Single IC Projects

R. A. Penfold

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PREFACE

There is now a vast range of I.C.s available to the amateur user, many of which are intended for some specific function, such as a stereo multiplex decoder or an audio amplifier of some kind.

However, the majority of I.C.s are not necessarily designed for use in a single application and can offer almost unlimited possibilities. The most obvious examples are the operational amplifier and the digital I.C. These represent very useful electronic building blocks which are used almost as frequently in amateur electronic designs as the transistor.

Devices of this type are featured in several of the projects described here, and all such projects have some proper application, and do not just demonstrate the properties of the device concerned.

All projects are relatively simple to construct and set up (where necessary), and are based on a single I.C. A few projects employ one or two transistors in addition to the I.C., but in most cases the I.C. is the only active device used.

A stripboard layout is provided for each project, together with any special constructional points and setting-up information, making this book suitable for beginners as well as more advanced constructors.

R. A. PENFOLD

CONTENTS

CHAPTER 1.

LOW LEVEL AUDIO CIRCUITS

MC3340P Electronic Attenuator (MFC6040)

This device is an electronic attenuator which is contained in a standard 8 pin DIL plastic package. It formerly had the type number MFC6040 and was contained in a 6 pin Quad-in-line package, but this version is now obsolete.

Principal Ratings

Supply voltage	9V. min. 18V. max.
Gain	13dB. (typical) 11dB. min.
Maximum input voltage	500mV. R.M.S. (typical).
T.H.D. at full gain	0.6% (typical).
Attenuator Range	90dB. (typical) 70dB. min.

Volume Expander

Recording and transmission systems in common use provide only a relatively low dynamic range (the difference between the noise level of the system and the maximum level that can be handled without serious distortion), and this inevitably leads to compromises in order to give acceptable results on difficult material having a very wide dynamic range. Without such compromises the signal would either be lost in the background noise when at or near its minimum dynamic level, or peak level signals would be clipped and seriously distorted. Possibly even both would occur.

the recording or sound engineer therefore has to manipulate the dynamic levels so that low level signals are raised above the noise, and (or) high level signals are attenuated to avoid clipping. However skilfully this is done there is no escaping the fact that the dynamic range of the signal has been compressed, and the resultant audio signal may lack impact and effect when compared to the original because of the reduction in dynamic range.

A volume expander is a device which boosts high level signals somewhat, but does not affect the gain on low level signals. Thus with the volume at its normal level on quiet signals, high level signals will be boosted beyond their normal volume, giving the signal more impact due to the increased dynamic range. As the noise level remains unaltered, but the maximum volume level is increased, a volume expander can give an improved signal to noise ratio. The noise will be boosted along with high level signals of course, but this tends not to be very noticeable since high volume signals mask background noise.

The Circuit

The MC3340P can readily be used in many types of audio control circuits and effects units, but apart from use as a remote volume control, some additional active circuitry is invariably required, and as can be seen from the circuit diagram Fig. 1, this circuit is no exception.

Pin 1 of the MC3340P is the input terminal, and the input signal is coupled to this by C3. The output is available at pin 7 and is taken to the output socket by D.C. blocking capacitor C5.

The gain of the device can be controlled by either connecting a control voltage to pin 2, or by using a control resistance between pin 2 and the negative rail. Full gain is available with a voltage of 3.5V. or less at pin 2, or a resistance of less than 4k connecting it to the negative rail. Gain is reduced by 10, 20, 40, 60, and 80dB. with nominal control voltages of 4.3, 4.7, 5.2, and 5.8 volts respecitvely. For the same attenuation levels, control resistances of 7k, 8.8k, 12k, 17k, and 25k respectively are required.

The series resistance of R3 and R4 form a control resistance, and R3 is adjusted to reduce the gain of the device by about 10 to 12dB. When used at full gain the MC3340P has a gain of about 13dB., and so this gives the unit a quiescent gain of fractionally more than unity.

Part of the input signal is taken via C2 to a simple common emitter amplifier using Tr1. The amplified output is coupled by C7 to a rectifier and smoothing circuit which is comprised of D1, D2, and C6. This generates a D.C. positive bias which is proportional to the input signal level. This bias is fed to the base terminal of Tr2 through R5, and will cause Tr2 to conduct. The higher the signal level, the more heavily Tr2 will conduct.

