BERNARDS

BUILD YOUR OWN

SOLID STATE HI-FI AND AUDIO ACCESSORIES

compiled by

M. H. BABANI B.Sc. (Eng.)

85p

BUILD YOUR OWN SOLID STATE HI-FI AND AUDIO ACCESSORIES

BUILD YOUR OWN SOLID STATE HI-FI AND AUDIO ACCESSORIES

compiled by

M.H. BABANI, B.Sc.(Eng.)

BERNARDS [publishers] LTD
The Grampians
Shepherds Bush Road
London W67NF
England.

The compiler of this book wishes to express his most sincere thanks to "ELECTRONICS AUSTRALIA", the leading radio, T.V. and electronics magazine published in that continent and to Sungravure Syndication, their agents, for permission to use some of the material in this book which appeared originally as articles in that magazine.

I.S.B.N. 0 900162 61 9

First Published – July 1976

Printed and Manufactured in Great Britain by
C. Nicholls & Co. Ltd.

CONTENTS

SWING OVER TO STEREO WITH THIS ADD-ON DECODER

This little add-on decoder unit should have wide appeal. Providing your existing mono FM tuner or receiver is adequate, it will enable you to produce stereo signals of very good quality. The circuit uses one of the latest "no coils" decoder ICs, and includes active filters to inhibit heterodynes when tape recording.

If you're an audio hifi enthusiast with a mono FM tuner or receiver, you've probably been wondering if it would be possible to adapt it so that you can take advantage of the experimental stereo transmissions now being radiated. The answer to this is yes, provided that your tuner or receiver has a reasonable IF response, and provided also that you tackle the conversion in the right manner.

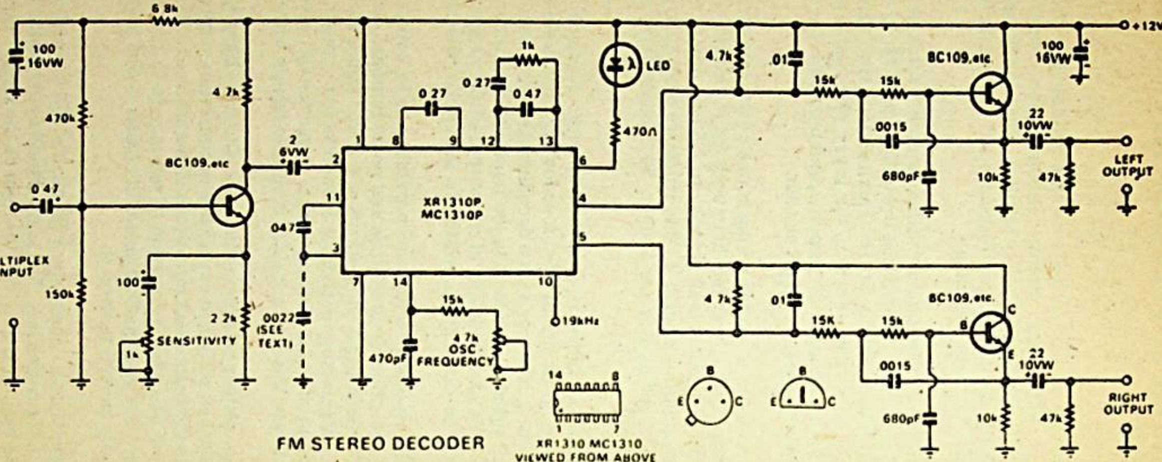
It isn't just a matter of simply wiring in one of the new decoder ICs, though – despite what you may have been led to believe. There are a number of pitfalls, and it is all too easy to fall into one of these and get very disappointing results. In this article we'll try to explain what the pitfalls are, and how to avoid them. So if you're keen to attempt the conversion, read on.

Before we proceed, however, a word of warning. It may not be possible to obtain good stereo signals from sone low-cost FM portable receivers. This is because satisfactory stereo decoding can only take place if the "difference" signal components in the transmitted stereo multiplex signal are reproduced faithfully at the output of the receiver detector – particularly in terms of phase, relative to the 19kHz pilot tone.

While the IF response of a low-cost portable may be sufficient for quite good mono reception, its amplitude and phase response may be such that the difference signal components of a stereo signal become too distorted for satisfactory decoding. It is possible to correct for a modest amount of phase shifting using a correction network, as we will explain, but even this may not give acceptable results with some sets.

At this stage, neither we nor anyone with whom we have discussed the problem have had sufficient experience in converting receivers to be able to state the brands and models which are not worth tackling. We can't even tell you what proportion of sets may be in this category. All we can do is warn you that the possibility exists – and hope that you aren't unlucky enough to meet it.

Broadly speaking, just about any high quality FM tuner should be capable of being converted. The same should apply to the better quality portables. It is only the "bottom of the range" types which



FM STEREO DECODER

are likely to prove disappointing, although even there our experience suggests that some sets can give quite acceptable results when "tweaked".

To begin, then, let us look briefly at the transmitted stereo multiplex signal – for unless you understand how the signal is made up, it won't be easy for you to follow how it is decoded.

To make stereo signals compatible, or capable of being received by mono equipment (in mono), they are transmitted not as the original "left" (L) and "right" (R) signals, but as two composite signals. One of these composite signals is made by adding the L and R signals together and dividing by two, to produce a "sum" or "mono" (M) signal. This is used to directly frequency modulate the station's carrier, and it is this component – only – which is detected by a mono tuner or receiver.

The second composite signal transmitted is the "difference" or "stereo" (S) signal, formed by subtracting the R signal from the L signal, and again dividing by two. It is basically this signal which is used by stereo receiving equipment to separate out the two original signals.

To keep it distinguishable from the M signal, the S signal is shifted in frequency by using it to amplitude modulate a 38kHz supersonic subcarrier. Then, to prevent the subcarrier energy from restricting the actual signal energy which could be carried in the final transmission sidebands, the subcarrier itself is suppressed. This leaves the S signal in the form of a suppressed-carrier double sidebonded components extending from 23 to 53kHz, and it is in this form that it is used to frequency modulate the stations' carrier along with the M signal.

To allow the receiving end to make use of the S signal components centred on 38kHz, in the absence of the 38kHz subcarrier, a low-level "pilot tone" signal is also transmitted. This is a continuous tone of 19kHz, derived from the subcarrier oscillator and therefore phase-locked to it.

The final stereo multiplex signal thus normally consists of the M signal, the S signal components, and the 19kHz pilot tone. It is this 3-in-one signal which is potentially available at the detector of a mono FM tuner or receiver, and necessary for proper stereo decoding.

Early FM stereo decoders operated in the following way. First, they used filters to separate the three signal components: a low-pass filter for the M signal (0–15kKhz), a sharp resonant filter on 19kHz for the pilot tone, and a 23–53kHz band-pass filter for the S signal components. Then the 19kHz pilot tone was used to regenerate the 38kHz subcarrier, and this was then fed with the S signal components into a synchronous demodulator, to recover the actual S signal.

The M (sum) and S (difference) signals were then fed through phase splitters into a resistive adding matrix, where they were combined to produce the original L and R stereo signals.

While this was a perfectly valid decoding technique, it had a number of practical drawbacks. Probably the main drawback was the filters required for separating out the three multiplex signal components; these tended to use fairly expensive L-C tuned circuits, and be critical of adjustment.

The drawbacks are avoided in modern decoders, including the design to be described here, by using an alternative decoding technique. The alternative technique is based on the fact that together, the M signal and the S signal components of the multiplex signal are equivalent to the main components of a signal produced by alternately sampling the original L and R stereo signals, each at a rate of 38kHz.

Because of this equivalence, it becomes possible to decode the original L and R signals in virtually a single operation, by performing a time demultiplexing operation. Thus in a modern decoder, the M and S signal components are not separated, but are left together. The 19kHz pilot tone is merely extracted, and used to regenerate the 38kHz subcarrier or sampling signal. This is then used to drive an electronic switching circuit, which alternately switches the composite M-plus-S signal between two output circuits. The bursts of signal arriving at the two output circuits are then simply filtered and de-emphasised, to produce the original L and R stereo signals.

At the heart of most modern decoders using this technique is a single IC, which performs virtually all of the operations just described. Not only this, but the IC usually performs automatic switching between stereo and mono modes, and drives a "stereo" indicator lamp into the bargain.

The particular IC used in our decoder design is the type XR-1310, made by Exar Integrated Systems of Sunnyvale, California, It is typical of the latest generation of these devices. Motorola make an equivalent device the MC1310, which may also be used.

The XR-1310 uses a phase-locked loop (PLL) system to regenerate the 38kHz subcarrier. The PLL uses an R-C oscillator, whose natural frequency is set simply by means of a potentiometer. This becomes the only "tuning" control in the decoder, and it is easily set up correctly on signal.

For good channel separation, it is necessary for the decoder to switch the signal between the L and R outputs with an accurate 1:1 mark space ratio. This means that the regenerated 38kHz subcarrier should be an accurate square wave. For accurate phase locking it is also desirable that the 19kHz feedback component which is compared with the pilot tone in the PLL comparator should also be an accurate square wave.

To allow these requirements to be satisfied, the XR-1310 actually runs the PLL oscillator at 76kHz, twice the subcarrier frequency. This is then divided by two twice using flip-flops, to obtain accurate square waves at 38 and 19kHz.

Three flip-flops are used in all, one to produce the 38kHz sub-carrier switching waveform and two to produce two separate 19kHz signals. One of these is used in the PLL comparator, to lock the loop with the pilot tone; the other is used in a second comparator whose output is used for automatic mono-stereo switching.

Two different 19kHz components are needed because a PLL normally locks with its feedback signal shifted 90 degrees from the input signal. To make sure that the 38kHz switching signal derived from the PLL is in phase with the original subcarrier, it is therefore necessary to use a 19kHz feedback signal which is shifted 90 degrees in the opposite direction, so that the phase shifts cancel.

On the other hand the comparator used for mono-stereo switching must be fed with an in-phase 19kHz signal, because the comparator is used purely to indicate coincidence between the internal 19kHz signal and the incoming pilot tone. Hence the use of two divider flip-flops, to generate the 90-degree shifted in-phase 19kHz signals.

A block diagram showing the various sections within the XR-1310 and the way they are interconnected, is shown in Fig.1.

When mono signals are being received, there is no pilot tone present in the signal fed to the XR-1310, so that the PLL loop cannot lock; the 76kHz VCO accordingly free runs. As a result there will be no coincidence registered by the in-phase 19kHz comparator, and the stereo trigger and switching circuit will remain off.

In this situation a single 38kHz switching signal is fed to the decoding switch, and this causes the switch circuitry to feed the mono input signal to both L and R outputs continuously.

When a stereo signal is being received, the presence of the 19kHz pilot tone at the input of the PLL phase comparator causes an error signal to be produced, and the PLL accordingly locks in. This causes the in-phase 19kHz signal to pull into phase coincidence with the pilot tone, so that the in-phase comparator produces an output to operate the trigger. And this in turn operates the stereo switch,

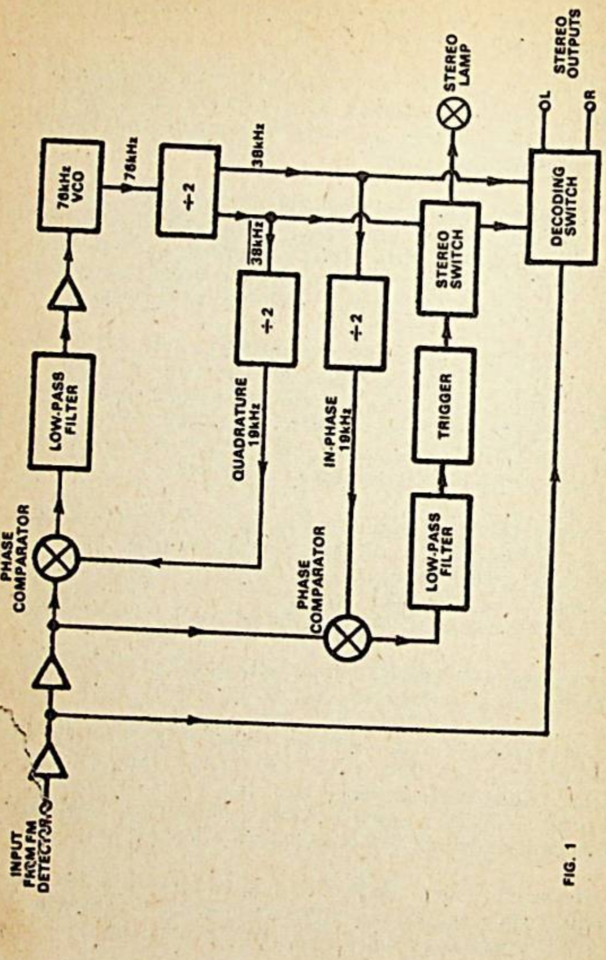

INPUT FROM FM DETECTOR
PHASE COMPARATOR
LOW-PASS FILTER
78kHz VCO
78kHz
÷2
38kHz
$\overline{38kHz}$
÷2
QUADRATURE 19kHz
÷2
IN-PHASE 19kHz
PHASE COMPARATOR
LOW-PASS FILTER
TRIGGER
STEREO SWITCH
STEREO LAMP
DECODING SWITCH
L
R
STEREO OUTPUTS

FIG. 1

turning on the stereo lamp and allowing a second (antiphase) 38kHz switching signal to be fed to the decoding switch.

The effect of the antiphase 38kHz switching signal is to cause the decoding switch to toggle the multiplex input signal alternately between the L and R outputs, performing the desired demultiplexing action. All that is necessary to recover the original stereo signals is de-emphasis and filtering.

Note that the existing de-emphasis network in the tuner or receiver feeding the decoder must be removed, because if retained it would seriously attenuate and distort both the pilot tone and the S-signal components. The stereo L and R signals are de-emphasised after decoding, using a time-constant of 50us for each.

Because of the switching action of the decoding circuit, the L and R outputs from the XR-1310 tend to have a significant 38kHz ripple component, even after de-emphasis, and despite internal rejection. While this generally causes no problems for direct listening, it can cause trouble when the signals are used for tape recording; a heterodyne tends to occur, due to interaction with the recorder bias oscillator. For this reason it is desirable for a decoder circuit to use additional filtering, to reduce the 38kHz ripple to a suitably low value.

The decoder circuit which we have evolved using the XR-1310 does incorporate such filtering, along with a number of other features. These are visible in the main circuit diagram, where you can see that along with the XR-1310 device we have used three BC109 or similar low-noise NPN transistor, plus a handful of minor components.

One of the transistors is used as an adjustable-gain input preamp. This has been provided so that the decoder can be arranged to work correctly with a wide variety of tuners and receivers including those whose detectors deliver only a low output. The preamp gain can be adjusted from a minimum of about 5 times to a maximum of about 100 times, and as the XR-1310 will operate with an input from about 200mV – 2.5V peak to peak, this gives the decoder the ability to cope with detector output levels from about 2mV to 500mV P-P.

For tuners with a detector output greater than 500mV P-P, the gain of the preamp can be reduced to 2 times, merely by omitting the 100uF capacitor and 1k tab pot in the emitter circuit. And if the gain is still too great the preamp can always be omitted altogether.

The components wired around the XR1310, are basically those recommended by the manufacturer. The R-C circuit connected to pin 14 is the VCO timing circuit, with the 5k pot for tuning. This is simply adjusted using an off-air stereo signal, as we will describe shortly.

The components connected between pins 12 and 13 form the PLL low-pass filter. Similarly the capacitor between pins 8 and 9, together with internal resistors, forms the low-pass filter for the in-phase comparator. The capacitor from pin 3 to pin 11 couples the 19kHz pilot tone from the input preamp to the two phase comparators. Pin 10 is an output from the quadrature 19kHz divider, to allow monitoring.

Pin 6 is the collector of the stereo lamp driver stage, and as can be seen, the LED used for stereo indication is connected in series with a 470-ohm resistor between this pin and the positive supply rail. A 40mA or 100mA incandescent lamp bezel may be wired in place of the LED and resistor, if desired, but no more than 100mA may be drawn through pin 6.

Pins 4 and 5 are the L and R outputs of the decoder, respectively. The R-C circuits from each pin to the positive rail are the de-emphasis timeconstants. The values have been chosen to give very close to the 50us timeconstant specified in the British FM stereo standards.

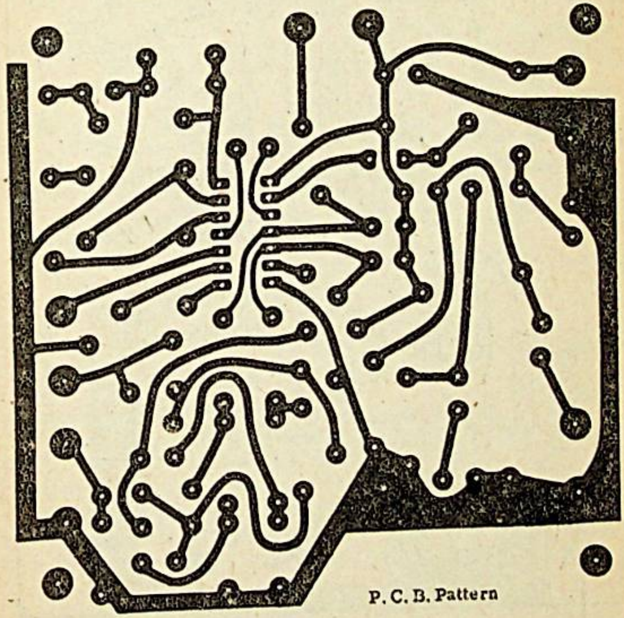
Following the XR-1310 decoder circuitry are the two remaining transistors, wired as simple low-pass active filters. These are for suppression of the 38kHz switching components in the decoder output, to prevent heterodynes and similar troubles when tape recording You can leave these stages out if you wish to save money and are certain that you will never want to record the decoder output signals, but only they involve a small component of the overall cost of the project. We therefore suggest that you wire them in, just in case.

Physically the decoder unit is built up on a small PC board, so that it can be either mounted inside your existing tuner or receiver, or alternatively housed in a small box of its own.

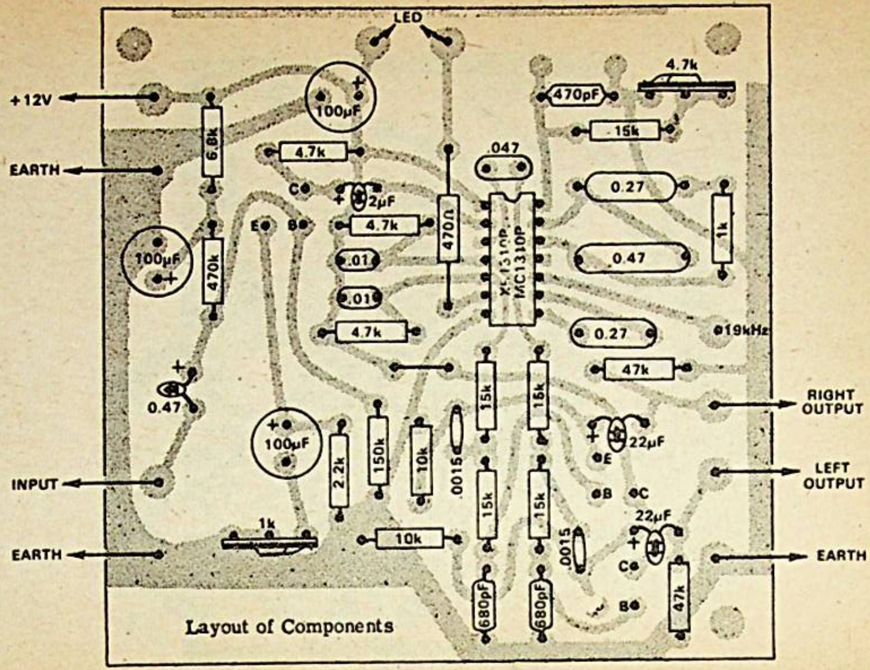
The board measures 91mm square and we have reproduced the pattern as shown, which of course, has to be enlarged to the correct size. Also shown is a diagram showing the position of all the components.

The main things to note when wiring up the PC board are that the IC, the transistors and the various polarised capacitors are correctly orientated. As you can see, there are not many components involved, and wiring the unit should be a simple job.

