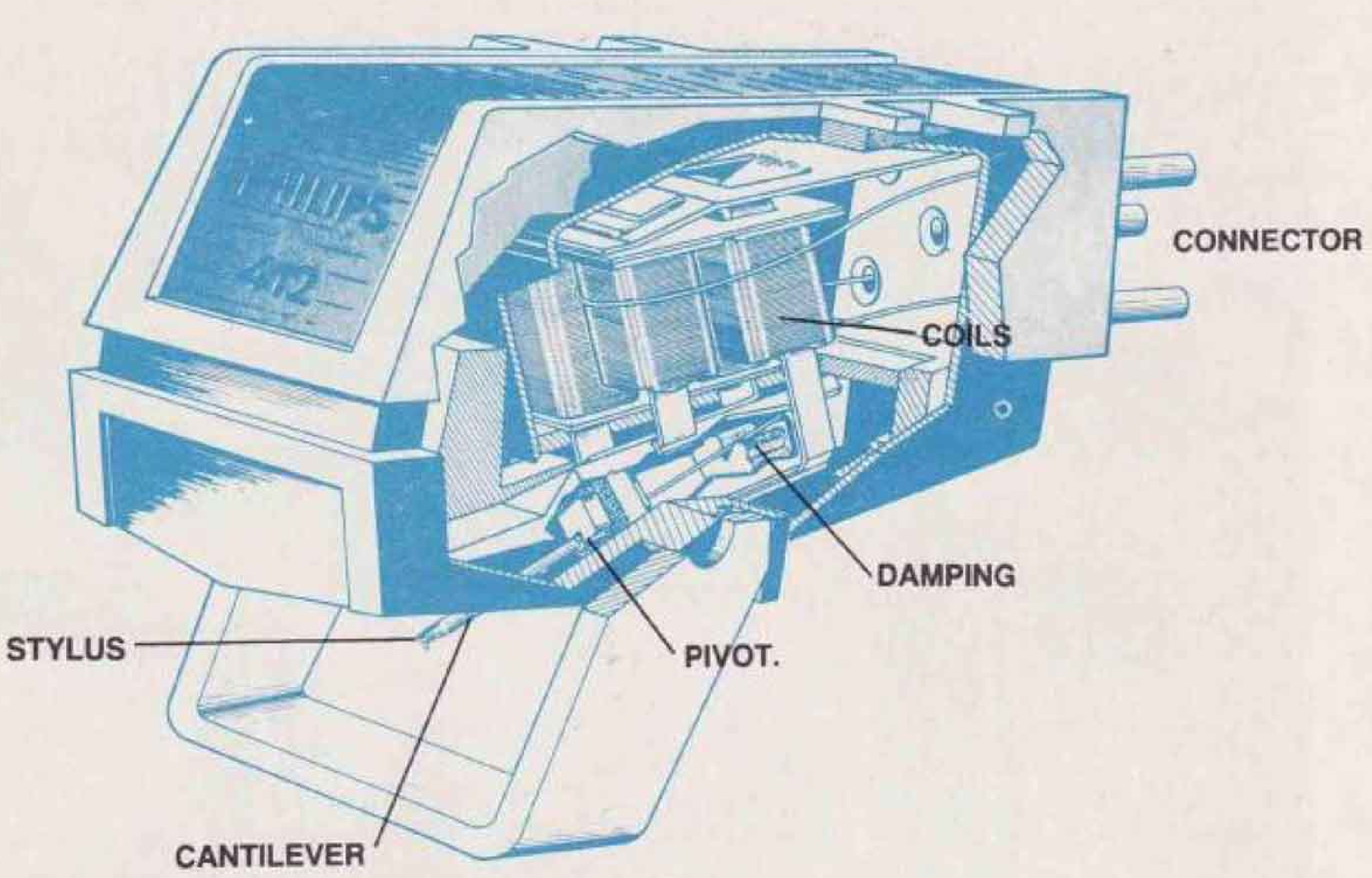
Until recently most stereo magnetic cartridges consisted of two fixed coils between the poles of a small magnet attached to the cantilever. Modulation of the record groove produces movement of the magnet, changing the magnetic flux and generating the signal voltage.

The coils usually have a large number of turns so that a reasonable signal voltage can be produced (typically in the order of 20 mV). The resistance of these coils usually ranges between 200-1000 ohms, but their impedance can be much higher, especially at high frequencies where the inductance of the coils becomes important. This type of cartridge is sometimes called a moving magnet cartridge to distinguish it from the more recently developed moving coil types. The relatively high reactive component of the cartridge impedance combined with the effects of the natural cartridge resonances makes it essential that the input impedance of the moving magnet (MM) input stage have well-defined characteristics if best performance is to be obtained from this type of cartridge. Most MM cartridges require a load impedance consisting of 47k of resistance shunted by several hundred picofarads. This capacitance is often provided by the shielded cable, but most cartridges require some additional capacitance across the MM input. In exceptional cases the input capacitance due to the shielded cable is too high. In these cases the length of the shielded cable used between the turntable and the MM amp should be decreased until the correct value is achieved.

In order to obtain the flattest frequency response possible from an MM cartridge it is essential that the load resistance be constant over the complete audio spectrum *and beyond.* For this reason measurements done on the input resistance of MM amps at one particular frequency (usually 1 kHz) are practically useless. Many input stages exhibit a characteristic of falling input resistance at high frequencies. The input resistance of a bipolar transistor, for example, even with a small amount of emitter current, is insufficient to ensure a constant resistive load to an MM cartridge. The common two or three transistor phono stages of a few years ago often suffered badly from this problem, degrading the top end performance of an otherwise good MM cartridge. The problem occurs because all bipolar transistors have decreasing gain at high frequencies.

The most common method used to increase the input impedance of a bipolar input stage is through the use of negative feedback. The decrease in gain of the individual transistors in the stage at high frequencies decreases the overall open loop gain of the stage, which in turn decreases the amount of negative feedback available. Furthermore, the negative feedback is often applied at the emitter of the first transistor. The problem with this configuration is that the phase response in the negative feedback loop can easily be affected by the complex reactances of the cartridge and connecting cables, producing unwanted frequency response variations, or even instability in some cases.

All these problems come under the general heading of 'cartridge impedance interaction', and represent the most important single reason for the difference in sound between preamplifiers. Most preamps suffer from some degree of cartridge impedance interaction and in *many* cases the effects are pronounced. ▶



View inside a typical high-quality moving magnet cartridge, the Philips GP412.

SPECIFICATIONS

ETI-478MM MOVING MAGNET INPUT STAGE

Gain:	74, 1 kHz
Frequency response:	Conforms to RIAA Equalisation ±0.2 dB. (This is the performance of the prototype. The actual figure obtained will be determined by the accuracy and long-term stability of the components used.)
Total harmonic distortion:	<0.001%, 1 kHz, 10 mV RMS input
Headroom:	>28 dB with respect to 5 mV RMS input signal, i.e: 135 mV RMS max.
Noise:	Total equivalent input noise, 122 nV 'A', input shorted, 216 nV flat, input shorted.

S/N ratio:

	1 mV	5 mV	10 mV
Flat	73 dB	87 dB	93 dB
A-weighted	78 dB	92 dB	98 dB

	IDEAL RIAA	MEASURED – SERIES 5000
Hz	dB	dB
2	−0.2	−0.2
4	+5.7	+5.7
8	+11.2	+11.2
16	+15.4	+15.4
20	+16.3	+16.2
30	+17.0	+17.0
40	+16.8	+16.8
50	+16.3	+16.2
80	+14.2	+14.2
100	+12.9	+12.8
150	+10.3	+10.2
200	+8.2	+8.1
300	+5.5	+5.4
400	+3.8	+3.7
500	+2.6	+2.6
800	+0.7	+0.7
1k	0.0	0.0
1k5	−1.4	−1.3
2k	−2.6	−2.4
3k	−4.8	−4.7
4k	−6.6	−6.6
5k	−8.2	−8.1
6k	−9.6	−9.6
8k	−11.9	−11.9
10k	−13.7	−13.8
15k	−17.2	−17.1
20k	−19.6	−19.5

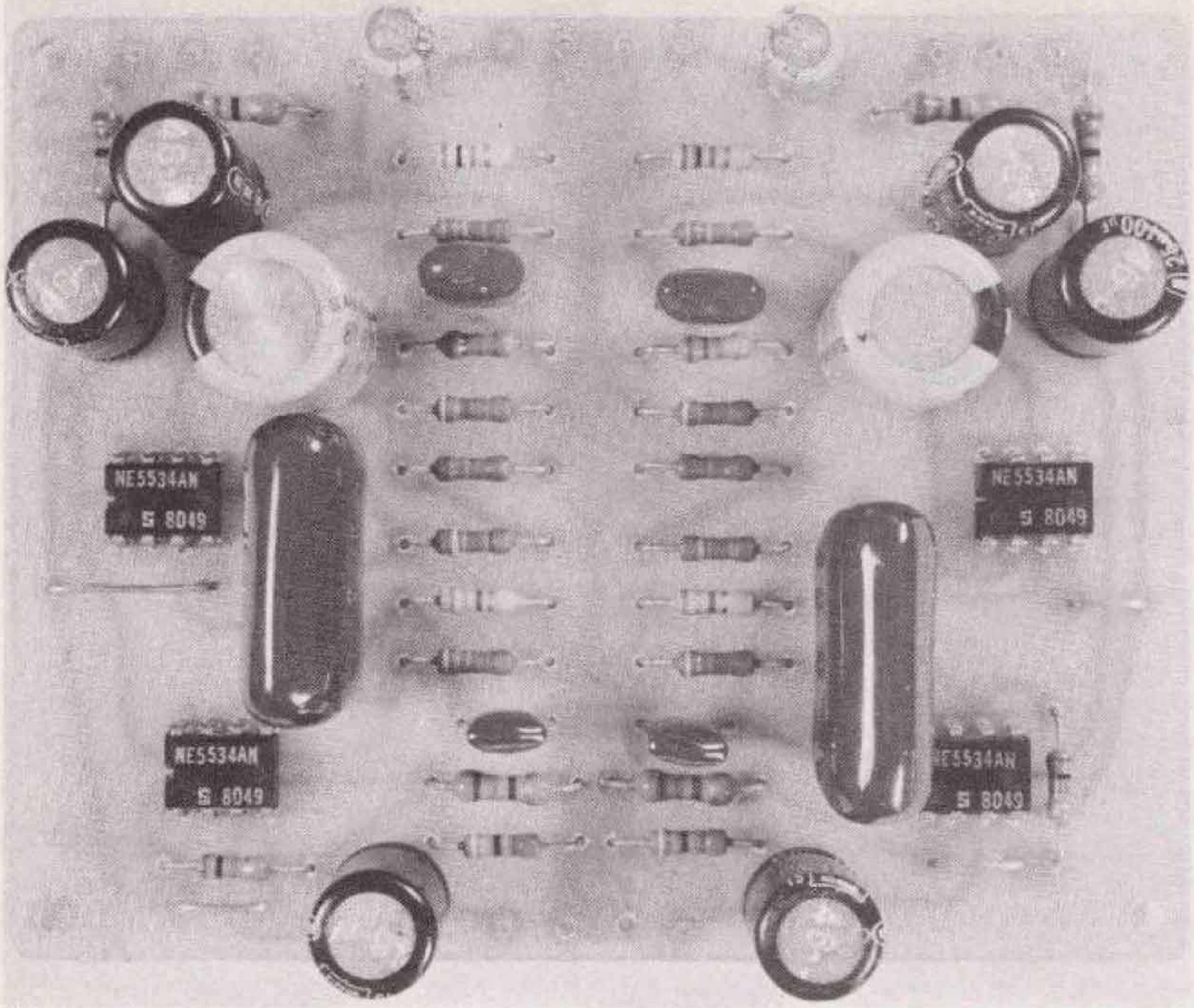
SPECIFICATIONS

ETI-478MC MOVING COIL INPUT STAGE

Gain:	24
Frequency response:	7 Hz-135 kHz +0, −1 dB
Total harmonic distortion:	<0.003%, 1 kHz, 30 mV input
Noise:	Total equivalent input noise: 83 nV flat, input shorted. 42 nV 'A', input shorted. 56 nV flat, after RIAA Eq., input shorted. 34 nV 'A', after RIAA Eq., input shorted.

S/N ratio of MC input stage after RIAA Equalisation:

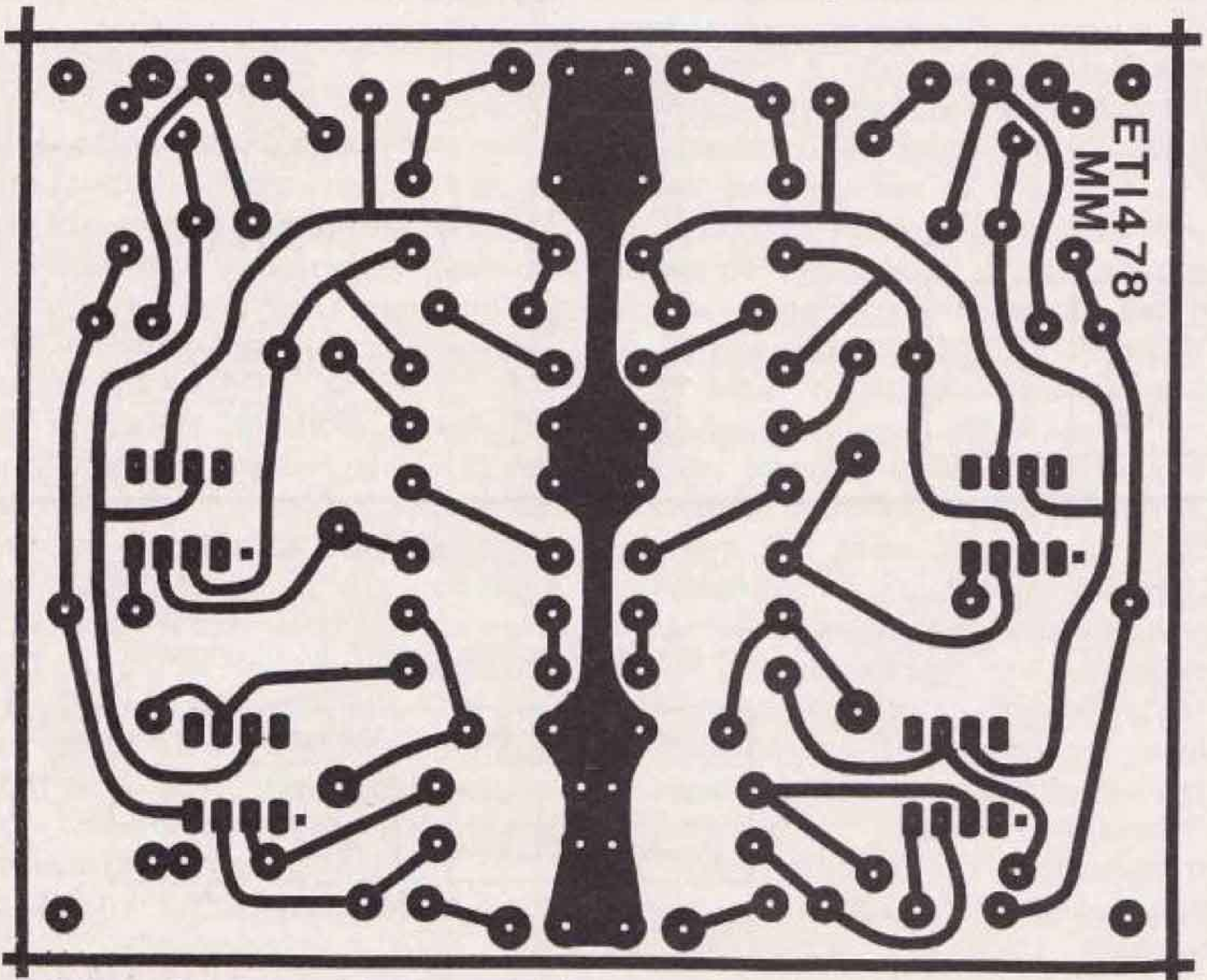
	60 μV	200 μV	500 μV
Flat	61 dB	71 dB	79 dB
A-weighted	65 dB	75 dB	83 dB



The completed moving magnet stereo input stage (ETI-478MM). Note that this project may also be incorporated in existing equipment if you wish.

The Series 5000 stereo control preamp has been designed specifically to overcome the problem of cartridge impedance interaction. This has been achieved by separating the MM input stage into two separate active stages (see Figure 1). The first stage consists of a single NE5534AN configured as a linear amplifier with a closed loop gain of around 8.3. The large amount of overall negative feedback increases the input impedance of the stage so that the measured input impedance is simply that of the 470k resistor, R2. Since the 5534 has a small signal bandwidth of around 10 MHz without additional compensation, the input impedance will remain unchanged over a very wide frequency range. The high input impedance of this stage would usually allow the input capacitor C2 to be conveniently small. However, for best noise performance the value must be increased substantially. This is covered in detail later in this article.

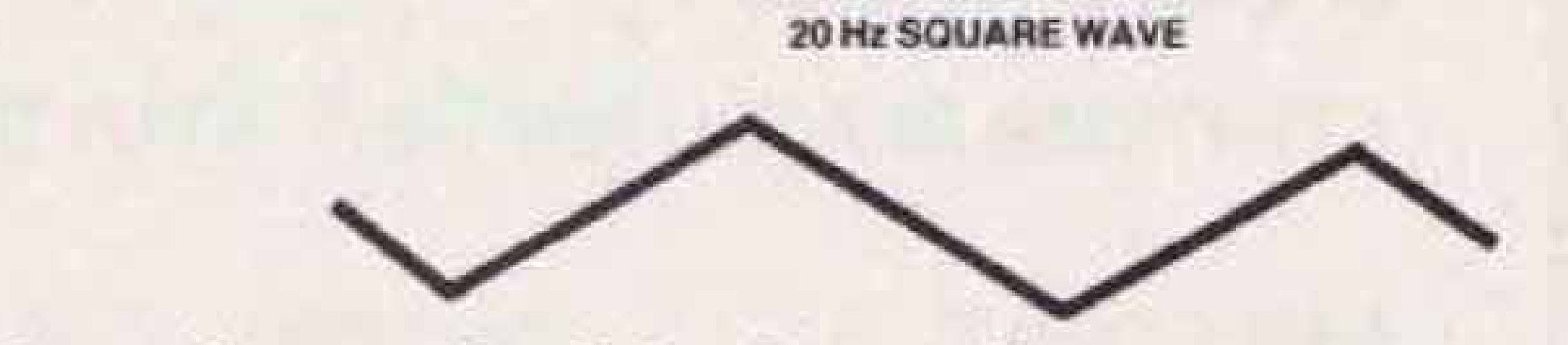
Capacitor C2 is necessary since it is not advisable to allow dc current from the first stage to flow through the cartridge. The value of C2 used here is 100μ, and this sets the lower -3 dB point well below 1 Hz. The upper -3 dB point of this stage is well above 100 kHz. An extended frequency response is necessary so that the accuracy of the RIAA equalisation is not affected by frequency response variations that might otherwise occur in the first stage.



Full-size pc board artwork for the moving magnet stage.

RIAA equalisation

We said earlier that the signal voltage produced by a magnetic cartridge is proportional to the velocity of the stylus. If a low frequency signal is to be reproduced by a magnetic cartridge, large excursions of the stylus are necessary. If for example a 20 Hz square wave is to be reproduced by the cartridge then the cartridge must produce a dc voltage at its output for a period of 25 ms.



In order to do this the stylus must move at a constant speed for this period of time, and therefore the waveform in the record groove is a triangle wave, as stated earlier.

Typical output voltages from moving magnet cartridges are in the order of 1 mV-2 mV for a stylus velocity of 1 cm/sec. So if the peak voltage required on the square wave was, say, 10 mV, a stylus velocity of 10 cm/sec would be required for a medium-sensitivity cartridge, so the stylus must move at a constant speed of 10 cm/sec for a 25 ms time interval. The stylus therefore moves a total distance of 2.5 mm! On a stereo record the channels are cut in opposite walls of the record groove. If a low frequency mono signal is to be produced, both sides of the record groove force the stylus away from its equilibrium position, and a large vertical stylus excursion results. In the case of our square wave, the vertical excursion would be roughly 3.5 mm, which is simply not possible. The record would have to be as thick as most turntable platters!

Two measures are used to overcome

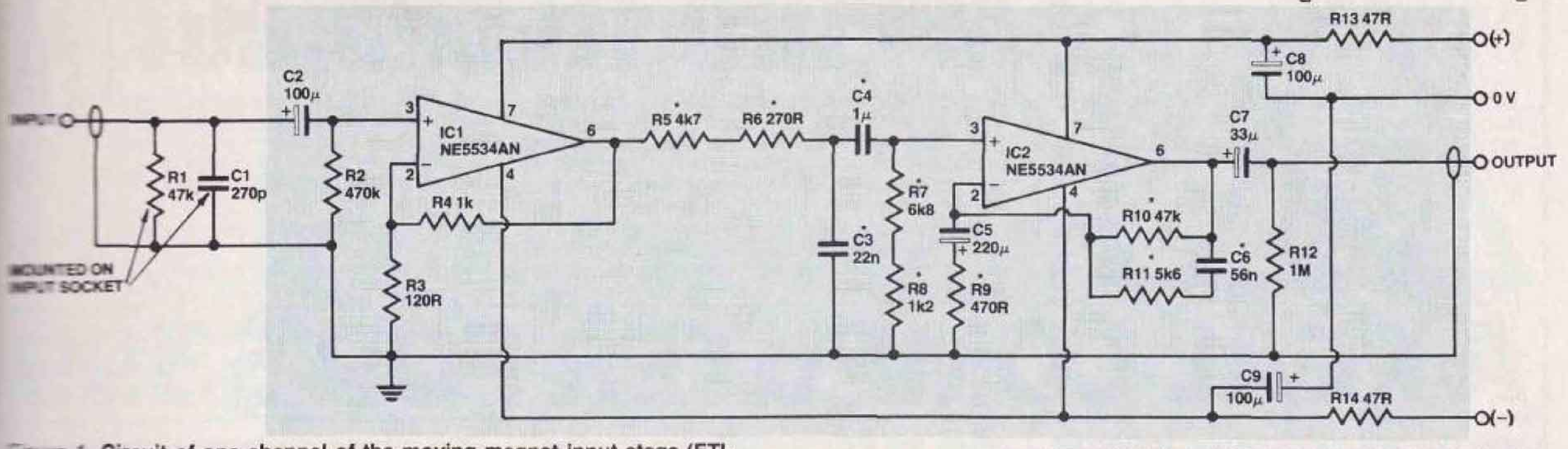


Figure 1. Circuit of one channel of the moving magnet input stage (ETI-468MM). Note that the RIAA equalisation is incorporated in this stage. Components for the other channel are designated R101, C101, IC101, etc.

* DENOTES COMPONENTS ASSOCIATED WITH THE RIAA EQUALISATION

this problem. Firstly the two channels are recorded on the record 180° out of phase, so that the large vertical excursion is replaced by a large horizontal excursion. Secondly, the low end of the frequency response is attenuated before the recording process, so the stylus excursions are decreased. The specific amount of low frequency attenuation is defined as that which would be caused by a first-order high-pass filter with a time constant of 318 μs (i.e: the filter would be formed by an ideal resistor/capacitor filter, in which $R \times C = 318\ \mu s$). To convert from these time constants into frequency, simply apply the equation:

$$f = \frac{1}{2\pi t} \quad (t = \text{time constant})$$

This is equivalent to a 6 dB/octave filter with a −3 dB point at 500 Hz. To prevent the low end from rolling off indefinitely a second 6 dB/octave filter is used to flatten the response again at 3150 μs or 50 Hz. After this equalisation is applied, the stylus excursion of the 20 Hz square wave, for example, is decreased to around 0.3 mm, which is manageable.

Similar problems occur at very high frequencies. If we consider now a 20 kHz square wave at the same output voltage and hence the same recording velocity, the stylus now only moves a total distance of 2.5 μm! Such minute distances are only a few orders of magnitude larger than the surface irregularities in the vinyl, so at these frequencies the signal to noise ratio is poor. To overcome this problem the top end is recorded at a higher level, which increases the stylus excursions and thereby improves the signal to noise ratio. The modifications to the recorded frequency response are referred to as RIAA pre-emphasis or equalisation (RIAA stands for Recording Institute Association of America), and must be corrected for by the input stage. The RIAA playback equalisation must boost the bass end and attenuate the treble end of the audio spectrum to return the overall frequency response to that of a linear system.

Since the low end is amplified most of all by the RIAA playback signal, any turntable rumble or cartridge/turntable resonances will be amplified. Modern power amps are quite capable of delivering full power to a pair of loudspeakers at 10 Hz or below, so appreciable amounts of subsonic content can be fed to the loudspeaker. This is potentially dangerous to the bass driver and decreases the clarity and accuracy of the low end.

In an attempt to overcome this problem the RIAA has proposed a change to its playback equalisation curve. The extreme bass frequencies are attenuated on playback by the addition of another time constant. This takes the form of a single-pole RC filter with a time constant of 7950 μs, i.e: a −3 dB point of 20 Hz. Since the frequency response is already flattened by the 3150 μs time constant, this new time constant gives a 6 dB attenuation rate below about 20 Hz. The resulting RIAA playback equalisation is shown in Figure 2. Note that there are four time constants associated with the proposed RIAA equalisation: 7950 μs, 3150 μs,

HOW IT WORKS ETI-468MM

The input from a moving magnet cartridge is connected to the non-inverting input of an NE5534AN via capacitor C2. R2 provides a dc current path to the input of the differential pair *in the op-amp. The gain of this stage is* determined by the ratio R4 to R3, which is around 8.3 in this case.

The resistor R1 provides a fixed resistive load necessary for best performance from an MM cartridge. Most cartridge manufacturers recommend that the input resistance be shunted by a certain amount of capacitance. This is the purpose of capacitor C1, the value of which should suit most cartridges. If you wish to optimise the value of this capacitor don't forget to allow several hundred picofarads for the shielded cable capacitance.

The best way to ensure that the cartridge is loaded correctly is with a test record containing a square wave track, and an oscilloscope. With the correct cartridge load and a good tonearm/cartridge combination, a good square wave can be obtained.

The value of resistor R1 at 47k is effectively in parallel with R2, giving an input resistance of 43k, slightly below the 47k normally used for MM input stages. This is unimportant, however, and will not affect performance of the cartridge. The important thing is that the value of this resistance remain constant over the full audio spectrum and beyond. In any case the value of the input resistance is easily changed by increasing the value of R1 to, say, 56k instead of 47k.

The output of the first stage is fed to two 6 dB/octave RC filters that provide one half of the RIAA equalisation. Resistors R5, R6 and capacitor C3 form a first-order low-pass filter set at the 75 μs time constant of the RIAA curve. At these frequencies (around 2122 Hz) the 1 μF capacitor appears as a short circuit connecting R7 and R8 in parallel with the capacitor C3. This must be compensated when choosing the value of C3 to ensure the correct RIAA equalisation. Similarly C4, R7 and R8 form a low frequency high-pass filter set at 20 Hz (the 7950 μs time constant).

The output of these two filters is fed to the input of the second op-amp stage. The remaining RIAA equalisation is accomplished by the feedback loop around this stage. At frequencies below 500 Hz the 56n capacitor C6 has a relatively high impedance. The voltage gain is therefore determined by resistors R9 and R10. At higher frequencies, however, where the impedance of C6 is less, both resistors R10 and R11 are in circuit. The capacitor C5 decreases the gain, at dc, of the second stage to unity, ensuring a low dc offset at the output and therefore symmetrical output stage clipping.

The 1M resistor R12 ensures that the dc voltage on the output remains at 0 V. This is important so that operation of the selector switch following the stage will not cause thumps in the output.

Resistors R13, R14 and capacitors C8, C9 isolate the supply to the stage to decrease the effects of interactions between stages and to ensure freedom from 50 Hz ripple.

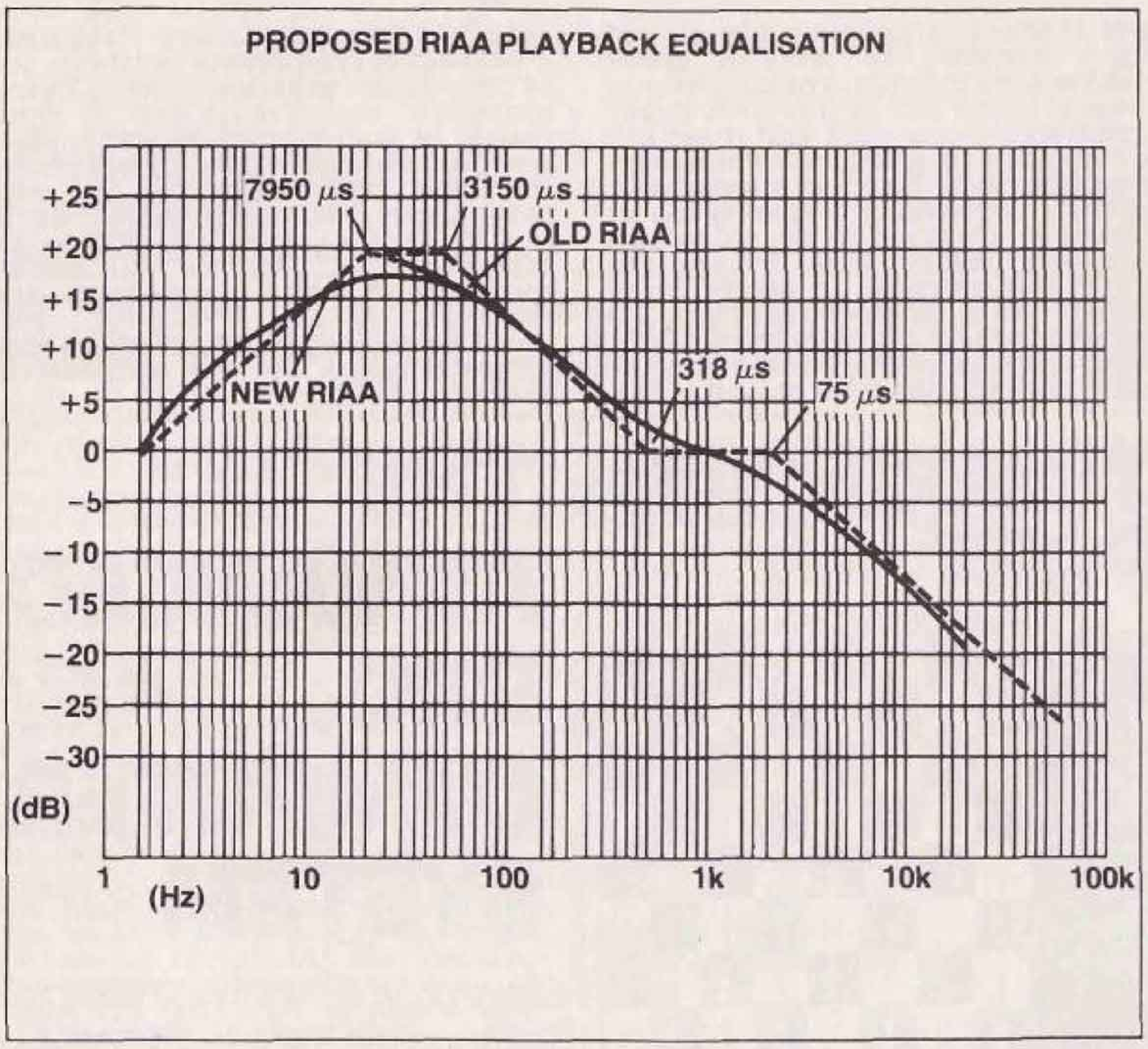
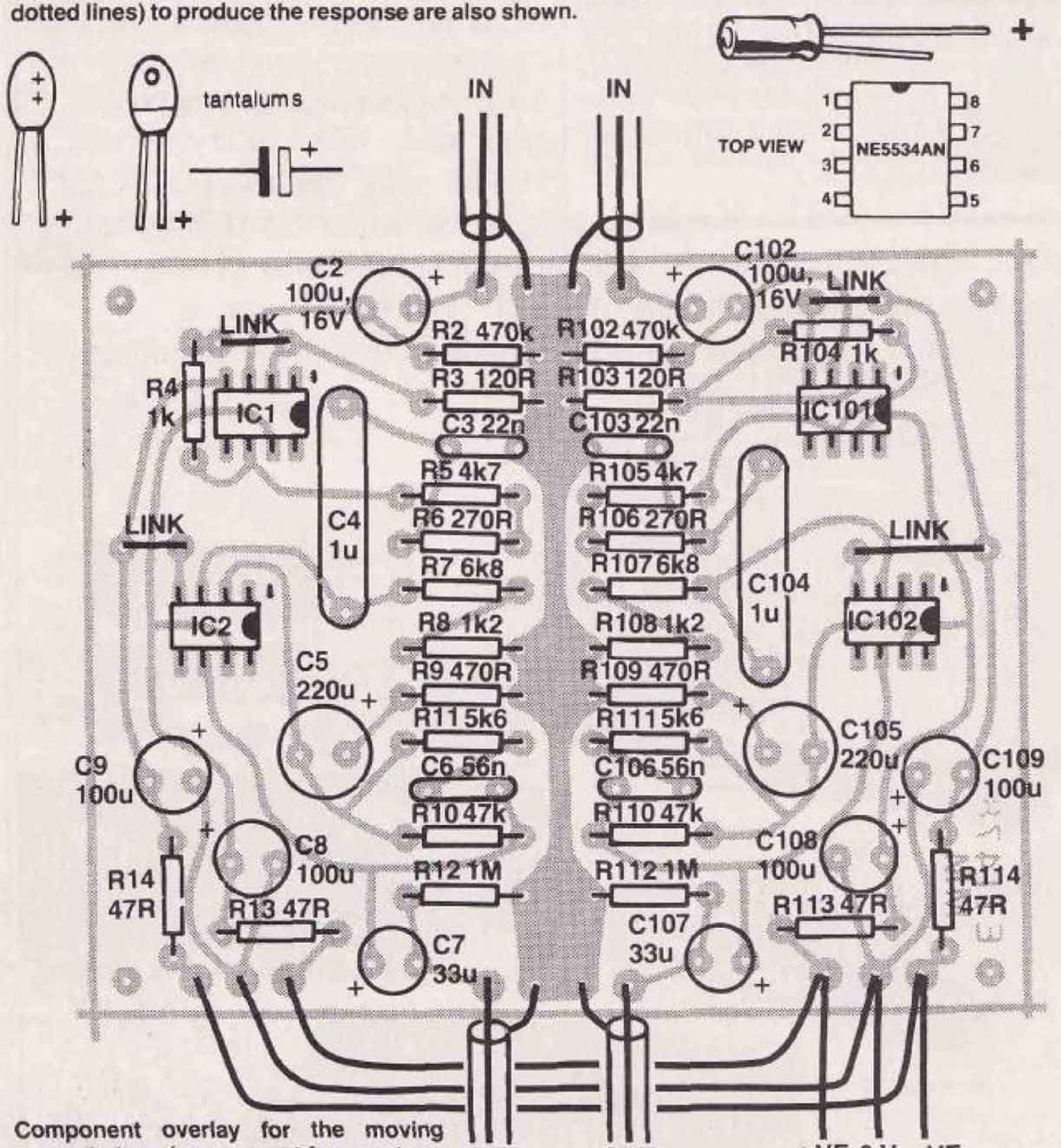


Figure 2. **Old and 'new' RIAA equalisation curves (solid line). The individual time constants (Bode plot — dotted lines) to produce the response are also shown.**



Component overlay for the moving magnet stage (see page 39 for construction hints).