Tr2 is connected so that its collector – emitter impedance shunts the control resistance. Therefore, on high level signals Tr2 conducts quite heavily and significantly reduces the control resistance, causing the gain of I.C.1 to increase. VR1 controls the effective voltage gain of Tr1, and this is adjusted so that the peak input signal level causes Tr1 to conduct sufficiently heavily to raise the gain of I.C.1. to maximum. Thus the required volume expansion action is produced.

The circuit has hysteresis, which simply means that it responds very quickly to increases in the dynamic level by boosting the circuit gain, but responds a little less slowly to decreases in dynamic level, although the decay time is still quite short. This technique is commonly employed in audio control circuits, and gives the best subjective results.

Construction and Adjustment

The obsolete MFC6040 device will also fit into the layout

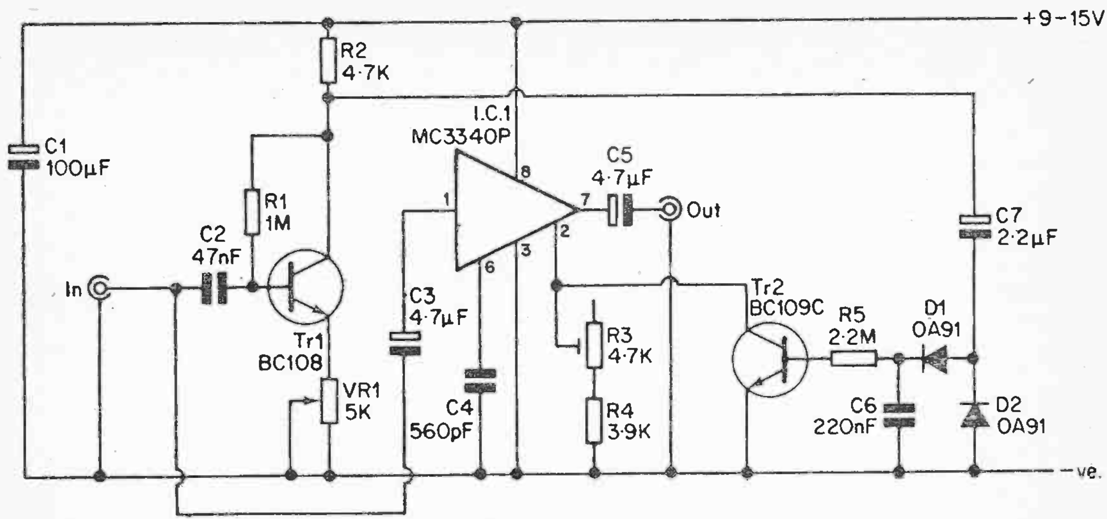

Fig. 1. The circuit diagram of the expander.

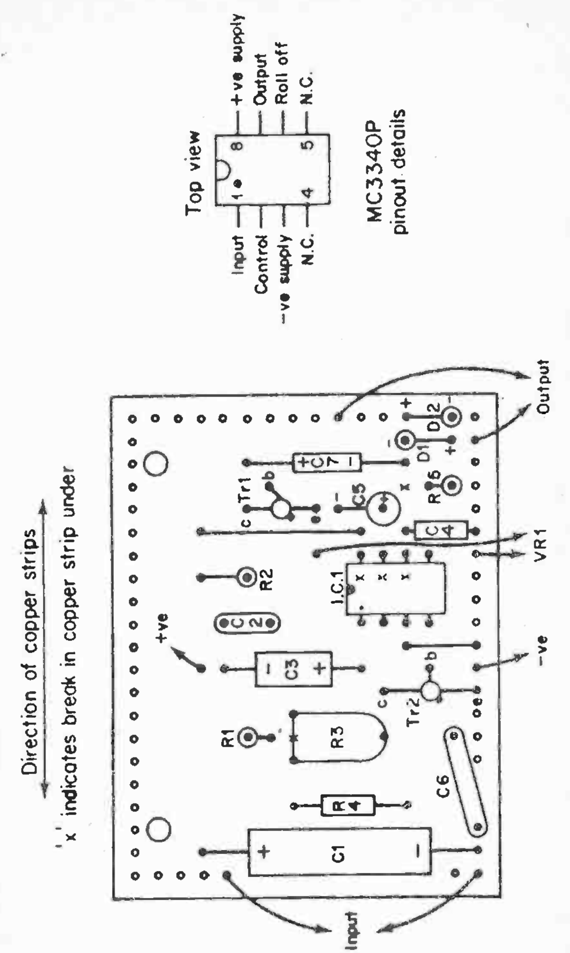

Fig. 2. The stripboard layout of the expander unit.

shown without difficulty if the constructor happens to have a suitable device to hand. For stereo operation two units are required, the two channels being processed separately. A nominal supply voltage in the range 9 to 15 volts is needed, and the current consumption of a mono unit is approximately 14mA.