With the decoder wired up, you are ready to connect it up to the FM receiver or tuner, and get it going. The first thing to do is organise a suitable source of power supply. The decoder needs around 12V DC, fairly well filtered; its drain is around 40-45mA, when receiving stereo signals (i.e., with the LED "on").



P. C. B. Pattern



Layout of Components

If a suitable power source is available in the tuner or receiver, use this by all means. Otherwise, you may need to build up a simple power supply using a 12V stepdown transformer, a bridge rectifier using four EM401 or similar silicon diodes, and a 470uF/25VW electrolytic as the main reservoir. A 100-ohm 1W resistor should be connected in series with such a supply, for filtering.

An alternative approach would be to run the decoder from a 9V battery. This is quite in order, but the load resistors connected to pins 4 and 5 of the decoder 1C will need to be reduced in value, from 4.7k to 2.7k. The shunt capacitors will also need to be changed, to preserve the de-emphasis timeconstants; increase them from 0.01uF to 0.022uF.

Having fixed up a source of power, the next thing to do is find the detector circuit of the tuner or receiver. The type of detector circuit employed will naturally vary, according to the vintage of the design; as a result, possibly the best plan is to work backwards from the audio output connector, in the case of a tuner, or from the volume control in the case of a portable receiver.

There are two things to be done, when you find the detector circuit. The first is to remove the original de-emphasis capacitor, so that the 19kHz pilot tone and 23-53kHz S-signal components become available for decoding.

Broadly speaking, you should find that there will be a series R – shunt C filter circuit between the FM detector output and the audio output jack or volume control. These will almost certainly be the de-emphasis components. Their actual values will vary, depending upon the impedance level, but their product in ohms times microfarads or or kilohms times nanofarads will generally turn out to be around 75 (corresponding to a 75 microsecond timeconstant).

Having identified the two components, clip out the capacitor. It will generally not be necessary to interfere with the resistor, unless it value is sufficiently high to produce appreciable attenuation with the 100k input impedance of the decoder – or appreciable phase distortion of the S-signal components due to cable capacitance. If in doubt, short it out!

The other thing to be done at this stage is to arrange for signal take-off, In the case of a tuner, this may mean nothing more than making up a suitable cable to connect the decoder input to the output connector of the tuner. With a receiver such as a portable, you may have to add a suitable connector. The exact arrangement you make will no doubt depend upon whether you intend building the decoder into the tuner or receiver case, or having it as an outboard unit.

Similarly you will need to connect up the stereo outputs of the decoder to your amplifier system, using twin-shielded stereo cable and a suitable plug or plugs. Generally speaking, the signals are of an amplitude and impedance level suitable for a normal "radio" or "tape" input, on most amplifiers and control units.

At this stage you can turn on the power to the various parts of the system, and tune the tuner or receiver to your local FM stereo station – we assume there is one, or you probably woundn't have bothered!

Don't be surprised if the sounds coming out of the speakers seem to be mono, and if the stereo LED doesn't light. You still have the two preset pots to adjust, so that if the decoder bursts into full life immediately, this will be good luck more than anything.

If you have an idea of the output level coming from your tuner or receiver, set the preamp gain pot (the one near the input) to an appropriate position. If the output is fairly high, as from an IC quad detector, turn the pot to near the minimum setting (full resistance); if fairly low, as from a ratio detector, advance it to say 1/3 its resistance, as a starting point.

Now turn the VCO tuning pot to one extreme, and slowly turn it back towards the other extreme, watching the indicator LED. Note where the LED comes on, indicating that lock has been achieved. Then continue turning, and the LED should extinguish again before the pot reaches the far extreme. Then reverse the procedure, turning back slowly and again noting where the LED comes on.

The correct setting for the pot is midway between the two settings where it comes "on" in each direction.

With this pot set, the LED should remain on continuously, and you should be aware that the program from the loudspeakers is in stereo rather than mono. But more about this in a moment. Strictly you should now adjust the preamp gain more accurately, to the setting between signal-to-noise ratio and distortion.

The idea is to adjust the preamp gain so that the decoder is handling the largest signal level possible, without running into significant distortion on signal peaks.

Probably the best way to do this is using the off-air signal from your local stereo station, waiting until it is broadcasting a typical loud passage. You can either use an electronic voltmeter to monitor the signal level at the input to the decoder IC (pin 2), setting it to about 2.5V P-P, or use an oscilloscope to monitor the L and R outputs of the decoder board and adjust the preamp gain until the signal is just short of the onset of peak distortion.

If you lack both an electronic voltmeter and an oscilloscope, the best idea would be to make the adjustment by ear. First turn up the pot until audible distortion is evident on signal peaks, then back it off until the distortion is no longer evident. This should be a good starting point, although you can modify it after further listening, if necessary.

If all is well, you should now be able to produce clean, well-separated stereo signals. But as we noted at the beginning of this article, this will depend very much on your mono tuner or receiver. If the IF response of this is not up to scratch, the results may be rather disappointing.

By the way, don't jump to the conclusion that all is well, simply because you were able to do the previous adjustments, and the decoder now lights up as it should whenever you are tuned to a stereo station. All this means is that the decoder is detecting the 19kHz pilot tone, is locking onto it, and attempting to decode the rest of the signal. If the signal reaching the decoder from the receiver or tuner is inadequate, you will won't be getting good stereo.

The best way to check if the system is really working well is to measure the actual channel separation being achieved, using an FM stereo signal generator. Failing this, though, another approach is to look at the two stereo outputs of the decoder using an X-Y presentation on an oscilloscope.

If you are able to do this, the pattern you get will give you a good idea of how well the system is working. You can interpret the pattern using Fig.2 as a guide. If the signal is being very badly distorted by the receiver or tuner, you may get little more than the 45-degree mono signal line in (a). On the other hand if the system is working well, and there is good stereo separation, the pattern will tend to expand into a complex 2-dimensional pattern like that in (c). A modest amount of distortion will produce an intermediate pattern, rather like (b).

It is best to monitor the signals for a while with this technique, because you may be misled by the station broadcasting a mono record, or one with very poor separation to start with!

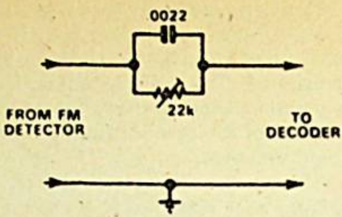


FIG 3

Fig. 3 (above): A suggested phase correction circuit for poor receivers. At left is the PC pattern, actual size.

Using the network of Fig.3 we were able to improve the separation obtained with a low-cost portable from a disappointing 6dB (!) to 30dB, so that it is capable of making quite a difference.

One final way of getting a minor improvement in separation is to connect a capacitor of 0.0022uF from pin 3 of the XR-1310 to ground. This corrects for a minor phase shift in the device, and in practice seems to be capable of giving an improvement of up to 5dB.

Well, there it is. All that remains is to wish you happy stereo listening!

IMPORTANT NOTE:–
For American and certain continental "Pilot Tone System" stereo multiplexing, the stereo L and R signals are de-emphasised after decoding, using a time constant of 75 usecs and not 50 usecs as used in the UK. To modify this time constant alterations in the RC network are required, the 4.7k resistor should be replaced by a 3.9k resistor and the .01 uF capacitor should be replaced by a .022 uF capacitor.

PARTS LIST FOR THE STEREO DECODER

1 board, 91mm square
1 XR-1310 or MC-1310 stereo decoder 1C
3 BC109 RS276-2031 or similar low noise NPN transistors
1 led, general purpose type
1 5k preset pot
1 1k preset pot
RESISTORS, ¼watt 5%
470 ohms, 1k, 2.2k, 3 x 4.7k, 6.8k, 2 x 10k, 5 x 15k, 2 x 47k, 150k, 470k.
CAPACITORS
1 470pF polystyrene
2 680pF polystyrene
2 1500pF polyester
2 0.01uF LV greencap
1 0.047uF LV greencap
2 0.27uF LV greencap
2 0.47uF LV greencap
1 2uF 6VW tantalum
2 22uF 6VW tantalum
1 100uF 6VW electro, single ended
2 100uF 16VW electro, single ended
MISCELLANEOUS
PC board stakes, mounting screws, connecting wire, solder, etc.

THREE CHANNEL STEREO MIXER

This describes a three channel stereo mixer which should appeal to many readers by reason of both its simplicity and its low cost. It should also provide an interesting constructional exercise for the beginner.

Mixer circuits appearing in the last few years have, generally, been complex, filling the requirements of pop-groups using very elaborate recording equipment. The cost of building a mixer using complex matching techniques is high in relation to the use such a mixer would have if it is being used only with domestic equipment, record players and cassette recorders.

The author's requirements were simple: a low-cost stereo mixer capable of handling three channels, with matching facilities for most portable equipment. It was to provide music and spoken accompaniment to a slide show, with the third channel to add a tone signal for the projectionist, indicating a change of slide. For club use, or for sending slides and a cassette to relatives abroad, a mixer adds a touch of professionalism difficult to obtain by any other means.

Before construction began it became obvious that potentiometers were the largest single cost, followed closely by sockets and matching circuitry.

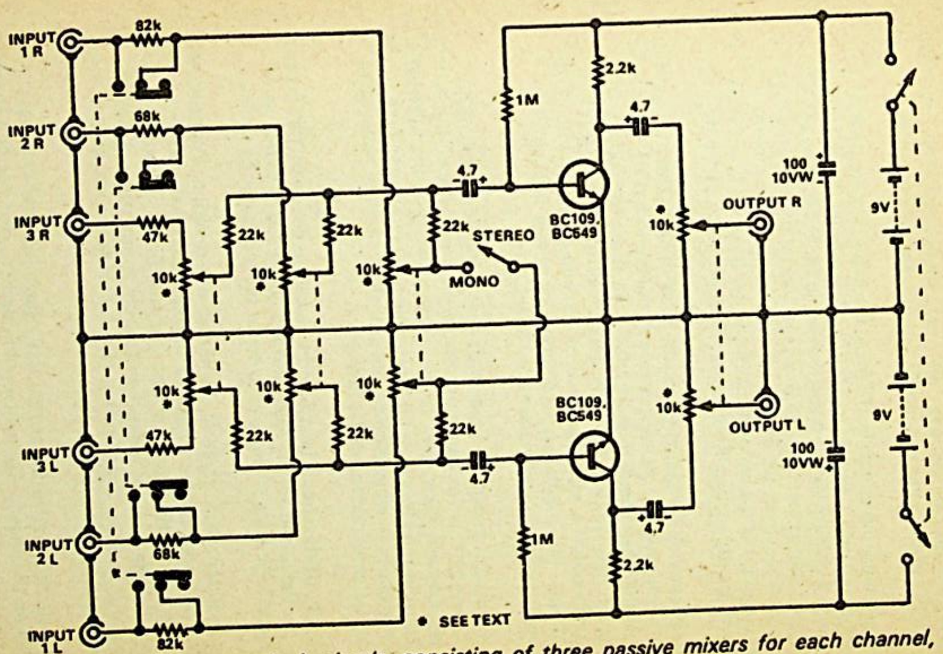
Sockets were limited to RCA type and 3.5mm jacks although DIN could also be added. Matching circuitry was limited to resistors switched in or out to give approximately the correct matching value.

It is possible to construct a passive mixer using potentiometers alone but two problems arise. Firstly, there is considerable interaction between the wipers of the potentiometers. This was overcome by adding a 22k resistor in series with the potentiometer wiper but introduced a second problem – significant signal loss – this was overcome by adding a very basic amplifier with a gain of about 20.

The amplifier uses the BC109 or equivalent. Whilst the circuit is not "hi-fi" the amount of distortion added is minimal and is unlikely to be greater than the inherent distortion in most cassette recorders.

The output could be taken directly from the 4uF capacitor at the collector but, in the prototype, a 10k output level control was added to function as a master control and give increased versatility to the unit.

There are three chassis on the prototype: one each for the input sockets, the circuit board and the output sockets. These were made from scrap aluminium, although the entire unit could be mounted in an aluminium box. But if the layout for the circuit board illustrated is used, the measurements given, particularly those for the potentiometer shafts, must be strictly adhered to.



The circuit is comparatively simple, consisting of three passive mixers for each channel, follower by an amplifier and a master volume control.

Cut and drill the chassis as indicated, drilling small pilot holes before drilling the 6mm (¼") holes for the shafts. Score the line where the chassis is to be folded several times with a tungsten carbide tipped marking pen or a sharp awl, then fold the chassis. When shaping the chassis fold a section through only a very small angle at any one time as aluminium stretches very easily and may become distorted.

Before drilling the circuit board mark the holes that are used by the trimpots and drill a slightly larger hole. The correct size is 1mm. Although a one-off circuit board was made for the prototype, Veroboard is also suitable as the hole size is usually correct for trimpots. If using Veroboard construct the mixer with the copper strips running in the same direction as the trimpot shafts and use links for the "earthy" side. This will greatly reduce the number of copper strips that have to be cut.

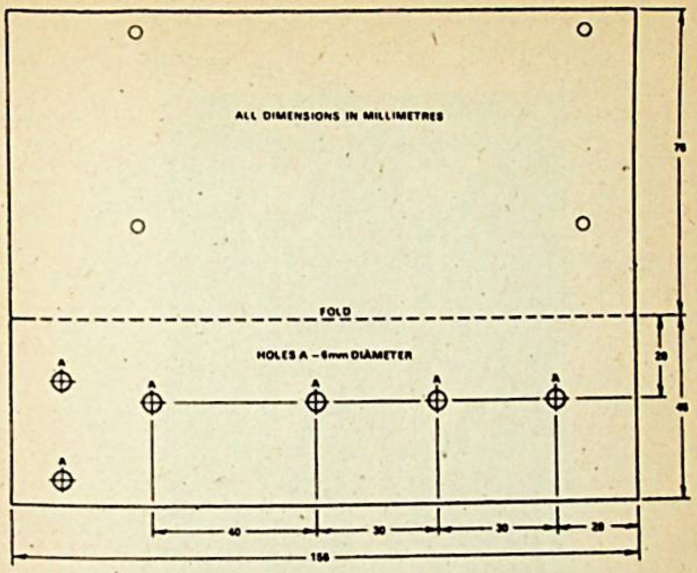
When assembling components, mount all of the pots first. Do not fold the mounting lugs on the copper side of the board but trim them very short so that they stand no more than 1mm above the copper. At a later stage the wiper arm has to be eased slightly away from the resistance strip and, if a wiper is moved too far and goes open circuit, the pot will have to be replaced. So it should be as easy as possible to remove from a completely assembled board.

After the circuit board has been assembled fit the input and output sockets and switches to their respective chassis and wire up the matching resistors. Connect the sockets to the board with the thinnest available shielded cable, and be aware that the cable must be long enough to rise above the shafts when assembled and also allow room to manoeuvre the small interconnecting shafts into place.

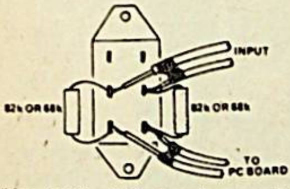
Temporally fit the circuit board to the chassis to determine the correct length for the power supply leads and the mono/stereo switch leads. After these have been soldered into place, the mixer can be tested.

With a small screwdriver, set all the pots to mid-position. Connect the batteries, and connect the output to an amplifier. Inject a known mono signal into each of the inputs in turn to ensure that a variation in component values has not cut a signal excessively in any channel. Switch the matching resistors in and out of circuit during the test as they too should be balanced.

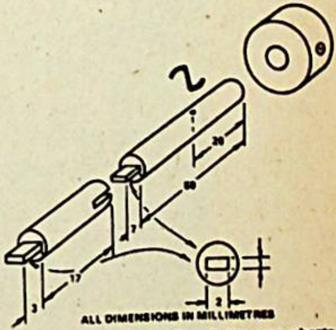
With the circuitry operating satisfactorily the board and switches can be mounted on the chassis. Connect an ohmmeter to one side of a trimpot and to the wiper. Very gently prise the wiper away from the resistance strip taking great care not to damage the carbon surface. Continually check the operation of the trimpot with a small screwdriver until a noticeable lowering of the friction has taken place but the wiper has not open circuited. A proprietary potentiometer lubricating spray is also a great aid in easing the operation of the pots. It will also assist in preventing noise from the pots.



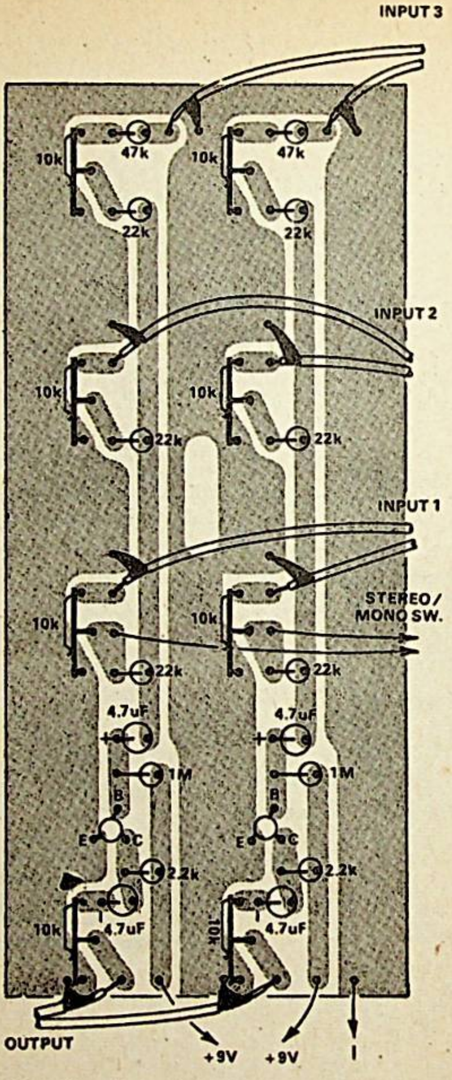
Drawing for the main chassis and panel. Panel holes must line up with the pot shafts.



Wiring details of the impedance switch.



Dimensions and details of the pot sharts.



Printed wiring board, viewed from the component side. Pattern is full size.

The pot shafts can now be made. Cut the knitting needles with a fine-toothed hacksaw to the overall lengths given in the drawing, 20mm and 57mm, and mark the area to be trimmed with a pen, preferably a solvent based pen.

The length to be trimmed is 3mm on the short shaft and 7mm on the longer one. Clamp the end of the shaft that is not to be trimmed between two pieces of wood and cut the 1 x 2mm rectangular section with a metal file or with the flat face of a fine grinding wheel mounted in a power drill. Check the dimensions frequently when grinding.

Cut the slot in the end of the short shaft with a hacksaw blade which, conveniently, should give a cut 1mm wide. Make sure that the cut aligns with the flat section filed on the other end of the shaft. The accuracy of this cut determines the stereo balance of the mixer.

The shafts can now be fitted to the mixer. Place the front shaft through the chassis hole and rest it in the slot of the front pot. The second shaft can now be fitted, preferably with both trimpots at minimum level setting, and held in place with tweezers as the front one is pushed in and slotted into the recess on the rear shaft. Slip a 10-15mm length of wire through the small hole drilled in the shaft to prevent it from being pulled out of the mixer.

After the operation of the shaft has been checked a small dab of fast-setting cement can be applied to the joint at the rear of the front pot. If the mixer is to receive a great deal of use, and after the operation of the mixer has again beech checked, more glue can be applied to the shafts where they butt on to the pots. Take care not to allow glue on any moving surface.

Once completed the mixer will invariably find uses for itself. The author makes cassettes for parties or cars using two turntables. This eliminates the pause between tracks on a record and offers the possibility of some variation in music rather than twenty minutes of one composer.

Although two batteries have been used, a battery drain rarely exceeds 2.5mA so their life should be quite long.

PARTS LIST

RESISTORS
2 2.2k 5% 1/8th W
6 22k 5% 1/8th W
2 1M 5% 1/8th W
2 47k 5% 1/8th W
2 68k 5% 1/8th W
2 82k 5% 1/8th W
8 10k potentiometers

SWITCHES
1 SPDT, 1 DPDT, 2 DPDT slider

CAPACITORS
2 100uF 16VW *4 4.7uF 10VW electrolytics*

SEMICONDUCTORS
2 BC549, BC109, RS276-2009

MISCELLANEOUS
One pair No 4 knitting needles.
Aluminium for chassis.
Veroboard or printed wiring board.