318 μs and 75 μs. These are shown on the Bode plot, which is the dotted line in Figure 2. It should be emphasised, however, that the introduction of this low frequency time constant is not sufficient to remove severe cases of turntable or tonearm resonance. Some preamps incorporate multiple-order subsonic filters that offer a very fast roll-off below 20 Hz. The problem with this, however, is that severe cases of tonearm resonance or rumble generate distortion harmonics well above 20 Hz, into the audio spectrum. The only real cure is to remove the problem at the turntable or tonearm.

Many different techniques are used to give the preamp the desired equalisation. The most common is to include the RIAA equalisation circuitry into the feedback loop of the first stage. Figure 3 shows a very simple MM input stage of the general type often found in medium-priced amplifiers.

Transistor Q1 functions as a standard common emitter amplifier offering a voltage gain that is determined by the total impedance from its collector to earth divided by the total impedance from its emitter to earth. Transistor Q2 is a PNP transistor but functions in an identical manner. The product of their two voltage gains is called the open loop gain of the stage. If a current path is now made available from the output of Q2 back to the emitter of Q1, the voltage

PARTS LIST — ETI-478MM FOR STEREO PC BOARD

Resistors	all ¼W metal film
R1, R101	47k
R2, R102	470k
R3, R103	120R
R4, R104	1k
R5, R105	4k7, 1%
R6, R106	270R, 1%
R7, R107	6k8, 1%
R8, R108	1k2, 1%
R9, R109	470R, 1%
R10, R110	47k, 1%
R11, R111	5k6, 1%
R12, R112	1M
R13,R113,R14, R114	47R
Capacitors	
C1,C101	270p ceramic
C2,C102	100μ, 16 V electro.
C3,C103	22n greencap
C4,C104	1μ greencap
C5,C105	220μ, 16 V electro.
C6,C106	56n greencap
C7,C107	33μ, 25 V electro.
C8,C108,C9,C109	100μ, 25 V electro.
Integrated circuits	
IC1,IC101	NE5534AN
IC2,IC102	NE5534AN

Miscellaneous

1 x ETI-478MM pc board; assorted mounting hardware; shielded cable.

gain will now drop to a new figure called the closed loop gain. This is negative feedback, and it has the effect of decreasing the distortion and increasing the input impedance of the stage. (See Series 5000 MOSFET Power Amp articles in ETI Jan., Feb., March 1981 for more information on negative feedback.)

The RIAA equalisation is introduced by applying the negative feedback via a network with a frequency dependent impedance. However, since this stage relies on the presence of negative feedback to ensure a satisfactorily high input impedance, the input impedance will vary as a function of frequency. The cartridge, however, must be loaded by a constant resistance if cartridge impedance interaction is to be avoided. Furthermore, since the negative feedback is coupled to the complex output impedance of the cartridge via the base-emitter junction of Q1, the negative feedback and hence the frequency response of the stage can be affected by the cartridge itself. As a result this type of stage can suffer badly from cartridge impedance interaction.

In the development of the Series 5000 preamp several input stage configurations were tested for noise, distortion and cartridge impedance interaction. When a medium-priced moving magnet cartridge was connected to a stage like that in Figure 3, severe cartridge impedance interaction was evident. The frequency response of the preamplifier peaked above 2 dB at 13 kHz and rolled off rapidly above 15 kHz. The same cartridge when connected to the Series 5000 MM amp exhibited quite a good frequency response to beyond 20 kHz, and the frequency response graph obtained was identical to that when a FET buffer amp was placed between the cartridge and the input stage, indicating almost total lack of cartridge impedance interaction in the 5000 stage. This is a result of the use of the separate linear gain stage formed by IC1 (see Figure 1) to isolate the cartridge from the RIAA equalisation.

The Series 5000 Preamp conforms to the proposed RIAA equalisation in Figure 2. The 75 μs and 7950 μs time constants are obtained by passive RC filters at the output of the first stage. Resistors R5, R6 and capacitor C3 form a simple 6 dB/octave low-pass filter with a -3 dB point at 2122 Hz, and

$$t = \frac{1}{2\pi f} = \frac{1}{2\pi(2122)} \doteqdot 75\,\mu s.$$

Capacitor C4, together with resistors R7 and R8, form a 6 dB/octave high-pass filter with a -3 dB point at 20 Hz, which is equivalent to a 7950 μs time constant. The two remaining time constants are introduced into the negative feedback of IC2 and are formed by the values of resistors R9, R10, R11 and capacitor C6.

This method of generating the RIAA curve offers a number of advantages over the more conventional method.

Firstly there is low interaction between the different time constants, so that the RIAA curve can be optimised for a particular cartridge more easily by changing the resistor or capacitor values slightly. If the 75 μs time constant is included in the negative feedback of a stage, the gain of the stage must decrease to unity at a suitably high frequency, so the stage must be compensated for unity gain to prevent instability. In the Series 5000 MM stage the gain of the second stage does not drop below 10; since the NE5534AN is internally compensated for gains of 3 or above, no additional compensation is required.

▶

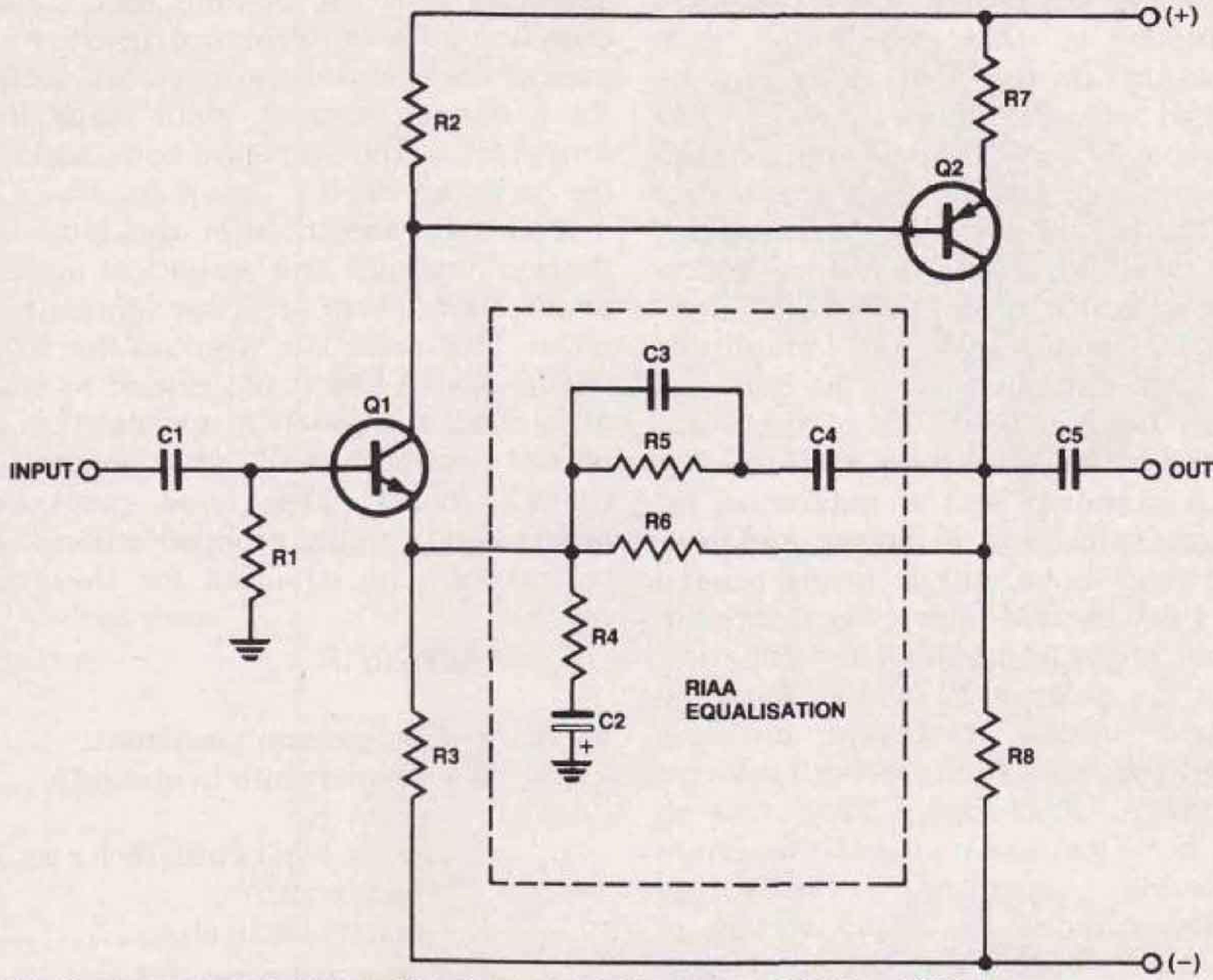
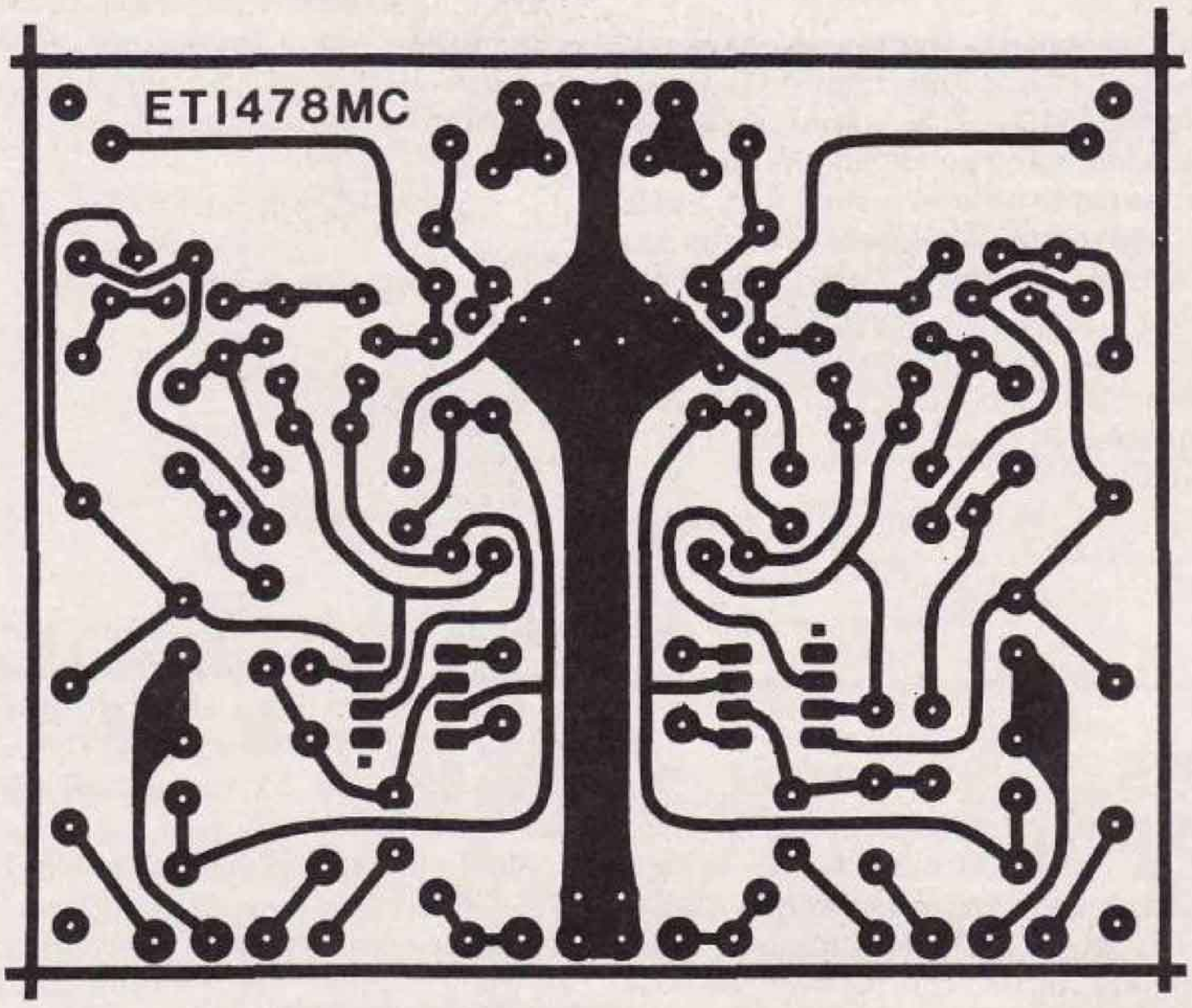


Figure 3. **Typical moving magnet input stage found in medium-priced amplifiers.**



Full-size artwork for the moving coil input stage.

Another advantage of the two-stage approach is that the total gain necessary in the MM stage can be divided between the two stages, so more negative feedback is available for each stage. This will have the effect of decreasing non-linearities in the stages, provided the stages conform to the criteria for the avoidance of SID (slew-induced distortion) and amplitude overload. Fortunately, in the case of a phono input stage, both of these are limited by the recording medium. The RIAA standard sets a maximum recording velocity of 25 cm/sec, and most cartridges have output levels around the 1 mV/cm/sec figure. So maximum output levels from such a cartridge will be in the order of 20-30 mV. Even the highest output cartridge produces signal voltages usually in the 5 mV/cm/sec range. Combining a worst case of, say, 5 mV/cm/sec with the maximum allowable recording velocity of 25 cm/sec yields an output voltage of 125 mV. To ensure that the input stage cannot be overloaded we simply set the gain of these stages so that this maximum input signal cannot drive the output of the input stages into clipping. The NE5534AN is capable of driving to within 2 V of the supply voltage, so a supply voltage of ±15 V gives the desired gain of around 75. We have divided this gain between the two input stages so that the first stage has a gain of 8.3 and the second stage a gain of 9 in the midband region (the actual gain of the second stage is of course a function of frequency due to the RIAA equalisation).

As a result the total harmonic distortion of this MM input stage is well under 0.001%. The actual measured distortion using an HP3580A spectrum analyser was around 0.0005% at 1 kHz. (At these distortion levels even the best distortion analysers are practically useless, since the distortion is well below the level of noise.) Similarly, intermodulation distortion (IMD) was measured at well below the 0.001% figure.

Noise

Another very important parameter for both MC and MM input stages is noise performance. Since an op-amp is used as the first stage of the MM input amp, we have only limited control over the noise performance of the stage. It is therefore essential that the op-amp used have excellent noise performance. In order to predict the necessary noise performance for a moving magnet input stage we must look at the sources of noise within the cartridge itself.

It can be shown from the laws of thermodynamics and statistical mechanics that every resistor generates noise. This noise is a result of the way nature works and is not caused by imperfection in a practical resistor (i.e: a perfect resistor will still generate thermal noise). This noise must be added to any signal dropped across the resistance. The equation for thermal noise is:

$$\overline{e_n} = \sqrt{(4kTR\triangle f)}$$

where k = Boltzmann's constant,
T = temperature in absolute units (K)
$\triangle$ = noise bandwidth (brickwall bandwidth)
R = resistance in ohms
$\overline{e_n}$ = average noise voltage

This equation predicts that thermal noise is raised by increasing resistance temperature or the bandwidth of the measuring equipment. So the frequency response of the apparatus used to determine thermal noise must be quoted if the figure is to be meaningful. Furthermore, the $\triangle f$ here refers to a 'brickwall frequency response', not the usual half-power bandwidth, although for many purposes this is sufficiently accurate. To overcome this problem noise performance is often quoted in the form of total equivalent input noise and expressed in units of $nV/\sqrt{Hz}$ ($1\ nV = 10^{-9}\ V$). This is justified by the equation for thermal noise, i.e:

$$\text{since } \overline{e_n} = \sqrt{(4kTR\triangle f)}$$

$$\text{then } \overline{e_n} = (\sqrt{\triangle f})(\sqrt{(4kTR)})$$

$$\text{or } \frac{\overline{e_n}}{\sqrt{\triangle f}} = \sqrt{(4kTR)}.$$

So the ratio:

$$\frac{\overline{e_n}}{\sqrt{\triangle f}}$$

depends only on temperature and resistance, and this is just what we want. In order to get from this figure to an actual total equivalent noise figure we simply multiply by the square root of the bandwidth.

Most moving magnet cartridges have a coil resistance around 500 ohms. This resistance will generate thermal noise, so the cartridge itself limits the best possible signal-to-noise ratio. Using the equation for thermal noise we obtain for the noise of the cartridge:

$$\frac{\overline{e_n}}{\text{Hz}} = \sqrt{(4 \times 1.37 \times 10^{-23} \times 290 \times 500)}$$

(assuming temperature of resistor is ÷ 290K).

$$\text{i.e: } \frac{\overline{e_n}}{\sqrt{\text{Hz}}} = 2.8 \times 10^{-9}$$

$$\text{i.e: } \overline{e_n} = 2.8\ nV/\sqrt{Hz}.$$

We can express this in more familiar terms by converting the cartridge noise figures into a signal-to-noise ratio figure. In audio we can regard the bandwidth in question to be around 20 kHz, i.e: $\sqrt{\triangle f} = 140$, and $140 \times 2.8\ nV/\sqrt{Hz} = 392\ nV$. If the average output level of the cartridge is around 5 mV, the signal-to-noise ratio is given by:

$$20 \log \frac{5 \times 10^{-3}}{392 \times 10^{-9}} \div 82\ \text{dB}$$

This figure represents the best signal-to-noise ratio possible with most moving magnet cartridges, since this is due to noise generated within the cartridge itself. A well-designed input stage should approach this noise figure as closely as possible without sacrificing performance in other equally important parameters such as distortion and frequency response.

The noise generated by an active device is determined by a number of factors, the most important of which is the current flowing through the device. However, since we have elected to use a high-quality operational amplifier for the input stage, we have no control over device current. All we can do is choose a low-noise op-amp and avoid degrading its noise figure as much as possible. The NE5534AN has a recommended equivalent input noise voltage around $4\ nV/\sqrt{Hz}$, only 3 dB above the noise generated by the cartridge itself! In order not to degrade this figure we must keep all resistances in series with the cartridge as low as possible. Any additional resistance will generate a thermal noise voltage of its own, which must be added vectorally to that generated by the cartridge. From the basic equation of thermal noise generated by two individual resistors R_1 and R_2, for example, we obtain:

$$\overline{e_{n_1}} = \sqrt{(4kTR_1\triangle f)} \text{ and}$$
$$\overline{e_{n_2}} = \sqrt{(4kTR_2\triangle f)}$$

Here we assume that both resistances are at the same temperature. Since these noise voltages are not correlated (i.e: they consist of 'randomly' changing voltage) we add them using the vector sum:

i.e: $\overline{e_{nT}2} = \overline{e_{n1}2} + \overline{e_{n2}2}$

where $\overline{e_{nT}2}$ is the square of the total equivalent noise voltage.

Therefore $\overline{e_{nT}2} = 4kT\triangle f(R_1 + R_2)$
or $\overline{e_{nT}} = \sqrt{(4kT\triangle f(R_1 + R_2))}$.

If R_1 now represents the cartridge resistance and R_2 the value of an added resistance equal to the value of R_1, we get:

$$\overline{e_{nT}} = \sqrt{(4kT\triangle f(2R_1))} = \sqrt{2}\sqrt{(4kT\triangle fR_1)}$$

$$\text{or } \overline{e_{nT}} = 1.4\overline{e_{nT'}}$$

equivalent to a 3 dB decrease in the signal-to-noise ratio.

Figure 4 shows the standard technique for connecting an op-amp to a signal generator such as a moving magnet cartridge. Most op-amps, and certainly the 5534, have input stages that consist of a differential pair, providing both inverting and non-inverting inputs.

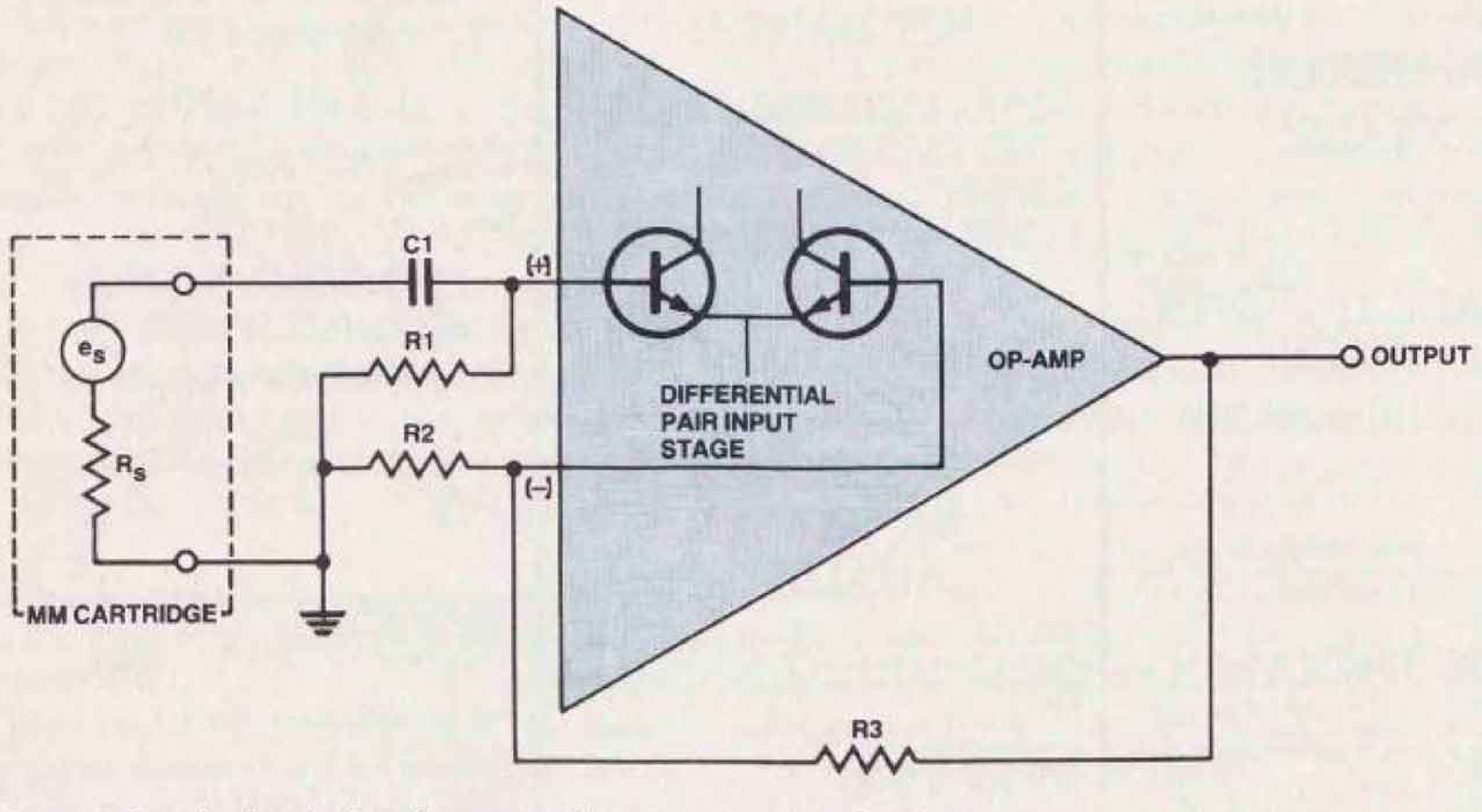


Figure 4. **Standard technique for connecting an op-amp to a signal source.**

The effective signal voltage generator of the cartridge is represented by e_s and the cartridge resistance by R_s. Resistor R1 in this case would be 47k, so that the cartridge would have the correct load resistance. (The input impedance of the op-amp is very high and can be ignored for this discussion.) Capacitor C1 prevents any dc current flowing through the cartridge from the non-inverting input. Since the combination of R1 and C1 forms a 6 dB/octave high-pass filter, the value of C1 would ordinarily be chosen so that the resulting −3 dB point was well below the audio spectrum, around 5 Hz for example. This will occur when the impedance of C1 is equal to that of R1, i.e: 47k. Since the reactance of the capacitor is given by the equation:

$$X_c = \frac{1}{2\pi fC}$$

we have:

$$C = \frac{1}{2\pi fX_c}$$

$$\text{In this case } C = \frac{1}{2\pi \times 5 \quad \times 47 \times 10^3}$$

$$\doteqdot 6.77 \times 10^{-7} \text{ Farads.}$$

So to obtain an adequately flat frequency response a suitable value for C1 would be 680 nF (0.68μF), which is convenient since a greencap could be used.

When noise considerations are taken into account, however, this value is entirely unsuitable. The increasing impedance of C1 at low frequencies, while not sufficient to cause gross frequency response errors, will seriously degrade the noise performance of the stage. At sufficiently low frequencies the impedance seen by the non-inverting input will be simply the value of R1. Using the equation for thermal noise given earlier, we can calculate the resulting signal-to-noise ratio. Since

$$\overline{e_n} = \sqrt{(4kT\triangle fR_1)},$$
$$\overline{e_n} = \sqrt{(4 \times 1.37 \times 10^{-23} \times 290 \times 20 \times 10^3 \times 47 \times 10^3)} = 3.87\ \mu V,$$

only 62 dB below 5 mV.

Furthermore, since the input stage is a noise generator, a low source impedance is necessary to minimise the resulting noise at the output of the op-amp. To overcome this problem we increase the value of C1 so that at worst its impedance at, say, 3 Hz is comparable to that of the cartridge, i.e:

$$C = \frac{1}{2\pi \times 3 \times 500} = 106 \times 10^{-6}\text{F}.$$

So a value around 100 μF should suffice. Notice that this capacitor would have to be an electrolytic or tantalum. Tantalum capacitors are not recommended, however, since their capacitance can be modulated by the input signal, producing considerable distortion at low frequencies.

The value of resistor R2 must also be low, so that the source impedance to the inverting input of the op−amp can be kept as low as possible. The limitation here is due to the minimum load impedance allowable on the output of the op-amp. Since the gain of the stage is given by the equation:

$$A_v = \frac{R2 + R3}{R2},$$

the ratio of R2 and R3 is determined by the desired voltage gain. At the same time, however, the total resistance R2 + R3 represents the load on the output stage of the op-amp. Since this must not be less than a certain specified resistance, determined by the individual op-amp used, a minimum value for R2 is predicated. In the Series 5000 MM input stage, for example, the required voltage gain in the first stage is around 8.3, so:

$$\frac{R2 + R3}{R2} = 8.3$$

The NE5534AN has a measured minimum load impedance of 600 ohms, and for minimum distortion it is desirable to increase this slightly, for example to around 1k2. Therefore:

$$\frac{1k2}{R2} = 8.3 \text{ or } R2 \doteqdot 144R$$

A suitable value for R2 would be 120 ohms, making R3 1k to give the required voltage gain. Fortunately this value for R2 is low enough not to have significant effect on the noise performance.

Similar measures must be adopted around the second stage. At low frequencies the non-inverting input of IC2 (see Figure 1) has an input source resistance determined by R7 and R8, i.e around 8k. The noise performance of the second stage would be improved if this value could be decreased. Unfortunately this would entail increasing the value of C4, which is not practical since this capacitor must be a greencap if the preamp is to conform accurately to the RIAA curve. This is not really a problem, however, since the voltage gain in the first stage increases the signal voltage at the input of IC2 to around 40 mV for a 5 mV input signal ensuring a sufficiently good signal-to-noise ratio in the second stage. ▶

The moving coil input stage

The subject of noise performance is particularly important for a moving coil input stage. The moving coil cartridge works on exactly the same principle as the moving magnet. The signal voltages produced are the result of relative motion between a coil of wire and a magnetic flux. In this case, however, the magnet assembly is mounted rigidly to the cartridge body and the coils are mounted on the cantilever assembly; hence the name 'moving coil'.

In order for the total mass and therefore the inertia of the stylus/cantilever system to be kept to a minimum, the coils are made with very fine wire and a small number of turns. Typical output voltages for moving coil cartridges vary widely from one manufacturer to another, but a figure of 40 μV/cm/sec is probably a reasonable compromise. A gain of 25 is therefore required to boost this voltage to that of a typical moving magnet cartridge. Once again we can calculate the best possible signal-to-noise ratio for a moving coil cartridge based on its thermal noise. The coil resistance of a moving coil cartridge with an output of 40 μV/cm/sec would be approximately 20 ohms (although this figure can vary widely, typcially 5-50 ohms).