If suitable test gear is available, R3 can be adjusted to reduce the voltage gain of the unit by about 10 to 12dB. from its minimum resistance setting. Tr2 base should be shorted to the negative supply rail to prevent Tr2 from affecting the measurements.

If suitable test gear is not available, this control must be set by trial and error. Too low a resistance will manifest itself as a lack of effectiveness, whereas an excessive resistance will give too much expansion which will be clearly audible and totally unsatisfactory.

The best setting for VR1 will need to be found by empirical means. Too much gain (set for too low a resistance) will result in the expansion being applied at a fairly low level and rather abruptly. This will give an expansion characteristic which will readily be heard in operation, and which will not be satisfactory. Full expansion will not be achieved if there is too little gain.

The expander can be connected between virtually any tuner or tape deck and an amplifier, and can be adjusted to suit a wide range of input levels. It can only be fed from a crystal or ceramic cartridge if a high input impedance buffer amplifier is added at the input. Similarly, it can only be fed from a magnetic pick-up via suitable preamplifier. If the amplifier or receiver has a tape monitor facility or something similar, the expander can be connected into this part of the system.

Components EXPANDER

Resistors.	All fixed values are miniature ¼ watt 5 or 10%
R1	1 Meg
R2	4.7k
R3	4.7k 0.1 watt horizontal preset
R4	3.9k
R5	2.2 Meg
VR1	5k lin carbon

Capacitors.

C1	100MFD 25vw
C2	47nF type C280
C3	4.7MFD 25vw
C4	560pF polystyrene or ceramic
C5	4.7MFD 25vw
C6	220nF mylar or type C280
C7	2.2MFD 25vw

Semiconductors.

I.C.1	MC3340P (or MFC6040)
Tr1	BC108
Tr2	BC109C
D1	OA91
D2	OA91

Miscellaneous.
Power Source
0.1in. matrix stripboard panel. (all projects in this book)
Control kobs, wire, solder, transistor and l.e.d. clips, i.c. holders, etc. to suit each project.

MC1310P MPX Stereo Decoder

The MC1310P and its many equivalent devices are contained in a standard 14 pin DIL plastic package and can be used as the basis of an MPX stereo decoder which is easy to set up for use due to the absence of any inductors.

Principal Ratings

Supply voltage	16V. max.	8V min.
Input impedance	50k typ.	20k min.
Maximum input level for 0.5% T.H.D. (1% mono)	560mV. R.M.S. min.	

Stereo separation (1kHz)	40dB. typ.	30dB. min.
Lamp current	75mA. max.	
Stereo switch level (19kHz pilot tone level)	20mV. max.	

The Circuit

This is shown in Fig. 3. The composite input signal is applied at pin 1 via D.C. blocking capacitor C2. In order to avoid the necessity for the coils and complicated alignment associated with earlier types of MPX decoder, the MC1310P utilizes a phase locked loop (P.L.L.) circuit. This has a voltage controlled oscillator operating at 76kHz, and it is fed to the phase comparator by way of two divide-by-two circuits, giving a 19kHz signal. This enables the 76kHz oscillator to lock onto the 19kHz pilot tone, and the 38kHz output from the first frequency divider stage is, in effect, the required frequency doubled pilot tone. This is then used to decode the MPX signal using what is basically just a switching type decoder circuit.

The operating frequency of the V.C.O. is determined by R1, R2, and C7. R1 is adjusted to give the correct V.C.O. frequency, and is the only adjustment that needs to be made to the finished unit. The P.L.L. filter components are R5, C8, and C9. C4 is the filter capacitor for the stereo switching circuitry, and C3 is a coupling capacitor between two stages within the MC1310P. D1 is the stereo indicator beacon and R6 is a current limiting resistor. The lamp current is about 10mA. or so, depending upon the supply voltage used. R3 and R4 are load resistors for the output stages of the I.C., and C5 plus C6 are the de-emphasis capacitors. C1 and C10 provide D.C. blocking at the outputs.