FET PREAMPLIFIER FOR CERAMIC PICKUPS

Here is an economical preamplifier for use with low-output ceramic cartridges. The following article discusses the principal of operation and the various ways in which the preamplifier can be used.

Many readers have asked for an article describing a simple, add-on preamplifier. This would enable lower-output ceramic cartridges to be used with amplifiers which previously were suitable for use with crystal cartridges only. This is a common situation where people have one of the earlier stereo amplifiers, designed before ceramic cartridges became available.

A preamplifier is also often desirable when modifying a mono record player to suit stereo records. Instead of merely substituting a high output crystal stereo cartridge with the two channels paralleled, it is better to add a preamplifier and use a better-quality ceramic cartridge.

As well as requiring a greater amplifier sensitivity, the substitution of a ceramic cartridge requires a higher input impedance than does a crystal type. Typically, most ceramic cartridges require a load of 2 megohms and in the case of the Decca Deram, a sensitivity of around 60mV RMS for full power from the amplifier.

There are several possible approaches to providing the necessary gain and high input impedance using a single transistor (one for each channel). One would be to use a high beta, low-noise bipolar transistor in a boot-strapped common emitter amplifier. (Beta is a measure of the direct current gain of a transistor and is approximately equal to the ratio of collector current to base current.) Bootstrapping refers to the technique of applying positive feedback—with less than unity gain—from the emitter to increase the effective input resistance provided by the biasing resistor network.

While a bootstrapped input stage could be arranged to give the required high input impedance the gain may not be sufficient, while the noise generated in the biasing resistors can be a real problem. The use of high-quality carbon film resistors will not always alleviate the problem as the noise generated in the resistors is regenerated by the positive feedback.

Another approach is to use one of the very high beta transistors now available, without bootstrapping, and rely on biasing resistors with values up to 10 megohms to obtain the high input impedance. Noise may still be a problem, though to a much lesser extent, such that the use of high quality carbon film resistors will keep it within acceptable limits. Unfortunately, this type of resistor is not always readily available over the counter, particularly in the high values required. Another problem is that transistors with a minimum guaranteed beta of say 500, tend to be expensive and often in short supply.

The approach we have taken is to use a field effect transistor, an N-channel device made by Motorola, the 2N5459 which supersedes the MPF106. This FET is available economically and its parameters are more closely controlled than the first economy FETs. A major advantage of using a FET circuit is that the input impedance required is obtained simply by "plugging-in" the desired value of resistor. Noise generated in the input resistor is not a problem since no gate current flows (under small signal conditions).

Reference to the circuit diagram will show that the configuration is very similar to that used in triode amplifier circuits which employ "cathode-bias." Indeed, the principal of operation is very similar. We will explain it for the benefit of our novice readers.

The gate bias voltage – the voltage between gate and source – is used to set the correct operating conditions for the FET, such as to provide the most linear amplification for the supply voltage used.

For N-channel FETs, the gate is required to have a negative voltage with respect to the source. This voltage is generated by the current flowing through a resistor connected between source and the negative supply rail, making the source positive with respect to the negative supply rail. The 2.2M resistor for the gate carries no current and thus the gate is at the same potential as the negative supply rail. This means that the gate is negative with respect to the source. Bias developed in this manner is known as "source bias" which is analogous to "cathode bias" in valve circuits.

Source bias for P-channel FETs is obtained by the same method as described above except that the gate voltage is positive with respect to the source.

As with valve circuits, the source resistor must be "bypassed" with a suitable value of capacitor in order that the maximum voltage gain can be realised. "Bypassing" refers to the practice of providing a low impedance path for AC signals, so that they do not develop an AC voltage across a resistor which is used for deriving a DC voltage. If the source resistor was not bypassed, the audio signal fed to the gate of the FET would reappear across the source resistor, reversed in phase. Thus the mechanism by which the source bias is developed would apply the signal in reverse phase between the gate and source (negative feedback) and the gain would be reduced. The value of bypass capacitor selected must be such that its impedance is low for the lowest frequency to be handled.

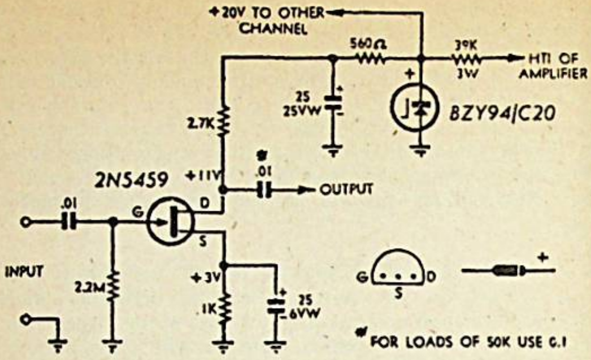
The preamplifier uses a supply rail of the order of 18 to 21 volts and can be run directly from an 18-volt battery. The 18-volt supply is necessary to ensure minimum variation in gain over the likely range of parameters of this FET and to give a high margin of overload with the expected range of input signals. The gain of the preamplifier will be between 5 and 6 times, which makes it ideal for augmenting the gain of amplifiers previously suitable for use with crystal cartridges only. The preamplifier will overload with an input signal of approximately 700mV RMS, although this will vary with the gain and the supply voltage.

The above order of overload capability is highly desirable as today's heavily recorded discs can result in the cartridge delivering a much higher output than its nominal output voltage would suggest. The preamplifier is suitable for ceramic cartridges with a nominal output voltage up to about 200mV or so. The BSR CI and equivalent cartridges in the Sonotone range are eminently suitable, as are the lower output ceramic cartridges such as the Decca Deram and Connoisseur. For the latter cartridges, the preamplifier may not have sufficient gain to drive some amplifiers to full power, although in most cases it should be adequate.

The circuit diagram shows a zener diode network to derive the pre-amplifier supply from the main supply rail (HTI) of valve amplifiers. The zener diode is necessary to protect the FET from the higher-than-usual voltages occurring in valve equipment just after switch-on before the valves begin to draw current. This characteristic is particularly noticable in valve amplifiers which have semiconductor rectifiers. The zener diode also has the advantage of rendering the decoupling network compatible with a wide range of likely HT supplies.

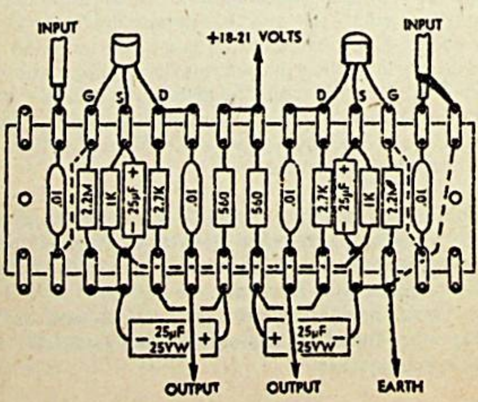
In the prototype the supply was derived from that of the amplifier via the 560-ohm resistor and 25uF capacitor shown on the circuit diagram. Here, the decoupling network is used to filter hum appearing on the amplifier's supply rail, and to eliminate the possibility of instability due to the increased overall gain of the amplifier plus preamp. If an 18-volt battery is used, the decoupling components can be dispensed with and the unit could be installed in a small metal box underneath the turntable. If this is done, care should be taken in the positioning of the box so that it does not pick up hum from the turntable motor or associated wiring.

The preamplifier was constructed on a 14-lug tagboard. If the zener diode and 39K resistor are required an extra two lugs will be needed. If only one channel if required the current drain of the zener diode network can be reduced by increasing the resistor to 82K. The layout we have used is not mandatory but it has been arranged so that the inputs are at either end of the tagboard to keep cross-talk between channels as low as possible.



the circuit diagram of the preamplifier. If only one channel is built, the 39K resistor can be increased to 82K to reduce current drain of the zener diode network. If used with a transistor amplifier the zener network can be omitted.

a suggested layout, using tagboard. The inputs are placed at either end of the board to keep the crosstalk between channels to a minimum. If the preamplifier is built to suit a valve amplifier, a longer piece of tagboard will be needed to accommodate the zener diode network.



Shielded cable should be used for the inputs, with the shields connected as shown on the wiring diagram. If the unit is installed in a separate metal box as mentioned above, shielded cable should also be used for the outputs.

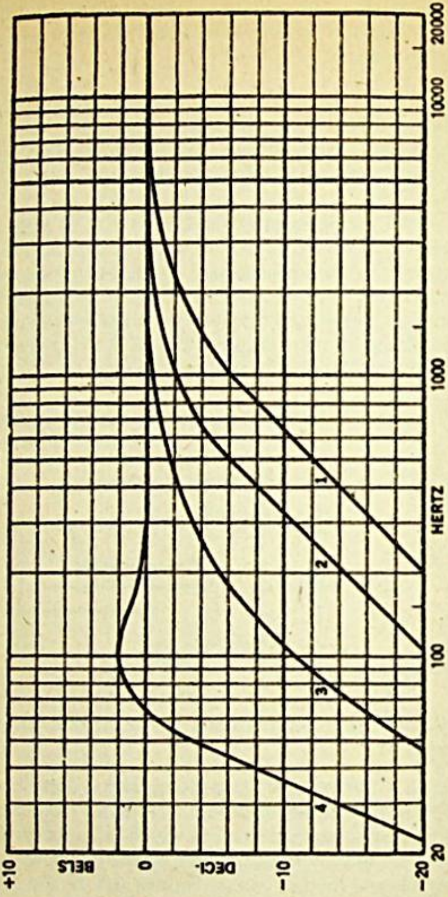
PARTS LIST

1 BZY94/C20 zener diode.
2 ZN5459 n-channel FETs.
1 14-tag panel.

RESISTORS
(½ or ¼ watt unless specified)
2 x 2.2M, 1 x 39K/3W, 2 x 2.7K, 2 x 1K, 2 x 560 ohms.

CAPACITORS
2 x 25uF/25VW electrolytic
2 x 25uF/6VW electrolytic
4 x ,01uF polyester (low voltage rating)

FIG. 3



Bass response curves for each of the feedback capacitor switch positions.

The Veroboard measures 2½ x 2⅛ in and is easily assembled. Lockfit transistors may be used is desired.

The preamp should be adequately shielded against hum and strong RF fields. If it is a self-contained unit with its own battery, it should be mounted in a metal case which is connected to the negative supply line. If the preamplifier is powered from an amplifier supply rail but is separately housed the DC return path could be via the shield of the output cable, if need be. This means that a figure-8 shielded cable or two wires with a common shield could be used instead of separate cables to the amplifier.

A modification can be made to the feedback loop to modify the bass response of the microphone. One of the problems encountered by "pop" singers is that the apparent bass response of a microphone changes depending on whether it is held very close to the mouth or further away. This has the effect of muddying the singer's voice when he is singing with a close mike. The problem is much reduced if the bass response of the preamplifier is rolled off below about 500Hz and this achieved by decreasing the size of the 10uF capacitor shown on the main circuit diagram.

Figure 2 is a switching arrangement of different capacitors which replace the single 10uF capacitor in the feedback loop. In the first position of the switch the feedback capacitor is 0.82uF and this corresponds to the maximum bass roll-off in the frequency response diagram, figure 3. The other three positions of the switch add capacitors in parallel to the 0.82uF to improve the bass response. Each frequency response plot corresponds to one of the switch positions.

Notice the 10K resistors associated with three of the capacitors on the

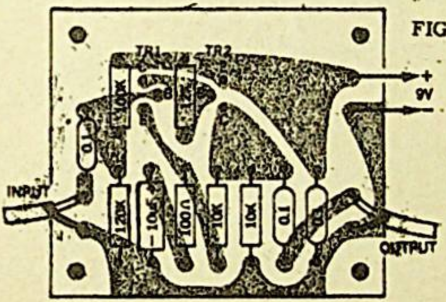


FIG. 4

switch bank. These maintain the capacitors at a DC potential equal to that across the 0.82uF capacitor and thus reduce switch clicks.

Two wires should be run from the feedback capacitor position on the board and the capacitors and resistors all mounted on the switch. Take care that the polarity of the electrolytic capacitors is correct.

Using the switch to modify the bass characteristic, the singer can choose the best position for good sound reproduction. With this arrangement, the singing can be piercingly clear instead of muddy and distorted.

PARTS LIST

1 printed board 2½ x 2⅛ in
1 silicon NPN transistor, BC109, BC149, or BC209
1 silicon PNP transistor, BC178, BC158, or 2N3638A

1 9V battery
1 SPST switch
1 shorting jack socket
3 0.1uF/100VW metallised polyester capacitors
1 10uF/6VW electrolytic capacitor
RESISTORS
(½ or ¼ watt rating)
1 x 120K. 1 x 100K. 1 x 12K. 2 x 10K.
1 x 100 ohms.

STEREO DYNAMIC NOISE FILTER

For many people, cassette decks have one big problem. During quiet passages hiss becomes obtrusive. The stereo dynamic noise filter presented provides a marked reduction in hiss while not affecting the bandwidth of high level signals.

SPECIFICATIONS

Frequency Response 30Hz to 100kHz between -1dB points with filters switched out.
Insertion loss: 1dB (0.9).
Input impedance: 70k.
Output impedance: less than 1k. Output load should be more than 10k for low distortion.
Distortion: Less than 0.1% up to 3V RMS input.
Separation between channels: Better than 40dB.
Input overload: 4V RMS.
Signal-to-noise Ratio: Better than 63dB with respect to 100mV out; Hum output: less than 74dB with respect to 100mV.

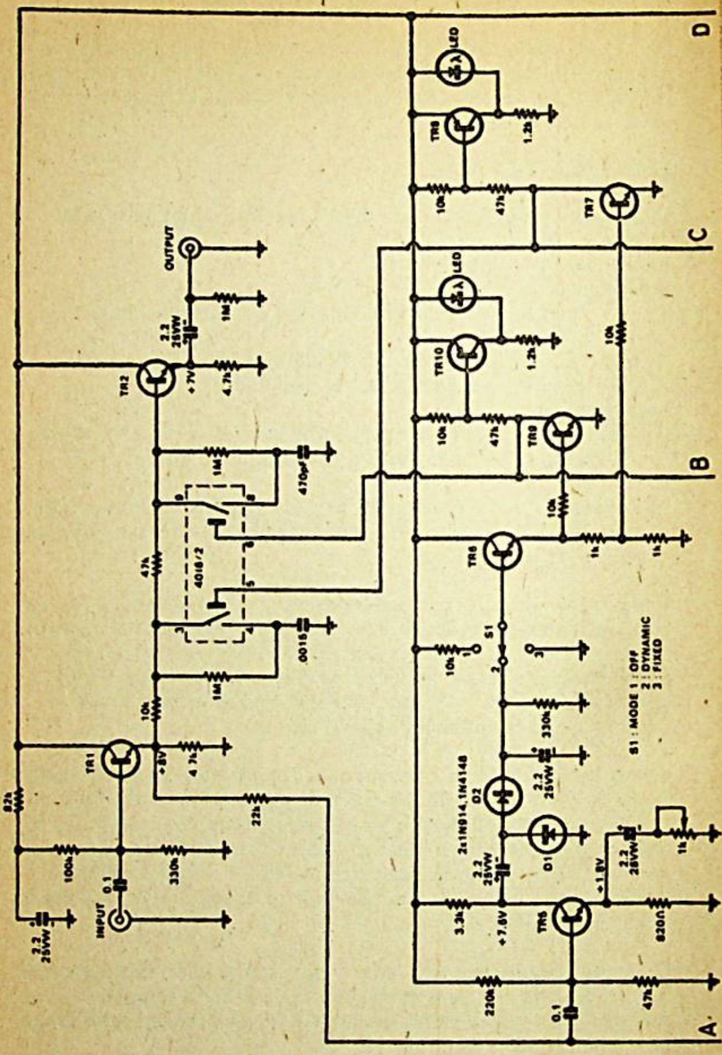
The dynamic filter unit described here is interposed in the signal line between a cassette deck playback terminals and the tape monitor inputs of typical stereo amplifiers.

Basically, it monitors the signal level in both channels and when the signal rises above a predetermined level, it progressively switches out two single-pole filters which otherwise roll off the response above 7kHz. Maximum rate of attenuation of high frequencies is 12dB/ octave. A CMOS quad bilateral switch is used to switch the filter sections in and out of circuit in both channels.

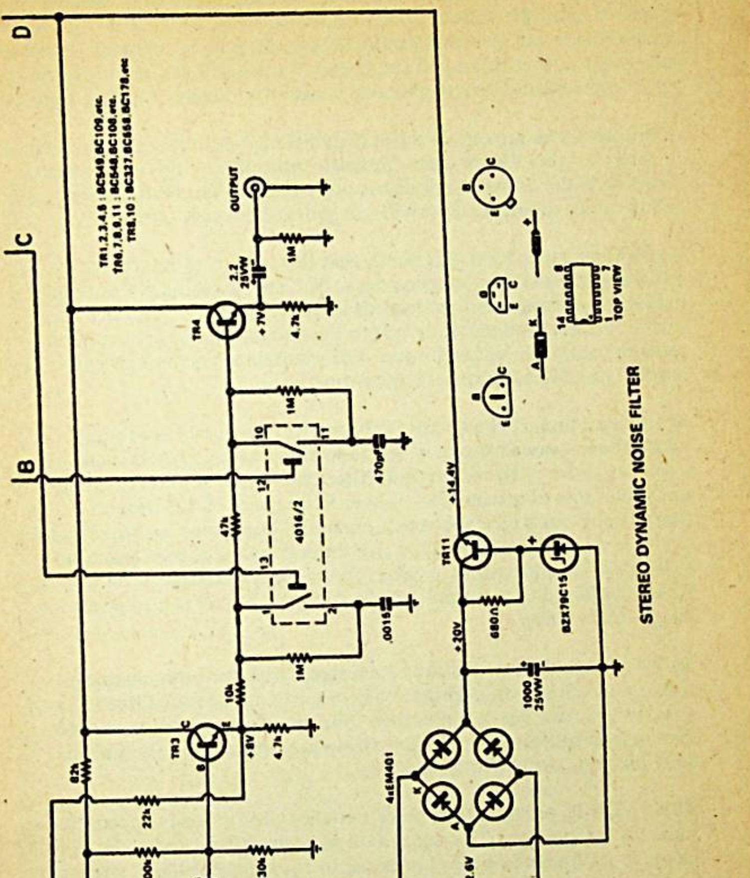
Let us state at the beginning that in our opinion most users of cassette decks do not need this unit. For a start, a major part of the population just are not aware of tape hiss as being a problem. If you are over 30 and/or cannot hear the normal 15,625Hz whistle of television flyback circuitry, or the somewhat more intense whistle of the switching regulator used in many colour television receivers, then you probably fall into this group.

Taking the comparison a little further, if you can hear no appreciable difference in quality between the average disc and a cassette copy of that disc, then you probably also do not have a problem as far as tape hiss is concerned.

For the few acute-eared readers, cats and dogs still interested in this article, let me state that I did not really want to alienate all those other readers. But at least they will not needlessly build this device!



Connects with drawing on opposite page



Connects with drawing on opposite page

Even for those of us with acute hearing tape hiss really only becomes obtrusive when the volume control is well advanced to give an appreciable sound level and particularly while playing softly recorded passages. At other times you can largely "tune out" the hiss using the brain/ear combination's remarkable capability to resolve a noisy signal.

Of course if you are one of those fussy listeners who tends to "tune in" the hiss, i.e. you always listen for faults rather than enjoy the music, then you really do need a dynamic noise filter of this kind. There is a subtle catch, though and we will talk more about that later.

We should also state, at this point, that this filter is of no use in recording noise or hiss on some discs. While we occasionally note that tape hiss is present on some records in our reviews it is very rare that this is obtrusive and certainly not to the same degree as present with cassette recorders. Nor is this unit really suitable for eliminating the residual hiss of Dolby cassette recordings.

Now have a look at the circuit. Tr1 (and Tr3) is an emitter-follower which feeds a cascaded pair of single-section low-pass filters each with a roll-off above 7kHz. When both filters are switched in, they give a maximum rate of attenuation of high frequencies of 12dB/octave. The first filter consists of a 10k resistor and .0022uF capacitor which results in a filter impedance of roughly five times less than the second filter (47k and 470pF) to avoid loading effects. Output signal of the second filter is buffered by a second emitter-follower, Tr2 (Tr4), to give a low output impedance.