From the equation for thermal noise we obtain:

$$\frac{\overline{e_n}}{\sqrt{Hz}} = \sqrt{(4kTR)},$$

$$\text{i.e.: } \frac{\overline{e_n}}{\sqrt{Hz}} = \sqrt{(4 \times 1.37 \times 10^{-23} \times 290 \times 20)}$$

$$\doteqdot 0.56\ \text{nV}/\sqrt{\text{Hz}}.$$

The total noise over a 20 kHz noise bandwidth is therefore:

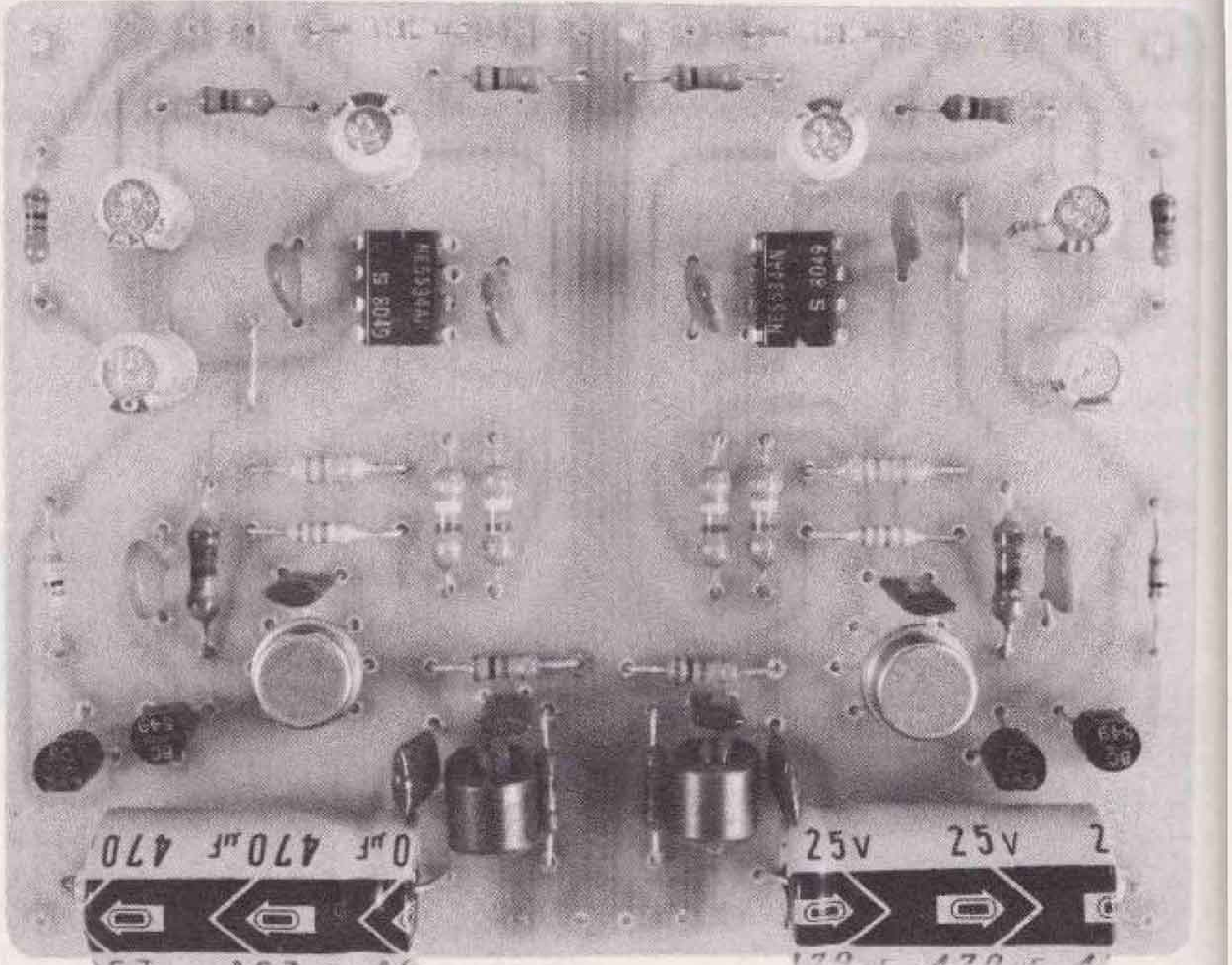
$0.56\ \text{nV} \times \sqrt{(20 \times 10^3)}$
i.e: $0.56\ \text{nV} \times 140 \doteqdot 78\ \text{nV}$.

Since the cartridge output voltage will be around 40 μV/cm/sec x 5 cm/sec, i.e: 200 μV for a recording velocity of 5 cm/sec, the resulting signal-to-noise ratio will be:

$$20 \log \frac{200 \times 10^{-6}}{78 \times 10^{-9}}$$

or around 68 dB unweighted.

This figure is only approximate, of course, but it is roughly correct and represents the best possible signal-to-noise ratio with a moving coil cartridge. The object is to design a preamplifier that will approach this noise figure and maintain a flat frequency response, low distortion and a constant resistive input impedance. At these noise levels we cannot use an NE5534AN in a circuit like the MM input stage. The total equivalent input noise in that case was around 4 nV/$\sqrt{\text{Hz}}$, i.e: 560 nV. The resulting signal-to-noise ratio would be only 51 dB with respect to an input signal of 200 μV.


The completed moving coil input stage (ETI-478MC). Note that this project could be used 'stand-alone' as an MC head amp.

In order to achieve a satisfactory noise performance it is necessary to look at the various sources of noise in bipolar transistors and decrease the total equivalent input noise through optimum biasing of the input stage and choice of the first transistor.

One source of noise in the transistor is of course thermal noise. We saw before that to minimise thermal noise it was necessary to ensure a low source resistance over as broad a frequency range as possible. In order to do this for the MC stage the total resistance in series with the source must be kept to a similar value to the source resistance, i.e: around 10 or 20 ohms, depending on the cartridge.

The problem is that the resistance of the base-emitter junction of most bipolar transistors, called the base spreading resistance, is usually much higher than this. One solution is to use a large number of low-noise transistors in parallel to form the input transistor, thus decreasing the base spreading resistance. This was the technique used in the Series 4000 moving coil preamp, published in ETI October 1979. Another solution is to use a power transistor, such as a 2N3055, as the input transistor, and the results using this method can be quite good! The third alternative, and the one we elected to use in this design, is to make use of an exceptional matched pair produced by National Semiconductor. This device, the LM394, has a low base resistance, very low noise and a high h_{FE} of around 500. (A data sheet for the LM394 will be included in next month's article).

Another source of noise in bipolar transistors is shot noise or base current noise. This is a white noise generator (i.e: the average amplitude of the noise current is constant with frequency), but the noise is increased if emitter current is increased. The base resistance, however, is also a function of the current flowing in the emitter, and is given roughly by the equation:

$$r_b = \frac{26}{I_{E(mA)}}.$$

The resistance of the base *decreases* with increasing emitter current, so noise voltage produced by thermal noise across the base resistance is decreased by increasing the emitter current. In a bipolar transistor, therefore, we have two distinct sources of noise, one increasing with emitter current while the other decreases. For this reason an optimum emitter current exists which represents the best compromise between these two noise sources. With an LM394 operated from source resistances typical of moving coil cartridges, the optimum emitter current is around 8 mA, much higher than would normally be used in an input stage. The result, however, is a very low

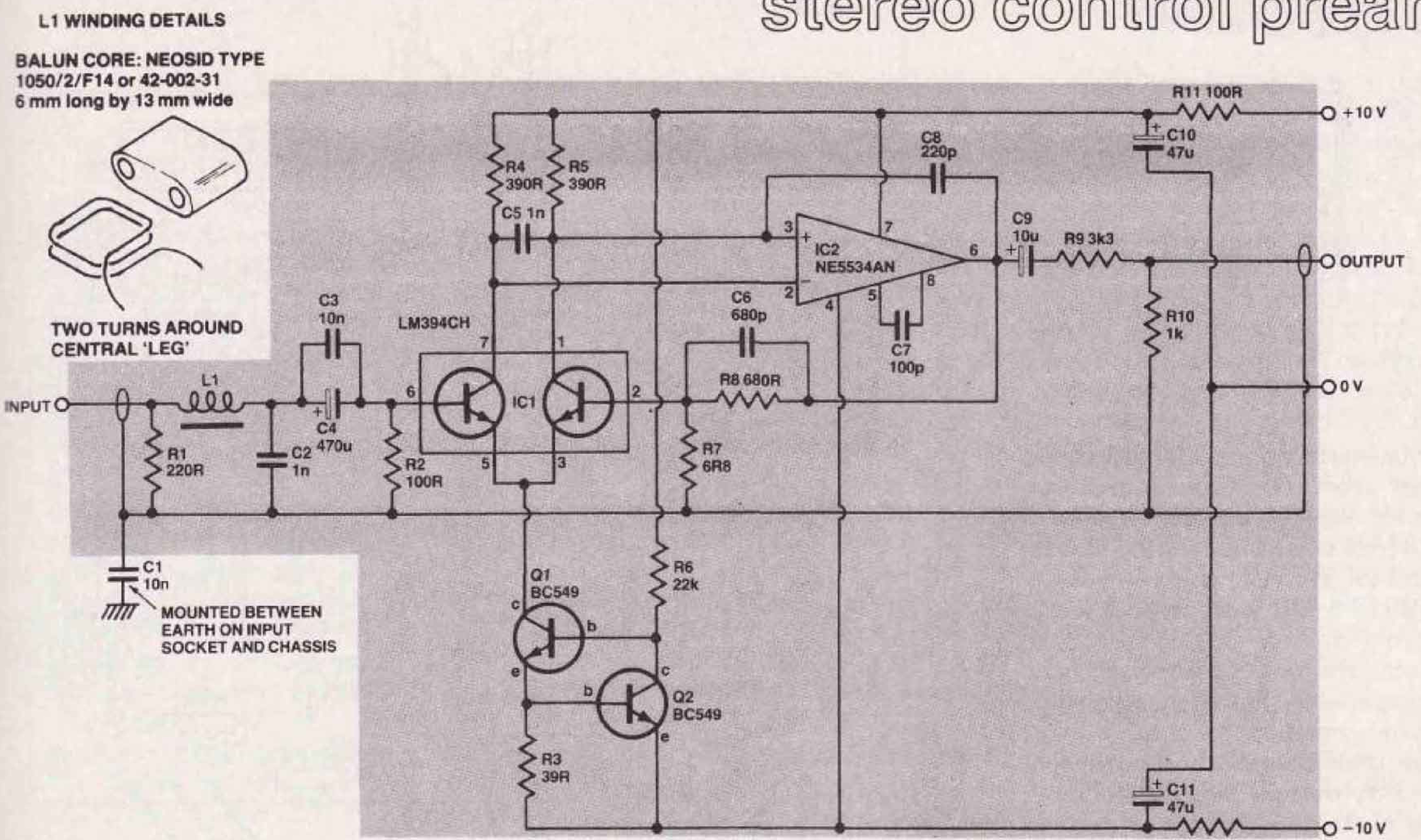


Figure 5. **Circuit of one channel of the moving coil input stage. Components for the other channel are designated R101, C101, IC101, etc. Note that data for the LM394 and NE5534 devices are included at the end of this article.**

value of input noise for source resistances around 10 ohms.

The complete circuit diagram for the moving coil input stage is shown in Figure 5. The collectors of the LM394 are connected to the input of an NE5534, which functions as a high-gain differential amplifier, providing adequate open loop gain to ensure low distortion and a flat frequency response when negative feedback is applied. The input choke is used to minimise the stage's susceptibility to RF noise.

The input impedance of the stage is determined by the parallel combination of R1 and R2, around 65 ohms for the values shown. This should be suitable for most moving coil cartridges, but is easily changed if required. The dc operating point of the LM394 is determined by the constant current source formed by Q1, Q2, R3 and R6. So the current in resistor R2 is determined by this constant current source and the dc current gain of the LM394. Hence the value of R2 can be increased, in order to increase the input impedance, over a fairly wide range of values without affecting the operation of the circuit.

Once again the input coupling capacitor C4 is used to prevent dc current from flowing through the cartridge. Capacitor C4 is shunted by C3, a 10n capacitor, so that the base of the first transistor in the LM394 is decoupled for RF, through C2. Capacitor C2 represents a shunt capacitance to ensure correct loading of the moving coil cartridge. The value shown should be suitable for most cartridges, but can be changed for optimisation with any particular cartridge.

To prevent loading the 5534A, the feedback resistor R8 is kept above 600R, i.e: 680R. Resistor R7 effectively increases with the cartridge and must be kept as low as possible for best noise performance. The value of 6R8 chosen gives the stage a gain of around 100, which is too high. This is corrected, however, by a simple passive voltage divider at the output, formed by R9 and R10. Capacitor C9 doubles as a feedback isolation capacitor to ensure that reactive components in the load cannot cause a phase shift sufficient to cause oscillation. ▶

HOW IT WORKS ETI-478MC

The input from a moving coil cartridge is fed via L1 and capacitors C3 and C4 to the base of one of the transistors in the LM394, which functions as a differential input stage.

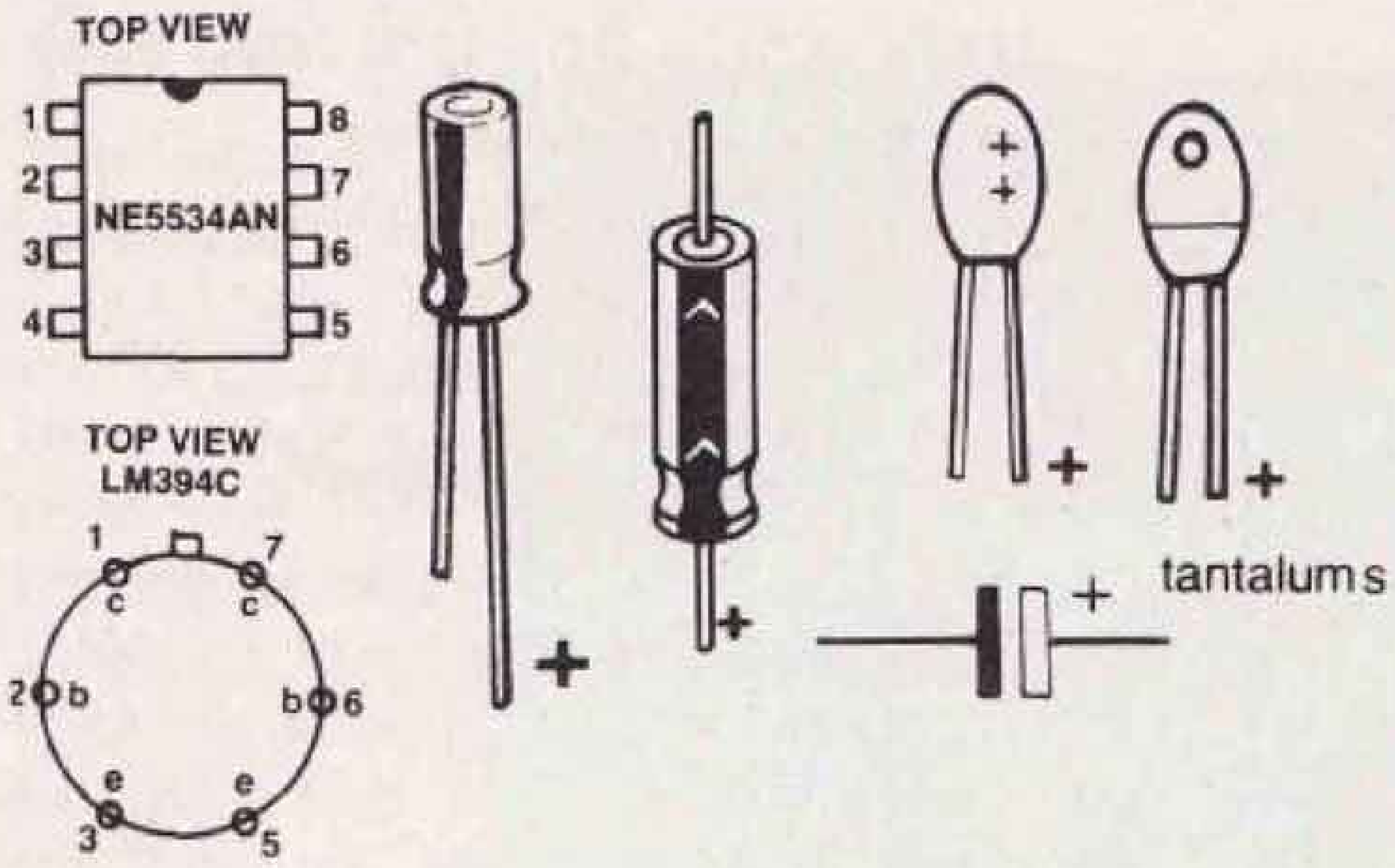
Q1 and Q2 form a constant current source, which stabilises the dc operating point and ensures a high impedance source to the emitters of the differential pair. The constant current source works by ensuring that a constant voltage is maintained across a fixed value of resistance. Resistor R3 is used for this purpose, with the base emitter voltage of Q2 expressed across it. If the current through R3 were to try to increase even slightly, the voltage on the base of Q2 would be increased, turning Q2 on harder. This causes the voltage on the collector of Q2 to decrease, decreasing the current through R3. So Q2 provides negative feedback acting to correct any deviations in the current flowing through the differential pair.

The collectors of the LM394 are shunted by the 1n capacitor C5. This decreases the gain of the first stage at high frequencies and helps to ensure stability (i.e: freedom from high frequency oscillations).

The input stage is operated in full differential mode by connecting both collectors to inputs of the NE5534AN. If this is not done the voltage gain of the input stage is decreased and the signal-to-noise ratio is degraded. Because differential pairs have two base-emitter junctions in the input circuit, their total equivalent input noise is inferior to that of a single transistor. However, since it is possible using a differential pair to obtain noise figures of the same order of magnitude as the thermal noise of the cartridge, the marginal decrease in the theoretically best signal-to-noise ratio is of little consequence. On the other hand the inherent linearity of a differential pair offers a significant advantage over a single transistor, improving both distortion and high frequency stability.

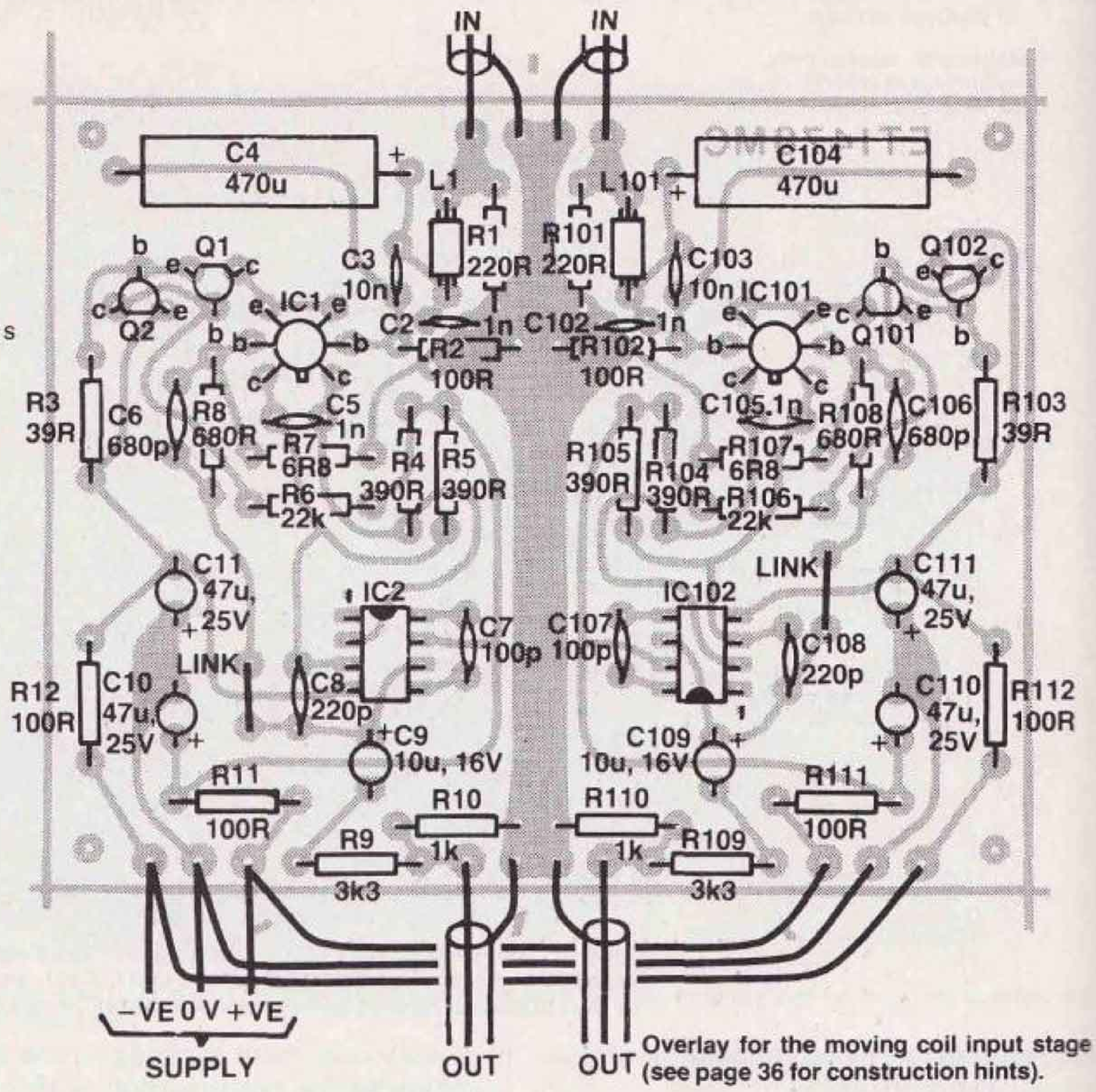
Capacitor C7 ensures stability of the op-amp by providing adequate compensation for the increased gain around the stage due to the differential pair. C9 provides dc isolation of the stage. The resistors R9 and R10 form a potential divider to decrease the signal level to that suitable for the MM input. If the particular moving coil cartridge used requires a different amount of voltage gain than is provided, the value of R9 can be changed accordingly. Replacing R9 with a short circuit (i.e: a piece of tinned copper wire in place of the resistor on the circuit board) increases the voltage gain of the stage to slightly over 100.

The two RC networks R11, C10 and R12, C11 provide isolation of the supply voltage from other stages using the same power supply. This decreases interactions between stages, thereby improving crosstalk and the overall stability of the preamplifier.



The noise performance of the stage is extremely good. The total equivalent input noise was measured at 83 nV over a 20 kHz noise bandwidth. This is equivalent to 0.6 nV/√Hz or a signal-to-noise ratio of 68 dB with respect to an input signal voltage of 200 μV. This might sound like only an average noise figure compared to that attainable with the moving magnet preamp, but it should be remembered that the noise generated by the cartridge itself is of this order of magnitude!

Another point worth mentioning here is that all the noise figures quoted so far in this article are flat or unweighted measurements. This means that the measurement was carried out with a noise and distortion analyser with a flat frequency response over the quoted noise bandwidth, usually 20 kHz. This is convenient and meaningful for the analysis of electrical circuits so long as the frequency distribution of the noise is also known. Probably the most useful way to quote noise figures at audio frequencies, however, is to graph noise circuits of nV/√Hz against frequency. The problem with flat noise measurements is that the human hearing mechanism does not detect all frequencies with equal sensitivity. For example, a noise generator with a high average noise voltage in the 1 kHz to 5 kHz region will be perceived as 'noisier' than one with an identical unweighted noise measurement but with higher average noise voltage in the 100 Hz to 1 kHz band. To overcome this problem the frequency response of the measured equipment can be modified to accent or 'weight' the appropriate frequency bands. The most common weighting curve used in audio measurement is shown in Figure 6. The use of A-weighting gives a better indication of the apparent loudness of a noise voltage than do unweighted ('flat') measurements, and this is the reason almost all manufacturers quote A-weighted noise figures. A-weighted noise measurements for the Series 5000 MM and MC stages are quoted with specifications elsewhere in this article.



Overlay for the moving coil input stage (see page 36 for construction hints).

ETI-478MC PARTS LIST FOR STEREO PC BOARD

Resistors	all ¼W metal film, 5% unless noted otherwise
R1,R101	220R
R2,R102,R11, R111,R12,R112	100R
R3,R103	39R
R4,R104,R5,R105	390R
R6,R106	22k
R7,R107	6R8
R8,R108	680R
R9,R109	3k3
R10,R110	1k
Capacitors	
C1,C101,C3,C103	10n greencap
C2,C102,C5,C105	1n greencap
C4,C104	470u 16 V electrolytic
C6,C106	680p ceramic
C7,C107	100p ceramic
C8,C108	220p ceramic
C9,C109	10u 16 V electrolytic
C10,C110, C11,C111	47u 25 V electrolytic
Semiconductors	
Q1,Q101,Q2,Q102	BC549
Integrated circuits	
IC1, IC101	LM394CH
IC2, IC102	NE5534AN
Miscellaneous	
L1	Two turns on ferrite balun core, Neosid type 1050/2/F14 or 42-002-31

ETI-478MC pc board; shielded cable; assorted mounting hardware, etc.

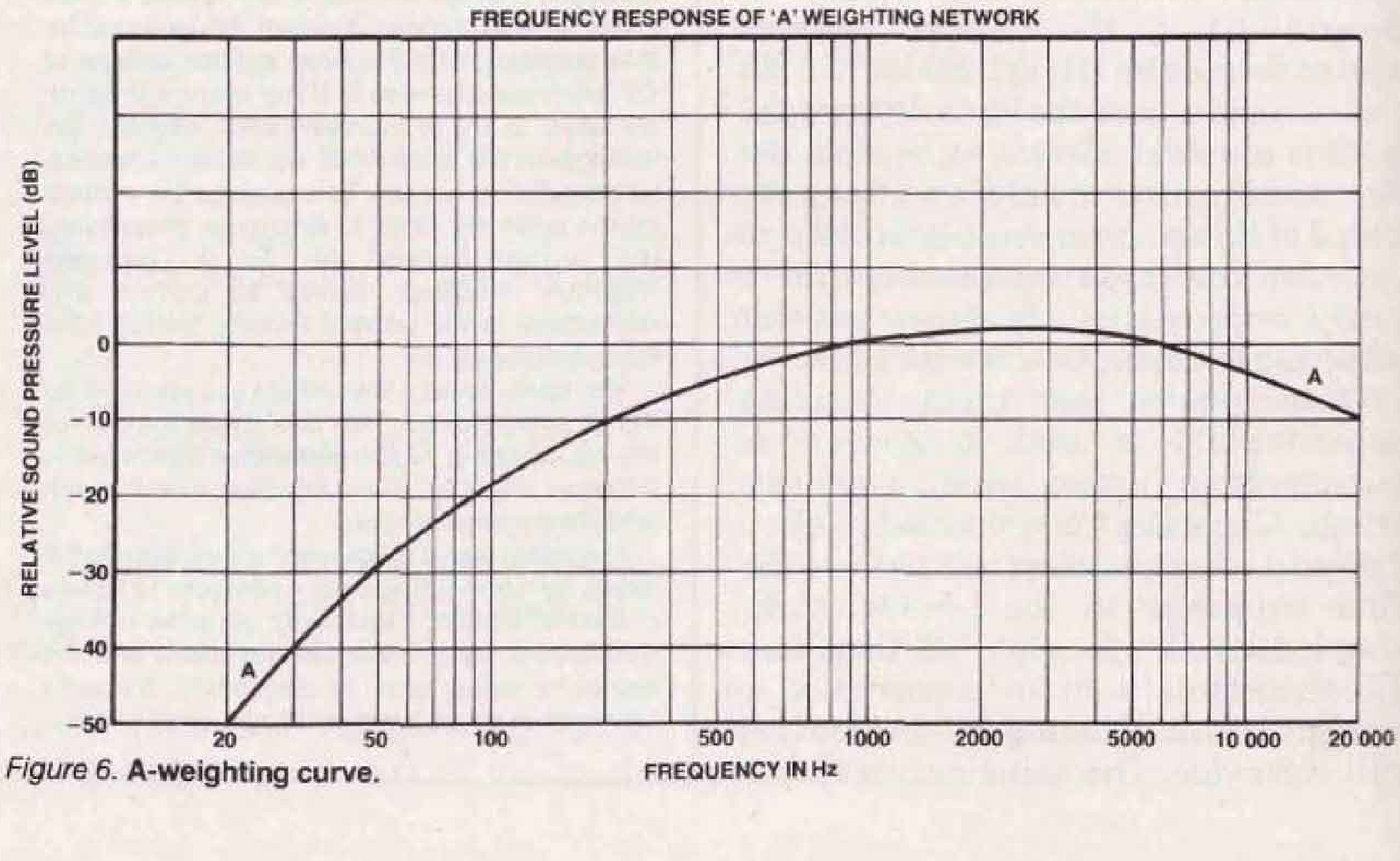


Figure 6. **A-weighting curve.**

Construction

Construction of both boards is relatively straightforward, since almost all the components are mounted on the pc boards. Resistor R1 and capacitor C1 on the moving magnet board are intended to be mounted directly across the back of the input socket. Order of construction is not critical, although it is probably easier to mount small components first, followed by the larger components such as the electrolytic capacitors and IC1. Be careful to watch the orientation of the electrolytic capacitors, ICs and transistors; these components will be damaged if the unit is powered up with them inserted incorrectly. Shielded cable should be used on all inputs and outputs. We have used mono shielded cable rather than the stereo type for ease of soldering.

Each of the pc boards is a stereo input amplifier, with each channel sharing a common input earth track running down the centre of the board. The power supply wiring from each channel on the MC board can be connected in parallel, so only three wires (+, 0, −) need to be brought out for power to the MC amp. The MM board is similar, but the power supply wiring for the two boards should be kept separate, since they run from slightly different supply voltages. If you are using these boards in a different application to that of the Series 5000 preamp and wish to run both boards from the same supply voltage (i.e: 15 V-0-15 V), the value of resistors R11 and R12 on the MC board should be increased to around 270R, to decrease the power dissipation on the LM394.

The input earth is *not* connected to the 0 V line from the power supply at any place on the pc boards. This means that without a separate 0 V connection added to the input stage they will *not* work. This has been done deliberately to ensure that hum present on the earth line, due to supply bypass capacitors for example, cannot modulate the signal earth, producing hum in the output. The 0 V line on the pc boards is in fact a separate supply bypass earth line and is not equivalent to the signal earth. Full details of the signal earth connections will be given in Part 3 of this series of articles. For the purpose of testing the stage, however, a separate wire should be run from the centre point (0 V point) of the power supply used to the signal earth at the input sockets.