Automatic mono/stereo switching is a feature of the MC1310P, but a manual stereo mute control (S1) is also fitted in this circuit. This can be useful when listening to a very weak stereo station which has a unacceptably high high background noise level. Closing S1 will force the decoder into the mono mode and will often drastically improve the signal to noise ratio.

Note that in practice there is always some slight reduction in signal to noise ratio when a stereo decoder is added to a tuner and it is in the stereo mode. Indeed, unless a reasonably strong aerial signal is obtained, stereo operation will produce a much lower signal to noise ratio, and in poor or mediocre reception areas this may well necessitate an improved aerial system. On mono signals or when in the mono mode the decoder will not significantly reduce the signal to noise ratio.

Construction And Use

Although the decoder is very simple to align and construct, it is recommended that inexperienced constructors should not attempt to add the unit to a tuner unit unless they are quite sure they know what they are doing.

The unit requires an input level of about 200mV. to 550mV. R.M.S., and most recent tuner circuits employing a quadrature type detector will supply a suitably high signal level. Tuners having other types of detector (ratio discriminator, etc.) may well have an inadequate output and require some audio preamplification, although in many cases it is likely that this amplification will be already present in the tuner circuitry.

Some mono tuners have an output for an MPX decoder, but if such an output is not available, the ordinary mono output can be used provided the de-emphasis component(s) in the tuner are removed. Otherwise the 19kHz pilot tone and ultrasonic components of the MPX signal would be seriously attenuated, and proper operation would not be possible.

If used with positive earth equipment, the polarities of C1, C2, and C10 should be reversed. A radio set for mains operation and not incorporating an isolation transformer should not be used as the tuner for this decoder for reasons of safety.

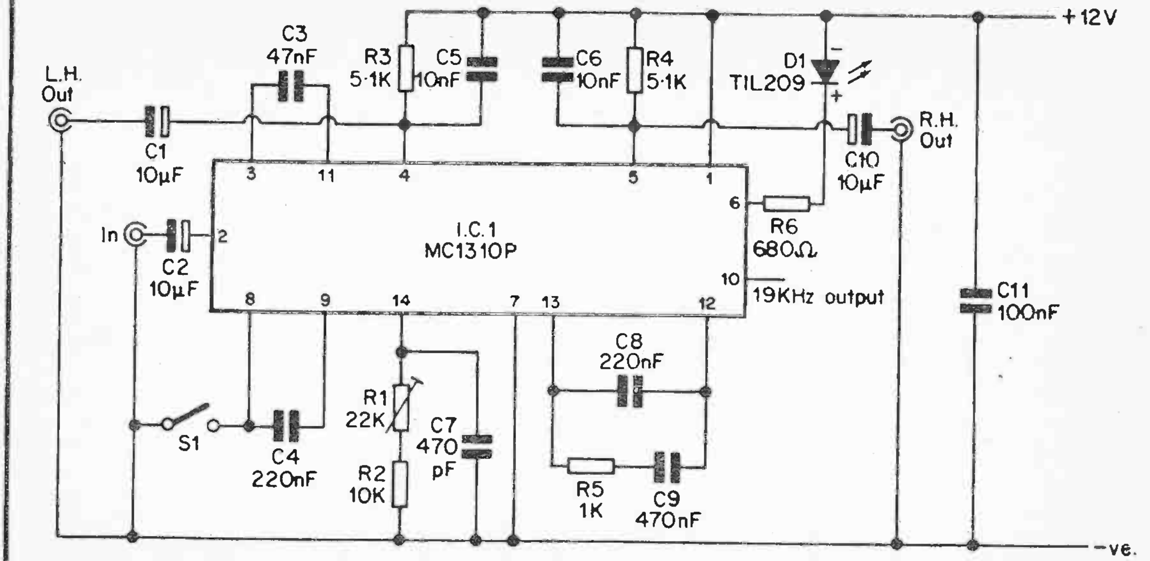

Fig. 3. The circuit diagram of the MC1310P stereo decoder.