So Tr1, 2, 3 and 4 provide the actual signal path for both channels, with an overall gain of slightly less than unity. Since four filter sections are involved, two per channel, then four switches are needed to place them in and out of circuit. The switches are provided by the CMOS quad bilateral switch IC, type 4016.

This CMOS IC provides four SPST switches which closely approach the ideal for our purpose. The signal path of each switch can handle signals of up to 15V peak-to-peak (i.e., equal to the supply voltage) with very low distortion and low cross-talk between other switches. The switches have very high OFF resistance and low ON resistance (typically 200 ohms), and all switches are closely matched. Isolation between control and controlled signals is extremely high, with resistance values in the region of 1 Tera-ohm being quoted. 1 Tera-ohm is 1 million megohms.

Finally, the feedthrough signal can be as high as 40MHz while toggle rates can be up to 10MHz. These figures add up to impressive performance.

There are two points about the circuit which should be noted at this stage of our discussion. First, Tr2 is biassed via the 10k and 47k signal path resistors from the emitter of Tr2 (similarly with Tr4 and Tr3) so that the signal swings symmetrically between the supply lines (ie, with reference of approximately half the supply voltage). Second, each CMOS bilateral switch is interposed between the signal path and the rolloff capacitors which are .0022uF and 470pF. Thus the signal applied to the CMOS switches swings symmetrically between the positive and zero supply lines, which is the condition for minimum distortion.

If the relative positions of CMOS switch and filter capacitor were transposed, the signal fed to the switch would swing symmetrically about the zero supply line. This would cause clipping of one side of the signal due to the protective diodes on the inputs of the CMOS switch.

Tr5 monitors signals from the emitters of Tr1 and Tr2 via 22k resistors. It acts to amplify these signals, with gain variable via the threshold potentiometer in its emitter circuit. Signals from the collector of Tr5 are rectified by D1 and D2 and filtered by a 2.2uF capacitor. The resulting voltage is fed to Tr6 which is merely another emitter-follower which reproduces at its emitter, the voltage fed to the base.

A voltage-divider in the emitter circuit of Tr6 feeds base current to Tr7 and Tr9 via 10k resistors. This means that Tr9 turns on before Tr7 as the voltage at the emitter of Tr6 increases.

The control pins of the 4016 are pins 5, 6, 12 and 13. Pins 5 and 13 are tied together to control the first filter section in both channels, while pins 6 and 12 are tied together to control the second filter sections in both channels. Pins 5 and 13 are switched by the collector of Tr7 while Tr9 controls pins 6 and 12.

At low signal levels, the resultant DC voltage at the emitter of Tr6 is low so that Tr7 and Tr9 are non-conducting. This means that all filter sections are in circuit and the high frequency response is attenuated. As the signal level rises, Tr9 conducts first and switches out the second filter section in both channels. Higher level signals allow Tr7 to conduct also, which switches out the remaining filter sections to achieve a flat frequency response. Reduction in signal level reverses this process.

Two LEDs are used to indicate the condition of the filters. When both filters are in circuit, both LEDs glow and so on. Thus the LEDs can be used to set the level of the threshold control and they also serve as an indication that power is applied at switch on.

So when Tr7 and Tr9 are non-conducting Tr8 and Tr10 are also non-conducting, which allows the LEDs to glow. When Tr7 and Tr9 conduct to switch out the filters Tr8 and Tr10 conduct also to extinguish the LEDs.

S1 is used as a Filter Mode switch. Position 2 gives "Dynamic" operation which is described above. Position 1 connects a 10k resistor to the base of Tr6 and so switches the filter sections out of circuit. Position 3 grounds the base of Tr6 to place the filter sections in circuit all the time regardless of signal level. Thus if the hiss level is not troublesome, the filters can be switched right out of circuit. Alternatively, if the hiss seems consistently bad, then the filters can be switched permanently into circuit.

Power supply requirements remain to be discussed. The 4016 has an absolute maximum supply voltage rating of 16 volts so we have arranged for this never to be exceeded by using a 15V zener diode and emitter follower Tr11. This results in a nominal supply voltage of 14.4V plus or minus 0.75V which is the normal zener tolerance.

While the combination of bridge rectifier, 1000uF/25VW filter capacitor and filter/regulator Tr11 result in a ripple voltage of only a few millivolts superimposed on the 14.4V supply rail, this is not quite low enough for hum free output. Accordingly, the bias networks for Tr1 and Tr2 are split with the 82k and 100k resistors and a 2.2uF/25VW capacitor bypasses ripple to the zero rail. This results in very low hum output, much lower than the total residual noise.

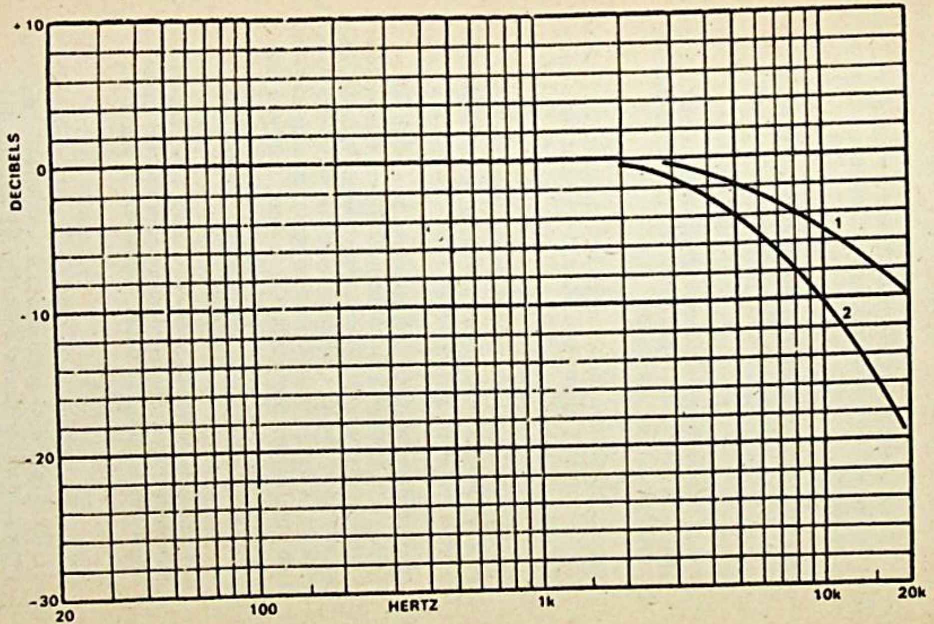
All the relevant performance details are shown in the specification panel and the filter response curves. Curve 1 shows the response with the first filter section switched in while curve 2 shows the resultant response with both filters switched in. The maximum rate of attenutation as shown in curve 2 is closer to 10dB/octave than the 12dB/octave that simple theory would suggest. This can be explained by loading effects on each filter section.

Frequency response with both filter sections switched out is very flat, with -1dB points at 30Hz and 100kHz. Gain or insertion loss is minus 1dB, which is negligible.

Harmonic distortion is very low under all conditions and in fact we were merely measuring the distortion of our oscillator at most times. Accordingly, for signal levels below 3V RMS and for all frequencies in the normal audio bandwidth, we have quoted distortion at less than 0.1%. Actual distortion will be very much lower than this.

Maximum input signal before clipping is 4V RMS. Separation between channels is better than 40dB for all frequencies of interest. This is measured with the unused input unloaded so actual results would be better again. Similarly, signal-to-noise ratio is quoted at –63dB with respect to 100mV RMS with open-circuit imputs and with a noise bandwidth of 25kHz. Loading the inputs with a low impedance source improves the figure to 65dB. In practice, the unit causes no discernible degradation of an amplifier's S/N ratio and is considerably better than the average or even above-average cassette deck in this respect.

Threshold settings to actuate the filter control circuitry are variable between 50mV and 300mV RMS. Attack time is about 7 milliseconds while decay time is about 200 milliseconds (mainly determined by the beta of Tr6).



Response curves for the filters. Curve 2 is the resultant curve with both filters.

Now to explain the compromise or "catch" we mentioned earlier in the article. Notice that each filter section begins it rolloff (−3dB point) at about 7kHz and the total rolloff begins at about 4kHz. While this sounds drastic, this compromise was necessary in order to obtain a useful reduction in hiss. Now the catch is that people with acute ears may be able to hear the hiss being "gated" on and off by the arrival of high level signals. This can be more disturbing to some people than a constant background hiss. To reduce this effect, the rolloff frequency should be raised by reducing the filter section capacitors. However this means a consequent lesser reduction in hiss

For those who wish to experiment with the rolloff frequencies, the .0022uF and 470pF capacitors are varied directly in proportion to the proposed rolloff frequency.

Whichever rolloff frequency you do decide upon, we think you will agree that this circuit is very attractive because it has no internal setting up adjustments and literally no distortion.

Our prototype was housed in a diecast metal box measuring 222 x 146 x 57mm, with the control panel on one side and the input and output sockets on the opposite side.

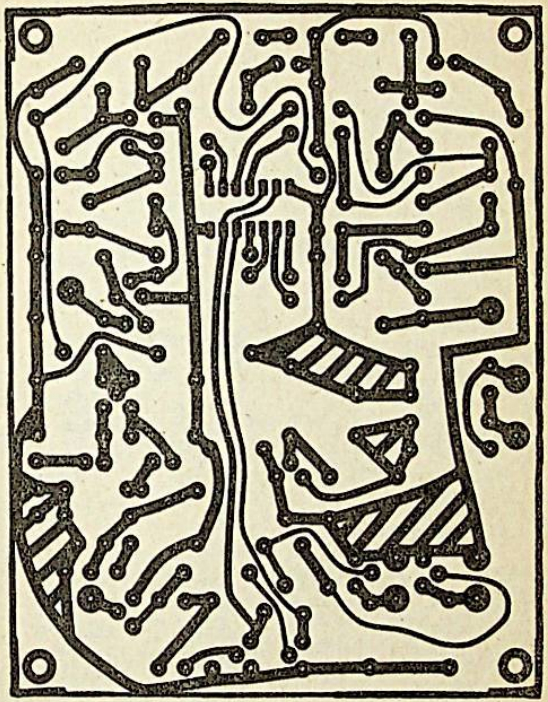
A PC board measuring 127 x 102mm accommodates the circuitry. Assembly of the PC board can begin by installing all the small components, leaving the CMOS integrated circuit till last.

General-purpose small-signal silicon transistors can be used in this circuit. Ideally, Tr1, 2, 3, 4 and 5 should be BC549 or equivalent low-noise silicon NPN types to ensure the best signal-to-noise ratio and high beta requirements set by the high resistance base bias networks. However, the performance we quote in the specification panel was obtained using BC548's which were selected for beta of 250 or more. So if you have means for measuring beta and have BC548's or other general-purpose silicon transistors to spare, they may be pressed into service here.

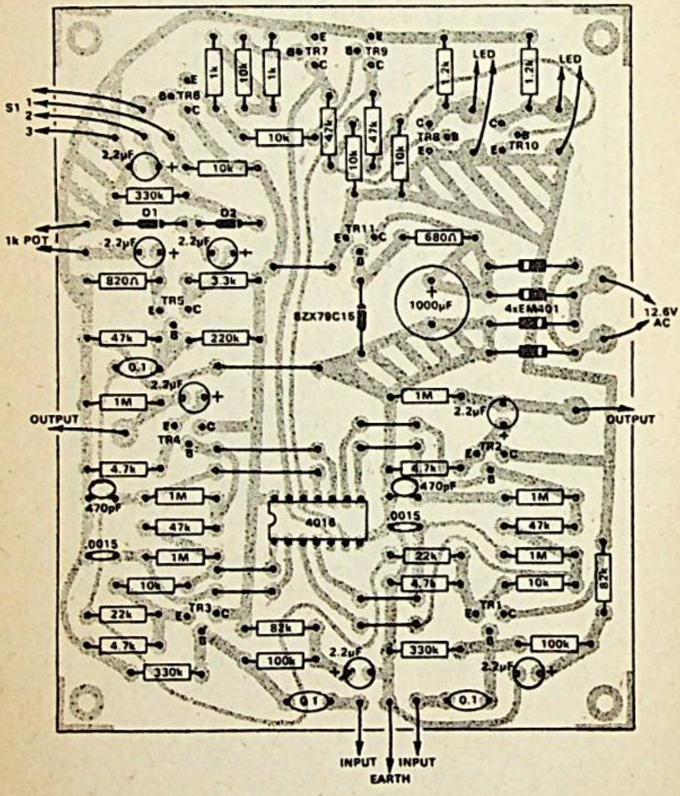
Tr6, 7, 9 and 11 can also be BC548 or any equivalent type. Tr8 and 10 are BC327, BC558 or equivalent general purpose silicon PNP types. D1 and D2 can be any small signal diodes, germanium or silicon or small rectifier diodes such as EM401.

Resistors can be ¼W or ½W types with 5% tolerance. Use low noise types such as cracked carbon or metal glaze for the high value resistors, particularly those used in the bias networks for Tr1 and Tr3. The smaller ¼W resistors are preferable since they are easier to insert.

All the electrolytic capacitors are end-mounting PC types and we have standardised on 2.2uF (for all but the main reservoir capacitor) in the interest of the economy and simplicity. Capacitors used in the single-section filters, .0022uF and 470pF, should have 10% tolerance or better. If low voltage ceramic capacitors are used for the 470pF, they should be checked to ensure they are within 10% of value specified.



P.C.B. Pattern - not full size



Layout of Components

Use PC stakes or pins to make inter-connections to the PC board. Seventeen will be required.

When all the small components have been inserted and soldered, you can deal with the CMOS integrated circuit. It will be supplied with its pins inserted into black conductive foam or wrapped in aluminium foil. Refer to the PC layout to determine the correct orientation for the IC and then insert and solder it, still with foil or foam shorting the pins. When soldering is complete, remove the foam or foil. Actually, if you have a low-voltage soldering iron with a grounded tip you can remove the foam or foil before soldering. This latter procedure is relatively safe for this particular CMOS IC since it has diode protection on all inputs.

This does not mean you can use your old 240VAC 100W soldering iron that your father used to solder galvanised guttering! That is asking for trouble. Care is still required.

Having completed assembly of the PC board, attention can be turned to the diecast case. Drill all the required holes and countersink where necessary. Now the hardware can be installed.

A four-way connector strip was used for the RCA phono input and output connectors. These require more filing and cutting of the case than if four individual sockets were used, but the final result is neater. Alternatively, if your system uses DIN sockets, then these are equally appropriate. The earth connections of the sockets are all connected together and then finally connected to the PC board earth but the signal earth does not connect to the case. This is earthed back via the three-core mains cord.

The transformer we used was a small unit with a 12.6V secondary.

The three-core mains cord should be passed through a grommetted hole in the rear of the case and anchored with a cord clamp. Terminate the earth conductor to a solder lug on the chassis and the active and neutral conductors to a three-way insulated terminal block. Connections to the power switch and transformer are then made via the terminal block.

We used a toggle for the mains switch. After wires are soldered to the switch, the connections should be covered with plastic sleeving to avoid contact with incautious fingers. The mode switch requires a single-pole, three-position rotary type but the one we actually used was a two-pole, three-position wafer. Both the rotary switch and potentiometer shafts should be cut to a length to suit the knobs before being installed. The potentiometer can be a logarithmic or linear type with value 1k.

Several choices are available as far as the LED indicators. We used LEDs with chrome bezels. A cheaper approach would be to use LEDs with plastic clip-lock bezels or simply glue the LEDs into holes in the front panel with a suitable adhesive such as Araldite.

When all the hardware is installed in the case drop the PC board into place and make all the connections. There is no need to use shielded cable for the inputs and outputs nor would there be any noticeable improvement gained by doing so. A short length of multi-coloured "rainbow" cable will provide all the necessary hook-up wire. It results in a neat job without the need for cable lacing. About 30cm of rainbow cable will be adequate.

Double-check your wiring and then apply power. Check voltages. These should be within 0.5V of those on the circuit. The main supply rail should be slightly less than 15V. If it exceeds 15V there is danger of "blowing" the CMOS IC. If all checks are OK you are ready to test the filter with a cassette deck and stereo amplifier.

First connect the output of the filter to the tape monitor inputs or high level inputs of the amplifier and listen for any hum or noise. Unless the amplifier has an exceptional signal-to-noise ratio, there should be no degradation, i.e., the amplifier should be no noisier than it normally is at a given volume control setting.

Now connect the line outputs from the cassette deck to the input of the filter and play a cassette. Adjust the threshold control so that the LEDs are extinguished when the playback signal exceeds about −15dB as indicated on the VU meters of the cassette deck. This adjustment will give you optimum results for most tapes but it can be varied to suit particular conditions.

It is important that the threshold control is not advanced to the point where the LEDs are flashing on and off with low level signals. In this condition the listener is more likely to be conscious that the signal is "gating" the noise on and off. As noted above, this effect could be more objectionable than continuous hiss.

In conclusion we note that this unit could be incorporated into a stereo amplifier with consequent savings on the case and power supply. If built as a free-standing unit like our prototype it would be possible to fit the PC board into a smaller case although more care would be required with layout and transformer location and orientation to ensure hum-free performance. It is also possible to replace the threshold control with a preset potentiometer soldered directly into the PC board, should you wish.

PARTS LIST

1 diecast case and lid.
1 front panel to suit (see text).
1 PC board, 127 x 102mm.
1 three-pin mains plug.
1 power transformer, 12.6V secondary.
1 SPST toggle switch.
1 four-way RCA phono connector strip.
2 decorative knobs.

SEMICONDUCTORS
1 SCL4016A, CD4016 or equivalent CMOS quad bilateral switch.
5 BC549 or equivalent low noise silicon NPN transistors.
4 BC548 or equivalent general purpose silicon NPN transistors.
2 BC327, BC558 or equivalent general purpose silicon PNP transistors.
1 BZX79/C15 15 volt 40mW zener diode.
4 EM401 silicon rectifier diodes.
2 1N914, 1N4148 silicon signal diodes.
2 LEDs with chrome bezel.

CAPACITORS
1 x 1000uF/25VW PC electrolytic.
7 x 2.2uF/25VW PC electrolytic.
3 x 0.1uF metallised polyester.
2 x .0022uF metallised polyester.
2 x 470pF polystyrene or low voltage ceramic.

RESISTORS
(¼W or ½W, 5% tolerance)
6 x 1M, 3 x 330k, 1 x 220k, 2 x 100k, 2 x 82k, 5 x 47k, 2 x 22k,
7 x 10k, 4 x 4.7k, 1 x 3.3k, 2 x 1.2k, 2 x 1k, 1 x 680 ohms.
1 x 1k potentiometer (log or lin.)

MISCELLANEOUS
17 PC stakes, length of three-core mains flex, 30cm of 1 flat rainbow cable, solder lug, grommet, rubber feet (if required), screws, nuts, lock-washers, solder, three-way insulated terminal block, cable clamp.

NOTE: Components with higher ratings may be used provided they are physically compatible. Lower rated components may also be used in some cases, provided their ratings are not exceeded.

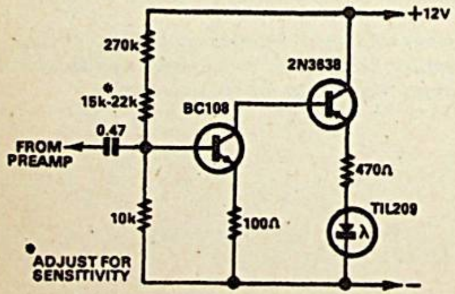
LED SOUND SOURCE INDICATOR

This LED Sound Source Indicator has applications in a multi-channel setup with microphones, pickups etc, in public address and many other applications. Ideally, there would be an indicator for each sound source, adjacent to its associated fader or switch. I have built eight of these units and they are installed in a public address system where I find them very useful.

The LEDs which I have used are type TIL 209, made by Texas Instruments. This type is an epoxy case 1/8in diameter. With 1/8in hole drilled in the panel, the LED may be inserted through the hole, resulting in a neat finish.