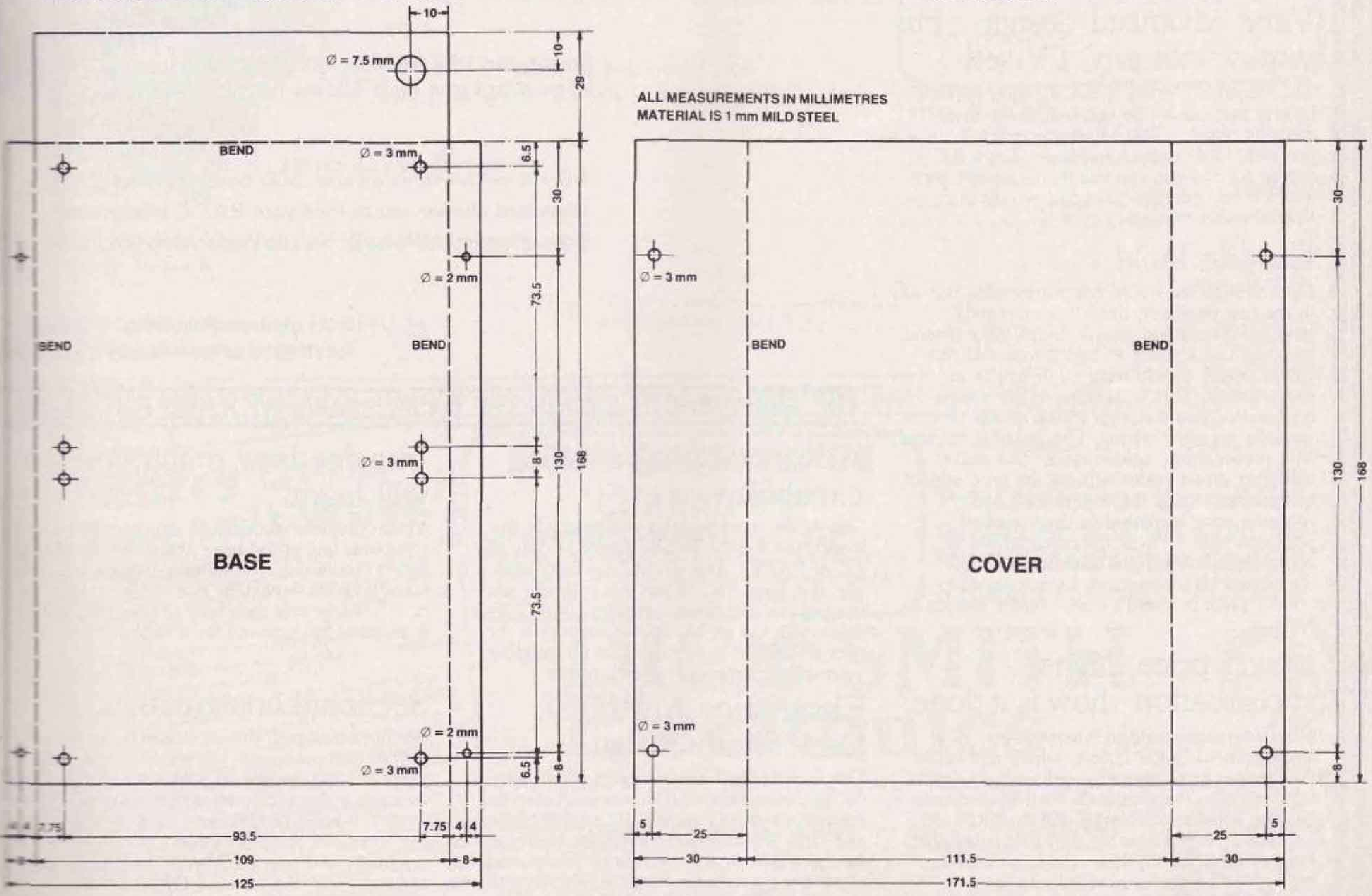
Both boards should be mounted in a *steel* box which can be mounted as a unit inside the main preamp chassis. This greatly improves the rejection to 50 Hz magnetic fields generated by nearby power transformers or 240 V cables. Details of the metalwork for this box are included with this article. Note that one end is left open to facilitate easy entry and exit of the shielded cable and power supply wiring. The MC stage should be mounted at the closed end of the box with its input end closest to the steel end panel. The input to the MC amp should be made through a small hole in this end of the box (see photo, p.24). The MM board is mounted closer to the open end, again with its input end pointing towards the solid end panel of the box (i.e: adjacent to the MC output).

Powering up

No setting-up procedure is required for either stage, but make a final check of all components before applying power to the unit. After a few minutes' operation the LM394s should be reasonably warm. These devices dissipate around 160 mW, so some heating should be expected. Similarly the NE5534s run slightly above ambient temperature. ●

(Next month we continue with the high level stages, completing the construction and data for the LM394 and NE5534s.)

Metalwork details for the low-level stages' shielding box. It may be constructed from tinplate or 1 mm galvanised mild steel plate. It *must be steel* to provide shielding from magnetically induced hum fields, such as from power transformers, turntable motors, etc.



Series 5000 stereo control preamplifier

Part 3

This is the final article describing the design and construction of the Series 5000 preamp. Last month we described the low-level amplifiers — the moving magnet and moving coil input stages. In this article we concentrate on the high-level switching, line and monitor amplifiers, muting and power supply, and complete the construction details.

David Tilbrook

A COMPLETE circuit diagram of the preamp is included in this article, with the sections described in previous issues shown simply as blocks (LED level meters were described in 'ETI-458 peak/average audio LED level meter', published in June 1981; overall block diagram and features were described in part 1 of the Series 5000 preamp, published in the July issue; moving coil and moving magnet input stages were described in the September issue).

As can be seen from the circuit diagram the preamp has three low-level inputs. The moving coil input is connected directly to the input of the MC head amp. The capacitors C17 and C18 are soldered between shield and active on each of the input sockets. The output of this amplifier is fed to the low-level selector switch on the front panel, together with shielded cables from the two moving magnet inputs. Once again resistors R13 to R16 and capacitors C19 to C22 are soldered on the input sockets. The output of the low-level selector switch is fed to the input of the MM input stage, which incorporates RIAA equalisation as described last month. The input of this stage has an input impedance around 470k, defined by resistor R2 in the MM circuit diagram (published last month). Since the input differential pair in the NE5534N requires approximately 200 nV into its bases, a voltage drop around 100 mV will appear across this resistor. Capacitor C2 (MM circuit diagram) is used to isolate this dc voltage from the cartridge. If the source resistance is changed rapidly, however, by unplugging the cartridge or otherwise open circuiting the source resistance, a rapid dc shift will occur, producing a loud thump in the loudspeakers. To overcome this problem the low-level input selector should be a make-before-break type and all unused low-level inputs should be shorted. The best way to do this is to construct shorting plugs by soldering the active and earth terminals together on an RCA plug. For convenience we have specified all switches in the preamp as three-pole four-position, make-before-break, rotary switches, manufactured by C&K. This was the switch used in the Series 4000 amplifier so availability should be no problem, although the most common type seems to be that with solder lugs rather than pc mount pins. If you are supplied with the solder lug type, the ends of the pins can be cut off with a pair ▶

SERIES 5000 PREAMPLIFIER — SPECIFICATIONS

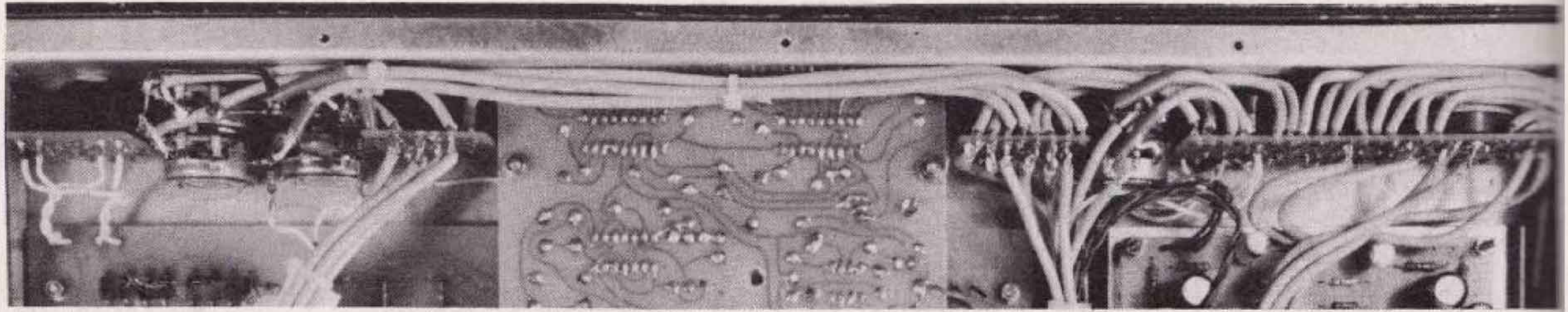
Frequency response:	High-level input: 15 Hz-130 kHz, +0, −1 db	
	Low-level input — conforms to RIAA equalisation, ±0.2 dB (see text).	
Distortion:	1 kHz ‹0.003% on all inputs (limit of resolution on measuring equipment due to noise limitation).	
S/N noise:	High-level input, master full, with respect to 300 mV input signal at full output (1.2 V):	›92 dB flat ›100 dB A-weighted
	MM input, master full, with respect to full output (1.2 V) at 5 mV input, 500 ohm source resistance connected:	›86 dB flat ›92 dB A-weighted
	MC input, master full, with respect to full output (1.2 V) and 200 μV input signal:	›71 dB flat ›75 dB A-weighted

UNUSED LOW LEVEL INPUTS SHOULD BE PROVIDED WITH SHORTING PLUGS

ALL SWITCHES ARE C&K 4-POSITION, 3-POLE, MAKE BEFORE BREAK, ROTARY, ADJUSTED (SEE PARTS LIST).

N.B. INPUT RESISTORS IN LED LEVEL METER SHOULD BE CHANGED TO 470k — SEE TEXT.

R90, R91 REPLACE THE 22k RESISTORS ON THE INPUTS OF THE LED LEVEL BOARDS



View of the rear of the sub-panel assembly, which holds all the switches and potentiometers. Wiring to the switches is from small pc boards mounted on the rear of each switch. This greatly simplifies the interconnecting wiring, which must be via shielded cable (I used 3 mm dia. cable). This also ensures that the correct signal earth is preserved throughout the wiring, avoiding hum and noise problems.

of side cutters. There is just enough pin left to fit through the switch pc boards, so cut as closely as possible to the solder eye.

All switches are soldered to pc boards to bring the necessary contacts to the top of the chassis to facilitate ease of wiring. The wiring in the preamp is reasonably complicated, although not difficult thanks to the switch pc boards. I tried it originally by soldering directly to the back of the switches, but the resulting maze of shielded cable would have made it extremely difficult to fault-find and placed excessive strain on the centre lead solder connections. The circuit boards overcome this problem and provide a secure anchor for both the centre lead and the shield on the shielded cable used for most of the wiring inside the preamp. Furthermore, these circuit boards connect the necessary shields together to maintain the integrity of the signal earth, but more about this later.

The output of the MM amplifier is fed to the 'low' position on the high-level input selector (i.e: selecting 'low' selects the low-level input selector), together with tuner, aux 1 and aux 2 inputs. The output of the switch is fed to the tape input selector on the switch pc board and appears at the switch position marked 'high' on the front panel. The third set of contacts on this switch are used to drive the tape 1 and tape 2 muting transistors Q5 to Q9. If tape 1 for example is selected as an input, pin 10 on the tape input selector is taken high, driving the bases of transistors Q5 and Q8 via diodes D12 and D15 and resistors R31 and R36. R43 acts as a pull-down resistor to ensure that the transistor base-emitter junctions cannot be forward biased by large signal excursions. The diodes prevent this reverse voltage from driving the base-emitter junctions into reverse zener action. The operation of the muting transistors is a little unusual since the transistors are used 'upside down'. It is not commonly known that bipolar transistors can be operated by forward biasing the base-collector junction and using this as the control junction of the transistor. This forms a low gain transistor that has the advantage of a lower on resistance, which is ideal for this situation.

The mute transistors for the line and monitor outputs are driven by the muting control circuitry that senses the presence of the 30 volt ac supply voltage from the power amp. When the amp is turned on, the circuit mutes the line and monitor outputs, turns on the main supply rails and then releases the muting. This eliminates the problem of turn-on thump, although a slight click will be heard as the muting transistors are switched. Similarly at turn-off the muting circuit mutes outputs until the main supply voltage has dropped sufficiently.

The output of the high-level selector is fed via the master level control to the line amplifiers. From the line amplifiers the signal is fed through the tape monitor switch to the balance and monitor level potentiometers, through the mode switch to the monitor amplifiers. When the mode switch is switched to the L-R position the left channel monitor volume wiper is connected to the output of the unity gain phase inverter. The output impedance of the inverter has been set to correspond to that of the left monitor pot when it is at full volume, so turn the monitor fully up when using this facility and use the master as the volume control.

The 400 Hz oscillator is based around the 741 op-amp IC6 and its associated circuitry. The design is a simple Wien bridge oscillator with amplitude stability achieved through the use of back to back diodes, D18 and D19. This results in an output waveform that is not really a sine wave although it is reasonably close and entirely adequate in this application.

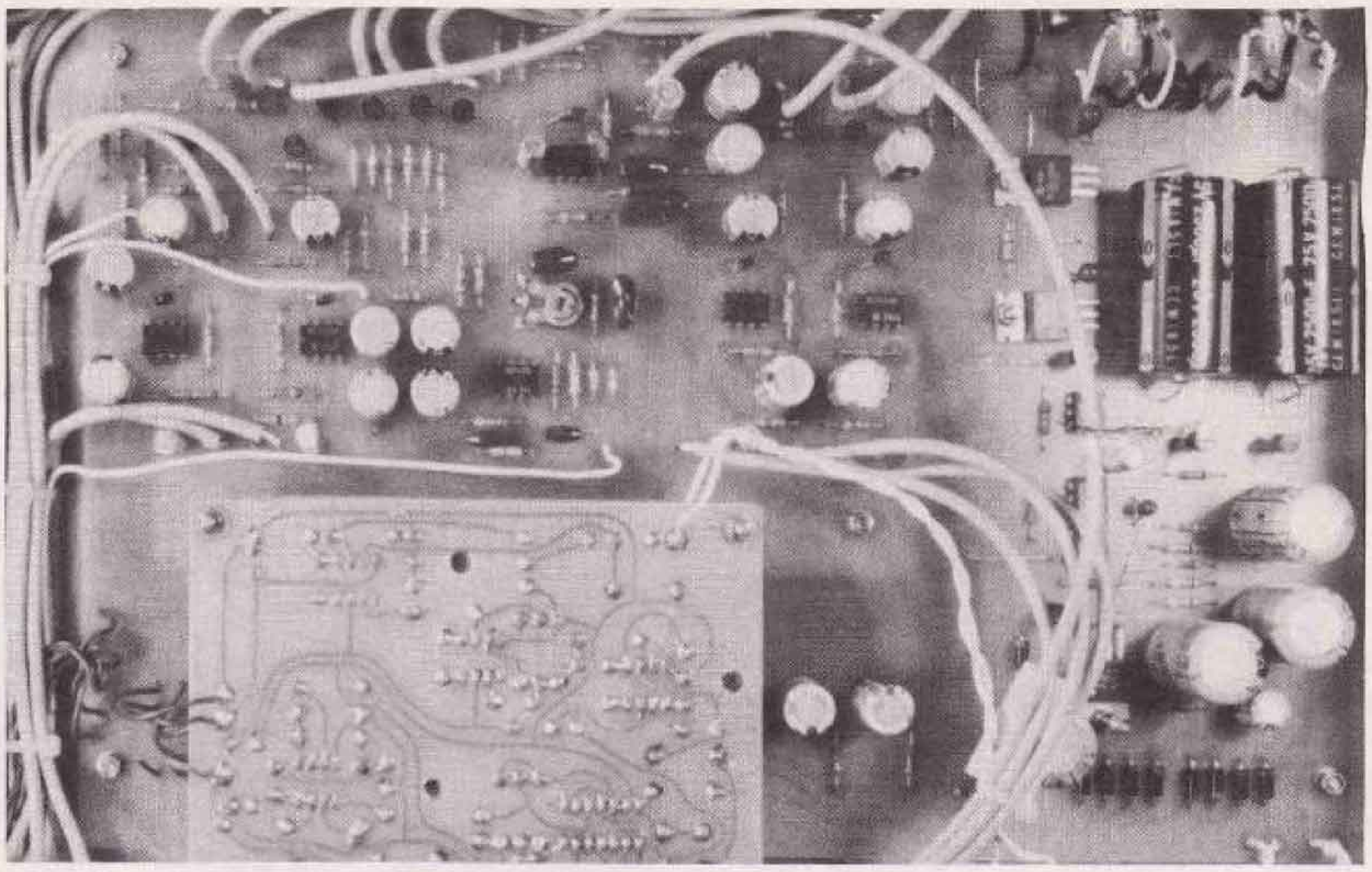
Construction

Commence the construction by assembling the LED level meters and the MM and MC input stages. Full construction details for these boards have been given in earlier articles (see above). Ensure that sufficient shielded cable has been soldered to the low-level amplifiers before they are mounted in the separate low-level amp sub-chassis. The bolts used to mount these pc boards are also used to mount the sub-chassis itself, so leave the mounting of the sub-chassis until later.

In this project the LED level meters are mounted with their track sides closest to the top of the preamp. In this way the LEDs run from left to right. This has the disadvantage, however, that calibration of the level meters must be done before mounting. Alternatively drill holes through the level meter pc boards, large enough to take a small screwdriver, immediately behind the three preset pots on each board. In this way adjustment of the level meters can be done after mounting, which is considerably easier. A second modification which must be done to the LED level meters is to increase their input impedance. This is done by removing the 22k parallel input resistor, R1 on the level meter circuit diagram, and replacing it with a 470k. These additional resistors are included on the main pc board parts list.

Next assemble the main pc board; a component overlay has been included to simplify this stage of construction. First make a visual inspection of the circuit board, checking for open circuits or short circuits between adjacent tracks. This is a reasonably complex board and any faults are best found at this stage. Check also that the holes are drilled to convenient sizes. I prefer to enlarge holes intended to take the shields from the shielded cables, and the three holes for the preset RV4 must also be large anough to accommodate the fairly wide pins. There are five mounting holes for the circuit board itself; these should be 6BA clearance (approx. 3 mm). Similarly, the mounting of the IC regulators is done with 6BA nuts and bolts. The LED level meters mount on their own pillars, two of which pass through the main pc board. These holes (see overlay) should be large enough to allow a 6BA bolt without interference from the main pc board.

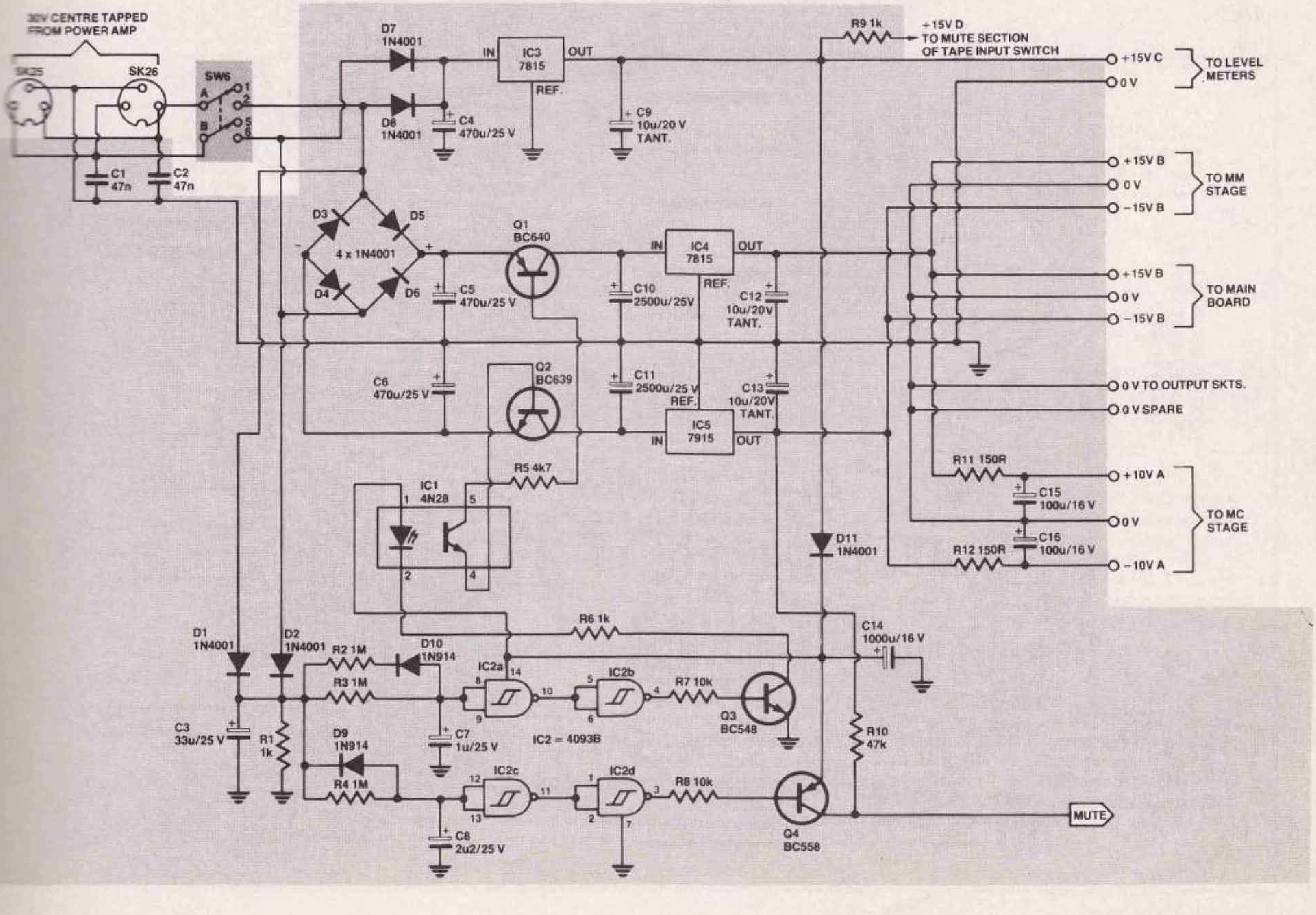
If all is correct mount the wire links, resistors and nonpolarised capacitors such as greencaps and ceramics. Next mount the transistors and diodes, ensuring that they are inserted the correct way around. Note that in the row of diodes near the power switch, diodes D3 and D4 are mounted in the reverse direction to the other diodes. Mount the integrated circuits, again making sure the orientation is correct. The voltage regulator ICs are best mounted by bending the leads with a pair of side cutters first, then inserting the pins through the pc board and securing the regulators with 6BA nuts and bolts. Pass the bolts through the pc board from the underside (i.e. nut on top). Finally solder the pins. The regulator IC3 runs the warmest of these regulators since it supplies the positive rail to the LED level meters.

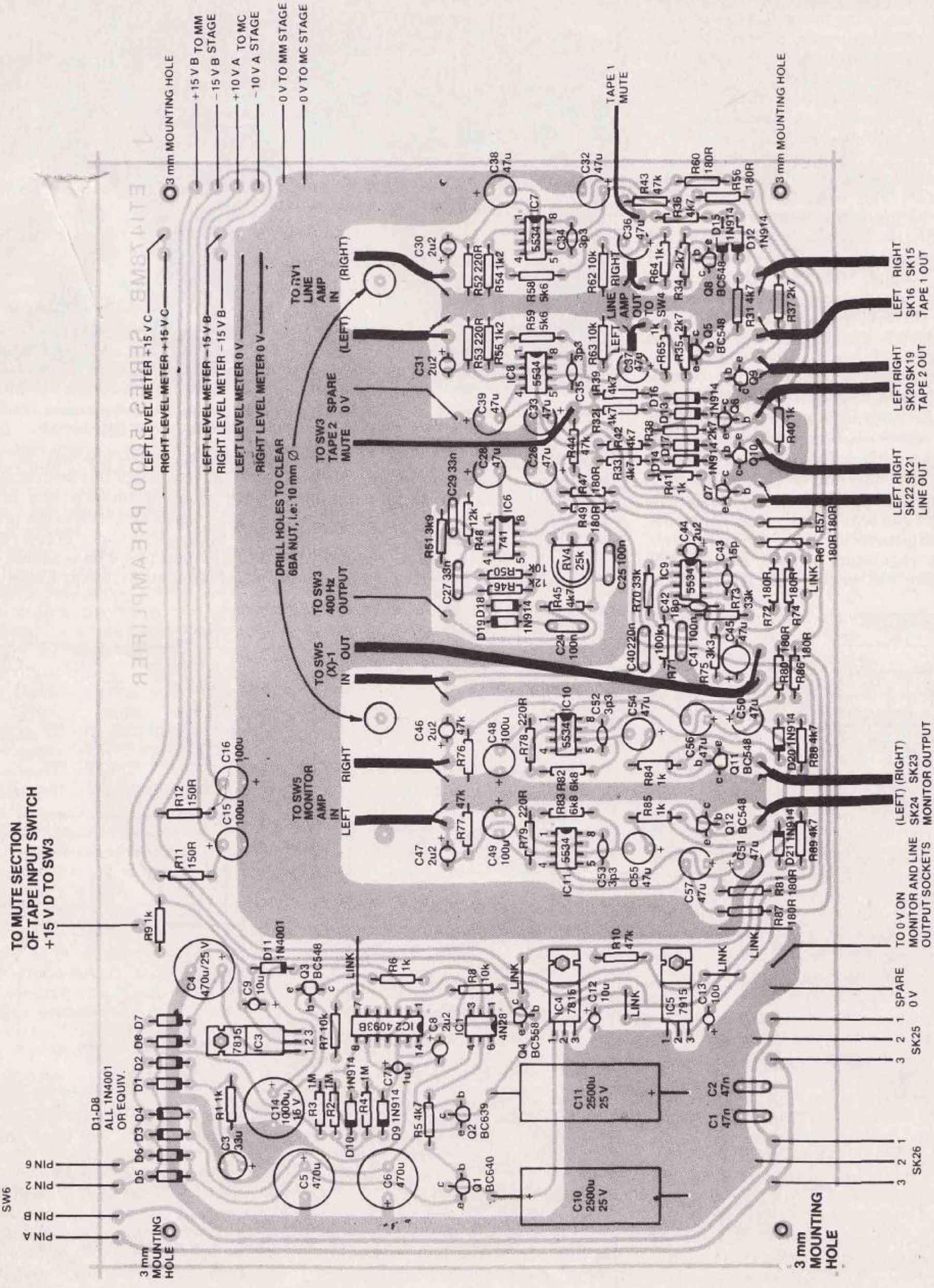


The completed motherboard (ETI-478 MB) assembled in the chassis. This board contains all the power supply circuitry, muting, line amplifiers and the unity gain inverting amplifier ('(x)−1 amp').

Mount the preset RV4. The last components to be mounted on the main pc board are the electrolytic and tantalum capacitors. Once again be careful of the orientation of these components.

The final stage in the construction of the main board is to solder the connecting cables. These are left as 'flying leads' at this stage but with sufficient length to allow them to run to their respective positions within the preamp. If the main board is positioned roughly on the bottom panel an estimate of the necessary lengths is easily made. ▶



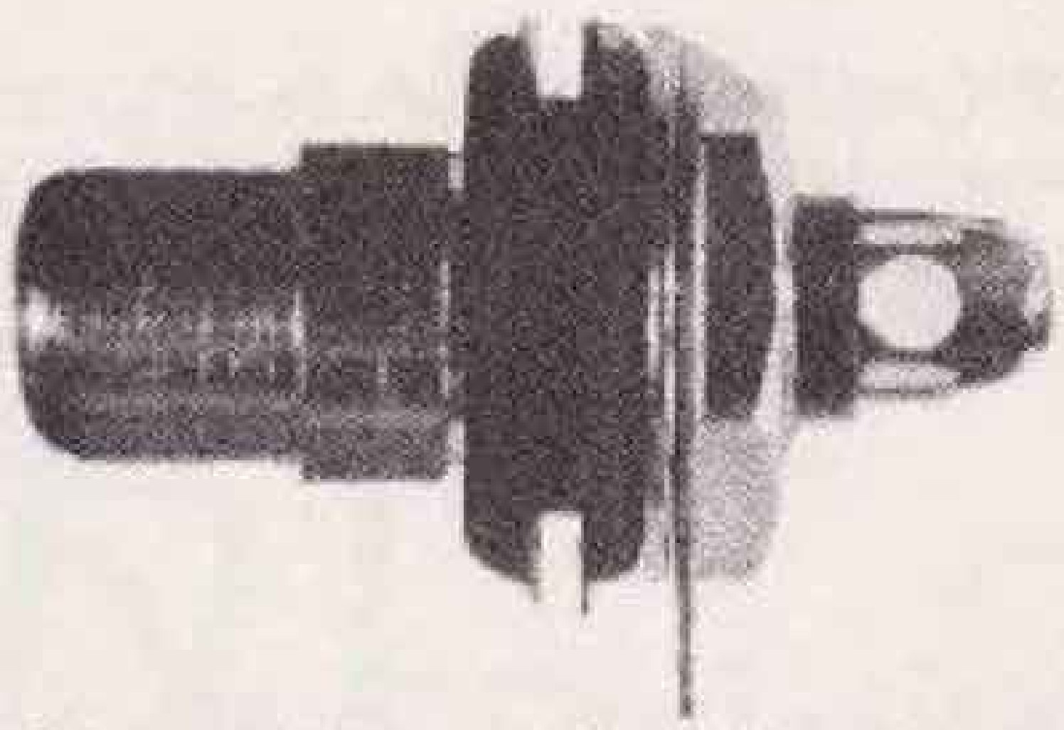


COMPONENT OVERLAY FOR THE MOTHERBOARD ETI-478 MB

The connections marked +VA, −VA, +VB, −VB and the two 0 V connections at the extreme right edge of the pc board supply power to the MC and MM input stages. These leads are already soldered to the MC and MM pc boards, so leave them empty at this stage. The power supply leads to the LED level meters, however, should be soldered to the main board and left flying for the time being. Notice that all signal-carrying leads are shielded cable, and provision has been made on the pc board to accommodate the shields. The connections to the mute lines and the 400 Hz oscillator are done with conventional hookup wire (see overlay and relevant photographs).

Next construct the rear panel assembly. Start by disassembling the chassis so that you can work on the panel without interference from the bottom panel or side bars. All inputs and outputs are done with RCA-type sockets, with the exception of the two three-pin DINs. The RCA sockets must be insulated from the chassis. This is done by first fitting rubber grommets to the drilled holes in the rear panel and then mounting the sockets through the grommets. Ensure that the earth lug points toward the top of the rear panel. This was the technique used for the input sockets to the Series 5000 power amp and forms an effective and inexpensive insulated socket.

Once all the RCA sockets have been mounted, fit the two three-pin DIN sockets. All the leads to the rear panel come from either the main board or the front panel assembly so no leads need to be soldered to the rear panel at this stage. Instead solder all the resistors and capacitors as shown in the rear panel assembly drawing. Note that all the sockets with the exception of the four tape outputs have parallel resistors and/or capacitors. The overlay drawing included shows the position of these components.



The RCA sockets mount through the hole of rubber grommets fixed to the rear panel, electrically isolating them from the panel.

The next stage is the assembly of the front panel. Once again start by disassembling this part of the chassis; the wiring is fairly complicated and is much easier to do with the sub-panel separate.