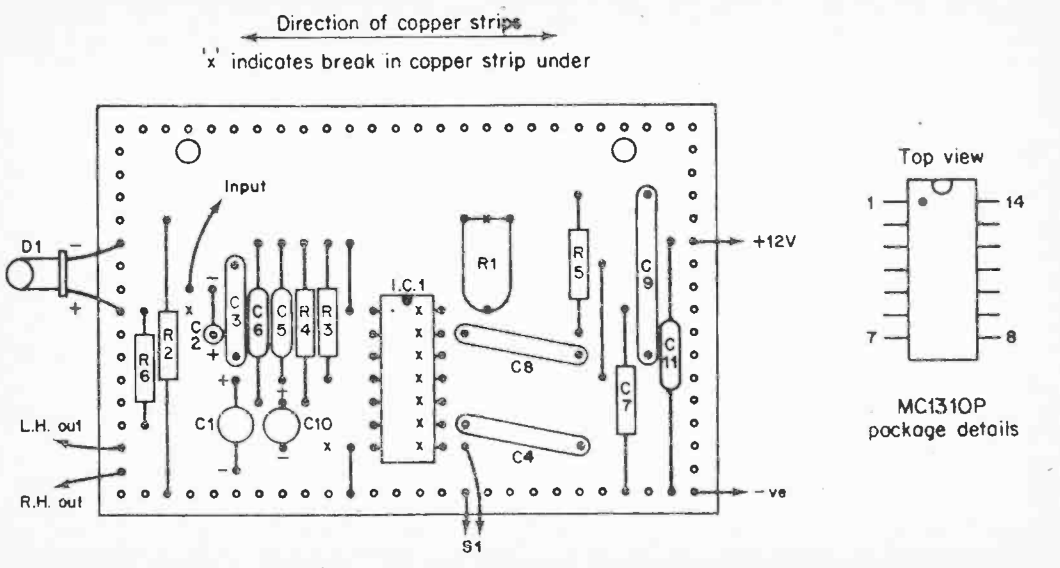

Fig. 4. The 0·1 in. pitch stripboard component panel for the stereo decoder.

Ideally the unit should be aligned with the aid of a frequency counter. With this connected to the 19kHz output at pin 10 of I.C.1, and no input coupled to the decoder, R1 is adjusted so that an output frequency of 19kHz is indeed obtained.

If a suitable frequency meter is not available, with the decoder coupled to a MPX signal it should be found that R1 has a small range of settings over which D1 comes on. Satisfactory results should be obtained with R1 set to approximately the centre of this range of settings.

Components **MPX STEREO DECODER**

Resistors.	All fixed values are miniature ¼ watt 5 or 10%
R1	22k 0.1 watt horizontal preset
R2	10k
R3	5.1k
R4	5.1k
R5	1k
R6	680 ohms

Capacitors.	
C1	10MFD 16vw
C2	10MFD 16vw
C3	47nF type C280
C4	220nF type C280
C5	10nF type C280
C6	10nF type C280
C7	470pF polystyrene 5% or better
C8	220nF type C280
C9	470nF type C280
C10	10MFD 16vw
C11	100nF type C280

Semiconductors.	
I.C.1	MC1310P or equivalent (KB4400 etc.)
D1	TIL209

Switch.
S1 S.P.S.T. toggle or sub-miniature toggle type.

Miscellaneous.
Wire, solder, clip for D1, etc.

LM382N Low Noise Audio Preamplifier

This device can be used in a number of preamplifier applications, but it is primarily intended for the amplification of very low level signals where equalisation is required. One example of such a circuit is the RIAA preamplifier described below, but there are other circuits of this general type, such as a tape head preamplifier.

Principal Ratings

Supply voltage	9V. min.	40V. max.
Open loop voltage gain	100dB. typ.	
Supply ripple rejection	120dB. typ.	
Total equivalent input noise	0.8μV. typ.	

Magnetic Cartridge Preamplifier

The LM382N consists basically of two low noise operational amplifiers contained in the same standard 14 pin DIL package. It also contains feedback resistors which can be used to set the closed loop voltage gain at certain levels, or as in this case, they can be used in conjunction with discrete capacitors to both determine the closed loop gain and provide the desired equalization curve.