If the controls are in the low level part of the mixer, additional amplification may be required over that shown in the circuit. The sensitivity of the amplifier is adjusted by the bias divider and when set correctly, the LED should give full brilliance with normal programme level.

Editorial note: The resistor value range of 15k to 22k for sensitivity adjustment may not be sufficient to cope with all parameter variations. It may be necessary to go outside this range to adjust the sensitivity to the required level.



A SOLID-STATE AF VOLUME COMPRESSOR

A fast-acting circuit with low distortion. Use it to "squeeze" the dynamic range of signals for recording, or to permit deep modulation of transmitters without risk of overmodulation distortion and "splatter."

Specifications:

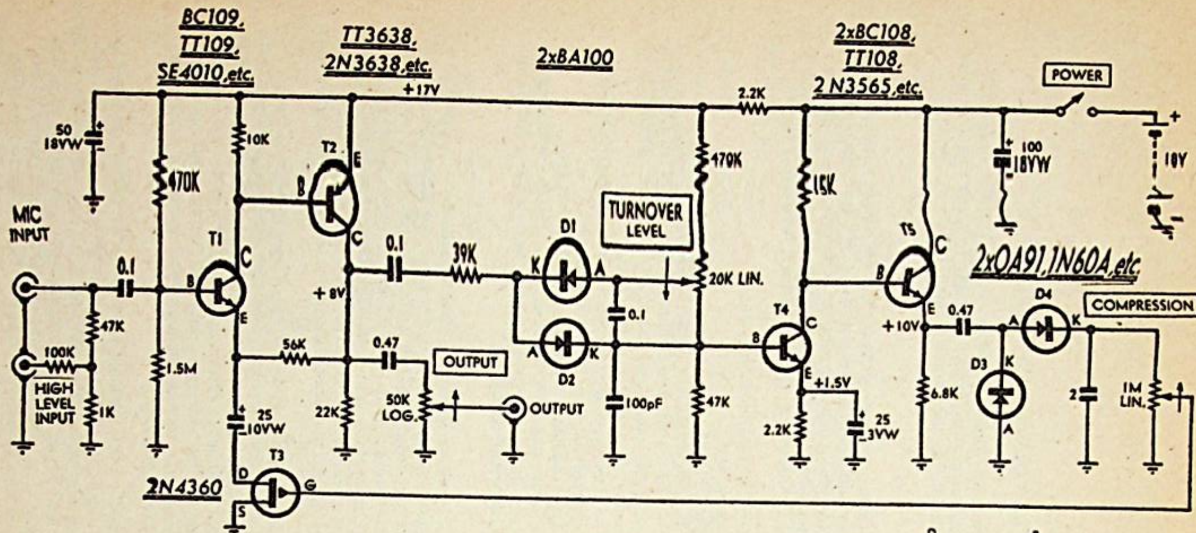
A low distortion AF volume compressor design whose performance makes it very suitable for recording and communications use. The circuit is fully solid state and uses five silicon transistors and four diodes, Compression threshold level or "turnover" is adjustable; the slope of compression is also adjustable from zero to full limiting. Nominal output level is 250mV.

LIMITING COMPRESSION: Approximately 46dB.
ATTACK TIME: Approx. 5mS.
DECAY TIME: Approx. 800mS.
FREQUENCY RESPONSE: Better than 30Hz–15KHz, ± 3dB.
DISTORTION: At 300mV output, zero compression, less than 1.5 per cent. at 300mV output with 16dB compression, less than 1.5 per cent. At 2.5V output, zero compression, approximately 4 per cent.
PRECOMPRESSION GAIN: Microphone input, approx. 46dB (47K impedance). High level input approx. 6dB (100K impedance).
NOISE OUTPUT: Approximately 40dB below 250mV output.
POWER SUPPLY: Circuit requires 18VDC at approx. 3mA, which may be supplied from either an internal or external battery or a zener divider from the power supply of associated equipment.

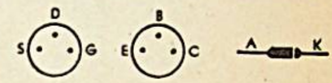
Although considerable advances have been made in the development of high quality sound recording systems, in many cases the attempt to record musical and other material with complete fidelity is still rather similar to the old problem of attempting to fit a quart of liquid into a pint pot. While frequency response and harmonic and intermodulation distortions can be reduced to a negligible level, there remains the problem of dynamic range, or the volume ratio involved between the loudest and softest program levels.

For faithful reproduction of the dynamic range of program material, a recording system should ideally have a dynamic capacity–as defined by the difference between the system overload level and the overall residual noise level–which exceeds the program dynamic range by about 10dB. This relationship ensures that the loudest program passages remain undistorted while the softest passages are reproduced at a power level approximately 10 times that of the system noise.

A significant proportion of symphonic music has a dynamic range as high as 70dB, which figures corresponds to a power ratio of 10,000,000 to 1. Most naturally occurring sounds also fall within this range, with the exception of exceptionally loud but infrequent sounds such as gunshots, sirens and explosions. As faithful reproduction of the amplitude of sounds of the latter category is neither necessary nor desirable (!), an ideal recording system should in theory therefore have a dynamic capacity of around 80dB.



SOLID-STATE AUDIO COMPRESSOR



The main circuit diagram of the compressor, which uses a silicon JFET as the gain control element. Features of the unit are low distortion, fast attack and adjustable turnover and slope characteristics.

Unfortunately at the present state of the recording art, of the three recording systems in common use—magnetic tape, gramophone discs and photographic film—as normally implemented none has a basic dynamic capacity better than about 55dB. Accordingly various techniques have been evolved to increase the effective dynamic capacity.

A technique which has been applied to all three recording systems is that of volume compression, which as the name suggests consists of the compression or "squeezing" of the dynamic range of the recording signals to a degree necessary to accommodate the signal variations within the dynamic capacity of the recording system. In simple terms volume compression is performed by varying the gain of the recording system so that louder sounds receive less gain than those which are originally softer.

In practice there are two general ways in which this may be done. The simpler and more common approach is to arrange that low level signals receive the full recording system gain, while high level signals receive less gain; however, interest has lately been focussed upon the complementary approach, whereby high level signals receive the normal system gain while low level signals receive additional gain. Although there are practical differences between the two approaches in terms of both implementation and performance, it should be realised that from the viewpoint of basic theory they are equivalent.

It is true that volume compression alone does not increase the effective overall dynamic capacity of a recording system; the recorded and reproduced sounds are not a replica of the original but a "squeezed" version, tailored to fit the narrower confines of the system range. Theoretically it is therefore necessary during reproduction to pass the signals through a complementary process of expansion in order to produce sounds which are a replica of the original.

Circuits and devices capable of being used to perform the operation of volume expansion have been developed, and have been used at various times. However, in contrast with basic theoretical requirements, such circuits are used far less frequently than compressors; hence in the majority of applications in which volume compression is employed, no attempt is made to perform subsequent expansion.
One reason for this is that, in practice, it proves difficult to match the operations of compression and expansion sufficiently accurately to produce an overall linear result. Another reason is that expansion circuits tend to magnify certain types of signal distortion produced by recording systems. However, more cogent than either of these reasons is the fact that, in many applications, expansion of signals to their original dynamic range has been found either unnecessary or inadvisable.

Examples of applications where expansion of compressed signals is unnecessary are in the recording of discussions, interviews and speeches, in monitor recording of two-way communication system traffic, and in certain public address applications. Examples of other applications where expansion is inadvisable are in motion picture sound tracks, where full expansion to the original dynamic range would involve either discomfort during loud passages or inaudibility of soft passages (due to auditorium noise), or both; and in background music systems, where expansion would defeat the intended purpose.

An application which may be regarded as a special case of volume compression is that of peak limiting, which is used in recording and (more especially) broadcast transmitters to permit deep and effective modulation without risk of over-modulation. As the name suggests, peak limiting involves restriction of the signal peaks so that they are not permitted to exceed a predetermined level; as such it may be regarded as a case where compression is applied sharply and to a degree which almost balances further increase in input signal amplitude.

It may be seen that volume compression is very often used in applications where subsequent expansion is not attempted, and that when so used it can be a very useful technique. However, it should be noted that compressed sound material which is reproduced without expansion is in theory accompanied by two forms of distortion additional to those distortion products contributed from other sources.

One of these distortion components is simply that involved in the compression process itself, whereby the original relationships between loud and soft passages is changed. However, the second component is a more subtle one which arises because, in general, it is not possible to compress only the high amplitude components within a signal; rather it is necessary to compress the whole of a signal whenever high amplitude components are present. As a result, low amplitude components of a complex signal are compressed "unjustly" whenever they happen to be accompanied by high amplitude components, and this amounts to a different type of distortion which may be called "compression modulation" distortion.

Recently developed compression systems have reduced this source of distortion to a degree by employing filters to split the input signal into a number of separate frequency bands which are compressed separately before being recombined. However, this and similar procedures can only reduce the "compression modulation" distortion; they cannot remove it entirely.

Practical compression circuits tend to introduce two further types of distortion. The first arises because most practical devices and circuit configurations which lend themselves for use as compressors also tend to exhibit a degree of nonlinearity in their behaviour at various signal amplitudes. This produces waveform distortion (usually second harmonic), to a degree which is generally proportional to the peak-to-peak amplitude of the signal applied to the compressor control element. Hence to reduce this source of distortion the signal swing across the control element must be kept to a minimum.

The second practical source of compressor distortion arises from the fact that, upon the arrival of a signal peak at the input of a compressor, a finite time must elapse before the appropriate compression control signal can be generated and acted upon. Hence unless elaborate and costly signal delaying equipment is used, the initial portions of signal peaks do not receive the full compression appropriate to their amplitude, and accordingly "sneak through" at increased amplitude.

To reduce this form of distortion to a minimum it is necessary to reduce the "attack time" of the compressor, so that it acts as rapidly as possible on the arrival of a signal peak. Additional improvement may be obtained by increasing the "decay time", so that the compression signal generated by each signal peak does not die away rapidly, but relatively slowly. As a result, a smaller increment in control signal is required upon the arrival of closely following peaks, and the effective attack time is reduced.

In general, if the attack time of a compressor is kept below about 10mS and the decay time made greater than about 600mS, transient distortion is reduced to a level which is acceptable for all but the most critical applications.

A number of techniques have been used in the past to perform volume compression, each technique differing from others mainly with respect to the particular device or circuit used for controlling the gain applied to the signal. Among the control elements used were voltage-dependent resistors, thermistors, incandescent lamps, light-dependent resistor/lamp combinations, remote cut-off pentode valves (usually used in push-pull pairs), and biased diodes.

Unfortunately most of these elements possess characteristics which make them less than ideally suited for this application. The majority introduce appreciable waveform distortion due to nonlinearity, while many are too slow in operation and cause considerable transient distortion. Again, others such as light-dependent resistor/lamp combinations have an undesirable control characteristic, while some elements require a DC polarising potential which makes it difficult to provide a rapid attack characteristic without also producing a transient effect known as "thump".

The volume compressor design to be described in this article is fully solid-state and employs as the control element a junction field-effect transistor (JFET). When used in the particular manner employed in the new design, the JFET displays control and signal handling characteristics which are particularly well suited for this application. Waveform distortion is low, attack time can be made very short (and decay time very long), while the control characteristic permits smooth and stable control.

The way in which the JFET is used as a control element is illustrated in figure 1. As may be seen from the diagram, the drain-source channel of the device is used purely as a resistor, whose value is controlled by the DC control voltage applied as a reverse bias to the gate electrode. Direct polarising voltage is blocked from the channel by a high-value series capacitor.

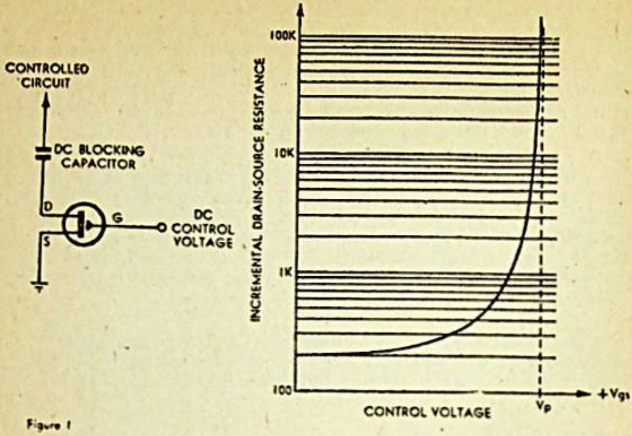


Figure 1

When used in this manner the effective channel resistance of the device varies with control voltage in an exponential fashion, as shown in the curve. At zero control voltage the resistance has a value of around 200 ohms, rising at an increasing rate with control voltage to a value higher than 100K at pinch-off. Hence the control resistance ratio available from the device is better than 500:1, or 54dB.

Spread variations in device parameters do not materially affect either the shape of this control characteristic or the overall control ratio. The only aspects subject to significant variation are the starting resistance at Vgs=0 and the control voltage at which pinch-off occurs (Vgs=Vp).

At first sight perhaps the most appropriate way of using the JFET as a compression element would be to connect it in series with a resistor to form a simple voltage divider across the signal source. However consideration of the nature of the control characteristic shown in figure 1 should reveal why this is not a feasible approach.

In order to ensure low distortion and a simple ground-referenced control circuit, the JFET would have to be the lower element of the divider. In this position it would have to be biased at Vp or beyond for low signal levels (no compression), with compression performed by reducing the gate control voltage. It will be seen that traversing the control characteristic in this direction implies an initially high control gain which falls away rapidly with increasing compression; as a result is becomes difficult to achieve even moderate static compression levels and attack characteristics before the control circuit develops serious transient instability.

From the foregoing it may be deduced that to realise the full compression range and response rate potential of the JFET it is necessary to ensure that the device is operated in the direction of increasing control voltage–i.e. in the direction of increasing resistance and control gain.

In the new design this requirement is met by using the device as the lower element in a negative feedback divider. As may be seen from the main circuit diagram an economy device type 2N4360 is used, connected in the input emitter circuit of a two-stage complementary feedback preamplifier using transistors T1 and T2.

For low amplitude signals, and when compression is inoperative, the preamplifier has a gain of approximately 200 (46dB) as defined by the 56K feedback resistor and the zero-bias JFET channel resistance. When compression is required a positive control voltage is applied to the JFET gate, increasing its channel resistance and accordingly increasing the negative feedback applied to the preamplifier. The gain of the latter therefore falls smoothly and prograssively as required dropping to unity or below if sufficient control voltage is applied. Almost all of the JFET control range may therefore be realised, with a maximum compression of greater than 46dB.

The JFET control voltage is derived from the preamplifier output by the remainder of the circuit, which employs silicon diodes D1 and D2, transsistors T4 and T5 and germanium diodes D3 and D4. The action of this section of the circuit is that of a peak-responding signal rectifier having adjustable amplitude delay.

Transistor T4 is used in a conventional common-emitter voltage amplifier stage, to provide control signal amplification. Transistor T5 is used as an emitter-follower stage directly coupled to the output of T4, providing impedance matching between this stage and the voltage doubling rectifier formed by D3 and D4.

The low output impedance of the emitter-follower permits rapid charging of the 2uF control voltage reservoir capacitor at the rectifier output, ensuring a compression attack time of better than 5mS. The compression attack time is almost wholly determined by the charging rate of this capacitor, because the JFET itself is capable of operation at very high speeds.

A voltage-doubling rectifier circuit is used in order to allow response to both positive and negative signal half-cycles.

The proportion of the rectifier output voltage applied to the JFET gate–and hence the compression slope, or rate of gain reduction relative to sighal amplitude increase–is adjusted simply by means of the 1M potentiometer across the reservoir capacitor. The potentiometer itself forms the decay circuit for the capacitor, having a value which ensures a compression decay time of longer than 800mS.

It is desirable that a volume compressor be adjustable in terms of the threshold point or "turnover level"—that is, the signal level at which compression is brought into operation. This may be seen by reference to the diagram of figure 2, which illustrates idealised compressor control characteristics.

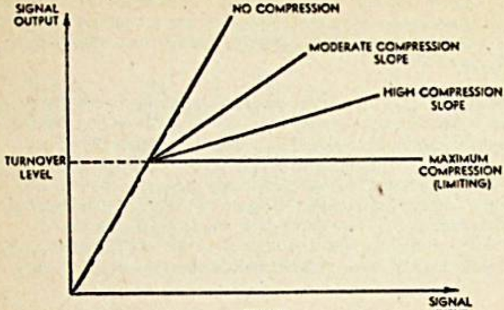


Figure 2

Idealised compressor control curves, illustrating the concepts of turnover, compression slope and limiting.

Fairly clearly the optimum turnover level in a particular application will depend upon the prevailing signal level and the system overload and noise levels. If the turnover level is lower than optimum, the compressor will act mainly as an attenuator because its transfer slope will be constant for all but the lowest signal levels; on the other hand a turnover level higher than the optimum will not permit a useful degree of compression.

A secondary reason for the provision of a turnover adjustment is that it allows compensation for variations in turnover level which occur normally due to parameter spread in the rectifier diodes and JFET.

If the control circuitry of the compressor were arranged to operate for all signal levels in a linear fashion, a moment's thought should show that the circuit will have an effective "turnover level" of zero. It will operate in the compression mode for all signals, and will behave mainly as a fixed attenuator. Hence to provide the circuit with finite and non-zero turnover level, it is necessary to arrange that compression does not commence until the signal reaches a preordained level. In short, the control circuit must be provided with "delay," and to provide turnover level adjustment the delay must be made variable.

Note in passing that the conventional use of the word "delay" in this context tends to be misleading, as it suggests a response time effect. For this reason it is better to think in terms of "amplitude threshold" or "turnover level." Although there will in fact be some relationship between turnover level and response time for typical signals, the relationship will be very much a second-order one.

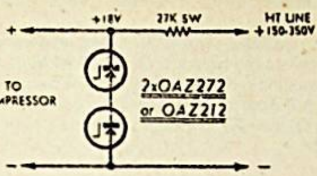


Figure 3 ALTERNATIVE POWER SUPPLY

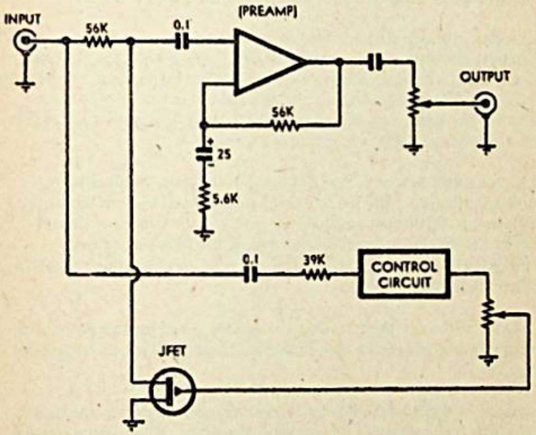


Figure 4 MODIFIED CONFIGURATION FOR VOLUME EXPANSION

For those who care to experiment with volume expansion, here is a modified circuit configuration which performs this operation.

Although the required control circuit "delay" could be provided by the conventional method of applying a reverse "hold-off" DC bias either to the rectifier or to the JFET itself, this approach has a serious disadvantage. A significant portion of the available peak-to-peak output voltage swing of transistors T4 and T5 would be used purely to overcome the hold-off bias, and would therefore be unavailable for control purposes. As the transistor supply voltage is relatively modest, this approach would therefore cause a significant reduction in the maximum available control voltage.

In contrast with the conventional approach the method used in the new compressor design provides adjustable turnover level without reducing either the gain or the maximum output of the control circuit, As may be seen from the main circuit diagram, this is achieved quite simply by using a pair of silicon diodes D1 and D2 in a signal threshold circuit at the input to the control amplifier.

The operation of this circuit depends upon the shape of the forward bias characteristic of a silicon diode. It may be remembered that at low values of forward bias voltage, the current drawn is small and the diode acts as a high resistance. However, as the voltage is increased to around 600mV the current rises sharply and the diode resistance falls to a low value.

Two such diodes connected in inverse parallel thus form a non-linear resistor which passes only slight current for low applied voltages of either polarity, but a proportionally much higher current at higher applied voltages of either polarity. When connected in series with the control amplifier input, as shown, the diodes therefore provide the required threshold action by forming a non-linear signal voltage divider.