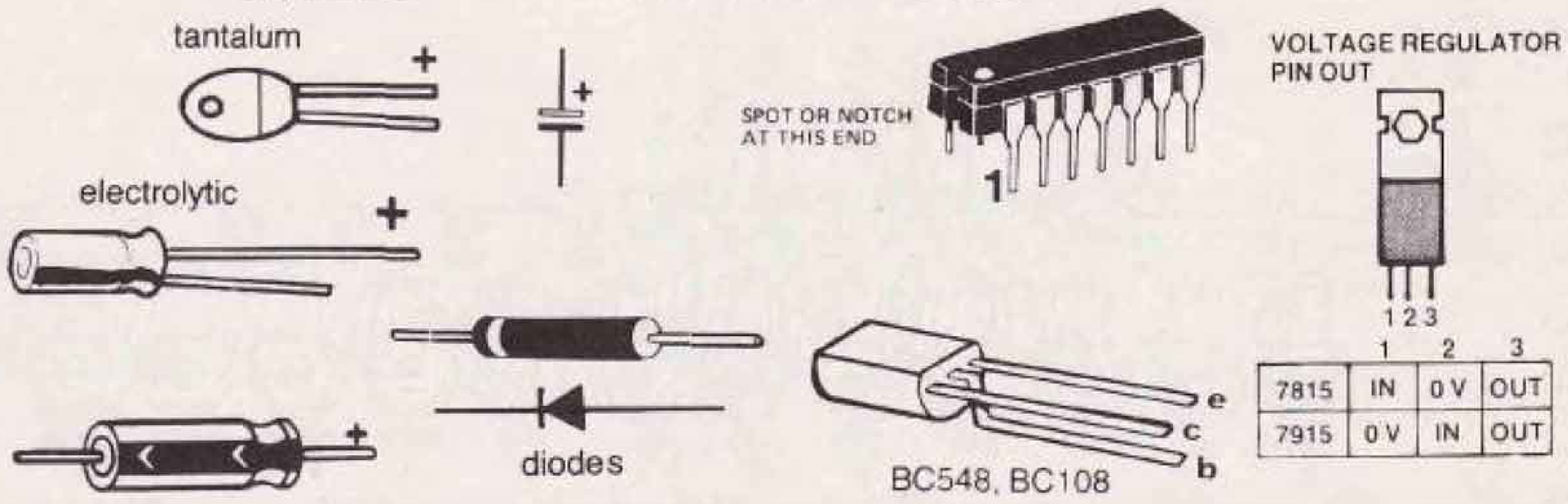
▶

PC BOARD ARTWORK AND CABINET DRAWINGS

We do not have sufficient room to reproduce the pc board artwork (boards are ETI-478 MB, SA, SB, SC, SD) and the cabinet metalwork drawings. A complete set may be obtained by sending a 300 x 250 mm stamped, self-addressed envelope to:

SERIES 5000 ARTWORK & DRAWINGS
ETI MAGAZINE
15 BOUNDARY ST
RUSHCUTTERS BAY NSW 2011

COMPONENT PINOUTS



	1	2	3
7815	IN	0 V	OUT
7915	0 V	IN	OUT

PARTS LIST ETI-478MB SERIES 5000 PREAMP MOTHERBOARD AND CASE ASSEMBLY

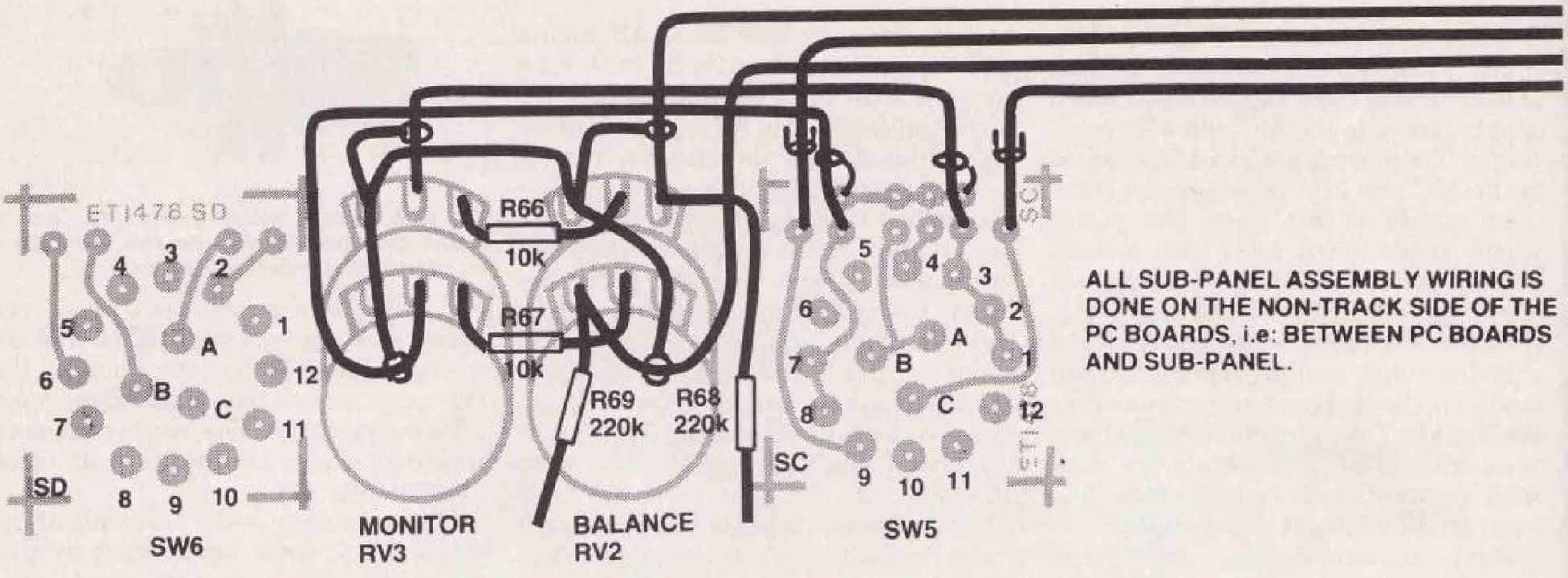
Resistors	all ½W, 5%
R1,6,9,40,41, 64,65,84,85	1k
R2,3,4	1M
R5,27 to 33,36, 39,42,45,88,89	4k7
R7,8,50,62, 63,66,67	10k
R10,13,14,43, 44,76,77	47k
R11,12	150R
R15,16,71	100k
R17 to 26,46,48	12k
R34,35,37,38	2k7
R47,49,56,57,60, 61,72,74,80, 81,86,87	180R
R51	3k9
R52,53,68,69	220k
R54,55	1k8
R58,59	5k6
R70,73	33k
R75	3k3
R78,79	220R
R82,83	6k8
R90,91	470k NOTE: R90,91 replace R1 (22k) in each ETI-458 LED level display.
RV1	100k/C dual log. pot.
RV2	10k/A dual linear pot.
RV3	10k/C dual log. pot.
RV4	25k trimpot.

Capacitors	
C1,2	47n greencap
C3	33u/25 V RB electro.
C4,5,6	470u/25 V RB electro
C7	1u/25 V RB electro.
C8,44	2u2/25 V RB electro.
C9,12,13	10u/20 V tantalum
C10,11	2500u/25 V axial electro.
C14	1000u/16 V RB electro.
C15,16,48,49	100u/16 V RB electro.
C17,18	4n7 greencap
C19,20,21,22	220p mica or styroseal
C23,24,25,41	100n greencap
C26,28,32,33,36, 37,38,39,50, 51,56,57	47u/25 V RB electro.
C27,29	33n greencap
C30,31,46,47	2u2/16 V RB electro.
C34,35,52,53	3p3 ceramic
C40	220n greencap
C42	18p ceramic
C43	15p ceramic
C45,54,55	47u/16 V RB electro. NOTE: electrolytic and tantalum capacitors have been specified with the minimum working voltage rating and the pc board has been laid out to suit. Higher voltage rating capacitors may not fit. RBLL types may be substituted where we have specified electrolytics, but where tantalums are specified no substitiution may be made.

Semiconductors	
D1 to D8,D11	1N4001, 1N4002 etc.
D9,10,12 to 21	1N914, 1N4148 etc.
IC1	4N28 opto isolator
IC2	4093B quad Schmitt NAND
IC3,IC4	7815 +15 V 3-terminal reg.
IC5	7915 −15 V 3-terminal reg.
IC6	741 op-amp.
IC7 to 11	NE5534N op-amp.
Q1	BC640
Q2	BC639
Q3,Q5 to 12	BC548
Q4	BC558

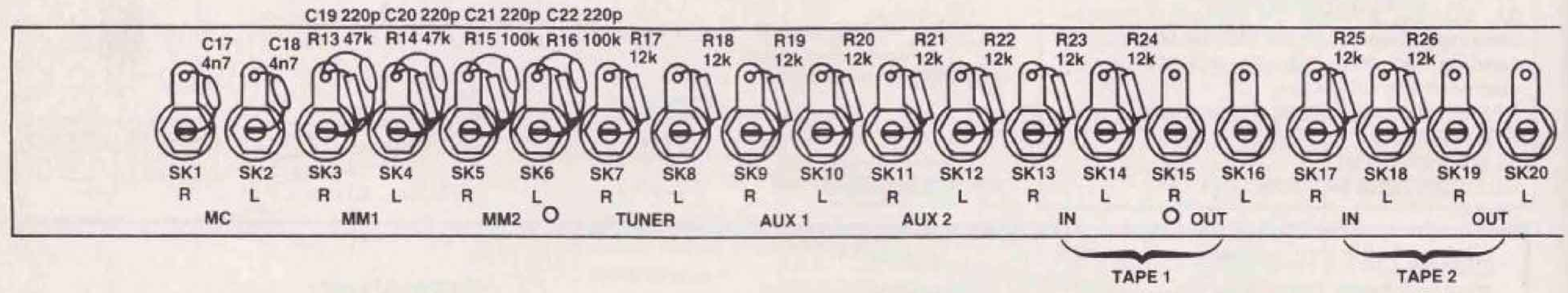
Miscellaneous	
SW1,SW4	3-pole, 3-position rotary
SW2,SW3,SW5	3-pole, 4-position rotary
SW6	3-pole, 2-position rotary All switches are C&K Lorlin 3-pole, 4-position rotary make-before-break types with stops set as

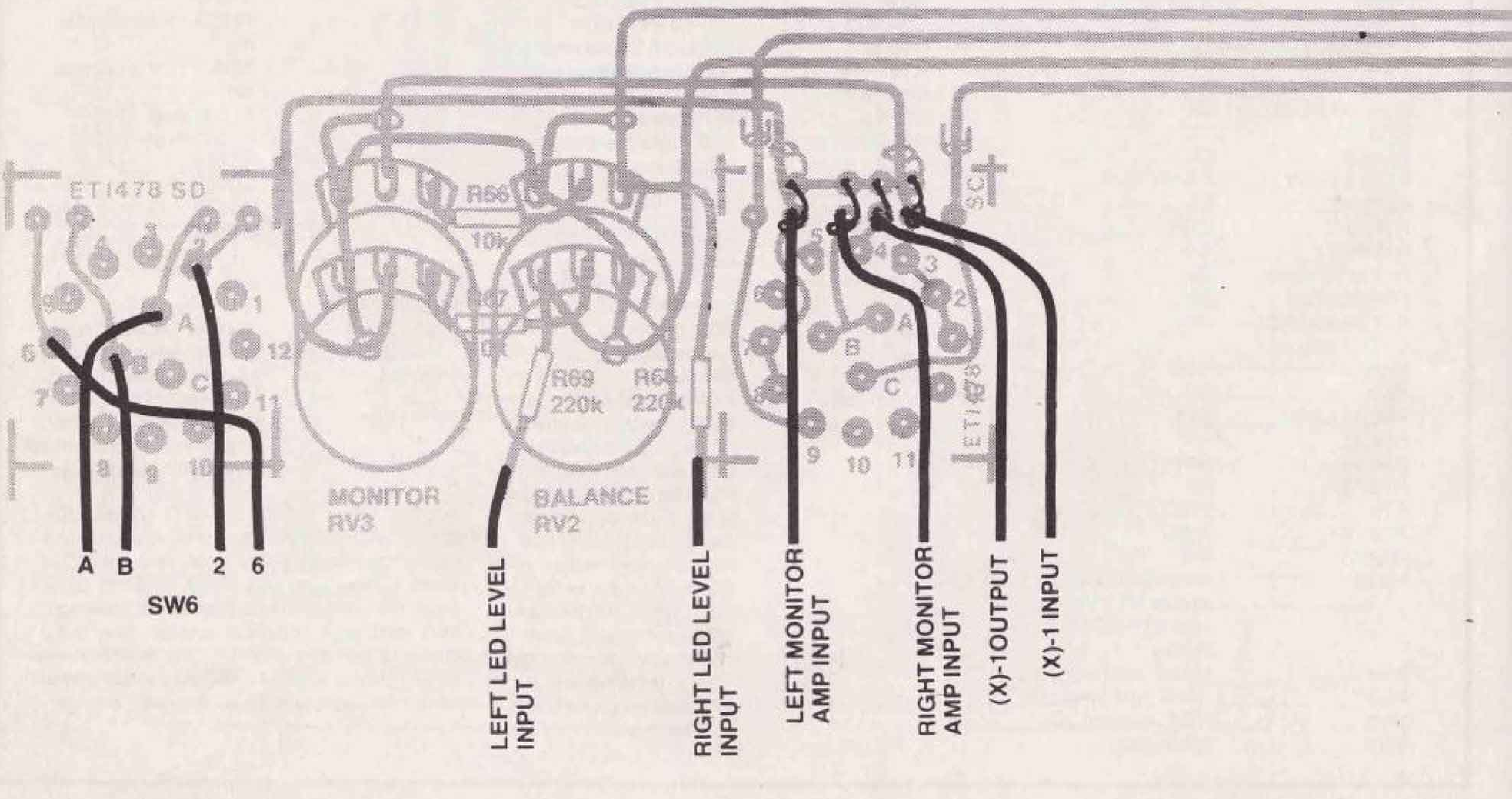
Printed circuit boards — ETI-478MB, SA, SB, SC, SD; 24 panel-mount RCA sockets; 24 rubber grommets 6 mm bore; two 3-pin DIN sockets; two 3-pin DIN plugs; shielded cable 4 mm dia.; metalwork as per cabinet drawings; front and rear Scotchcal panels; nine fancy knobs to suit; two ETI-458 LED level meters; ETI-478MM and ETI-478MC stages and metalwork; nuts, bolts, solder, hookup wire, etc.

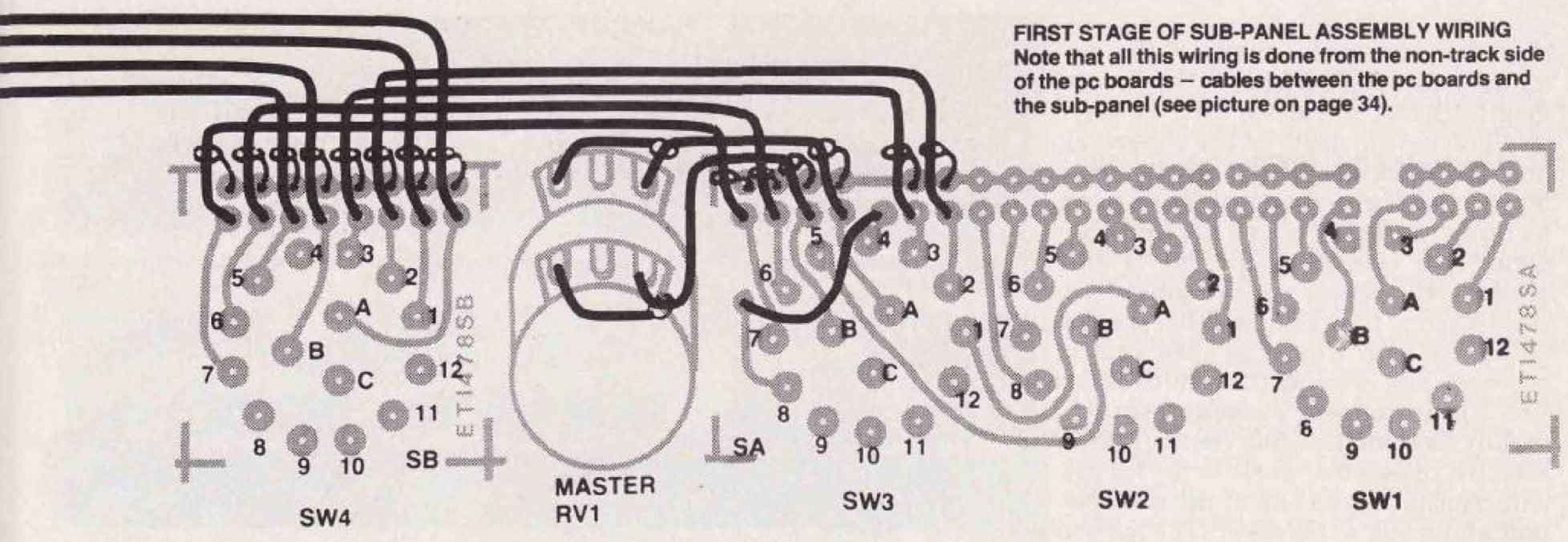


ALL SUB-PANEL ASSEMBLY WIRING IS DONE ON THE NON-TRACK SIDE OF THE PC BOARDS, i.e: BETWEEN PC BOARDS AND SUB-PANEL.

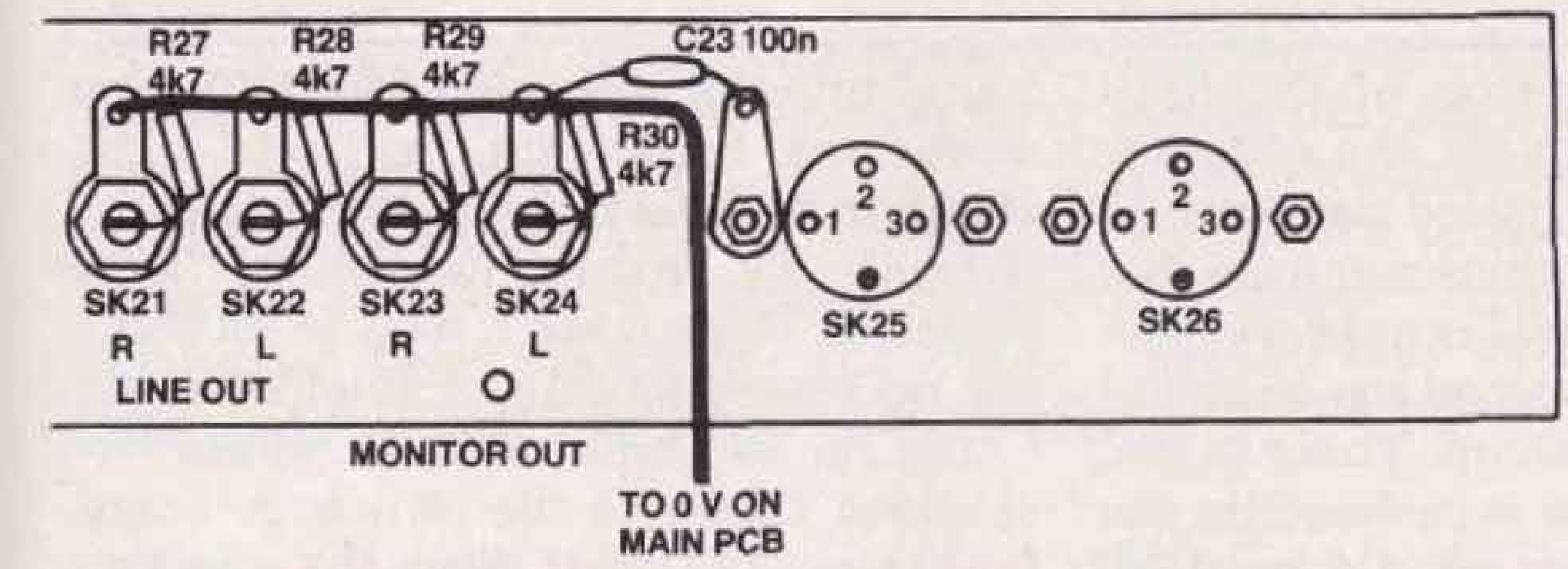
REAR PANEL ASSEMBLY

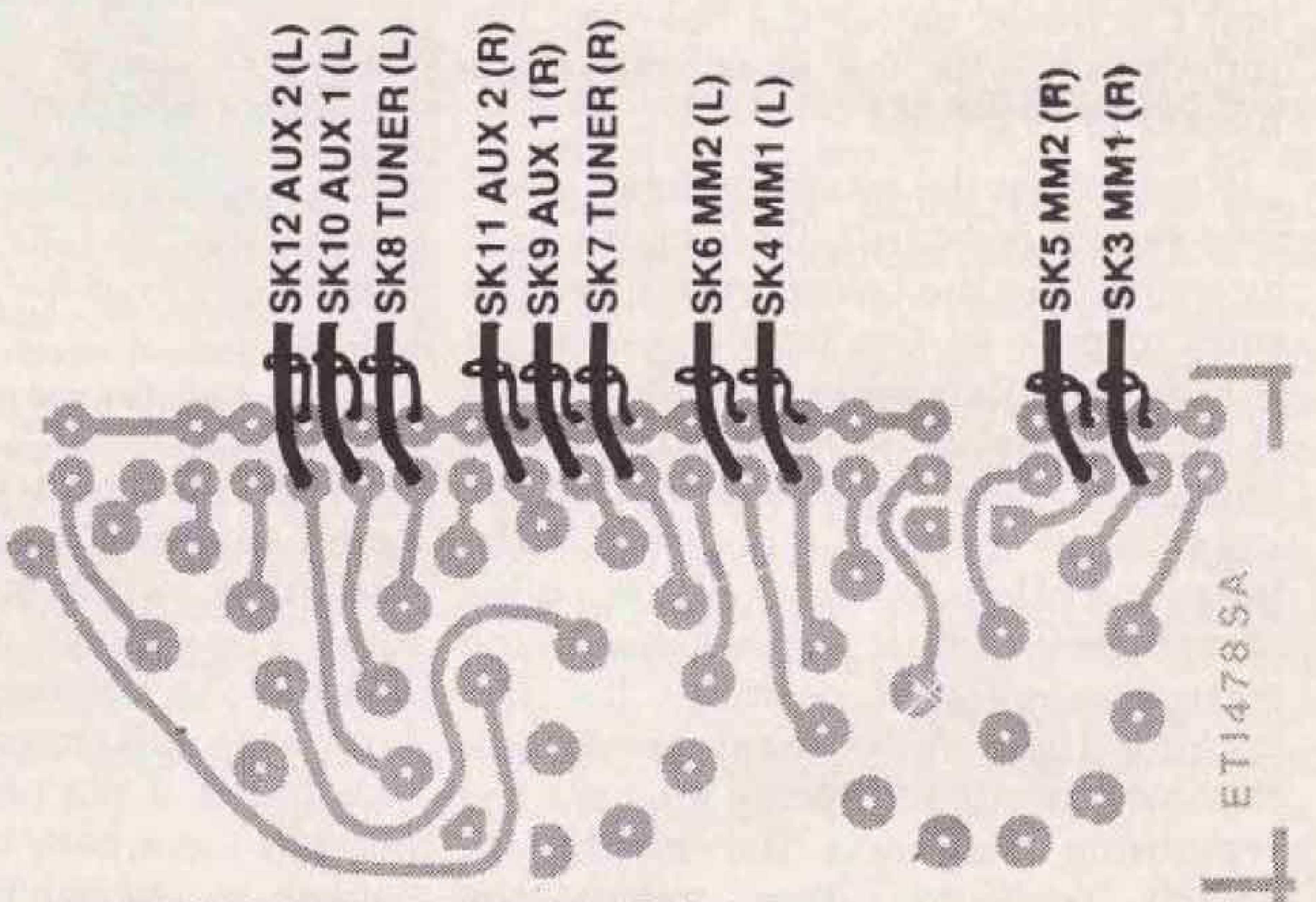




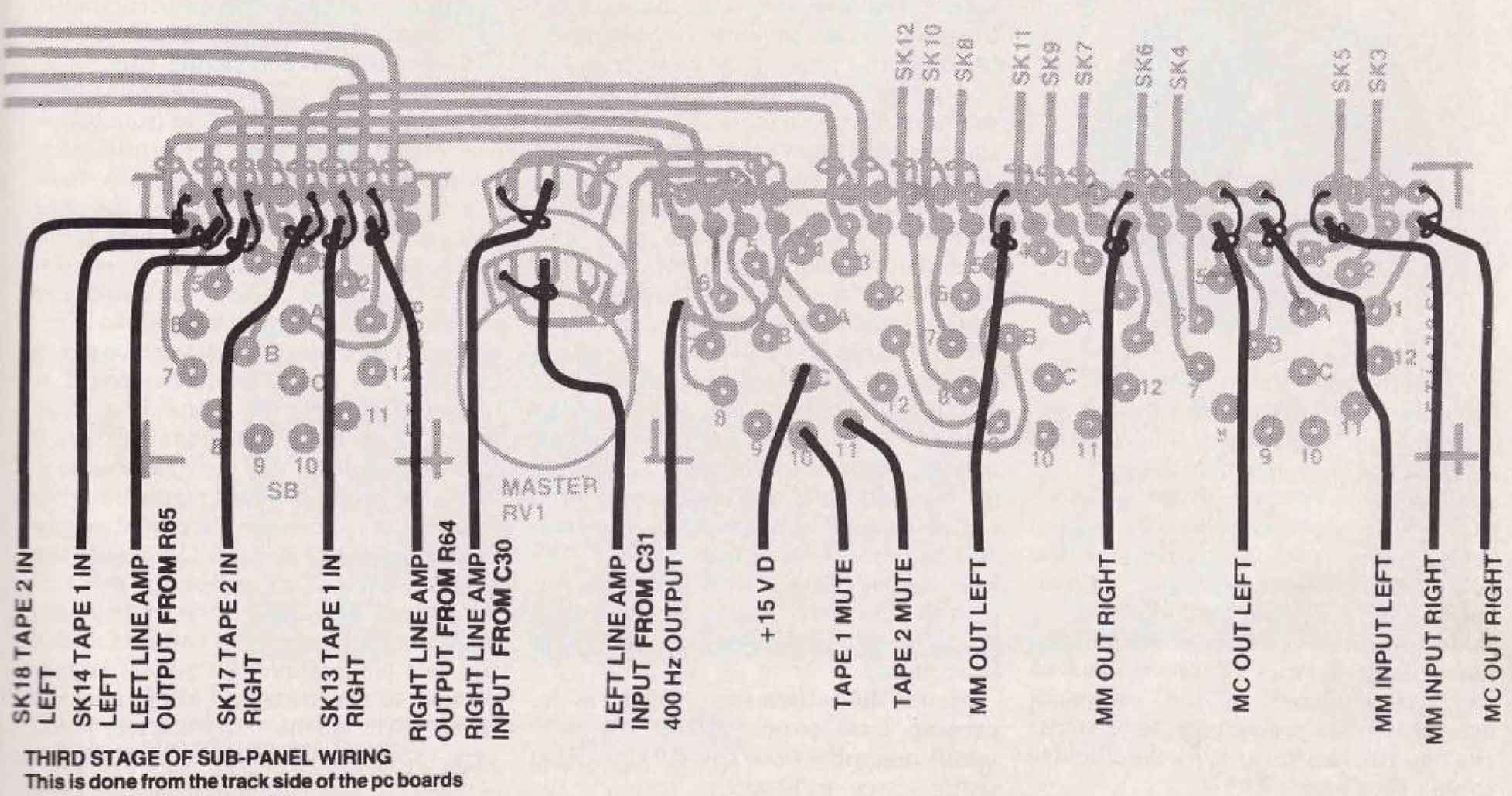


FIRST STAGE OF SUB-PANEL ASSEMBLY WIRING
Note that all this wiring is done from the non-track side of the pc boards — cables between the pc boards and the sub-panel (see picture on page 34).





SECOND STAGE OF SUB-PANEL WIRING
This too is done from the non-track side of the board (refer also to picture on page 34).

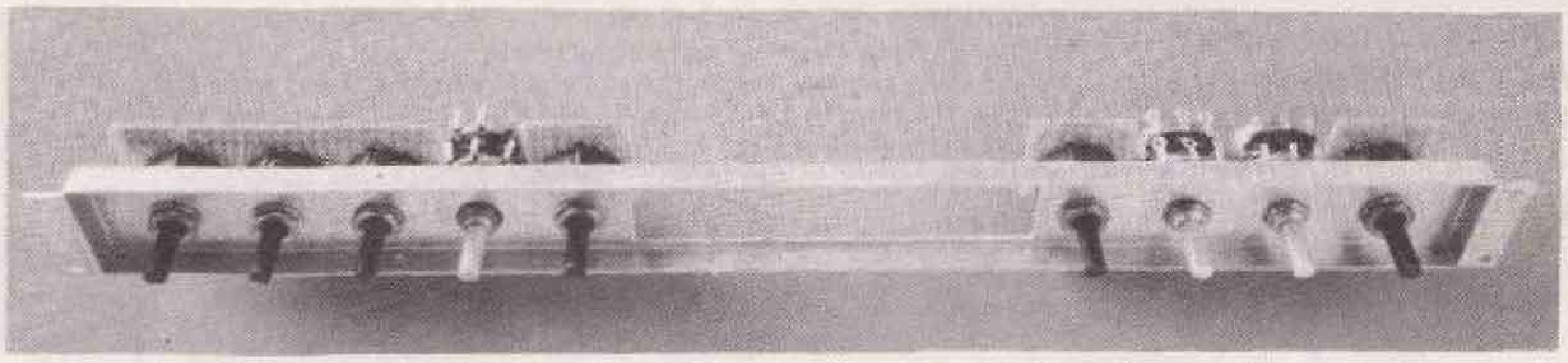


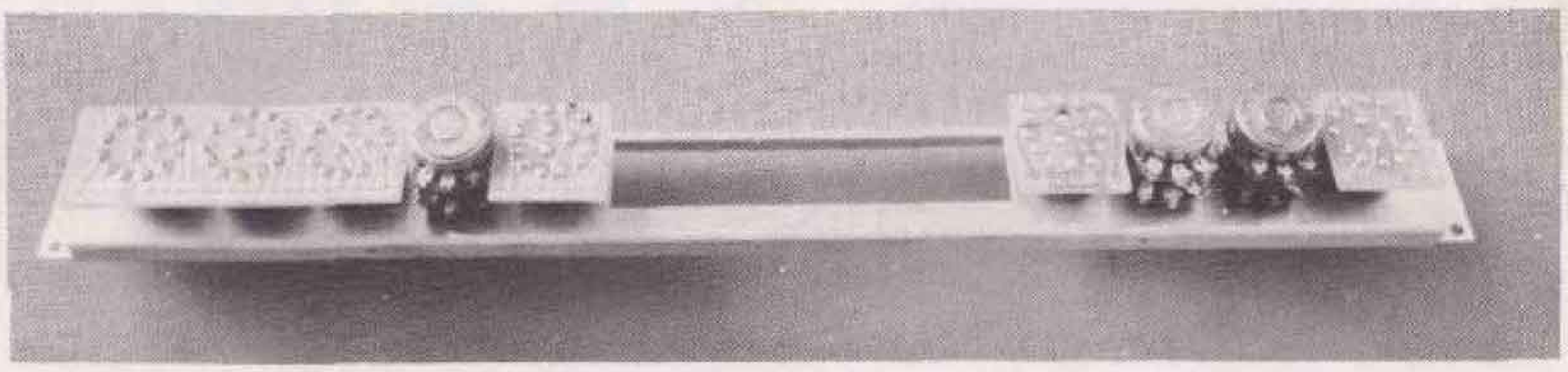
THIRD STAGE OF SUB-PANEL WIRING
This is done from the track side of the pc boards (see also picture on page 34).

Start by cutting the shafts of the potentiometers and switches to the correct lengths to suit the knobs used. Be sure to allow for the depth of the sub-panel and the thickness of the front panel. Mount the three pots with their pins pointing toward the top of the sub-panel, i.e: closest to the lid of the preamp. Be careful not to confuse the two 10k pots. If in doubt which is the log pot, place both wipers at their centre positions and check with a multimeter. The linear pot will measure approximately 5k from the wiper (centre pin) to both the outer pins, whereas the log pot will measure 1k to one of the contacts and around 9k to the other. The log pot is used for the monitor volume control and the linear pot as the balance. The remaining 100k log is of course the master level control.

Next mount the rotary switches to the four switch pc boards. Once again start by inspecting the boards carefully. The rows of pads at the top edge of these boards are intended to take the shields of the shielded cable, so enlarge them if necessary. Solder the switches, ensuring that the correct pin is closest to the top of the sub-panel assembly (see table of switches). Adjust the switches to the correct number of positions by first turning them fully counterclockwise. Remove the nut and spring washers; the remaining ring sets the number of switch positions. Now mount the switches to the sub-panel.

With all the switches mounted to the sub-panel, proceed with the interconnecting wiring. This is all done with shielded cable, as shown in detail on the sub-chassis assembly drawing. Note that all the wiring done at this stage is between the switch boards and the sub-panel. Later connections to the sub-panel assembly can then be done by soldering directly to the track sides of the switch pc boards. There are also four resistors soldered to the sub-panel assembly; it is probably wise to solder the two 10k resistors before the shielded cables. (See pages 40-41).


Switch assemblies and potentiometers, with their shafts cut to length to suit the knobs used, mounted to the sub-panel.


Rear view of the sub-panel assembly, prior to wiring.