Fig. 5 shows the circuit diagram of the LM382N magnetic

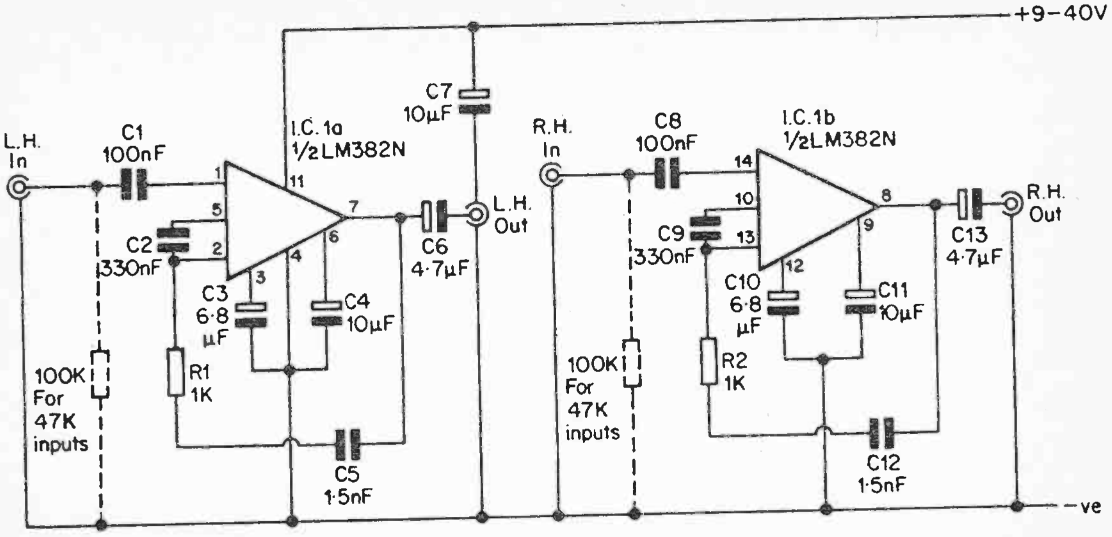

Fig. 5. The circuit diagram of the LM382N magnetic cartridge preamplifier.

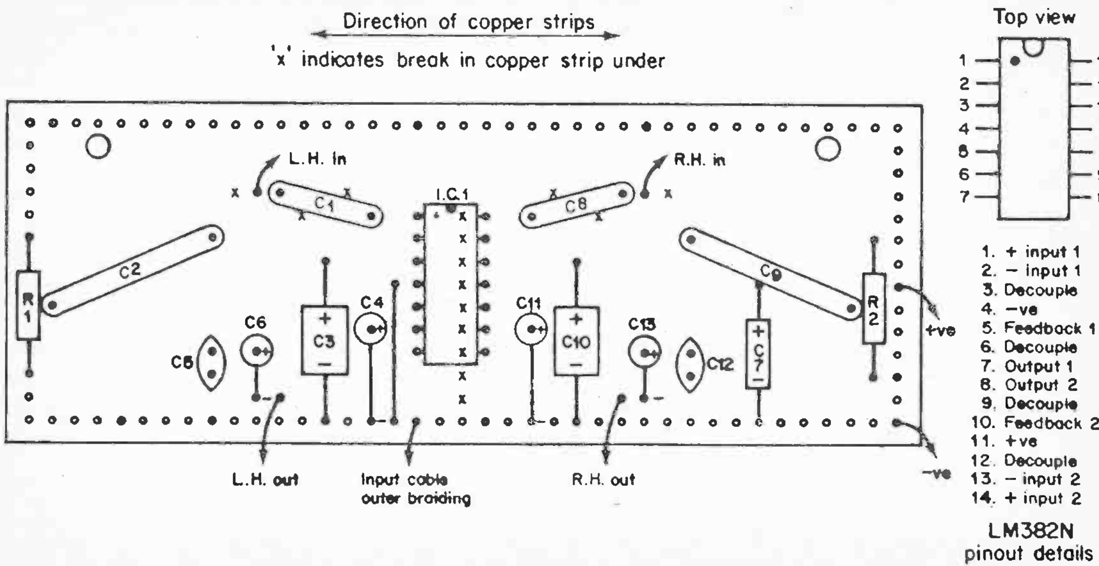

Fig. 6. The 0·1 in. matrix stripboard layout for the LM382N magnetic cartridge preamplifier.

cartridge preamplifier. The left hand input signal is coupled to the non-inverting input of one section of the device by C1, and the output signal is extracted via D.C. blocking capacitor C6.

The external frequency shaping components are C2, R1, and C5, and these give the appropriate treble cut and bass boost to give a nominally flat frequency response from a standard magnetic pick-up. C3 and C4 are decoupling capacitors which are connected in series with internal shunt feedback resistors of the LM382N.

As will be apparent from Fig. 5, the right hand amplifier is essentially the same as the left hand channel, but is based on the other section of the I.C. The two amplifiers share common supply pins and decoupling capacitor, the latter being C7.