The exact signal amplitude at which the threshold occurs is adjusted by applying a small adjustable DC forward bias to the diodes by means of the 20K potentiometer. With zero applied bias, the threshold is a maximum determined by the diodes themselves; increasing the bias lowers the threshold by reducing the signal amplitude required to reach the "knee" in the diode characteristics.

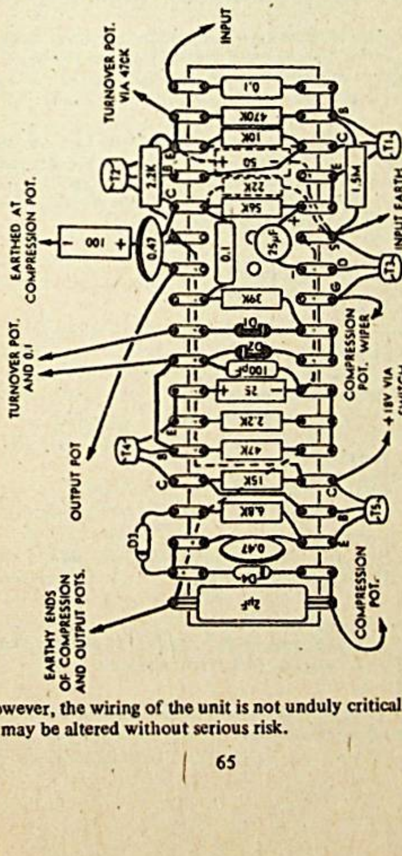
Note that the circuit is arranged so that the diodes are in inverse parallel as far as the signal is concerned, but are in like-oriented series with respect to the DC bias.

The 39K resistor in series with the signal input is used to swamp the non-linear diode input characteristic and prevent distortion of the signal output from T2. The 100pF capacitor shunting the input of T4 prevents over-compression of high-frequency signal components by compensating for diode junction capacitance.

The power requirement of the compressor is 18V DC at approximately 3mA. This may be supplied from a pair of small 9V batteries as used for the prototype, or from a suitable source of well-filtered 18V within associated equipment. With valve equipment the most convenient approach will probably be to use a zener diode voltage divider connected to the HT line, as shown in the diagram of figure 3, or to derive the supply from the cathode circuit of an audio output valve.

The prototype compressor was built as a self-contained unit in a small utility box. However, there is no reason why the circuitry could not be built into an existing amplifier, mixer, recorder or transmitter modulator if so desired, the only stipulation being that the wiring must be fully shielded to prevent pickup of hum and noise.

To aid those constructors who wish to duplicate the original wiring and parts layout a wiring diagram is shown



However, the wiring of the unit is not unduly critical and if necessary or desired may be altered without serious risk.

When completed and in operation it may be found that the compressor has a tendency toward occasional oscillation immediately following switch-on if the compression control is set for peak limiting (full compression); this is due to a blocking situation created by switch-on transients. The effect may be prevented by reducing the compression control setting prior to switch-on. If this is not done and oscillation occurs it may be stopped by temporarily reducing the degree of compression.

Before concluding this article it may be worthwhile to note that the circuit configuration of the compressor may be re-arranged to perform the complementary process of volume expansion. Figure 4 shows the new configuration, which uses the same control circuitry as before but with the JFET connected into a simple voltage divider. Note that the expander configuration uses "open-loop" control, as closed-loop operation would be completely unstable. The preamp remains in circuit but has reduced gain (20dB) to allow it to cope with the higher signal levels produced from the expander divider.

It would be quite feasible to make up a dual function compressor-expander along these lines. However, to obtain accurate balancing of the two characteristics, it will be necessary to ensure that the signal delivered to the expander is at exactly the same level as the original output from the compressor.

As expansion tends to magnify signal amplitude and frequency response variations it will be found that such a system will give satisfactory results only with modest degress of compression-expansion--i.e., less than about 16dB. With greater expansion slopes the magnified signal variations generally far outweigh any improvement in dynamic capacity.

Parts List:

1 Case, 6½in x 4½in x 2in, with flanged front panel.
1 18-lug length of miniature resistor panel.
1 Miniature slider switch.
3 Small control knobs.
2 BA100 or similar silicon diodes.
2 OA91, IN60A or similar germanium diodes.
2 BC108 or 2N3565 NPN silicon transistors.
1 BC109 or SE4010 low noise silicon transistor.
1 2N3638 or similar PNP sillcon transistor.
1 2N4360 or similar junction FET.
1 Microphone connector.
1 Phone jack.

RESISTORS
(All half-watt 5 per cent type)
1K, 2 x 2.2K, 6.8K, 10K, 15K, 22K, 39K, 2 x 47K, 56K, 100K, 2 x 470K, 1M; 20K WW pot, 50K log pot, 1M linear pot.

CAPACITORS
(LV plastic except where marked)
100pF, 3 x 0.1uF, 2 x 0.47uF ceramic, 2uF, 25uF 3VW electro, 25uF 10VW electro, 50uF 18VW electro, 100uF 18VW electro.

MISCELLANEOUS
2 x small 9V batteries and connectors; rubber feet for case; screws, nuts, washers, connecting wire, solder, etc.

LOUDSPEAKER PROTECTOR

Ever had the misfortune to "blow" an output transistor in an amplifier without coupling capacitors? Blow your loudspeakers too? You can guard against this possibility by building the Protector circuit described here. It also eliminates switch-on "thumps" from the loudspeakers.

Many hifi fans do not realise that amplifiers with direct-coupled outlets to the loudspeakers can pose a real hazard – to the loudspeakers! By direct-coupling, we are referring to those amplifiers without output-coupling capacitors. Japanese manufacturers refer to them as OCL or "output capacitorless".

There are several advantages in having an amplifier with direct-coupling to the loudspeakers. It results in better damping factor and improved power output at low frequencies. To the designer it enables elimination of at least one large electrolytic capacitor, with a consequent cost saving. And it also eliminates one possible cause of switch-on transients.

But all these advantages add up to zero if a failure occurs in the power amplifier and applies a large DC voltage across the loudspeaker(s). The most likely result of this is that the loudspeaker voice coils are burnt out before the owner realises that anything is amiss.

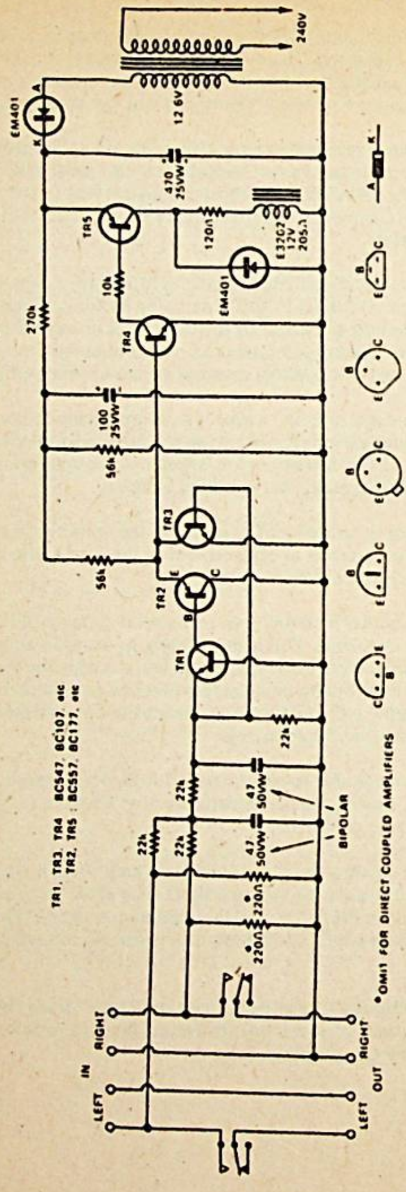
While the transistor or component that fails in the amplifier may be replaced at a low cost, repair or replacement of the loudspeakers can be very much more expensive.

Another problem which is common to many solid-state amplifiers is that of switch-on transients. This is more likely to occur in amplifiers with output coupling capacitors – when the output capacitors charge up there can be a loud thump emitted from the loudspeakers. Usually the large DC charging pulse is not likely to damage the loudspeakers, but its audible effect can be annoying.

Both of these problems can be eliminated with the loudspeaker Protector featured here. Indeed, similar circuits are now featured in many expensive high-power amplifiers.

Another problem common to many solid-state amplifiers is that they can cause the loudspeakers to thump a short time after being switched off. The Loudspeaker Protector will also eliminate most of this problem, particularly where the thump occurs several seconds after switch-off.

Some amplifiers also occasionally give a sharp "crack" from the loudspeakers at the instant of switch-off. However, that is a problem which cannot be cured by this simple circuit.


TR1, TR3, TR4 BC547, BC107, etc
TR2, TR5 : BC557, BC177, etc
LEFT
IN
RIGHT
OUT
220Ω
47 50VW
22k
BIPOLAR
TR1
TR2
TR3
TR4
TR5
56k
100 25VW
270k
10k
120Ω
470 25VW
EM401
E3202 12V 205Ω
12.6V
240V
*OMIT FOR DIRECT COUPLED AMPLIFIERS

Refer now to the circuit. It is simpler than it appears at first sight. Basically it consists of a relay which normally connects the loudspeakers to the amplifier a few seconds after switch-on. If a DC voltage is subsequently applied across the loudspeakers, the relay disconnects them.

Five general purpose transistors are used in the circuit. Tr5 drives the relay direct. A diode in the collector circuit protects Tr5 against the inductive kick-back from the relay when it is de-energised. Tr4 controls Tr5 via the 10k resistor. When Tr4 conducts, so does Tr5.

Base bias for Tr4 is provided by a network consisting of two 56k resistors, one 270k resistor and the 100uF capacitor. At initial switch-on the 100uF capacitor has zero charge and so no forward bias is applied to Tr4 and the relay is off. After about two seconds, the capacitor is charged sufficiently to allow Tr4 and Tr5 to turn on and energise the relay which connects the loudspeakers to the amplifier.

Tr1, Tr2 and Tr3 form a rather incestuous triple which monitors the amplifier outputs for DC fault conditions. They function as follows:

Both channels of the amplifier in question are monitored by Tr1, 2, 3 via a low-pass filter consisting of four .22k resistors and two 50uF capacitors. In a typical amplifier with direct-coupled output there is a normal "offset" DC voltage at the output which may be anywhere from about 20 millivolts to perhaps 200 to 300 millivolts. These normal offsets must not affect the monitoring network.

If one of the amplifier outputs goes positive by more than two volts, Tr3 is forward biased and it conducts to remove the base bias from Tr4. Hence Tr4 and Tr5 turn off and the relay disconnects the loudspeakers. Similarly, if the amplifier output goes negative by more than two volts, the emitter of Tr1 is pulled negative with respect to its base. Tr1 then conducts as does Tr2, and so Tr4 and Tr5 are turned off as before.

So all the transistors function as simple switches which are only controlled by DC signals. AC signals have no effect due to the input filter.

The two 50uF capacitors in the input filter are non-polarised electrolytics. They have to be, since DC voltages of either polarity may be applied to them. The capacitors we used are "non-polarised".

The 220 ohm resistors on the circuit are marked with asterisks. These should be included where the Protector is used to eliminate switch-on thumps from amplifiers with output coupling capacitors. The resistors allow the output capacitors to charge in the delay period before the loudspeakers are connected. If the resistors were omitted there would be an awful bang from the loudspeakers when the relay is energised.

While the 220 ohm resistors are essential where the Protector is to be used with amplifiers having output capacitors, they should be omitted where used with amplifiers having direct-coupled outputs. If they are included there is a strong likelihood that they will be burnt out in the event of an amplifier fault.

As it stands, the Protector circuit can be built in a number of forms. First, it can be built into the amplifier it is to work with, and powered from it. The supply rail may be anywhere in the range from 12 to 45V DC. The only change necessary to adapt to differing supply rails is that the 120 ohm resistor should be varied so that no more than 12V is applied to the relay. Coil resistance of the relay should be about 205 ohms.

Current drain of the circuit with the relay energised is close to 60 milliamps. If it is run from the main positive DC rail in the amplifier the diode and 470uF filter capacitor may be omitted. Note that the zero volt rail (ie, earth) of the Protector should connect to the main earth point of the amplifier.

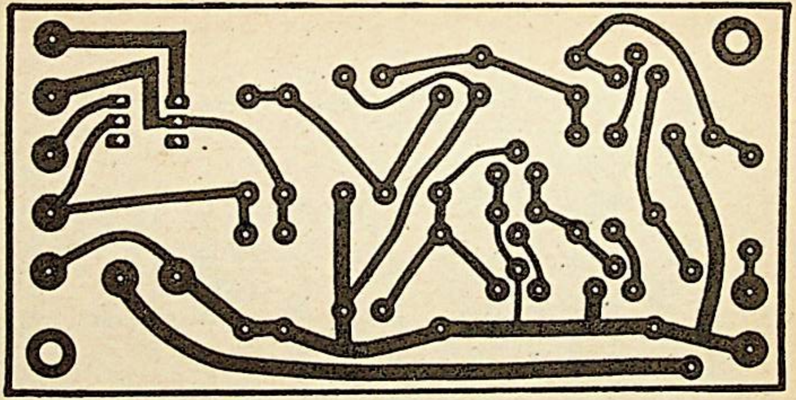
If it is inconvenient to power the Protector from the main positive DC rail of the amplifier it is possible to run it from an AC winding on the power transformer provided that one side of the winding is, or can be connected to the main earth of the amplifier. AC input voltage to the rectifier of the Protector may be in the range of 9 to 30VAC. If the resultant DC voltage is more than 25V, the voltage rating of the 470uF filter capacitor should be increased accordingly.

Where the constructor does not wish to incorporate the Protector into an existing amplifier it will be necessary to construct it as a separate unit with its own small transformer to provide the DC supply. Here again it can be built in one of two versions. Note that the Protector transformer must be energised at the same time as the amplifier.

Where the amplifier in question has switched 240VAC outlets the Protector can be plugged into the rear of the amplifier and controlled by the amplifier power switch. If the amplifier does not have a switched 240VAC outlet, the Protector will be required to have its own power switch and a 3-pin mains socket into which the amplifier can be plugged. The amplifier is then turned on and off with the Protector power switch. Our prototype is the latter version.

Our prototype was housed in a neat little case which measured 136 x 60 x 104mm (W x H x D).

A flush mounting 3-pin outlet and two sets of four-way screw terminals are mounted on the rear of the case. The sets of terminals are for connection of output wires from the amplifier and wires to the loudspeakers. The front panel is bare except for the power switch.



Here is the full size copper pattern of the PC board. However this accommodated our prototype relay. Check the mounting details of the relay you use and if necessary modify pattern to suit.

When operation of the circuit has been checked, the unit can be connected to the amplifier and loudspeakers.

You can then sigh with relief, because your precious loudspeakers are now safe from damage if your amplifier pops an output transistor.

There is one other advantage of the Protector. It enables you to quickly kill the sound of an objectionable program, rather than letting it fade away after normal switch-off.

PARTS LIST

1 Case and lid.
1 PC board, 102 x 51 mm.
1 PC board bracket (see text).
1 flush-mounting mains socket.
1 three-pin mains plug.
1 power transformer, 12.6V secondary.
1 SPST toggle switch.
2 four-terminal connector strips.
3 BC547 BC107 or equivalent NPN silicon transistors.
2 BC557 BC177 or equivalent PNP silicon transistors.
2 EM401 silicon power diodes.
1 relay, 12V double-changeover contacts.
1 470uF/25VW PC electrolytic capacitor.
1 100uF/25VW PC electrolytic capacitor.
2 50uF/50VW PC non-polarised capacitors.

RESISTORS
(¼ or ½W or 10% tolerance)
1 x 270k, 2 x 56k, 4 x 22k,
1 x 10k, 2 x 220 ohms, 1 x 120 ohms.

MISCELLANEOUS
7 PC stakes, length of three core flex, solder lug, grommet, rubber feet, screws, nuts, lockwashers, hook-up wire, solder.

NOTE: Resistor wattage ratings and capacitor voltage ratings are those used for our prototype. Components with higher ratings may be used provided they are physically compatible. Lower rated components may also be used in some cases provided their ratings are not exceeded. See notes on components in the text.

CRYSTAL MICROPHONE IMPEDANCE TRANSFORMER

The circuits will enable devices requiring high loading impedance, such as crystal microphones and crystal/ceramic pickups, to be fed into low and medium impedance inputs.

Silicon transistors have become so readily available and cheap,that they have almost entirely taken over from germanium. The major advantages of silicon devices being the high gain, low leakage and good temperature stability. The low power audio devices, which we now use, have current gains,of, typically, between 150 and 600.

Consequently, we have decided to present the impeaance transformer idea for the benefit of those who may desire to use a crystal microphone or pickup, or ceramic pickup, with transistor amplifiers.

As presented, the public address amplifier had two medium impedance input facilities, with provision for mixing the two. If it is contemplated using any sort of high impedance device with this amplifier, it will be necessary to use an impedance transformer such as is being presented.

While the transistor situation has changed over the past few years, the same is true for economy type microphones. It is now possible to buy a quite reasonable dynamic microphone for a price which compares more than favourably with crystal types. A probable reason for this is the wide acceptance and general use of transistorised amplifiers which, by virtue of their inherently lower input impedance, are more compatible with dynamic microphones.

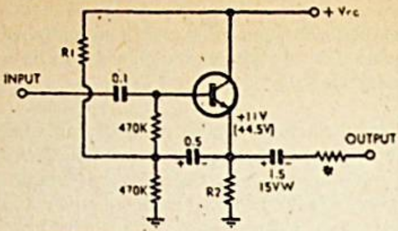
However, crystal microphones have characteristics which may appeal to some users, including possible price advantage, higher output, small size, or simply the fact that one is already on hand. Should one be already on hand, it would certainly be logical to make this transformer in order that it may be used. On the other hand, there may be no point in buying a crystal microphone for use with a transistor amplifier, and then having to make a special device to marry the two.

Even if the readers have both types of microphone, crystal and dynamic, the presently described transformer would enable the crystal microphone to serve either as a suitable standby microphone or as a second microphone with the P.A. amplifier mixing facility, as required. Alternatively, the transformer could be used with a crystal or ceramic pickup for mixing with a microphone input.

In any case, accepting that the decision has been made to use crystal or ceramic transducers of some kind, the circuit diagrams indicate how to achieve various impedance transformations.

We have presented two circuits, one using a single NPN transistor and one using two transistors for greater impedance multiplication. As

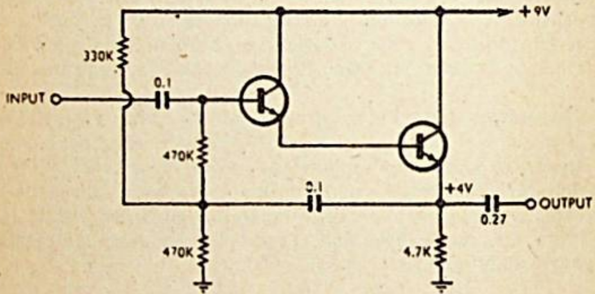
BC109,TT109,SE4010 etc.



SUPPLY VOLTAGE	R1	R2	INPUT IMPEDANCE
9V	220K	22K	4M
22V	330K	47K	8M

* SEE TEXT

2xBC109,TT109,SE4010,etc.



presented there are two voltage versions of the simpler device giving roughly the same multiplication factor, but intended to work into amplifiers of different input impedance.

The actual input impedance of the transformer, as the name implies, is dependent upon the load which is connected to the emitter output. In operation, the device multiplies the effective emitter load impedance—a combination of the DC load resistor and the AC coupled impedance—by a factor roughly equal to the transistor current gain.

As we have said, the transistors which we have used have a current gain of between 150 and 600, which means that the actual input impedance of a particular device will be somewhat dependent upon the transistor which is used. However, we have been able to make a worthwhile device using only one transistor as an emitter follower with bootstrapped bias resistors.

We have presented two versions of the single transistor device, for applications presenting wither medium or high load impedance at the emitter output. This has been done by simply increasing the emitter load resistance, re-arranging the bias, and increasing the supply voltage to maintain the same transistor current.

Values for the two arrangements are given in a table on the circuit diagram. Also we have assigned nominal input impedances of 4M and 8M which were measured with the emitter output unloaded. Consequently the input impedances, in practice, will be something less, depending upon the output loading.