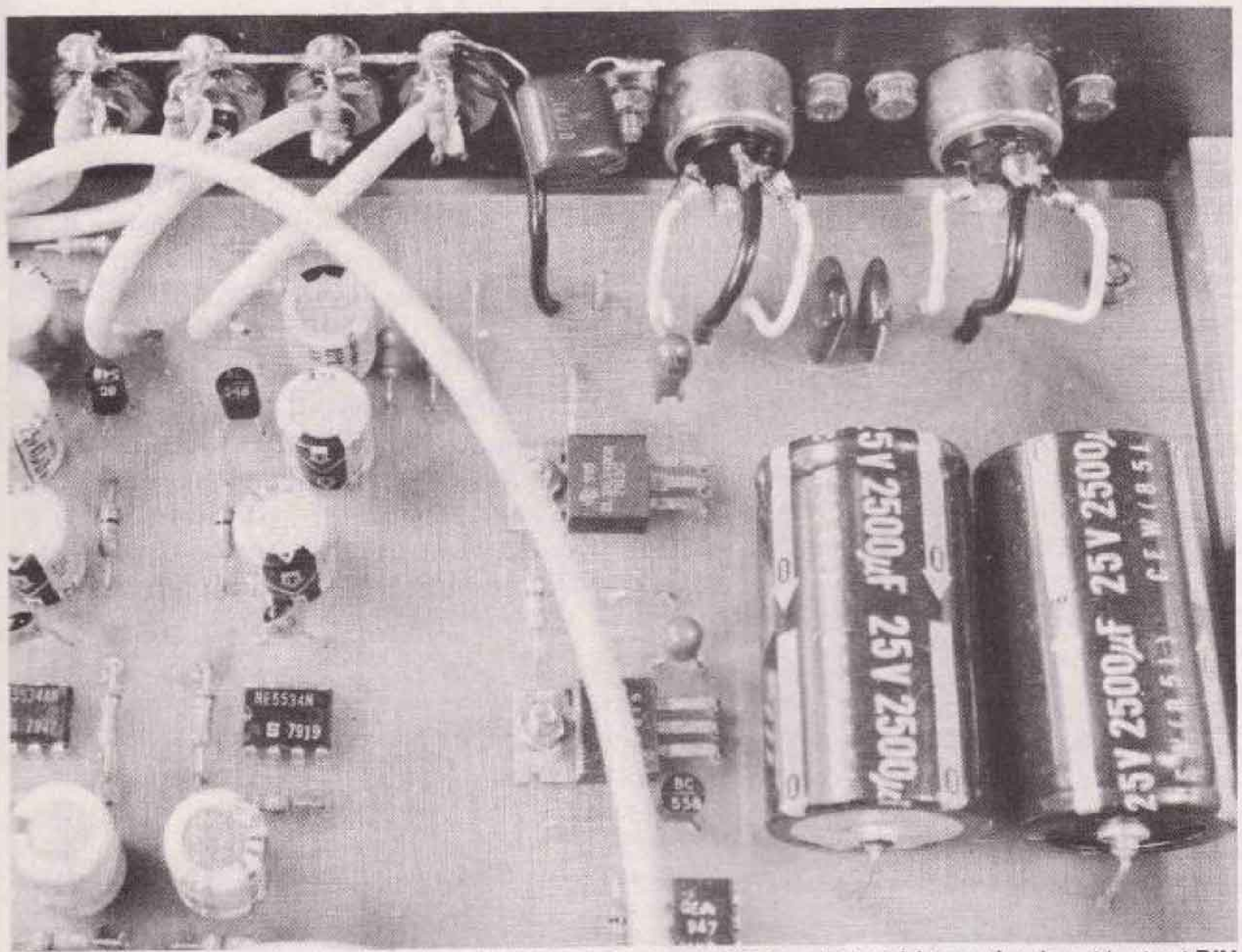
Switch	Number of Positions	PC Board	Pin Closest to Top
Low level input selector	3	SA	3,4
High level input selector	4	SA	4
Tape input selector	4	SA	4
Tape monitor	3	SB	3,4
Mode	4	SC	4
Power	2	SD	3

Table 1. Switches and positions.

With the construction of the front panel assembly complete, the wiring to the input switch pc board can now be done. The shielded cables run from this board to the rear panel between the low-level amp sub-chassis and the left hand side panel of the preamp. There is just enough room here to accommodate the cables, so the wiring should be neat, avoiding twists or crossovers between cables. The best way to do this is to first mount the low-level sub-chassis onto the preamp bottom plate. Use 6BA bolts through the preamp base from below and secure them with eight nuts and washers. The base plate of the low-level sub-chassis is now placed on these bolts with its open end closest to the front panel of the preamp, and secured with another eight nuts. These nuts also double as standoffs for the MM and MC pc boards. This does not leave a great deal of height between the track side of the pc boards and the base plate of the sub-chassis, so ensure that all wires on the track side are trimmed as closely as possible to the solder connections. If all is well, mount the low-level amplifiers to the eight bolts and secure with 6BA nuts. As stated in last month's issue, the MC pc board is mounted closest to the rear of the chassis with its input end against the rear. Mount the MM amp with its output end closest to the front of the preamp.

Mount the bottom two side bars to the preamp base panel. Mount the sub-panel assembly onto the bottom panel using three self-tappers through the panel into the sub-panel assembly, and with two bolts into the side bars. Position the rear panel at the back of the preamp in approximately its final position. This makes it easy to estimate the necessary lengths of shielded cable to cut for the input wiring. Solder the shielded cables to the switch pc board first, run the cables down the side and behind the sub-chassis, trim and solder to the input sockets.

The next stage in the construction is to mount the main pc board to the bottom panel. Before doing this, however, pass four 1.25 inch (32 mm) 6BA bolts through the base plate from below and secure with a nut to act as pillars for mounting the LED level meters. Pass five shorter 6BA bolts through the base plate and secure with nuts. Once again these nuts act as standoffs, so ensure that wire ends on the main pc boards are trimmed close to the solder joints. If required, a second set of nuts can be fitted to the bolts before the main board is mounted in order to space it a little further from the bottom panel, although this is probably not really necessary. With the main pc board placed roughly in position, solder the six power supply leads from the MM and MC stages to the main board, ensuring that the polarity is correct. Now secure the main board with nuts and washers. Cut and solder the six leads from the power supply section to the three-pin DINs and the three wires to the mute section of the tape input selector, as well as from the output of the 400 Hz oscillator (see



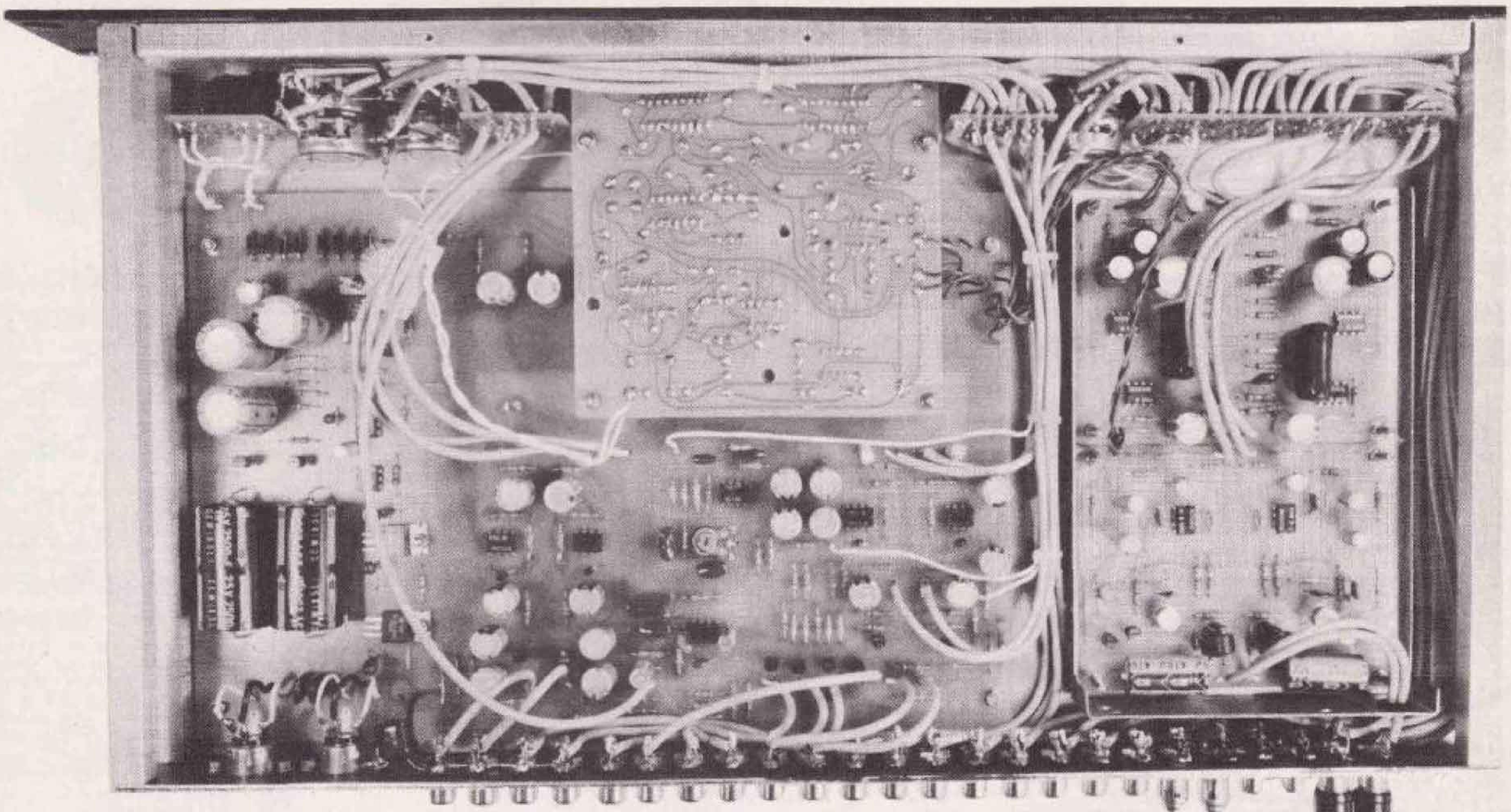
View of the rear panel and motherboard showing the output RCA sockets wiring and ac input/output DIN sockets wiring.

overlay diagram). Solder the leads to the power switch and the 0 V connection to the output socket earth.

The remaining wiring to the pc board consists of shielded cables. Before soldering these, however, run cable from the rear panel tape input sockets to the tape monitor switch. These cables are terminated directly to the track side of the switch pc board. The connections are shown in the rear panel and sub-panel assembly(p.40-41), followed by the monitor amp inputs and the unity gain inverter's input and output leads. As above, all of these cables solder to the panel assembly drawings. Then solder the line amp inputs and outputs to the track side of the switch boards. Cut and solder the output leads from the main pc board to the rear panel. Pass the input leads to the MC stage through the hole in the rear of the sub-chassis, which should be fitted with a rubber grommet, and solder to the MC input sockets. With all the wiring done to the rear panel it can be bolted to the chassis, together with the top two side bars. Solder the remaining shielded cables to the low-level amps to the track side of the input switch pc boards.

The final stage in the construction of the preamp is to mount and calibrate the LED level meters. If the added holes have been drilled to allow calibration through the pc board, the first level meter can be mounted. The height of the pc board is set by four nuts and washers, which can be adjusted to the correct height on the bolts. Alternatively, screw a further three nuts and a washer onto each mounting bolt. This is close to the correct height and ensures that strain is not placed on the pc board by different-height nuts. Secure with a further four washers and nuts. The mounting holes should be drilled well oversize so that

▶



The completed project, prior to installing the low-level stages' top shield cover and the cabinet cover. The wiring looks complicated, but it's not as bad as it looks! Note the holes drilled in the top LED level meter board so that the adjustments may be easily reached (see text).

final adjustment of the position of the LEDs can be carried out.

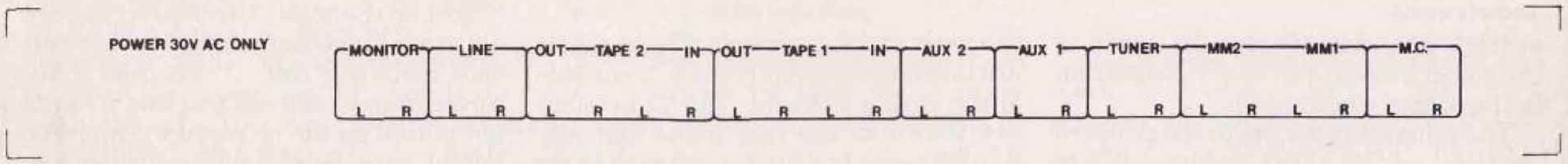
Now mount the front panel. If the LEDs are not in a perfectly straight line it is extremely difficult to get the front panel on, so it is well worth the effort of getting these as straight as possible. Solder the power supply wiring to the board and the input connection to the appropriate resistor on the sub-panel assembly.

Powering up

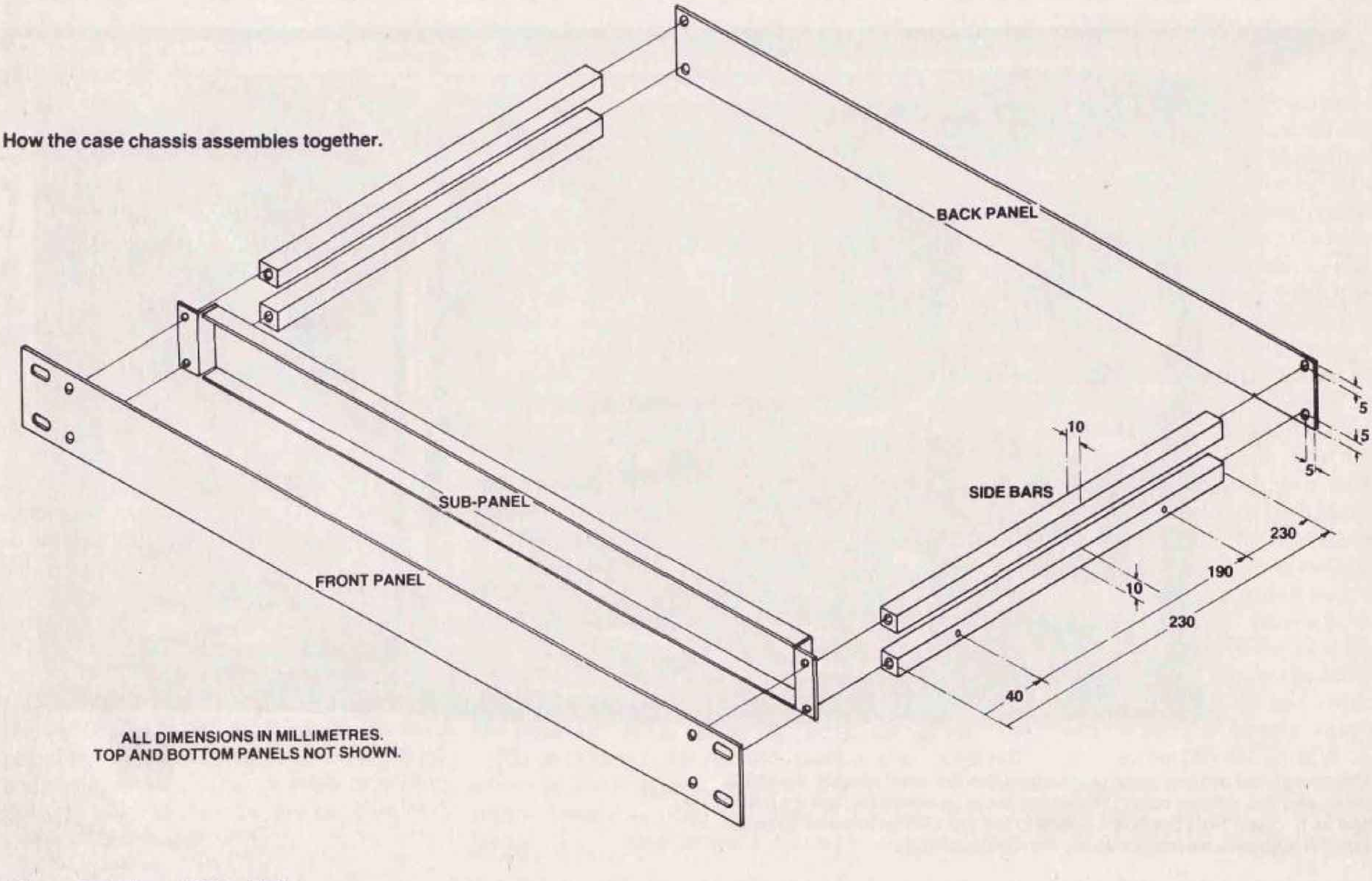
At this stage the preamp must be powered up, so check as much as is possible. In particular check the polarity of diodes, transistors and electrolytic and tantalum capacitors. Check also the power supply connections, especially those to the MM and MC preamp stages. If all is well construct a three-pin DIN lead using twisted hookup wire and apply power to the preamp from the 30 volt centre-tapped supply on the rear of the Series 5000 power amplifier. If you are not using the preamp with an ETI power amp, a separate 15-0-15 volt transformer must be used. Switch the tape input selector to the 400 Hz position, the tape monitor switch to the source position and the mode switch to stereo. Centre the balance pot and turn both master and monitor volume control fully on. Ensure that the three flying leads that will take power to the second level meter are not touching each other or anything else in the preamp. If the preamp is now turned on, the LED level meters should indicate the presence of the 400 Hz tone by moving swiftly to the right. If all is well, turn the monitor volume fully down and adjust the LED level meter dc offsets as described in the original article on the level meters. Ensure that the monitor level control is fully up, and by using a multimeter and

▶

Reproduction of the front panel artwork (above) and rear panel artwork (below).



How the case chassis assembles together.



LM394

GENERAL DESCRIPTION

The LM194 and LM394 are junction-isolated ultra-well-matched monolithic NPN transistor pairs with an order of magnitude improvement in matching over conventional transistor pairs. This was accomplished by advanced linear processing and a unique new device structure.

Electrical characteristics of these devices such as drift versus initial offset voltage, noise, and the exponential relationship of base-emitter voltage to collector current closely approach those of a theoretical transistor. Extrinsic emitter and base resistances are much lower than presently available pairs, either monolithic or discrete, giving extremely low noise and theoretical operation over a wide current range. Most parameters are guaranteed over a current range of 1 μA to 1 mA and 0 to 40 V collector-base voltage, ensuring superior performance in nearly all applications.

To guarantee long term stability of matching parameters, internal clamp diodes have been added across the emitter-base junction of each transistor. These prevent degradation due to reverse biased emitter current — the most common cause of field failures in matched devices. The parasitic isolation junction formed by the diodes also clamps the substrate region to the most negative emitter to ensure complete isolation between devices.

The LM194 and LM394 will provide a considerable improvement in performance in most applications requiring a closely matched transistor pair. In many cases, trimming can be eliminated entirely, improving reliability and decreasing costs. Additionally, the low noise and high gain make this device attractive even where matching is not critical.

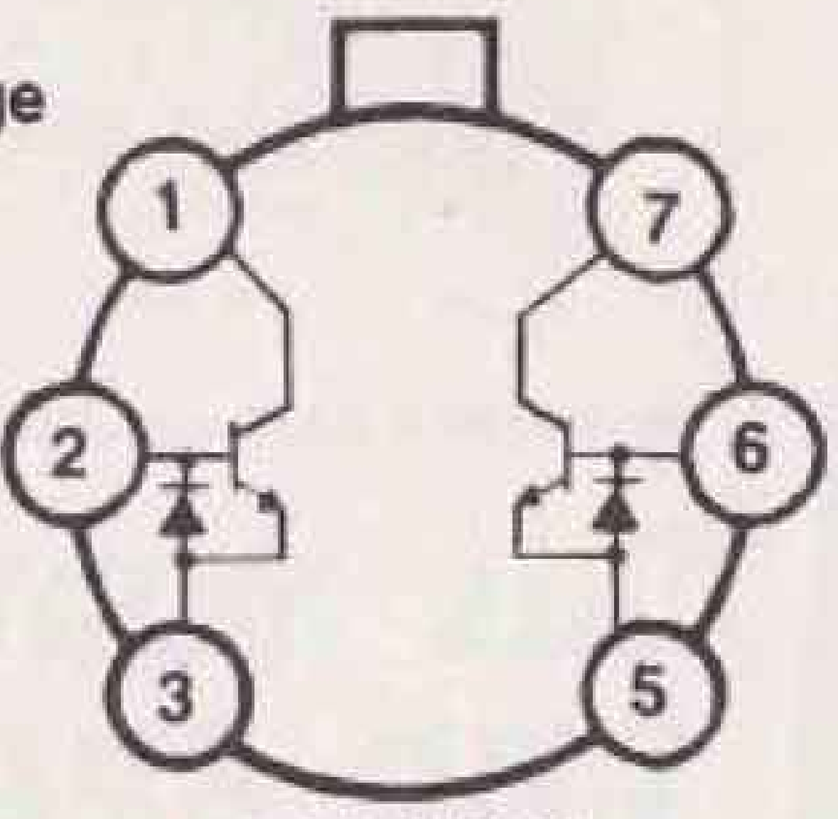
The LM194 and LM394/394B are available in an isolated header 6-lead TO-5 metal can package. The LM194 is identical to the LM394 except for tighter electrical specifications and wider temperature range.

FEATURES

- Emitter-base voltage matched to 50 μV
- Offset voltage drift less than 0.1 μV/°C
- Current gain (h_{FE}) matched to 2%
- Common-mode rejection ratio greater than 120 dB
- Parameters guaranteed over 1 μA to 1 mA collector current
- Extremely low noise
- Superior logging characteristics compared to conventional pairs

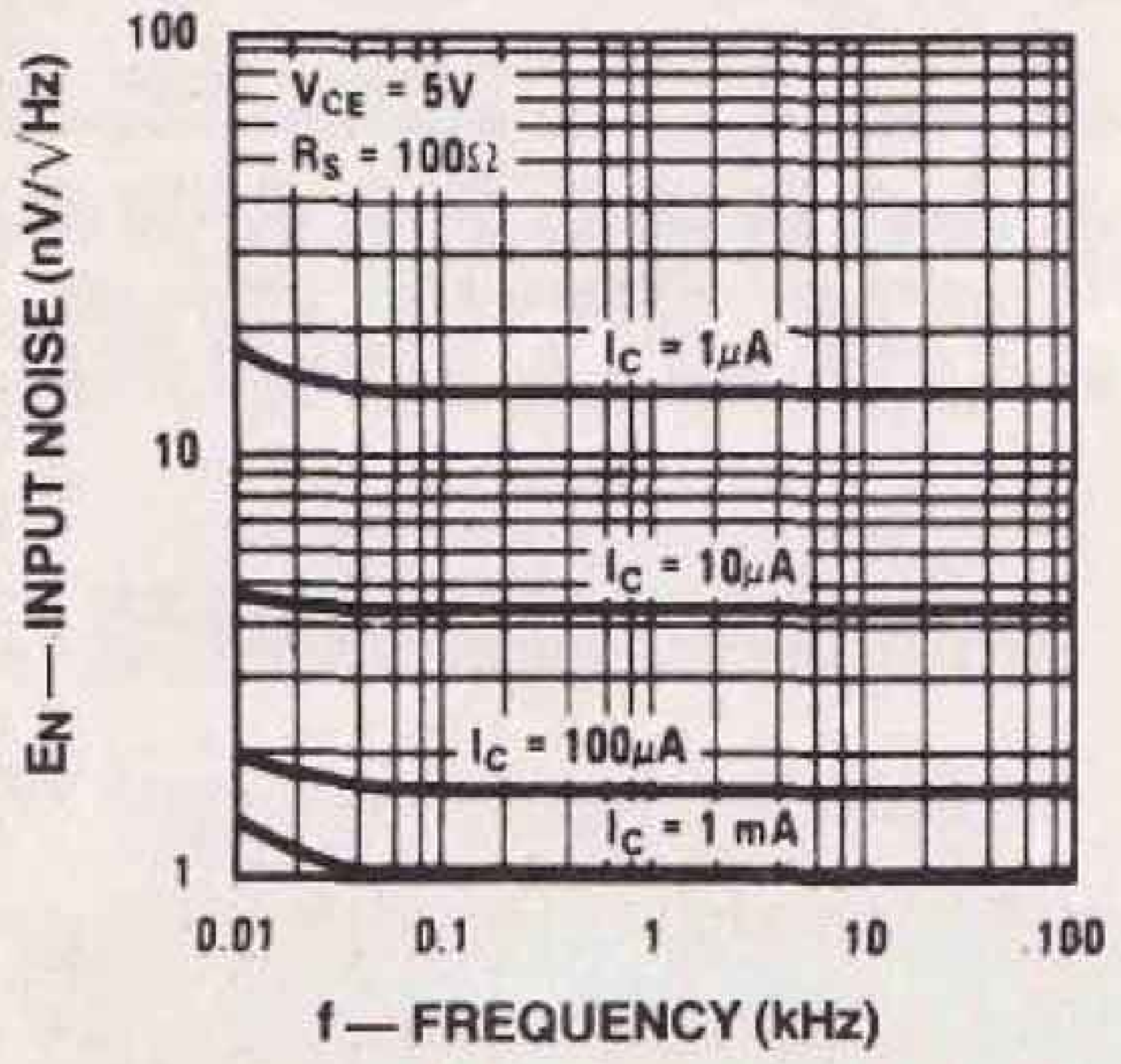
CONNECTION DIAGRAM

Metal Can Package

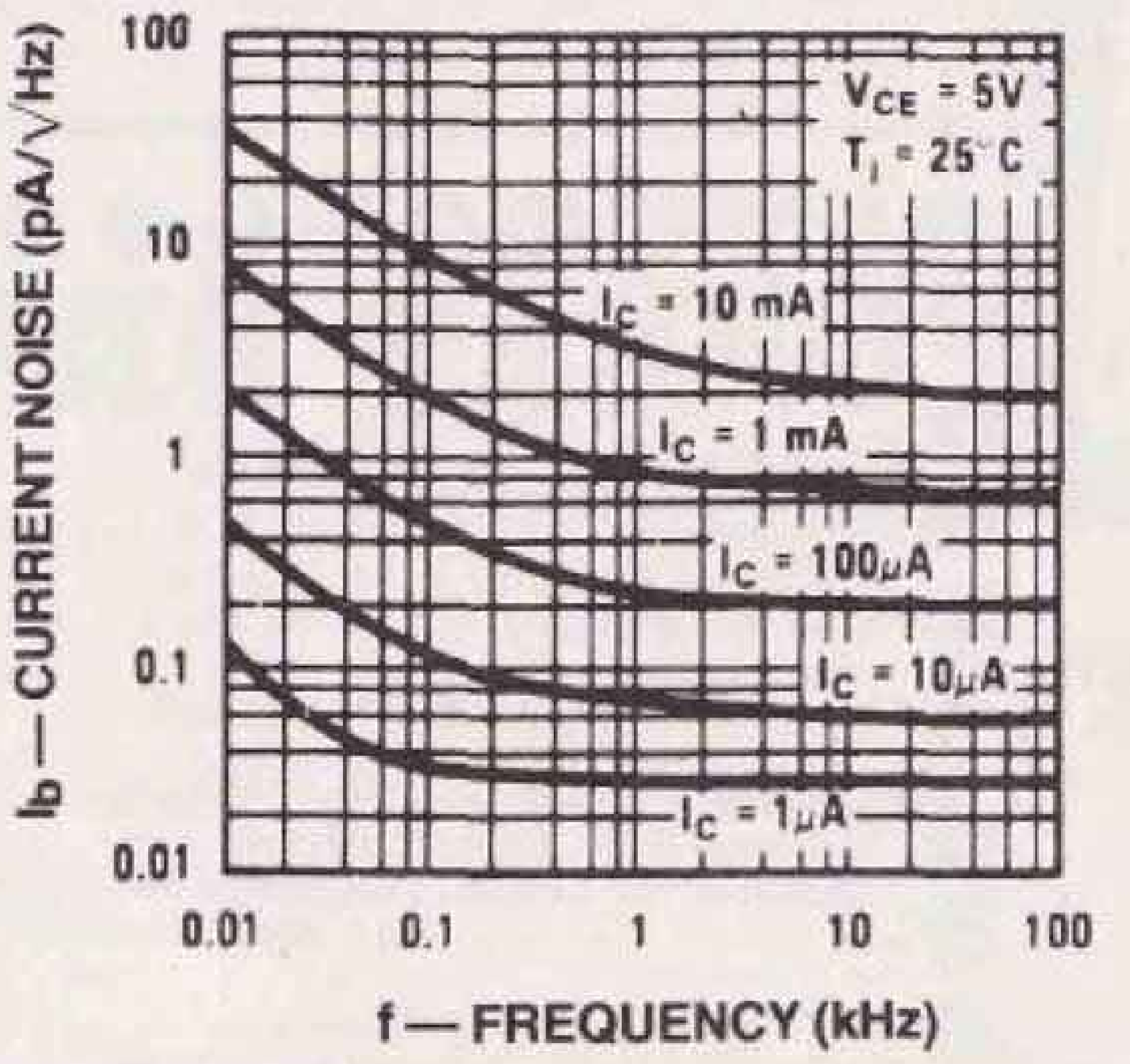


TOP VIEW

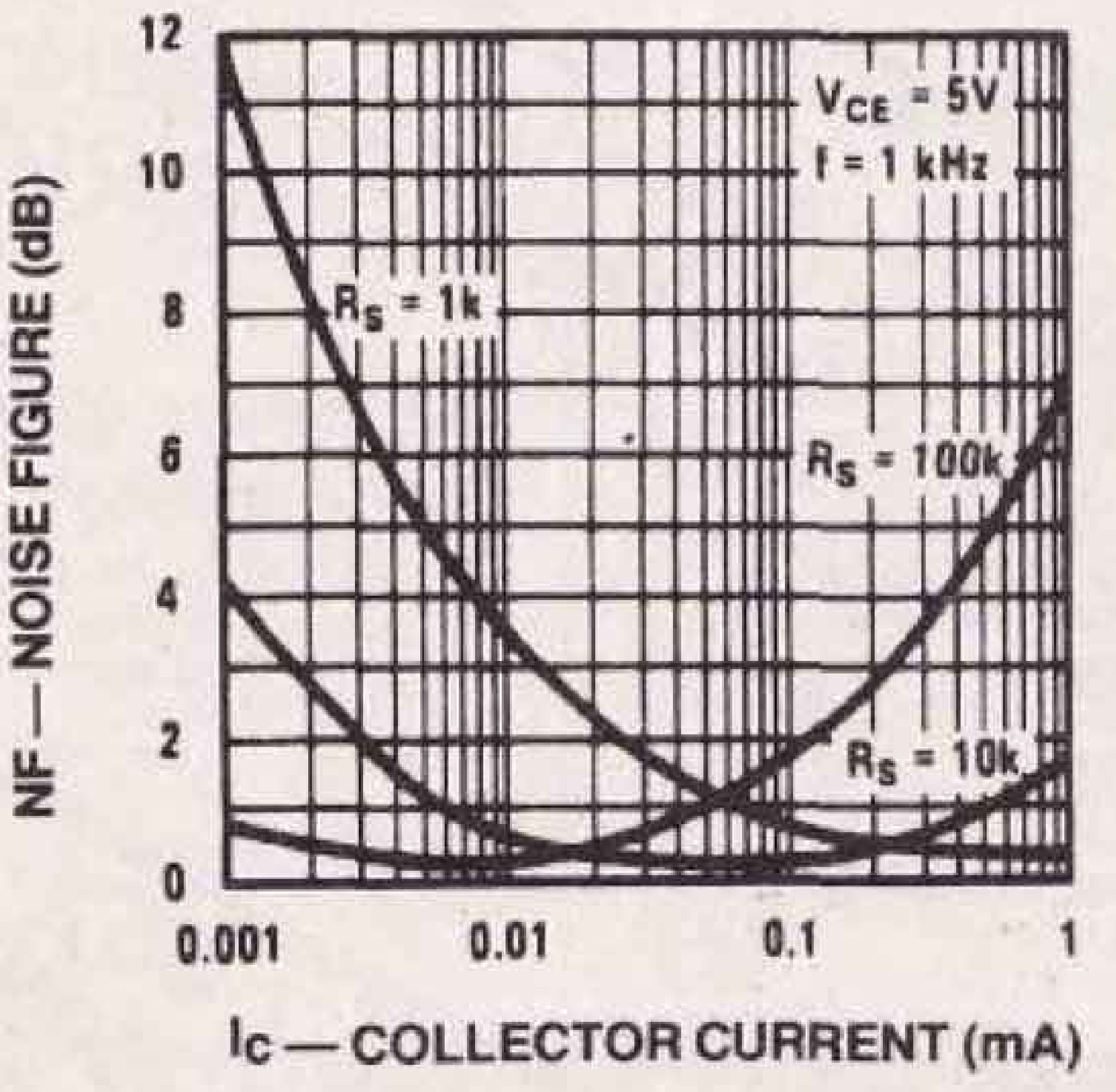
INPUT VOLTAGE NOISE vs. FREQUENCY



BASE CURRENT NOISE vs. FREQUENCY



NOISE FIGURE vs. COLLECTOR CURRENT



NE5534

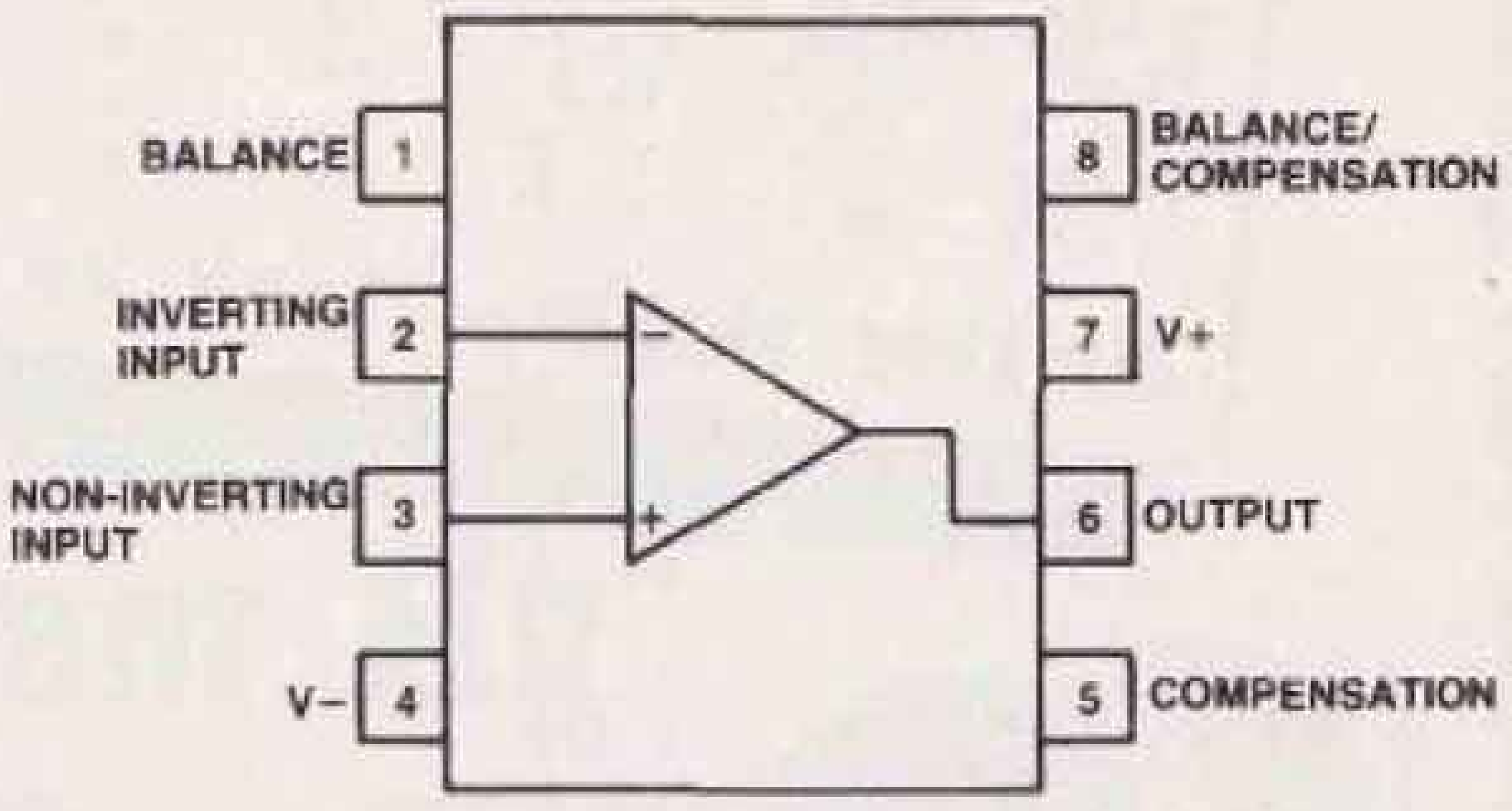
DESCRIPTION

The 5534 is a single high-performance low-noise operational amplifier. Compared to other op-amps, such as TLO83, it shows better noise performance, improved output drive capability and considerably higher small-signal and power bandwidth. This makes the device especially suitable for application in high quality and professional audio equipment, in instrumentation and control circuits and telephone channel amplifiers. The op-amp is internally compensated for gain equal to or higher than three. The frequency response can be optimised with an external compensation capacitor for various applications (unity gain amplifier, capacitive load, slew rate, low overshoot, etc). If very low noise is of prime importance, it is recommended that the 5534A version be used, which has guaranteed noise specifications.

FEATURES

- Small-signal bandwidth: 10 MHz
- Output drive capacility: 600R, 10 V (RMS) at V_S = ±18 V
- Input noise voltage: 4 nV/√Hz
- dc voltage gain: 100 000
- ac voltage gain: 6000 at 10 kHz
- Power bandwidth: 200 kHz
- Slew rate: 13 V/μs
- Large supply voltage range: ±3 to ±−10 V
- Pinout: 741
- Configuration: Single

N PACKAGE



ABSOLUTE MAXIMUM RATINGS

	PARAMETER	RATING	UNIT
V_S	Supply voltage	±22	V
V_{IN}	Input voltage	±V supply	V
V_{DIFF}	Differential input voltage[1]	±.5	V
T_A	Operating temperature range		
	SE5534/5534A	−55 to +125	°C
	5534/5534A	0 to 70	°C
T_{STG}	Storage temperature	−65 to +150	°C
T_J	Junction temperature	150	°C
P_D	Power dissipation at 25°C[2]		
	5534N	500	mW
	5534T	800	mW
	Output short circuit duration[3]	indefinite	
	Lead temperature (soldering 10 sec)	300	°C

NOTES

1. Diodes protect the inputs against over-voltage. Therefore, unless current-limiting resistors are used, large currents will flow if the differential input voltage exceeds 0.6V. Maximum current should be limited to ± mA.
2. For operation at elevated temperature T package must be derated based on a thermal resistance of 150° C/W junction to ambient, 45° C/W junction to case. Thermal resistance of the N package is 240° C/W junction to ambient.
3. Output may be shorted to ground at V_S = ±15V, T_A = 25° C. Temperature and/or supply voltages must be limited to ensure dissipation rating is not exceeded.

ELECTRICAL CHARACTERISTICS $T_A = 25°$ C, $V_S = \pm15$ V unless otherwise specified.

PARAMETER	TEST CONDITIONS	5534			5534A			UNIT
		Min	Typ	Max	Min	Typ	Max	
Input noise voltage	f_o = 30 Hz f_o = 1 kHz		7 4			5.5 3.5	7 4.5	nV/√Hz nV/√Hz
Input noise current	f_o = 30 Hz f_o = 1 kHz		2.5 0.6			1.5 0.4		pA/√Hz pA/√Hz
Broadband noise figure	f = 10 Hz — 20 kHz, R_S = 5 kR					0.9		dB
Channel separation	f = 1 kHz, R_S = 5 kR		110			110		dB

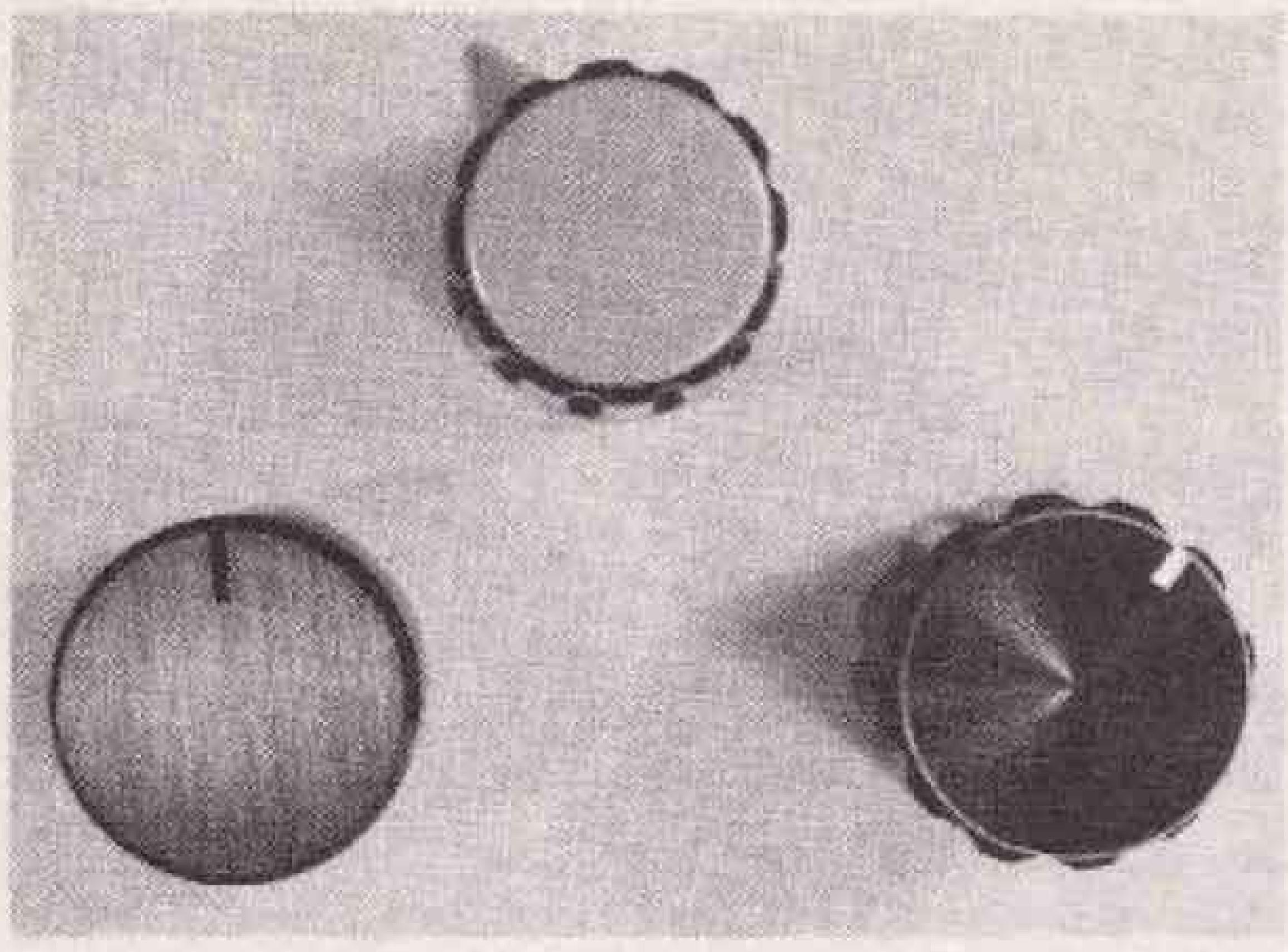
AC ELECTRICAL CHARACTERISTICS $T_A = 25°$ C, $V_S = \pm15$ V unless otherwise specified.

PARAMETER		TEST CONDITIONS	SE5534/5534A			5534/5534A			UNIT
			Min	Typ	Max	Min	Typ	Max	
R_{OUT}	Output resistance	A_V = 30 dB closed loop f = 10 Hz, R_L = 600R, C_C = 22 pF		0.3			0.3		R
Transient response		Voltage follower, V_{IN} = 50 mV R_L = 600R, C_C = 22 pF, C_L = 100 pF							
t_R	Rise time Overshoot			20 20			20 20		ns %
Transient response		V_{IN} = 50 mV, R_L = 600R C_C = 47 pF, C_L — 500 pF							
t_R	Rise time Overshoot			50 35			50 35		ns %
[illegible]	Gain	f = 10 kHz, C_C = 0 f = 10 kHz, C_C = 22 pF		6 2.2			6 2.2		V/mV V/mV
	Gain bandwidth product	C_C = 22 pF, C_L = 100 pF		10			10		mHz
	Slew rate	C_C = 0 C_C = 22 pF		13 6			13 6		V/µs V/µs
	Power bandwidth	$V_{OUT} = \pm10$ V, C_C = 0 $V_{OUT} = \pm10$ V C_C = 22 pF $V_{OUT} = \pm14$ V, R_L = 600R C_C = 22 pF, $V_{CC} = \pm18$ V		200 95 70			200 95 70		kHz kHz kHz

DC ELECTRICAL CHARACTERISTICS $T_A = 25°$ C, $V_S = \pm15$ V unless otherwise specified. 1.2

PARAMETER		TEST CONDITIONS	SE5534/5534A			5534/5534A			UNIT
			Min	Typ	Max	Min	Typ	Max	
V_{OS}	Offset voltage			.5	2		.5	4	mV
		Over temperature			3			5	mV
I_{OS}	Offset current			10	200		20	300	nA
		Over temperature			500			400	nA
I_B	Input current			400	800		500	1500	nA
		Over temperature			1500			2000	nA
I_{CC}	Supply current Per op amp			4	6.5		4	8	mA
		Over temperature			9				mA
V_{CM}	Common mode input range		±12	±13		±12	±13		V
CMRR	Common mode rejection ratio		80	100		70	100		dB
PSRR	Power supply rejection ratio			10	50		10	100	µV/V
A_{VOL}	Large signal voltage gain	$R_L \geqslant$ 600R, $V_O = \pm10$V Over temperature	50 25	100		25 15	100		V/mV V/mV
V_{OUT}	Output swing	$R_L \geqslant$ 600R $R_L \geqslant$ 600R $V_S = \pm18$V	±12 ±15	±13 ±16		±12 ±15	±13 ±16		V V
R_{IN}	Input resistance		50	100		30	100		kR
I_{SC}	Output short circuit current			38			38		mA

NOTES
1. For 5534/5534A. T_{MIN} = 0°C, T_{MAX} = 70°C
2. For SE5534/5534A. T_{MIN} = −55°C. T_{MAX} = + 125°C



You can make your choice of knobs to suit yourself. A wide variety is available and just three different styles are shown here. At left is an anodised aluminium knob with a slot – this is the type we used, simply because we had nine on hand! At top is a plastic collet knob with pointer (C&K and Associated Controls handle collet knobs), and at right is a fluted, anodised aluminium knob with white indicators, which is very common.

adjusting the master and balance controls obtain a voltage of 1.2 Vac at the monitor output sockets. Adjust the LED level meters to read 0 dB. The preamp can now be turned off and the other level meter fitted. Once again the height of the pc board can be set by nuts on the bolt, adjusting to give the correct height. Alternatively, fit two nuts and another washer to all four bolts. This should give the correct height. Solder the input and power supply connections. Note that this board can be difficult to get through the slots on the front panel unless the mounting holes have been drilled large enough.

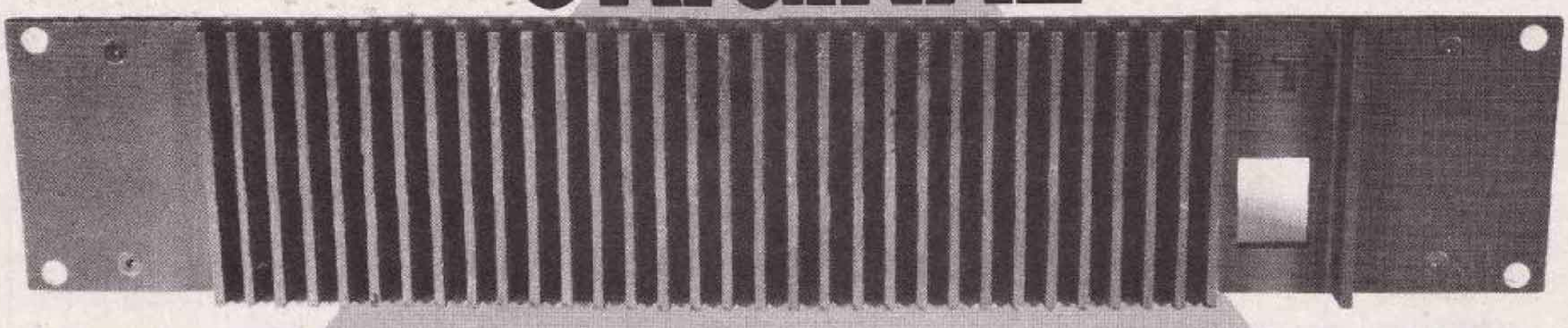
Power up the preamp again, and with the master turned fully down adjust dc offsets as before. Once again adjust the master and balance pots to achieve 1.2 V at the monitor output sockets with the monitor volume control set at full. Adjust the second level meter calibration control to read 0 dB. This aligns the two level meters approximately only. If the master is now varied slowly the LEDs on each display will probably turn on at slightly different times. Adjust the top level meter calibration preset so that the LEDs come on at the same time. With the preamp set in this way the level meters indicate dB below full power when the monitor volume is set at full and the master is used as the volume control. Although this is not the usual mode of operation it is a useful feature, especially when running power amps near their maximum output powers. The usual mode of operation is to adjust the master level to give a reading on the level meters around 0 dB and then use the monitor as the volume control.

All that requires to be done at this stage is to fit the sub-chassis top panel and the preamp top panel. Don't forget to use shorting plugs on all unused low-level inputs in order to avoid thumps in the loudspeakers when the low-level input selector is switched.

Performance

The aim of this project has been to design a high-quality preamplifier suitable for home construction that will not degrade the performance of the best available power amps. To do this the conventional parameters of frequency response, noise and distortion must be good. In these respects the Series 5000 is extremely good, as can be seen from the specifications quoted elsewhere in this article. Of equal importance, however, are the less-known parameters such as cartridge impedance interaction. This problem is overcome through the use of a separate linear gain stage at the input to the MM amplifier.

The final test of any audio amplifier is of course subjective, but for me the Series 5000 preamplifier offers a significant improvement over many other designs, offering a detail and clarity that is seldom heard. ●



LED level meter features simultaneous peak & average display plus 60 dB dynamic range

David Tilbrook

This project is, in effect, the first part of the construction articles for the Series 5000 Control Preamplifier. The LED level meter described here, though originally designed for the Series 5000 Preamp, is ideal for any application requiring a wide dynamic range level display. Naturally, two are required for stereo applications.

THE MOST common instrument used to measure audio signal level is the VU meter (VU stands for volume unit). Before the introduction of the VU standard however, ordinary meter movements were used. A full-wave rectifier converted the applied audio signal to dc suitable for driving a voltmeter, usually fitted with a dB scale. Although this is completely suitable for steady sinewave measurement it is entirely unsatisfactory for measurement of constantly changing voltages such as audio signal level. The biggest problem is overshoot of the meter movement. If a 1 kHz sinewave, for example, is applied to this type of meter, the movement can overshoot the correct reading by nearly 80%, indicating a transient that is in fact not present. The VU standard was introduced to overcome these problems. It does this by defining the 'ballistics' of *any meter movement to be used in audio* signal level measurement. A comparison of VU and ordinary meter movement ballistics is shown in Figure 1. The amount of overshoot of the VU meter is specifically defined by the standard to be not less than 1% and not greater than 1.5%. This characteristic is achieved by carefully modifying the shape of the meter pole pieces and counterweighting the pointer. These techniques ensure that the movement stabilises in the shortest possible time, around 0.3 s (300 ms) for the case shown in Figure 1. The VU meter still displays dB (i.e: 1 VU = 1 dB), but its reaction to transient signals is significantly better than the ordinary meter movement.

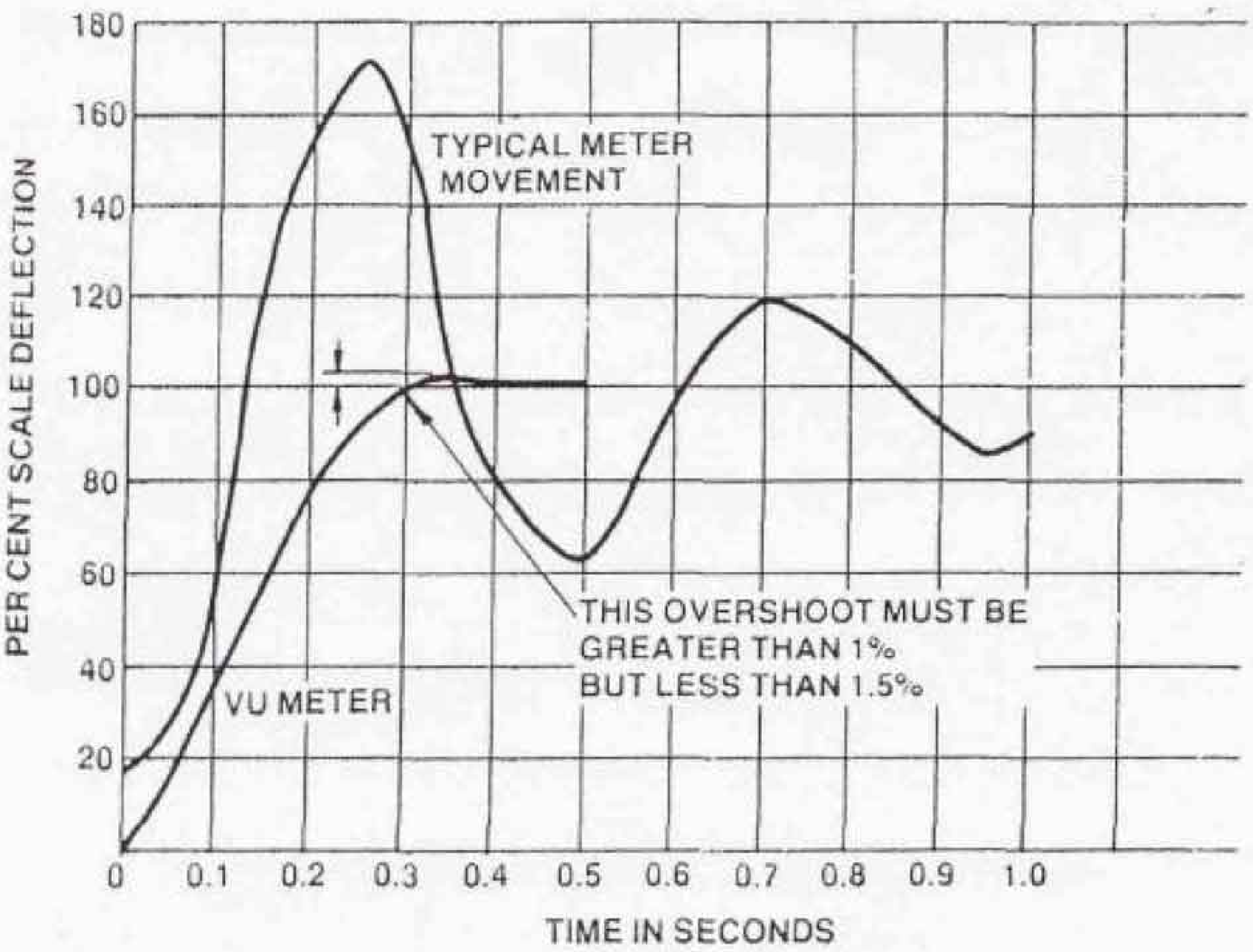
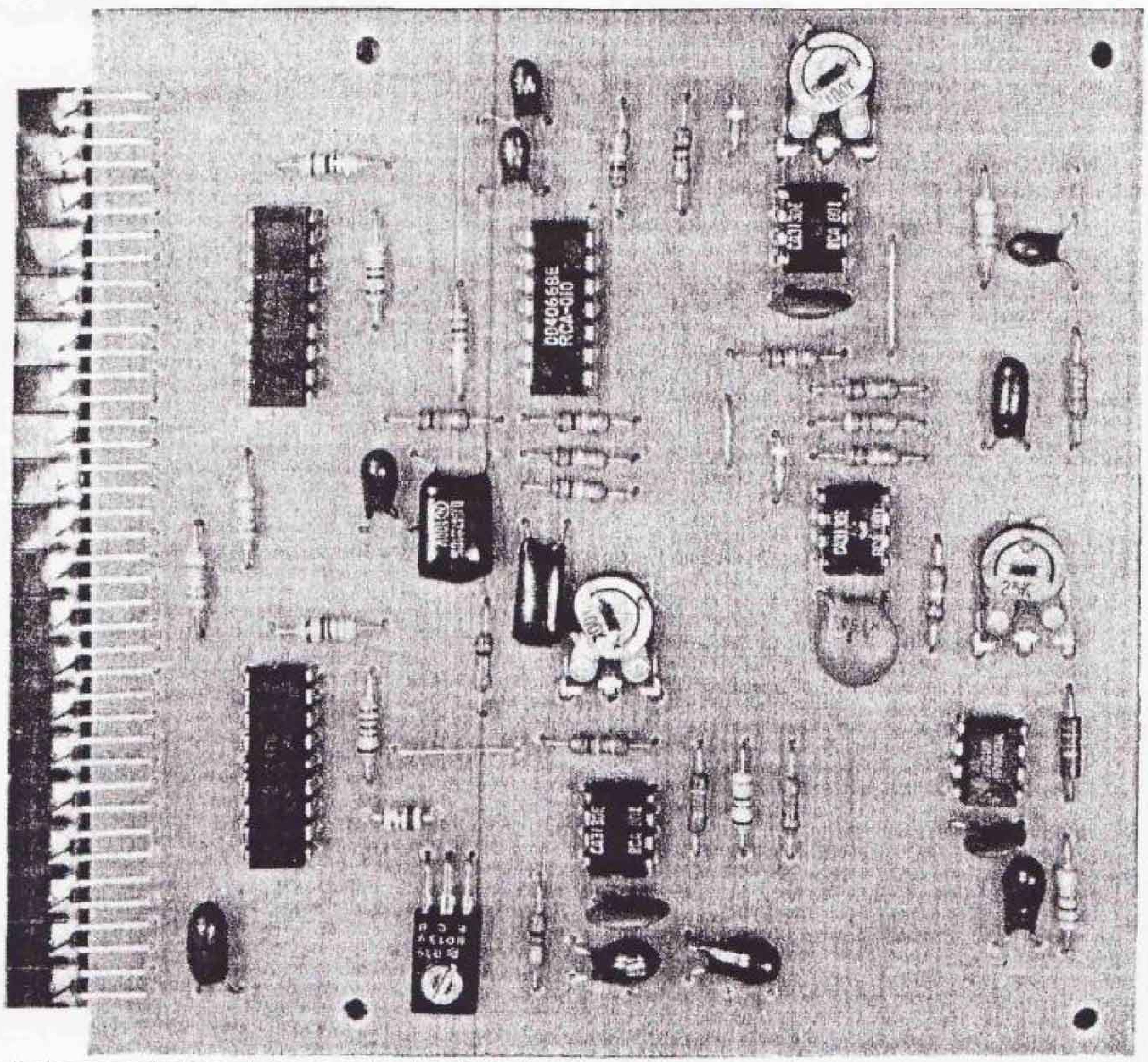


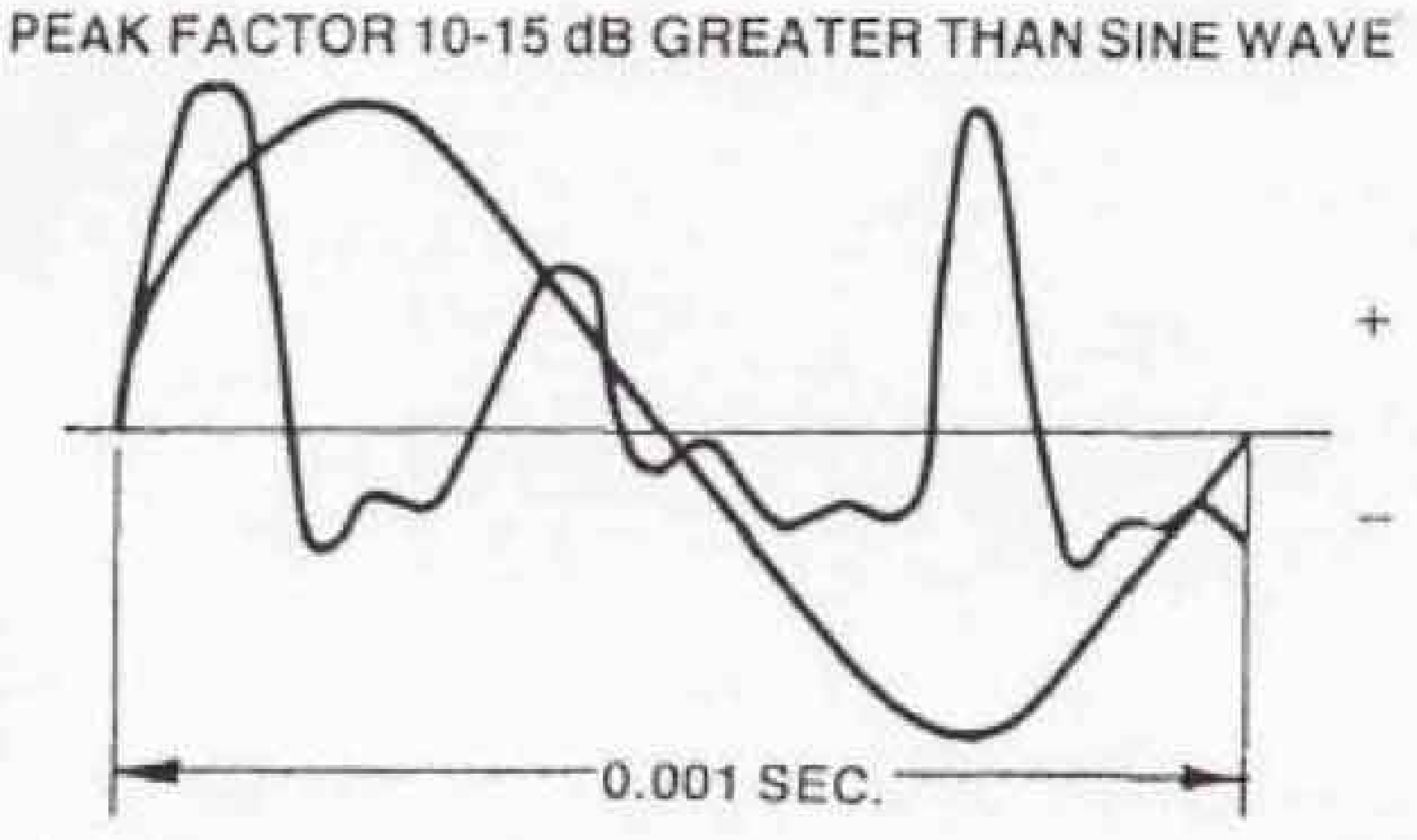
Figure 1. 'Ballistics' of a VU meter compared to conventional moving-coil meter.

Nevertheless, the VU meter is still very slow. It indicates something between the average and the real peak of the signal voltage depending on the complexity and transient nature of the particular input signal. The 0.3 s rise time of the meter will hide all but the most repetitive peaks, so a VU meter could be indicating a signal voltage of say −15 dB when the peaks of the signal are actually overloading an amplifier. Another disadvantage of most VU *meters is their limited dynamic range.* Usually they display only the 'top' 23 dB of the total range (i.e: −20 to +3 dB) and with the ever increasing dynamic range of modern recording techniques this is not sufficient.

The ETI-458 overcomes these problems by replacing the meter movement



Full-size reproduction of the completed project. Note the components are laid flat to permit close stacking of two boards for a stereo display.