The output impedance is approximately equal to the value of the emitter load resistor for both cases using a single transistor. Consequently, a load of the same value capacitively coupled to the output will reduce the input impedance to about half the unloaded value. In other words, a load impedance of about 22K connected to the output of the low voltage device will reduce its input impedance from 4M to 2M.

Similarly, for the higher voltage version, a load of 47K will reduce the input impedance from 8M to about 4M. It may be seen that the reduction in input impedance is proportional to the reduction in effective emitter load impedance. Halve the load and you halve the input impedance, quarter the load and you quarter the input impedance. The fractional reduction of the effective load may be found by the resistors-in-parallel calculation.

However, there is a limit to the amount by which the effective emitter load may be reduced, notwithstanding the acceptability of reduced input impedance. With the emitter resistor unloaded, the maximum permissible input voltage swing (P-P), before the onset of distortion in the form of clipping, is almost equal to the supply voltage. But, as the emitter is loaded, the maximum permissible voltage swing is reduced in roughly the same proportion as the input impedance.

For example, the maximum permissible voltage swing, with an applied load equal to the emitter resistor, will be a little less than half the supply voltage. In the case of the 9-volt unit it will be about 4V P-P, and for the 22-volt supply the maximum will be about 10V P-P.

When the transformer is used with a High Power Public Address amplifier, we suggest that a resistor be included in series with the output of the transformer and the input of the amplifier. The exact value of the resistor will depend upon the type of input source used, either crystal microphone or crystal/ceramic input.
There are two reasons for having such a resistor, first to reduce the loading effect of the amplifier input and secondly to reduce the overall input sensitivity. For use with a crystal microphone the resistor should be 82K giving an input sensitivity of about 25mV, and with a pickup input the resistor should be 330K giving an input sensitivity of about 100mV.

With either 82K or 330K the input impedance of the transformer was still about 4M. Under the same conditions the input impedance of the

22V version would be reduced slightly with an 82K resistor, but it would still be well in excess of 4M.

However, we feel that there is little point in using the higher impedance version with the P.A. amplifier as an impedance of about 4M is more than adequate for most microphones and pickups. Incidentally, manufacturers usually specify a load impedance of between 1.5 and 2M for crystal and ceramic pickups, but higher load impedances are perfectly satisfactory.

We actually built up the 9-volt version in a completed form, for use with the prototype public address amplifier. The components were arranged on a small piece of Vero board. Tantalum type electrolytics were used because of their small physical size.

The completed wiring side of the Vero board was covered with a layer of plactic insulation tape, so that it could be accommodated on top of the 9-volt battery. The complete assembly was housed in a small metal container about 2¼in long and 1¼in diameter.

A coaxial microphone connector was fitted in the end of the cylinder, and a small push-button on/off switch in the lid. The switch may be considered as an optional extra, because, with the extremely small transistor current drain the battery will probably last its normal shelf life. An output lead, with a coaxial plug attached to one end, passes through the lid adjacent to the switch.

The second circuit has been included for the general interest of readers, and for possible use as an impedance matching device in hi-fi applications. The configuration employs two transistors in a compounded emitter follower configuration, again with bootstrapped biasing resistors.

The use of two transistors, with a very high compounded current gain, understandably produces a very high input impedance with quite a low output impedance. With lower output impedance, this circuit tends to present an input impedance which is relatively independent of output loading.

GLIDE TONE GENERATOR CHECKS AUDIO EQUIPMENT

Here is a project using a new quad op-amp IC, it generates a smoothly gliding audio tone which is just the thing for checking the response of your amplifier and loudspeaker system. Low in cost, it is very easy to build.

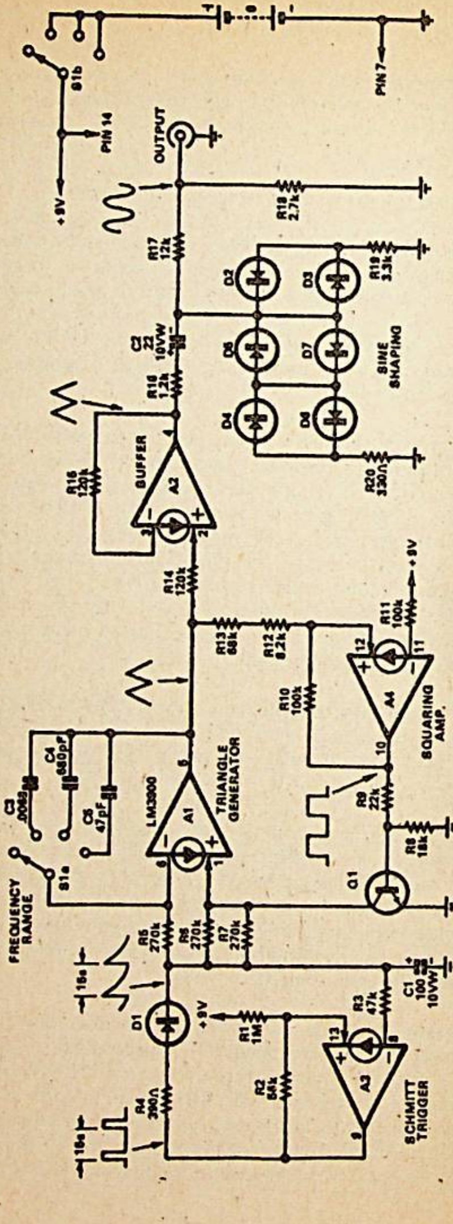
Recently I faced the problem of replacing my tweeters in loudspeaker cabinets where the crossover network was not easily accessible. There was doubt on correct phasing, so I used a function generator with a swept tone facility. By sweeping through the crossover frequency, the correct phasing was quite obvious – the wrong phase caused a null. Similarly, a friend had serious doubts on the performance of his tweeters. Again the sweep generator was used to confirm that they were worth their weight in scrap metal. So my thoughts turned to a cheap glide tone generator.

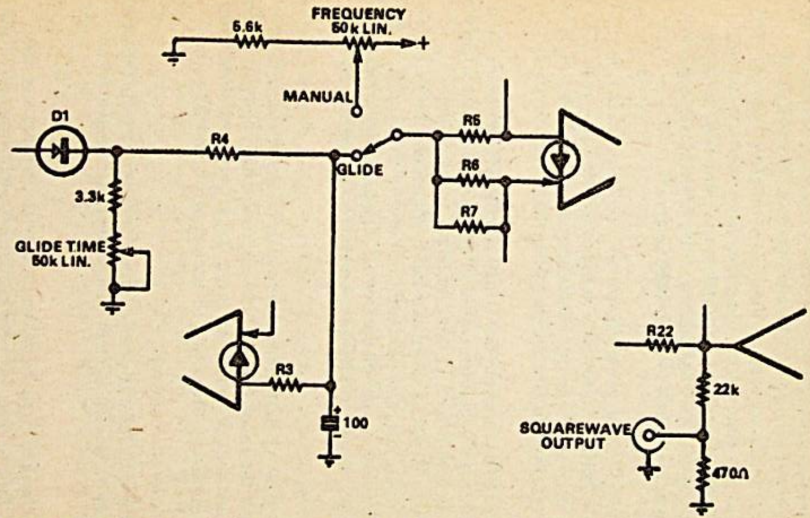
A linear sweep through the spectrum is too unbalanced for listening tests. A logarithmic sweep, which glides smoothly down the musical scale, is far superior. This simple generator covers the audio spectrum in three ranges – 24kHz gliding down to 1.6kHz, 2.4kHz to 160Hz, and 240 to 16Hz. Each range covers a 15 to 1 ratio of frequency, or 3.9 musical octaves. There is an overlap of 50pc between ranges, so that crossover regions in loudspeaker systems can be covered on one range. It takes about 15 seconds for the tone to glide through each range, or about 4 seconds per octave. Any of these characteristics can be changed by varying a particular component.

The glide tone generator is switched on by setting the selector switch to the appropriate range. It then glides through the range, starting at the high end. It repeats the glide or "sweep" ad infinitum or until you get sick of it.

The generator is light and small. It is powered from a 9 volt battery, and the drain of 6mA is so slight that the listerner can be expected to wear out before the battery does. The output is a 200 millivolt RMS sine wave from a 2.2k source, which is suitable for plugging into an auxiliary input of an amplifier. Therefore the glide tone tests the tone controls, filters, amplifier, and loudspeakers as a complete system in the home listening environment.

The first question in the reader's mind will be "then how do I measure the sound level at my favourite chair?" An enterprising hobbyist with a good quality microphone and preamplifier will be able to make a sound level meter but for most purposes the ear is quite adequate. If the tone glides smoothly, with no abrupt or obvious changes in sound level, the system should reproduce music smoothly. If there is some doubt about the quality of the glide, then there is probably something wrong.


FREQUENCY RANGE
S1a
S1b
C3 .0068
C4 680pF
C5 47pF
LM3900
A1
TRIANGLE GENERATOR
BUFFER
A2
R16 120k
R15 1.2k
C2 22 10VW
R17 12k
R18 2.7k
OUTPUT
+9V
PIN 14
PIN 7
D2
D3
D4
D5
D6
D7
R19 3.3k
R20 330Ω
SINE SHAPING
R14 120k
R13 68k
R12 8.2k
R11 100k
R10 100k
A4
SQUARING AMP.
R9 22k
R8 18k
Q1
R5 270k
R6 270k
R7 270k
D1
R4 390Ω
R1 1M
R2 56k
R3 47k
A3
SCHMITT TRIGGER
C1 100 10VW



Three simple elaborations on the basic device are shown in the above diagram. These include a voltage divider and output terminal for square wave output, a manual-glide switch and frequency potentiometer to allow manual control of frequency, and a shunt control across C1 to allow control of the glide repetition rate. Note that D1 and R4 should be transposed (as shown) if these modifications are incorporated.

Since the tone contains noticeable distortion, any changes in the tone quality would be significant. For example, if the tone seems to become "mellow", then the upper harmonics are not being reproduced. Similarly, if the tone seems to become harsher, then the upper harmonics for that particular range are being accentuated.

The next question will probably be "at what frequency did I hear those irregularities?" Even if you do not have access to a direct-reading frequency meter, there is an easy method for estimating the approximate frequency at any time during the glide. The generator sweeps through 3.9 octaves on each range and takes the same time to cover each octave. If the glide time was measured at 15½ seconds, say, then the generator takes 4 seconds to cover each octave.

So on the middle range, the generator starts at 2.4kHz, takes one second to reach 1.2kHz, two seconds to reach 600Hz and so on. In this way, if you keep your eye on the sweep second hand of your wrist watch, you can readily estimate the frequency at a given time.

A practical method for checking out a high fidelity system could be as follows: The glide tone generator would be plugged into an auxiliary or tuner input on the amplifier and then set to the middle range (ie, 2.4kHz to 160Hz). All filters, presence and loudness controls would be switched out of operation. Connect a multimeter to the loudspeaker terminals and switch it to one of the low AC voltage ranges. Advance the volume control to give a suitable sound level and reading on the multimeter.

Set the tone controls so that the multimeter is constant as the tone glides throughout the range. Then switching to the high and low ranges will either confirm that the amplifier has a flat response or the degree of any roll-off. Similarly, the action of tone controls and filters can then be checked.

With the tone controls reset to give the flattest response over the whole audible frequency range, the system is now ready for loudspeaker listening tests.

Phasing can be checked first. Place the loudspeaker systems so that they are almost face to face, forming a narrow V, with the top of the V towards the listening position. With the generator on the lowest range and the amplifier switched to mono operation, correct phase is easily identified. If the phase of one of the loudspeaker systems is reversed, bass cancellation will occur. This checks the phase of the woofers only.

The loudspeakers can now be returned to their normal positions and all ranges swept in mono as before. The sound should appear to emanate from a point located centrally between the loudspeakers and if it shifts noticeably during the glides, either the loudspeakers could be unbalanced (eg, the phase of one of the tweeters could be reversed) or the room acoustics are not symmetrical. Relocating the loudspeakers in the room may improve the situation in the latter case.

Each channel can be individually checked by switching the amplifier to "stereo". One loud sweep is recommended first to check for cabinet buzzes, then at normal levels. As long as the response is smooth, then the tone controls can compensate musical imbalance to suit the listener's taste.

Particular attention should be paid to the loudspeaker crossover points. If the sound dips at crossover, try reversing the phase of the tweeter. If this does not help, the crossover network may be inaccurate, and the easiest remedy is to increase the capacitor in series with the tweeter. Conversely, if the sound peaks at crossover, reduce the filter capacitor. Before embarking on detailed modifications, it would be wise to move the loudspeaker well away from walls and recheck to see if room acoustics are causing interference at this frequency.

The low range of the generator is then useful for checking woofer resonance and cabinet tuning in the case of a bass reflex system. If the bass is boomy and peaky, damping may be improved by filling the enclosure with an acoustic damping material such as bonded acetate fibre or even egg cartons. Frequency doubling can also be checked for at low frequencies – this is a sure sign of an overdriven woofer or perhaps a small diameter port in which the air velocity is too high.

In its basic form, then, the glide tone generator can perform quite a number of important tests on audio equipment. Combined with an oscilloscope, it can give visual as well as audible indications of amplifier performance.

Heart of the glide tone generator is the National Semiconductor integrated circuit, LM3900. This comprises four independent, internally compensated current amplifiers which operate from a single rail supply. The four amplifiers are interconnected to form a voltage-controlled oscillator in which the control voltage waveform is a sawtooth function. Refer now to the circuit.

Amplifier A1 is connected as an integrator which generates a triangular waveform. Amplfier A2 functions as a buffer so that the loading on A1 is reduced and distortion at high frequencies minimised. A3 is connected as a Schmitt trigger which recharges C1 at the end of each sweep. And A4 is a squaring amplifier which interacts with A1 to produce the reversal in slope of the triangular waveform.

The key to the gliding nature of the tone is the potential on C1, which falls exponentially, and is reset every 15 seconds by the Schmitt trigger. When the current through R3 into pin 8 falls below that flowing into pin 3 through R1, the potential on pin 9 rises, and increases the current into pin 3 through R2. C1 is rapidly recharged through R4 and D1.

Without R4 the recharging was found to be too rapid, and the internal resistance of C1 prevented it taking a full charge before the trigger reset. When C1 reaches full potential, the current through R3 exceeds that through R1 and R2, and the potential on pin 9 drops. The exponential discharge then recommences.

The potential on C1 feeds current into pins 1 and 6 of the triangle waveform generator. For symmetry of the triangle waveform, these currents must be in the ratio 1:2 so consequently R5, R6 and R7 all must have the same value, 270k. When Tr1 is conducting, current through R6 and R7 is shunted to the negative supply rail. Current into pin 6 causes the voltage at pin 5 (output of A1) to ramp down, at a rate controlled by the switched capacitor, C3, C4 or C5.

When the voltage at pin 5 is sufficiently low, the squaring amplifier A4 switches and turns Tr1 off. This enables current to flow into pin 1 via R6 and R7 in parallel. Since the current into pin 1 is double that into pin 6, the voltage at the output of A1, pin 5, starts to ramp up (increase linearly).

When the voltage at pin 5 rises to a sufficently high value, A4 switches its output to high which enables Tr1 to conduct again and the output of A1 ramps down again.

The high and low voltage levels of the ramp, at which A4 switches, are controlled by R10, R11, R12 and R13. R12 and R13 could be replaced by a 75k resistor if available.

In the unity gain buffer amplifier, A2, pin 4 follows the potential on pin 5, which creates equal currents in R14 and R15. It can also be called a "voltage follower", although it works by balancing equal currents. Its low impedance output is necessary for driving the diode shaping network, R16 to R20, and D2 to D7.

A triangular waveform can be "rounded over" to approximate a sine waveform quite closely. This is done by feeding the triangular waveform through a suitable resistor-diode network. As the voltage level of the triangular waveform rises above a selected value, the dioes conduct to slightly "round off" the waveform.

Each diode actually changes the slope of the ramp (of the triangle waveform) to make it more gradual. If enough stages of correction are used the approximation to sine wave can be very close, ie 1pc distortion. But there is always a slight discontinuity at the peaks of the resultant waveform.

It is this slight discontinuity at the peaks of the wine waveform plus the slight inflections where the diodes conduct, which give the resultant tone a roughness or "edge". In fact, if you were not expecting it, you could put it down to cross-over distortion in the amplifier or even a faulty tweeter.

Two stages of diode shaping are used in this circuit. When the potential across D2 or D3 exceeds 0.65 volts, R16 and R19 form a potential divider which changes the slope of the ramp. When the potential across D4 to D7 exceeds 1.3 volts, R20 creates a further potential divider.

Since the current drain of the circuit is light at 8mA, a small battery such as the Eveready 216 will be adequate.

The circuitry is obviously capable of elaboration, all the way up to a full function generator. Some features could be developed further if you felt this were justified. A potentiometer in series with a resistor and diode across the 100uF capacitor would give variable sweep rates.

A switch and a potentiometer could be included to give manual control over the frequency, so it could be set to a steady tone. The potentiometer could then be calibrated to give direct indication of frequency. A square wave and triangular wave output could also be added for the cost of the extra terminals and isolating resistors.

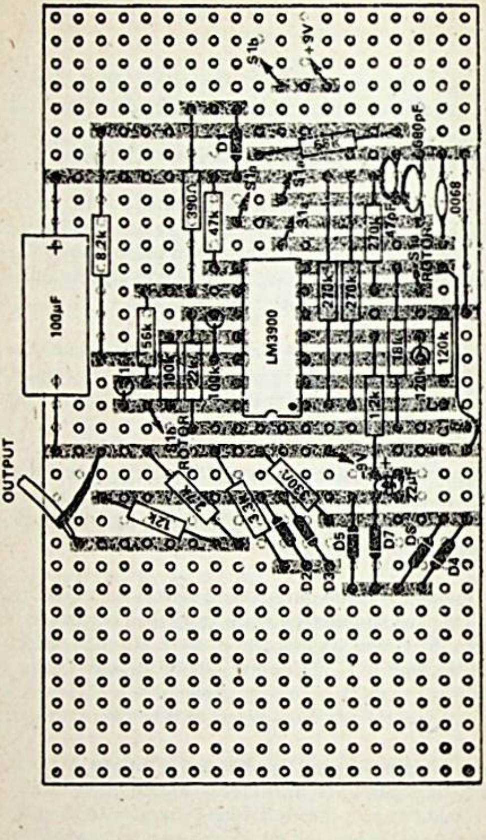
Readers who have access to a distortion meter may be able to improve the waveform shaping network or alternatively, better results might be obtained by placing the waveform shaping elements in the feedback network of the output buffer amplifier, A2.

The time taken on each glide can be varied widely to suit the user's preference by selecting different values for C1. The frequency range in each sweep can be varied by changing R3, with the proviso that R2 is always one preferred value higher.

The actual frequency of the output tone can be corrected by trimming C3, C4, and C5. One simple technique, if a piano is close, is to plug the generator into the stereo system, and observe the tone range. The top range should descend to G, 2½ octaves above middle C. The middle range should glide from D, 3 octaves above middle C, down to E below middle C. The lowest range should start at B below middle C.

Alternatively, Lissajous figures can be used. Connect AC from a low voltage mains transformer to the external sweep input of an oscilloscope, and connect the glide tone generator to the normal Y input. On the lowest range, the last figure seen before the glide resets should be a 1:3 figure – the "ABC trademark", but rotated 90 degrees. Then by measuring the selected C3 on a bridge, C4 and C5 should be one tenth and one hundredth the value, less 20pF, which is the distributed capacity of the circuit. The nominal design figures are 6800, 660, and 48pF respectively.

For most uses, calibration is not really necessary, and most hi-fi enthusiasts will find the generator quite useful when just assembled from unselected components.


OUTPUT
100µF
8.2k
56k
390Ω
47k
LM3900
270k
68k
6800pF
.0068
18k
120k
12k
ROTOR
S1b
+9V
D1
D2
D3
D4
D5
D7

Construction of the glide tone generator is quite straightforward. All components, even the battery holder, are mounted on a small section of Veroboard with 0.1in conductor spacing. Refer to the wiring diagrams for details.