A typical 'music' signal may have a completely different peak-to-average ratio compared to a sinewave, and the peaks are often not symmetrical in amplitude about the zero axis. The duration of peaks may be as short as 50 microseconds.

with a row of light emitting diodes driven by a pair of dB LED display drivers. Twenty LEDs are used, with 3 dB between each LED, so the total dynamic range displayed is 60 dB. The circuit monitors both the true peak and the average signal level and displays both simultaneously. The difference between the peak and the average voltages of a sinewave is around 3 dB, so with a sinewave applied consecutive LEDs will light. With music applied however, the difference between the two LEDs will be substantially greater, depending on the transient nature of the signal applied. ▶

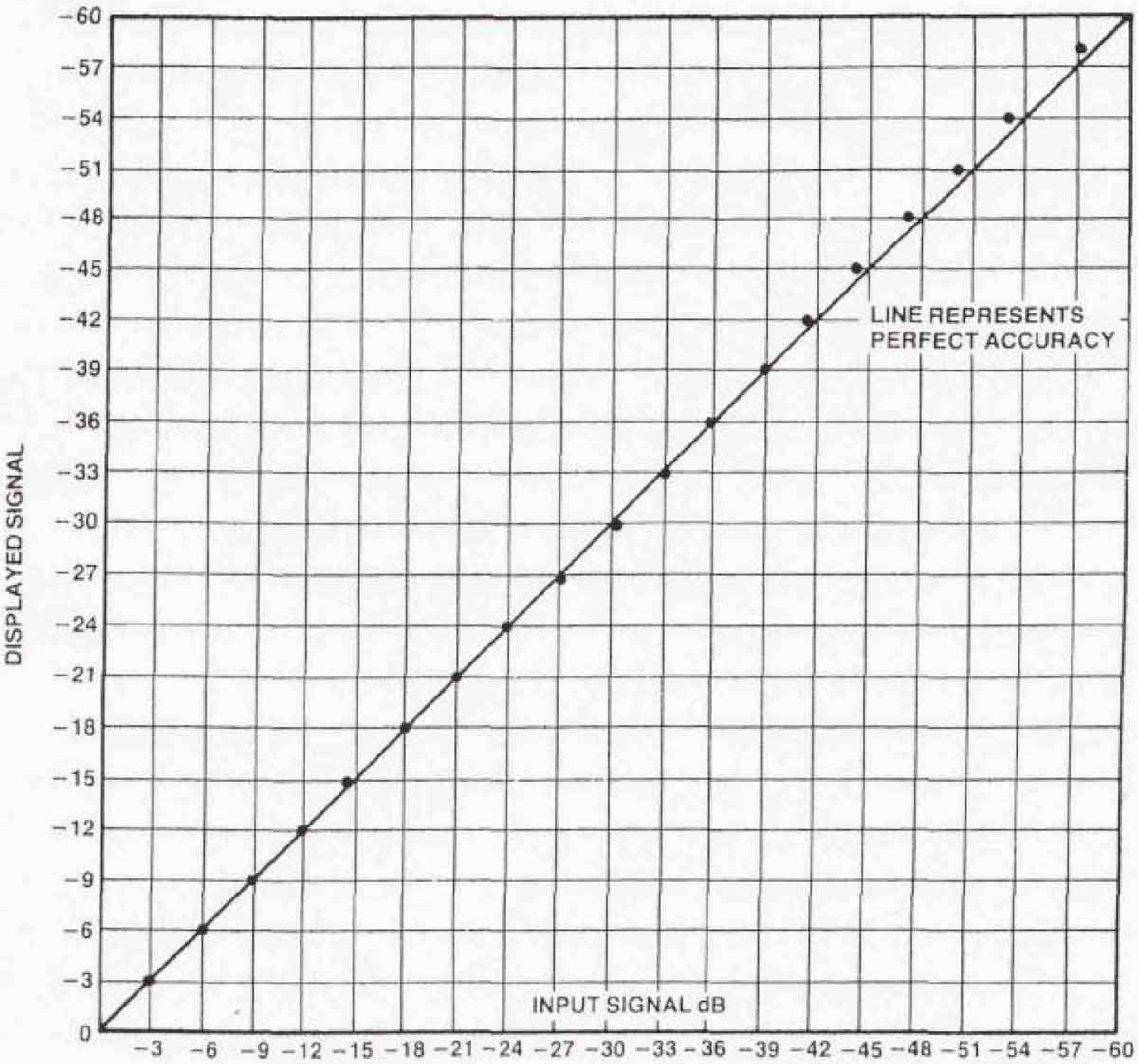


Figure 4. Accuracy of the ETI-458 LED level meter display (dots) compared to 'perfect accuracy' (line).

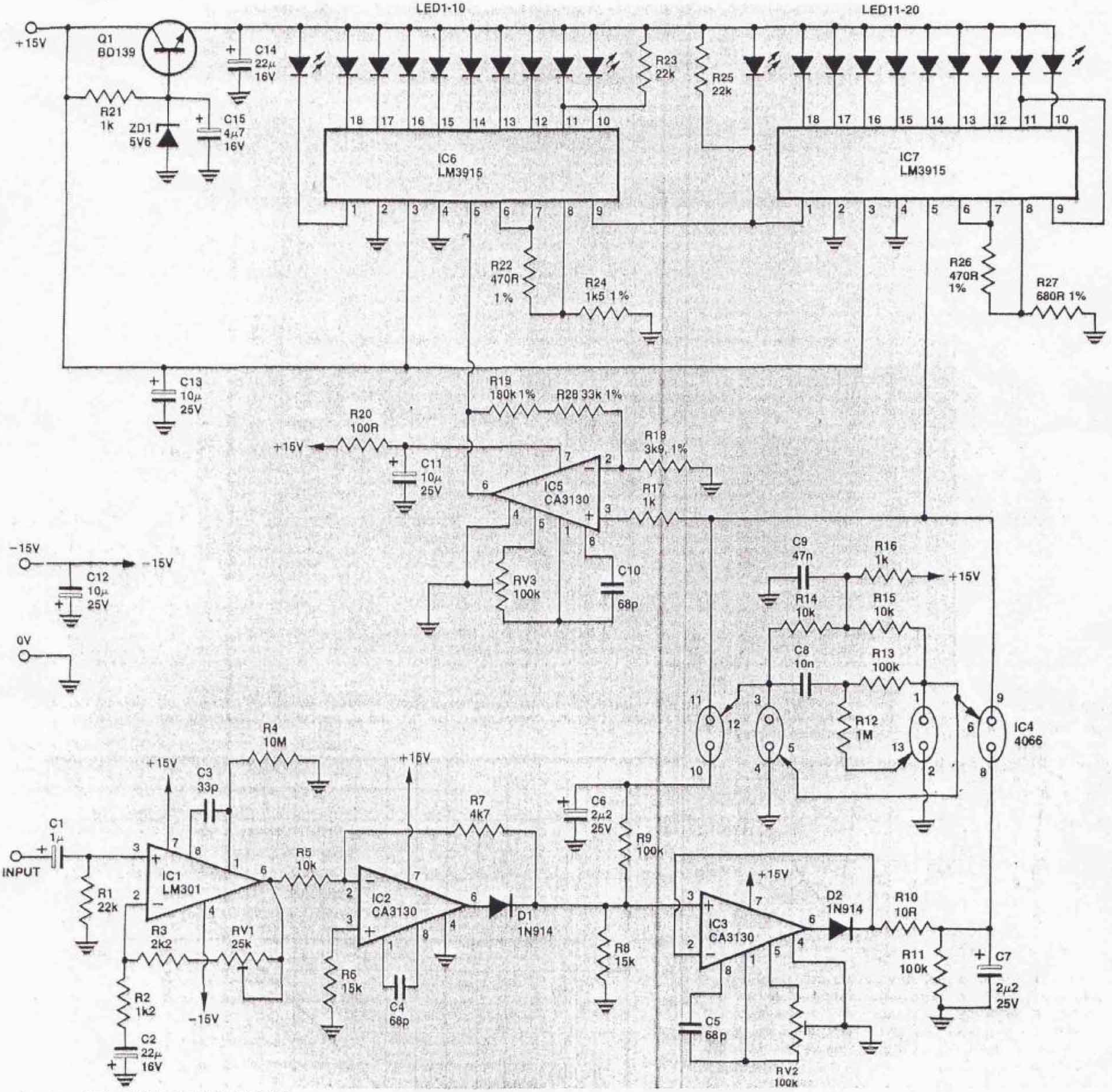


Figure 2 Circuit diagram of the ETI-458.

Figure 2 shows a complete circuit diagram for the LED level display. The input is fed first to a *prescaling* amplifier formed by an LM301 op-amp, IC1, and the associated passive components. This stage has adjustable gain, set by the preset RV1 that allows the 0 dB point to be set to the desired reference voltage. This will be covered in greater depth later, in the setting up procedure. The output of the prescaling stage is connected to the input of a full wave rectifier formed by IC2 and its associated components. Most full wave rectifiers use several op-amps so this circuit is considerably simpler. A detailed description of its operation is included in the 'How it Works' later in this article and should be of interest to anyone needing a full wave rectifier with a minimum number of components. The output of the full wave rectifier is fed to an averaging filter formed by R9 and C6, and to a peak follower formed by IC3 and associated components. The peak follower has a rapid attack/slow decay characteristic so that it responds quickly to any transients but decays slowly so the transient can be seen easily on the display. The outputs from the peak follower and the averaging filter are connected to the inputs of two CMOS analogue switches. The outputs of these switches are connected together and go to the input of the LED display. Two more CMOS switches are used to form a square wave oscillator. This oscillator has out of phase outputs used to drive the signal-carrying analogue switches alternately off and on at a relatively high frequency. When the switch connected to the output of the averaging filter is on, the average signal voltage is connected to the input of the LED display. This switch is subsequently turned off by the oscillator and the other analogue switch turned on, connecting the output of the peak follower to the LED display. So, only one of the two LEDs is on at any instant, but the rapid switching speed between them and the persistence of vision make them both appear to be on.

Input signals to the LED display portion of the circuit are fed simultaneously to the LM3915 driving the upper 30 dB display and via a voltage amplifier to the lower 30 dB display. The biggest problem in the design of an audio level meter with a 60 dB dynamic range arises from the fact that 60 dB below typical 0 dB input voltages could be around 2 mV. This is well below the dc offset voltage of most op-amps so special precautions have been taken in the design to ensure that dc offset errors can be reduced to negligible levels. This is the purpose of the presets RV2 and RV3. These are dc offset controls. Adjustment of these is covered in the setting up procedure. The sensitivity of the LM3915 can be adjusted by changing the voltage between pins 6/7 and ground. The IC maintains a voltage of 1.25 V across R22. The current through R22 will be 1.25/470 or approximately 2.67 mA. A further 75 uA is supplied from pin 8 of the device, so the total current through resistor R24 to ground will be 2.67 mA + 75 uA or approximately 2.73 mA. The voltage drop across R24 will therefore be around 2.73 mA x 1.5k or 4.1 V. Adding the 1.25 volts across R22 gives a total of 5.35 V between pins 6/7 and ground.

This means that the topmost LED driven by IC6 will light when the input voltage to the device is 5.35 V. Now, 30 dB below this is:

$$\frac{-30}{20} = \log \frac{x}{5.35}$$

or 1.17 V, which is well above the voltage expected on the output of IC5 due to the dc offset. The reference voltage used was chosen specifically to ensure that this would be the case. Now the easiest way to cascade two LM3915s would be to simply set the reference voltage of the second LM3915 the same as that of the first and precede the first one by a 30 dB gain amplifier. However, with the recommended supply voltage of +/−15 V the maximum peak signal voltage that can be delivered by IC1 will be around 6 V. The operation of the absolute value generator (full wave rectifier, IC2) further divides this by two, so the maximum peak signal voltage available will be around 3 V and the top several LEDs would never be lit. To overcome this problem the reference voltage of IC7 is decreased so that the top LED will be lit by a 3 V input signal, and the gain of the amplifier formed by IC5 is changed accordingly.

The resistors R26 and R27 set the reference voltage of IC7 at 3.1 V and 30 dB below this voltage is

$$\frac{-30}{20} = \log \frac{x}{3.1}, \text{ or } 98\text{ mV}.$$

Now, the top LED driven by IC6 must correspond to this voltage, so the required gain around IC5 is 5.34/98 mV or 54.6. The values of the resistors R19 and R18 set this gain at (180+33+3.9)/3.9 or around 56 which is a good enough approximation, amounting to an error of less than 0.5 dB.

▶

HOW IT WORKS — ETI 458

The input stage consists of a variable gain amplifier formed by IC1 and its associated components. This is a conventional IC amplifier circuit in which the gain is determined by the values of the components RV1, R3 and R2. Specifically:

$$A_V = \frac{R2 + R3 + RV1}{R2}$$

So the bigger the value set on RV1, the greater the gain. Capacitor C2 has the effect of decreasing this gain for very low frequencies, or dc, decreasing the dc offset on the output.

The second stage is the full wave rectifier or 'absolute value generator'. As mentioned in the text, most full wave rectifiers require more than a single op-amp, so this stage will be of use in any application requiring a full wave rectifier with minimum component count. For negative-going signals the stage functions as an inverting amplifier with a gain of 0.5. This is determined by the values of R5 and R7. When the input signal goes positive the output is driven hard against its negative supply voltage, which in this case is 0 V. So the output stage is turned off, and has a relatively high output impedance. In this state the resistors R5, R7 and R8 form a potential divider and connect the input signal to the output directly. Again, the output voltage is one half of the input voltage. In order for this circuit to work, the output stage in the op-amp must be CMOS so that the output can go completely to 0 V and have an output impedance high enough not to short out the signal voltage from the potential divider. This is the reason the CA3130 is used. Furthermore, this is a relatively fast device which ensures that the full wave rectifier will have a frequency response that covers the entire audio spectrum. The one disadvantage of the circuit is that it requires a high load impedance since the output signal for positive-going input signals is obtained from the potential divider and not from the op-amp itself. In this application the load is around 100k (R9) which causes negligible error.

The output of the full wave rectifier is fed simultaneously to an average filter formed by R9 and C6, and to the peak hold circuit formed by IC3 and its associated components. The peak hold circuit is really nothing more than a 'precision diode' that charges a capacitor to the peak voltage. The precision diode is formed by including a conventional signal diode in the feedback loop of a fast op-amp. If an input signal is applied which is less than the forward voltage drop of the diode, the stage is effectively in open loop gain (around 320 000 for the CA3130). The output voltage will rise very quickly, turning the diode on. Since the output of the diode is connected to the inverting input of the op-amp, the stage functions with unity gain once the diode has been turned on. Capacitor C5 ensures stability of the stage while preset RV2 allows adjustment of dc offsets due to this stage. The output of the peak hold circuit charges capacitor C7 through resistor R10. The combination of R10 and C7 defines the attack rate of the peak detector.

As shown, the value of R10 is 10 ohms and this is small in comparison to the output impedance of the CA3130, but is included in case some applications require the peak detector to have a slower attack rate. With the values shown, the LED level meter will display single 50 uS pulses accurately and this is entirely adequate for any audio application.

Resistor R11 discharges the capacitor and its value of 100k dictates a decay rate of around one second. This gives the level meter its rapid attack, slow decay characteristic and enables even short transients to be spotted.

As explained in the text, both the average and the peak levels of the signal are displayed simultaneously. This is accomplished by multiplexing the outputs of the peak and average detectors. This is done by switching between the output of these two circuits at a relatively high frequency (say a few hundred Hertz). In the circuit, this is done with CMOS transmission gates. The 4066 was chosen mainly because its on resistance is a little lower than the older 4016 and this enables the remaining two gates in the package to be used as the driving oscillator. The oscillator is formed by resistors R12 to R15 and capacitor C8, with the associated two transmission gates. The frequency of the oscillator is determined by the values of R13 and C8 at around 150 Hz.

IC5 functions as an amplifier stage as discussed in the text. Once again dc offset adjustment is provided, this time by RV3. Capacitor C10 provides the necessary compensation to ensure stability. Details of the two LED drivers and the amplifier formed by IC5 are in the main text.

The transistor Q1 and the associated components R21, C15 and ZD1 form a simple 5V regulator to power the LM3915s. Capacitor C16 is essential for stability of the LED drivers and must be mounted close to the LEDs.

Internally, the LM3915 consists of a string of comparators; each one compares the input signal to a reference voltage it derives from a ten-way potential divider (see Figure 3). The accuracy of the LM3915 is determined by these internal resistors and is therefore very good. To ensure the display is accurate over the entire 60 dB range it is only necessary to ensure that the changeover from one LM3915 to the other is accurate. Resistors R18, R19, R22, R24, R26, and R27 have been specified as 1% tolerance types for this reason. This is probably unnecessary for most applications. I have built the unit using 5% types and the error was only around 1.5 dB which is effectively hidden by the 3 dB increments between LEDs. Figure 4 shows the accuracy of one of the prototype units built with 1% resistors in the places specified. If the accuracy were perfect, all the dots would lie on the straight line. The deviation from the line is only small, so the unit is very accurate over the entire 60 dB dynamic range.

Transistor Q1 forms a simple voltage regulator delivering 5 V to the LEDs. This decreases the power dissipation in the LM3915s. The current consumption from the positive rail is around 100 mA while the negative rail needs only several milliamps. If the display is to be used from an existing power supply in a preamplifier for example, care should be taken to ensure that the relatively high positive rail current does not upset the preamplifier performance. In the Series 5000 preamp a separate positive rail is used for the display to decrease any possibility of interaction between the display and the audio signal voltages in the preamp.

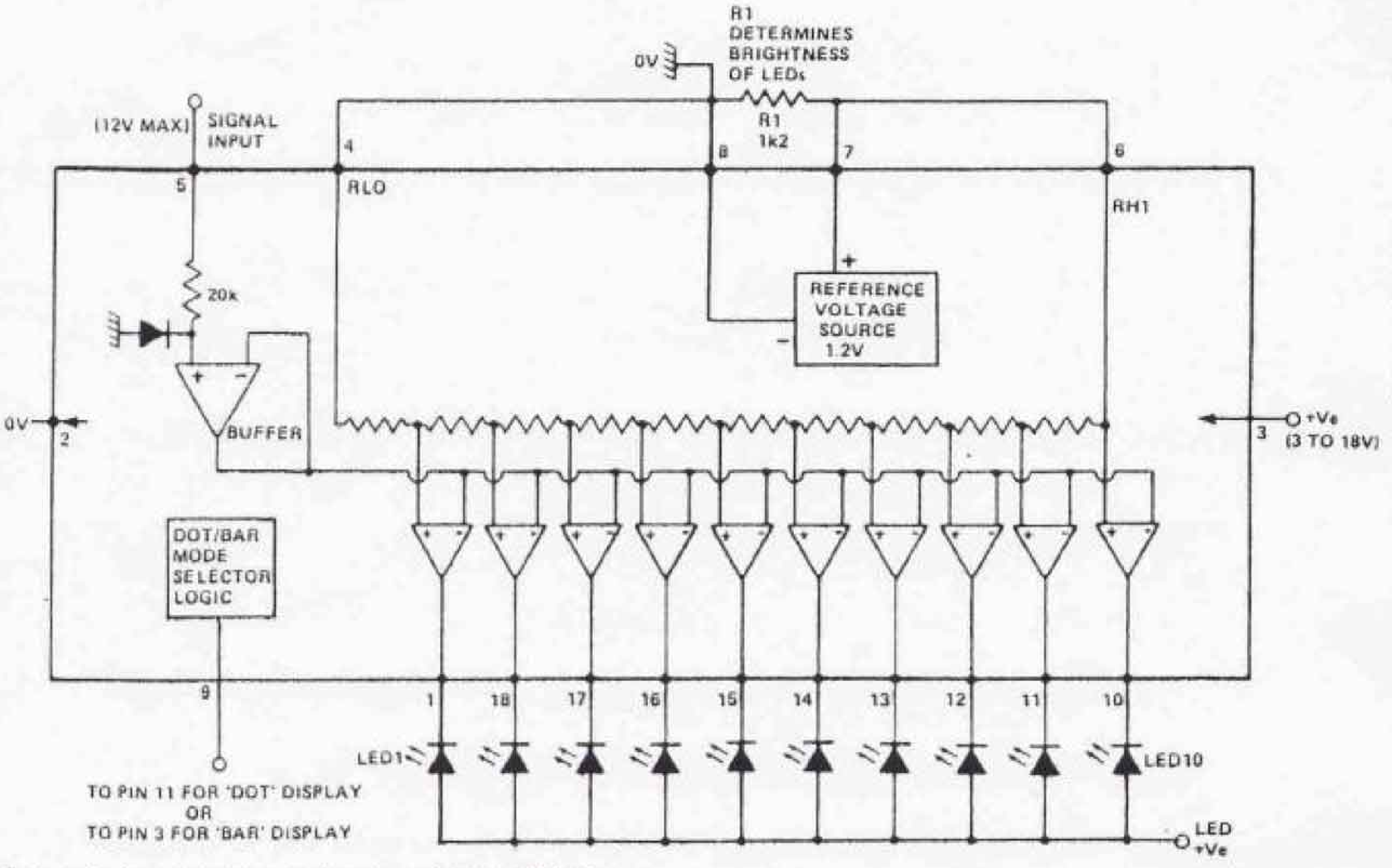
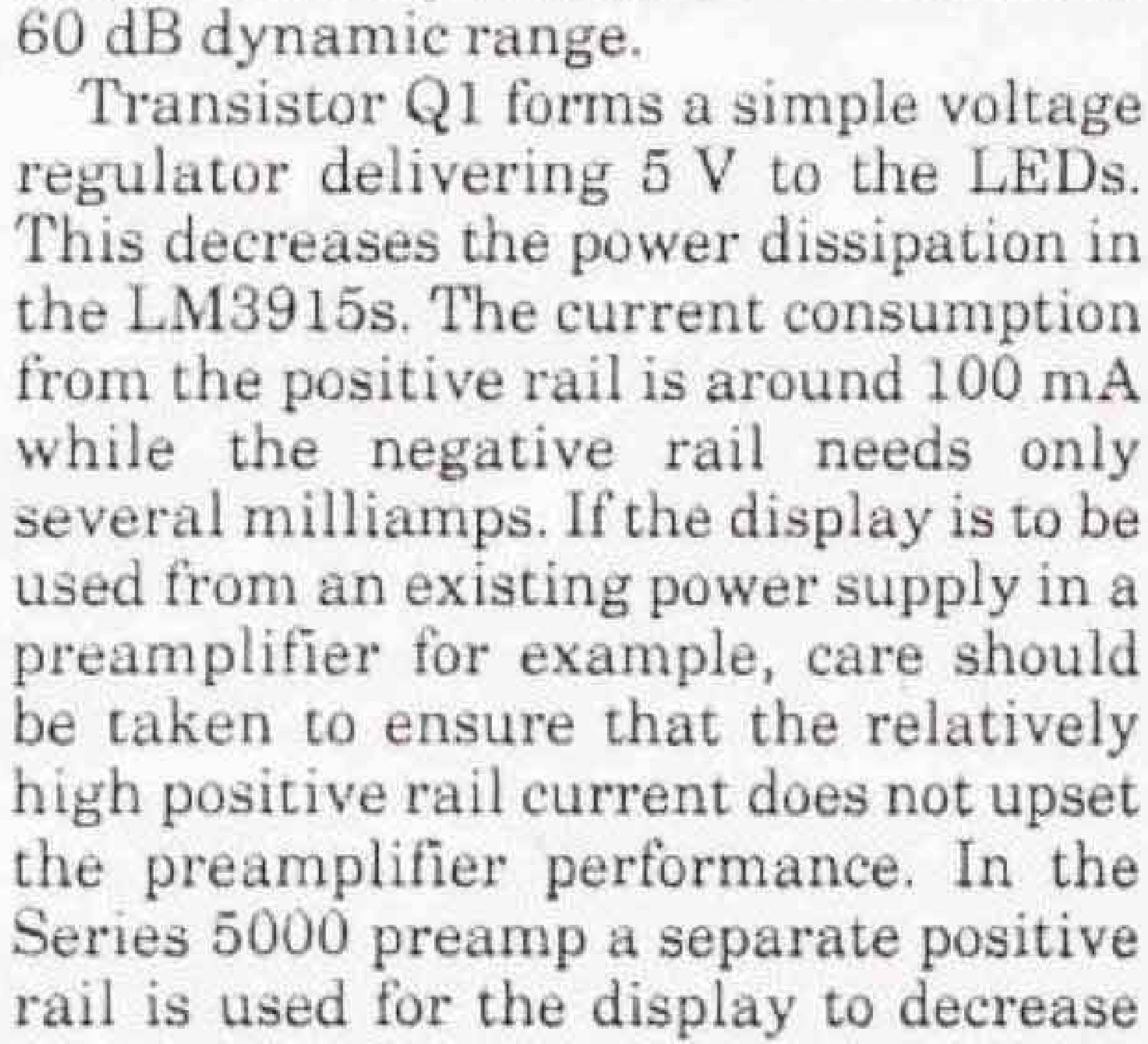


Figure 3. Internal block diagram of the LM3915.



Close-up of the pc board showing orientation of the LEDs. IC7 at lower right.

Construction

The pc board is virtually essential for this project, particularly if you are constructing it as part of the Series 5000 Control Preamp.

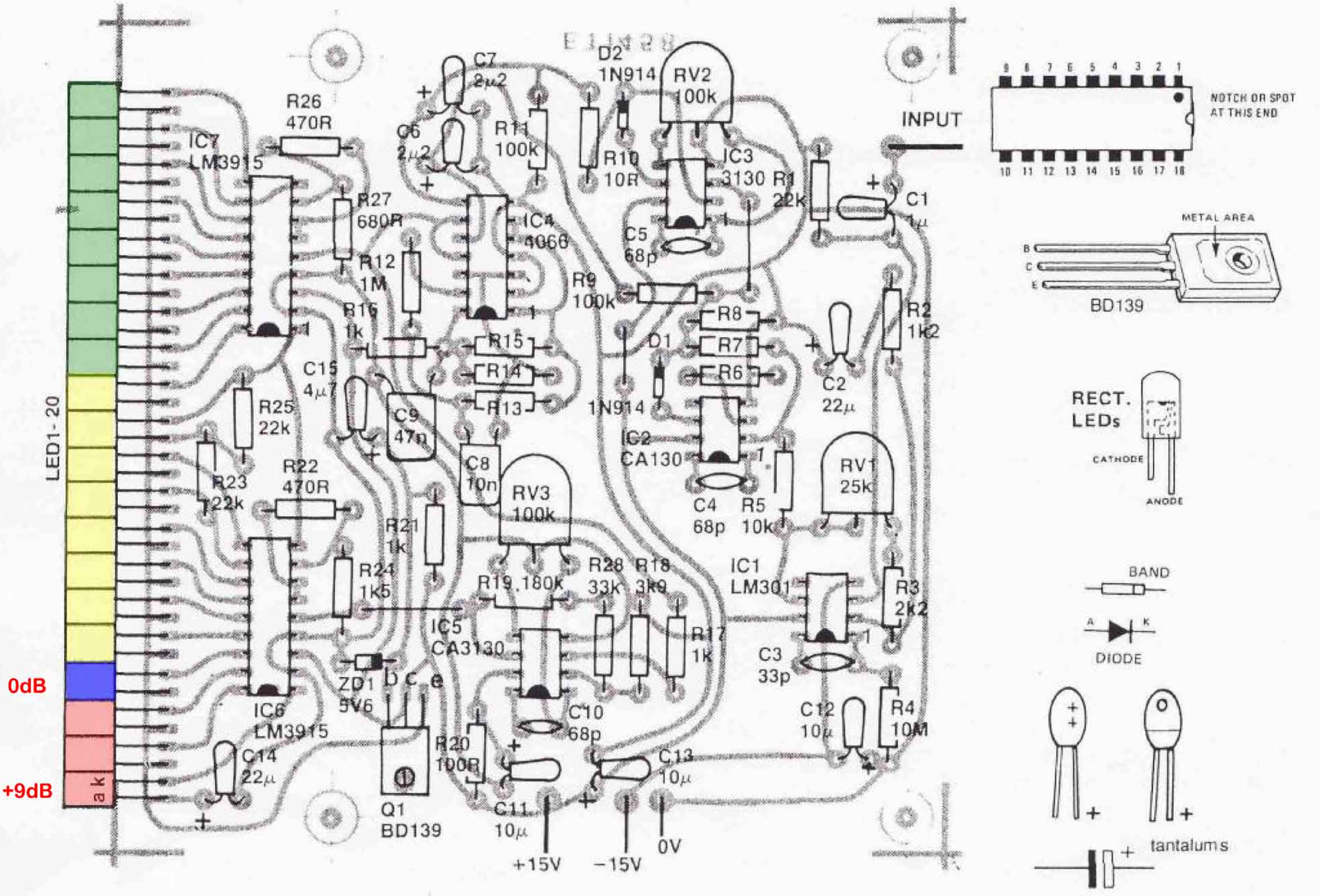
Start construction by mounting the LEDs. This is by far the most difficult part of the project. The LEDs must be inserted evenly and with equal heights, and this is *not* easy. Furthermore, the LEDs must be inserted the right way around. The longer of the leads represents the anode of the LED. Check the orientation of each LED against the overlay, before soldering. The best way I found to mount the LEDs is to start by inserting the first LED on one end of the display. Bend this LED flush against the *edge* of the pc board. Now solder the leads and bend the LED upright again. Insert the next LED and ensure that its height on the board is identical to the first. Now solder the second LED into position. Continue like this for the remaining eighteen LEDs, checking the orientation of each one as you go. After all the LEDs are soldered into position check that the heights are all even and make any adjustments needed now by reheating the appropriate solder joints. Be careful when soldering the LEDs that you do not overheat the leads; this will damage the device and is very easily done. Once all the LEDs are even bend the whole line down against the circuit board as shown in the photographs.

Now all the other components can be mounted. The order of mounting is not really important although it is good general practice to solder the passive components first (resistors and capacitors). And then solder the ICs and transistors. In the Series 5000 Preamp the LED level displays are mounted directly above one another, so all components should be mounted as close as possible to the pc board. The presets are mounted against the circuit board and this is best done by bending their leads at right angles first, and then soldering. Similarly, many of the larger capacitors, such as the greencaps and ceramics, may have to be folded against the board. Leave sufficient lead on the components so that this can be done. Alternatively, bend the component over before soldering. Be careful with the orientation of all polarised components, such as transistor Q1 and the electrolytic and tantalum capacitors. Tantalum capacitators, for example, are very intolerant of reverse biasing.

Setting up proceedure

Once all the components have been mounted on the pc board and checked, the unit can be switched on. Ensure that the power supply you are using has sufficient current capability for the positive rail and that it is correctly connected to the supply points on the circuit board. If the input is touched with a finger two LEDs should light and move up the display. If all is well the dc offsets can now be adjusted. The preset RV2 adjusts the dc offset of the peak follower. This will be adjusted to equal the dc level of the average filter, i.e: that from the output of the full wave rectifier. The overall dc offset can be nulled by RV3.

First connect the input of the LED level meter to earth on the board. This ensures that no signal voltage will be present when the adjustments are made. Now turn both RV2 and RV3 fully clockwise; both LEDs should run off the bottom of the display. Turn RV3 slowly anticlockwise until the second LED from the bottom has just turned on. If RV2 is now turned anticlockwise also, a second LED will light on the display. This is the peak level LED. Adjust RV2 to superimpose this LED



onto the second bottom LED. Now adjust RV3, turning it clockwise again until the LED has just run off the bottom of the display.

The final stage in the setting up procedure is to align the meter for the appropriate 0 dB level. Preset RV1 varies the gain of the prescaling amplifier stage formed by IC1. Adjustment of this preset will vary the input voltage required to light the top LED between 260 mV and 2.5 V. If your application requires 0 dB to be a higher voltage than 2.2 V, use a potential divider at the input to decrease the input signal voltage. If more gain is required increasing the value of the preset from 25k to 100k will decrease the necessary input voltage to around 70 mV, which should be sufficient for most applications.

In the Series 5000 amplifier the top LED is designated +9 dB, so the fourth LED from the top is 0 dB. Calibration of the 0 dB reference is best left until the preamp is finished and the procedure will be described in the Series 5000 Preamp construction article, coming soon. ●

PARTS LIST — ETI 458

Resistors all ½ W, 5% unless marked otherwise

R1, 23, 25	22k
R2	1k2
R3	2k2
R4	10M
R5, 14, 15	10k
R6, R8	15k
R7	4k7
R9, 11, 13	100k
R10	10R
R12	1M
R16, 17, 21	1k
R18	3k9 1%
R19	180k 1%
R20	100R
R22, R26	470R 1%
R24	1k5 1%
R27	680R 1%
R28	33k 1%
RV1	25k min. trimpot
RV2, RV3	100k min. trimpot

Capacitors

C1	1μ/6V tant.
C2, C14	22u/16 V tant.
C3	33p ceramic
C4, 5, 10	68p ceramic
C6, C7	2u2/25 V tant.
C8	10n greencap
C9	47n greencap
C11, 12, 13	10u/25 V tant.
C15	4μ7/16 V tant.

Semiconductors

IC1	LM301, 8-pin DIL
IC2, 3, 5	CA3130, 8-pin DIL
IC4	4066
IC6, IC7	LM3915
D1, D2	1N914 or sim.
ZD1	5V6 zener diode
Q1	BD139
LED1 - 20	Siemens LD80-2 or sim.

Miscellaneous

ETI-458 pc board (double-sided); one 6 BA bolt and nut.

Price estimate

We estimate the cost of purchasing all the components for this project will be in the range:

$38 — $44

Note that this is an **estimate** only and **not** a recommended price. A variety of factors may affect the price of a project, such as — quality of components purchased, type of pc board (fibreglass or phenolic base), type of front panel supplied (if used) etc — whether bought as separate components or made up as a kit.