The board plus the range-cum-on-off switch are comfortably accommodated in an aluminium Minibox measuring 102 x 57 x 42 mm. Fit a suitable length of shielded cable for the output signal plus an appropriate input connector for your amplifier. Output signal is 200mV RMS with an output impedance of 2.7k.

PARTS LIST

1 Small utility box, type AMB-6 or similar.
1 Small instrument knob.
1 2-pole 4-position rotary switch.
1 LM3900 quad op-amp IC.
7 1N4148 or similar silicon diodes.
1 Transistor, silicon NPN, BC108 or similar.
1 100uF 10VW electrolytic.
1 22uF 10VW electrolytic.
1 .0068uF ceramic capacitor.
1 680pF ceramic capacitor.
1 47pF ceramic capacitor.

RESISTORS
¼ W: 330 ohms, 390 ohms, 1.2k, 2.7k, 3.3k, 8.2k, 12k, 18k, 22k, 47k, 56k, 68k, 2 x 100k, 2 x 120k, 3 x 270k, 1M.

MISCELLANEOUS
Piece of Veroboard 2in x 3¾ in, battery and connector clip, output cord and connector.

VOICE-OPERATED RELAY

This is a multi-purpose circuit. It can be used as a voice-operated relay to control a tape recorder, or as a VOX circuit for a transmitter. It can be used with any low impedance or high impedance microphone or a high level source such as a tuner.

Quite a few uses can be imagined for this circuit. An obvious use is as a VOX control in a transmitter. It could also be used to control a tape recorder when the material being recorded is of short and spasmodic nature. Another possible use is as a basis for a muting circuit or indeed any sound – controlled function requiring short attack time and slow decay.

The circuit consists of three parts, a microphone preamplifier, a Schmitt trigger and a relay driver. We will start at the left of the circuit and begin with the microphone preamplifier. It is a direct-coupled pair employing two low-noise silicon NPN transistors. Gain is adjustable to suit 600 ohm or 50k dynamic microphones.

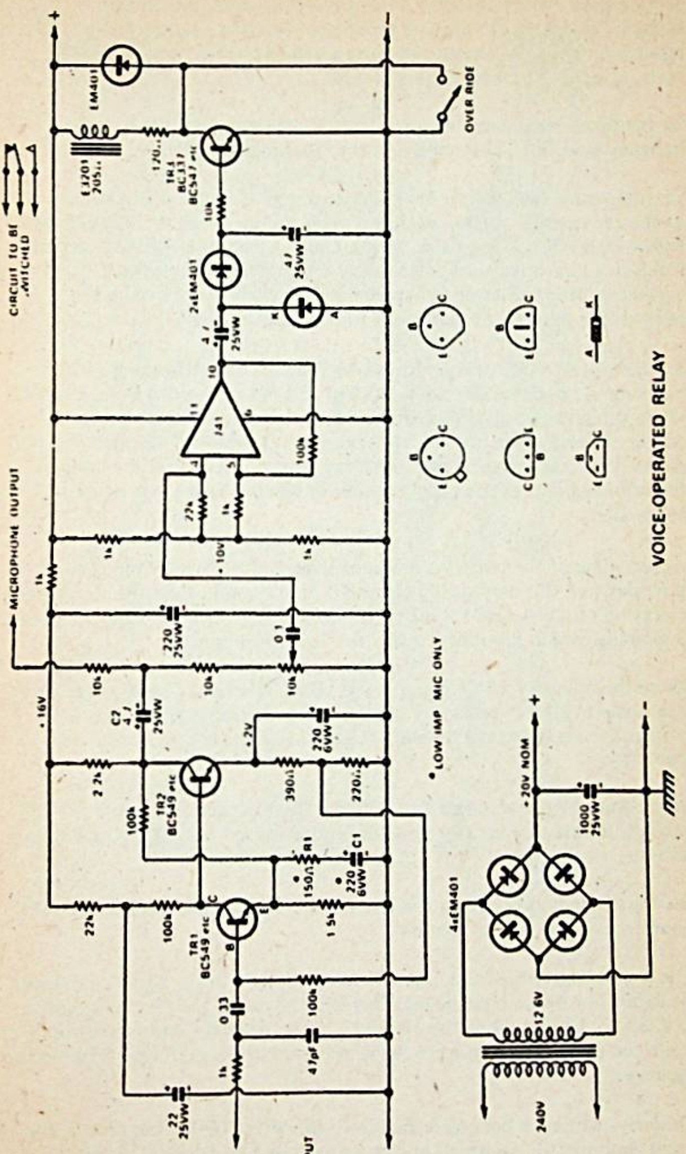
There are two feedback loops in the microphone preamplifier. That from the emitter of the second transistor to the base of the first is DC only and sets the DC conditions in the preamp. The second feedback loop is from the collector of the second transistor to the emitter of the first. This is predominantly AC feedback and sets the gain of the preamplifier.

With R1 and C1 included the gain is high enough to suit 600 ohm microphones. In fact it may be too high for some and may have to be reduced. To halve the gain, simply double R1. With R1 and C1 omitted, the gain is sufficient for normal 50k dynamic microphones.

Output signal from the preamplifier is coupled via C2 to a 10k resistor and 10k trimpot which feeds the op amp Schmitt trigger. The microphone output signal is also available via an additional 10k isolating resistor, to drive a high level input on a tape recorder or whatever.

The Schmitt trigger configuration may seem a little unusual at first sight. A voltage divider consisting of two 1k resistors sets the bias for the inverting and non-inverting inputs of the 741 operational amplifier and thus allows it to function with a single rather than the usual balanced supply rails.

Signals from the trimpot are fed to the inverting input. Notice that there is no negative feedback to the inverting input.Instead, there is a positive feedback to the non-inverting input. This has the effect of setting the hysteresis of the Schmitt trigger. For small signals to the inverting input there is no AC output and in fact the output terminal is at almost the positive rail potential.



VOICE-OPERATED RELAY

When the signal to the inverting input rises above about 170 millivolts peak-to-peak (which is 60mV RMS for a sine-wave) the output suddenly jumps to the limiting condition which is a square wave at the input frequency with amplitude just a little less than the full supply voltage, ie, about 18 volts peak-to-peak.

So the op amp suddenly changes from a zero gain condition to the limiting condition. This contributes to the fast attack time.

An interesting sidelight to the Schmitt trigger is that it does not revert to the zero gain condition until the signal to the inverting input drops below 20mV RMS (for a sine wave). Output from the Schmitt trigger is fed to a half-wave voltage doubler rectifier which charges a 47uF capacitor. This capacitor (when charged) provides base bias to the relay driver transistor to enable it to energise the relay.

So the overall mode of operation is as follows: Input signals to the microphone preamplifier are amplified and fed to the threshold trimpot. When the selected threshold is exceeded, the output of the Schmitt trigger suddenly rises to 18 volts peak to peak which is rectified and fed to the 47uF capacitor to turn the relay driver on and thereby energise the motor circuit of the tape recorder or whatever device is being controlled.

Attack time of the circuit is inherently limited by the closing time of the relay and this is typically about 10 milliseconds. Delay time after the cessation of input signal is set by the size of the 47uF capacitor to about 3 seconds, which should be ample for most purposes.

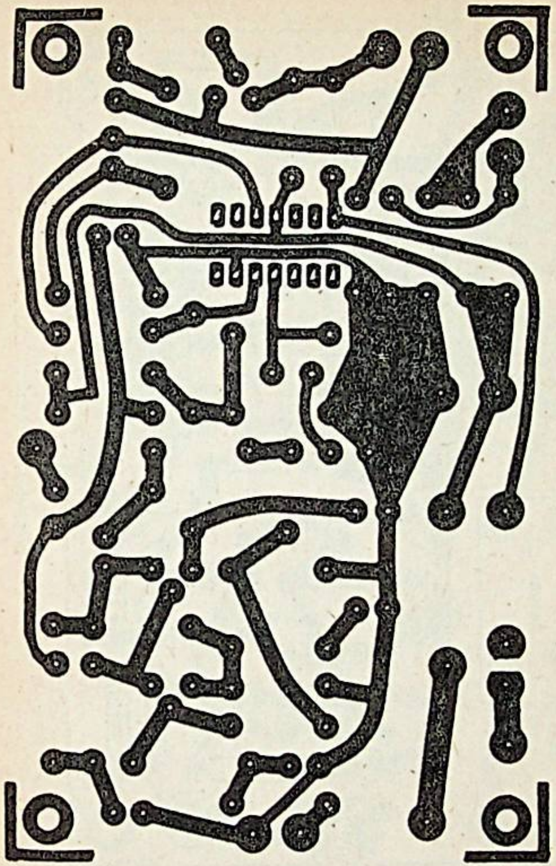
Since the relay is a 12V unit, it is fed via a 120 ohm resistor to obtain the correct voltage across it. A diode across the series combination protects the relay driver transistor against inductive kick-back from the relay.

A full-wave bridge rectifier and 1000uF filter capacitor provide the DC rail requirements from any transformer having a 12 to 13VAC secondary winding.

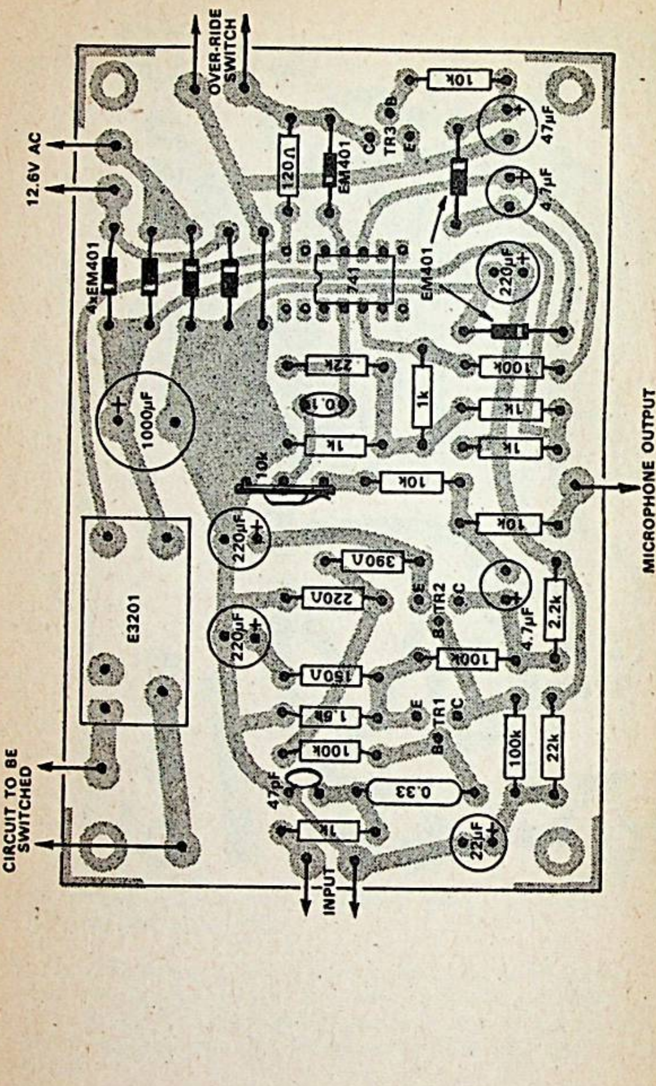
All the components, minus the transformer, are accommodated on a PC board measuring 110 x 70mm.

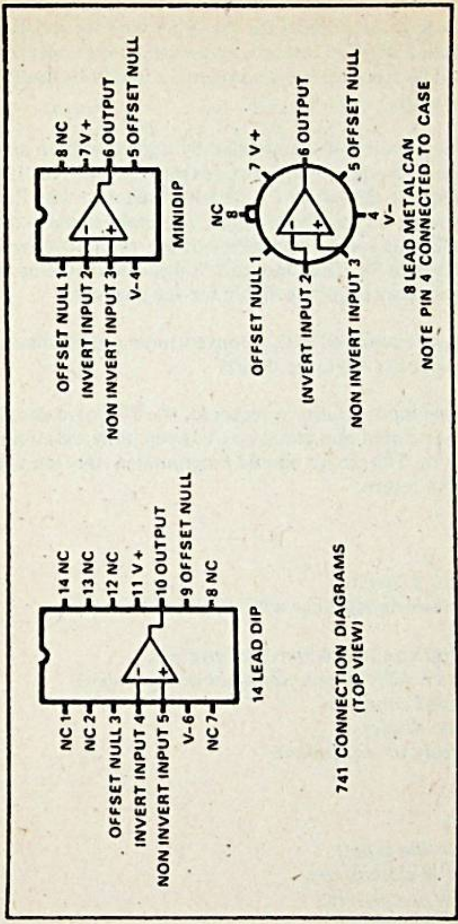
The copper pattern of the board is compatible with the three commonly available 741 op amp packages. This is by virtue of the fact that pins 1, 2, 7, 8, 12, 13 and 14 of the 14-lead package have no internal connection while the remaining pins have the same orientation as those of the smaller packages.

Similarly, while we have specified only BC series TO-92 transistors, the board is compatible with all currently available transistors. Check the transistors you have against the appropriate base diagram on the circuit.



Here is an actual size reproduction of the PC board.


12.6V AC
OVER-RIDE SWITCH
4xEM401
EM401
120Ω
741
TR3
10k
47µF
4.7µF
220µF
1000µF
22k
0.1
1k
100k
10k
E3201
220µF
390Ω
220Ω
150Ω
1.8k
100k
TR2
TR1
2.2k
22k
4.7µF
47pF
0.33
22µF
CIRCUIT TO BE SWITCHED
INPUT
MICROPHONE OUTPUT



741 CONNECTION DIAGRAMS
(TOP VIEW)

In the interests of uniformity we have specified EM401 silicon diodes throughout. If you have some small signal diodes on hand, they may be pressed into service to rectify the output of the Schmitt trigger.

If you substitute a different relay for the types we have specified and intend it to switch a 240VAC circuit, make sure it has appropriate contact ratings. The relay we used is soldered directly to the PC board, no socket being used.

If you intend the circuit to be controlled by signals with an amplitude of 100mV or more, the preamplifier stages may be omitted. Just omit all the components to the left of C2. Then substitute a wire link for the 10k resistor between C2 and the trimpot. The input signal is then coupled in via C2. The source impedance of the signal to drive C2 in this way must be less than 5k. The "monitor" output available on many cassette and tape decks is quite suitable for the purpose.

The trimpot may be replaced with a conventional potentiometer as a panel-mounting control if you so desire.

If the microphone input facility is required, the PC board should be well-shielded or enclosed in a metal box to keep hum and noise to an absolute minimum. The circuit should be grounded only via the microphone input earth return.

PARTS LIST

1 PC board, 110 x 70mm.
1 miniature power transformer with 12.6VAC secondary.
1 SPST switch.
2 BC549 NPN silicon transistors or equivalent.
1 BC547, BC337, NPN silicon transistor or equivalent.
1 741 operational amplifier.
7 EM401 silicon diodes.
1 12V SPDT relay or equivalent.
9 PC stakes.

CAPACITORS
(all PC end-mounting types)
1 x 1000uF/25VW electrolytic.
1 x 220uF/25VW electrolytic.
2 x 220uF/6VW electrolytic.
1 x 47uF/25VW electrolytic.
1 x 22uF/25VW electrolytic.
2 x 4.7uF/25VW electrolytic.
1 x 0.33uF metallised polyester.
1 x 0.1uF metallised polyester.
1 x 47uF ceramic.

RESISTORS
(all ¼ or ½W, 5% tolerance)
4 x 100k, 2 x 22k, 3 x 10k, 1 x 2.2k, 1 x 1.5k, 5 x 1k, 1 x 390,
1 x 220, 1 x 150, 1 x 120 ohms.
1 x 10k trimpot.

NOTE: Components with higher ratings may be used provided they are physically compatible. Lower rated components may also be used in some cases, provided their ratings are not exceeded.

Notes

BERNARDS & BABANI PRESS RADIO AND ELECTRONICS BOOKS

BP1	First Book of Transistor Equivalents and Substitutes	40p
BP2	Handbook of Radio, TV and Industrial & Transmitting Tube & Valve Equiv.	60p
BP3	Handbook of Tested Transistor Circuits	40p
BP4	World's Short, Med. & Long Wave FM & TV Broadcasting Stations Listing (International Edition)	60p
BP5	Handbook of Simple Transistor Circuits	35p
BP6	Engineers and Machinists Reference Tables	30p
BP7	Radio and Electronic Colour Codes and Data Chart	15p
BP8	Sound and Loudspeaker Manual	50p
BP9	38 Practical Tested Diode Circuits for the Home Constructor	35p
BP10	Modern Crystal and Transistor Set Circuits for Beginners	35p
BP11	Practical Transistor Novelty Circuits	40p
BP12	Hi-Fi, P.A., Guitar & Discotheque Amplifier Handbook	75p
BP13	Electronic Novelties for the Motorist	50p
BP14	Second Book of Transistor Equivalents	95p
BP15	Constructors Manual of Electronic Circuits for the Home	50p
BP16	Handbook of Electronic Circuits for the Amateur Photographer	60p
BP17	Radio Receiver Construction Handbook using IC's and Transistors	60p
BP18	Boys and Beginners Book of Practical Radio and Electronics	60p
BP22	79 Electronic Novelty Circuits	75p
BP23	First Book of Practical Electronic Projects	75p
BP24	52 Projects using IC741 (or Equivalents)	75p
BP25	How to Build Your Own Electronic and Quartz Controlled Watches & Clocks	85p
BP26	Radio Antenna Handbook for Long Distance Reception & Transmission	85p
BP27	Giant Chart of Radio Electronic Semiconductor & Logic Symbols	60p
BP28	Resistor Selection Handbook (International Edition)	60p
BP29	Major Solid State Audio Hi-Fi Construction Projects	85p
BP30	Two Transistor Electronic Projects	85p
BP31	Practical Electrical Re-wiring & Repairs	85p
BP32	How to Build Your Own Metal and Treasure Locators	85p
100	A Comprehensive Radio Valve Guide – Book 1	40p
121	A Comprehensive Radio Valve Guide – Book 2	40p
126	Boys Book of Crystal Sets	25p
129	Universal Gram-Motor Speed Indicator (Combined 50 & 60 c/s model)	10p
138	How to Make Aerials for TV (Band 1–2–3)	25p
143	A Comprehensive Radio Valve Guide – Book 3	40p
150	Practical Radio Inside Out	40p
157	A Comprehensive Radio Valve Guide – Book 4	40p
160	Coil Design and Construction Manual	50p
161	Radio, TV and Electronics Data Book	60p
170	Transistor Circuits for Radio Controlled Models	40p
177	Modern Transistor Circuits for Beginners	40p
178	A Comprehensive Radio Valve Guide – Book 5	40p
183	How to Receive Foreign TV Programmes on your Set by Simple Modifications	35p
196	AF-RF Reactance – Frequency Chart for Constructors	15p
200	Handbook of Practical Electronic Musical Novelties	50p
201	Practical Transistorised Novelties for Hi-Fi Enthusiasts	35p
202	Handbook of Integrated Circuits (IC's) Equivalents and Substitutes	75p
203	IC's and Transistor Gadgets Construction Handbook	60p
204	Second Book of Hi-Fi Loudspeaker Enclosures	60p
205	First Book of Hi-Fi Loudspeaker Enclosures	60p
206	Practical Transistor Circuits for Modern Test Equipment	60p
207	Practical Electronic Science Projects	[illegible]
208	Practical Stereo and Quadrophony Handbook	[illegible]
209	Modern Tape Recording Handbook	[illegible]
210	The Complete Car Radio Manual	[illegible]
211	First Book of Diode Characteristics Equivalents and Subst[illegible]	[illegible]
213	Electronic Circuits for Model Railways	[illegible]
214	Audio Enthusiasts Handbook	[illegible]
215	Shortwave Circuits and Gear for Experimenters and Radi[illegible]	[illegible]
216	Electronic Gadgets and Games	[illegible]
217	Solid State Power Supply Handbook	[illegible]
218	Build Your Own Electronic Experimenters Laboratory	[illegible]
219	Solid State Novelty Projects	[illegible]
220	Build Your Own Solid State Hi-Fi and Audio Accessories	[illegible]
RCC	Resistor Colour Code Disc Calculator	[illegible]

The Grampians, Shepherds Bush Road, London W6 7N[illegible]
Telephone: 01–603 2581/7296