Performance

Although not the ultimate in low noise preamplifiers, the LM382N has an extremely low noise level, and in practice it is likely to be the pick-up of stray hum that determines the signal to noise ratio of the unit.

The circuit has an input impedance of approximately 100k, and this can be reduced to about 47k by adding a 100k resistor across the input, if it is to be used with a cartridge having a recommended load impedance of 47k rather than 100k. The output impedance of the preamplifier is only about 30 ohms, and normal power amplifier or tone control circuits will not have any significant loading effect on the circuit.

The mid-band gain of the circuit is a little over 40dB., and this will provide an output level of about 500mV. R.M.S. from most magnetic pick-ups.

Construction

A common problem with units of this type is the pick-up of mains hum, and for this reason it is recommended that the unit is housed in a metal case to provide screening, whether it is to be used as a self contained add-on unit or as an integral part of a complete amplifier system. Of course, screened input leads must be used, and provided these are individually screened, a channel separation of 40dB. or more should be attained. Due to the low output impedance of the circuit it is not essential to use screened output leads, although it would probably be as well to do so if these leads are to be more than a few inches long.

The circuit can be powered from any supply voltage in the range of 9 to 40 volts, and due to the high ripple rejection of the LM382N it is not necessary to have a very well smoothed and decoupled supply. In the interest of obtaining a good overload margin from the circuit it is advisable not to use a supply voltage of less than 12 volts.

Components **MAGNETIC CARTRIDGE PREAMPLIFIER**

Resistors.	Both miniature ¼ watt 5%
R1	1k
R2	1k

Capacitors.	
C1	100nF type C280
C2	330nF type C280

Capacitors.	
C1	100nF type C280
C2	330nF type C280
C3	6.8MFD 25vw
C4	10MFD 25vw
C5	1.5nF ceramic or polystyrene 5%
C6	4.7MFD 25vw
C7	10MFD 40vw
C8	100nF type C280
C9	330nF type C280
C10	6.8MFD 25vw

C11	10MFD 25vw
C12	1.5nF ceramic or polystyrene 5%
C13	4.7MFD 25vw

Integrated Circuit.
I.C.1 LM382N

Miscellaneous.
Metal case.
Phono input and output sockets (or other preferred types).
Power source.

MC1312P(Q) SQ Quadrophonic Decoder

This device contains all the active circuitry necessary for a SQ quadrophonic decoder, and only requires the addition of disrete R – C phase shift networks plus input and output D.C. blocking capacitors. This gives only a basic SQ decoder which does not have the logic circuits to enhance the apparent front to back separation that are used in some commercial amplifiers, but a basic decoder of this type gives good results and the logic enhancement circuitry is not essential. The MC1314 and MC1315 devices together with the appropriate discrete circuitry can be used to provide this enhancement.

Principal Ratings

Supply voltage	20V. type 25V. max.
Input impedance	3 Meg. typ. 1.8 Meg. min.
Channel balance	± 1dB. (0dB. typ.).
Voltage gain	± 1dB. (0dB. typ.).
Min. input voltage for 1% T.H.D.	2V. R.M.S.

Typical T.H.D. for 500mV. R.M.S. input	0.1%
Signal to noise ratio referred to 500mV. R.M.S.	−80dB. typ.
Output impedance	5k typ.

The Circuit

The SQ quadrophonic system was developed by C.B.S., and it uses a system of phasing to encode the four quadrophonic channels into two channels. These can be carried by any normal two channel medium, although the system is only commonly used in disc recording. The signal is compatible with ordinary stereo equipment and will produce a conventional stereo image, but with a suitable decoder using the appropriate phase shift, summing, and differentiator arrangement, four quadrophonic channels can be recovered. In common with all quadrophonic systems using this general principle, there is only a fairly low level of separation between certain channels of the system, but quite a good quadrophonic effect can be obtained with only a modest level of channel separation. The circuit appears in Fig. 7.

The unit consists basically of a summing network, a differential network, two buffer amplifiers, and two 90 degree phase shift networks. Most of the discrete components are the R – C arms of the phase shift networks which are of the Wien type, and give a 90 degree phase shift (± 8.5 degrees) from 100Hz to 10kHz. The only discrete components which are not part of the phase shift networks are input D.C. blocking capacitors C12 and C14, output D.C. blocking capacitors C3, C4, C6, and C7, and supply decoupling capacitor C1. All the other circuitry is contained within the MC1312 device.

Construction And Use.

The layout is designed for the MC1312PQ device, which is

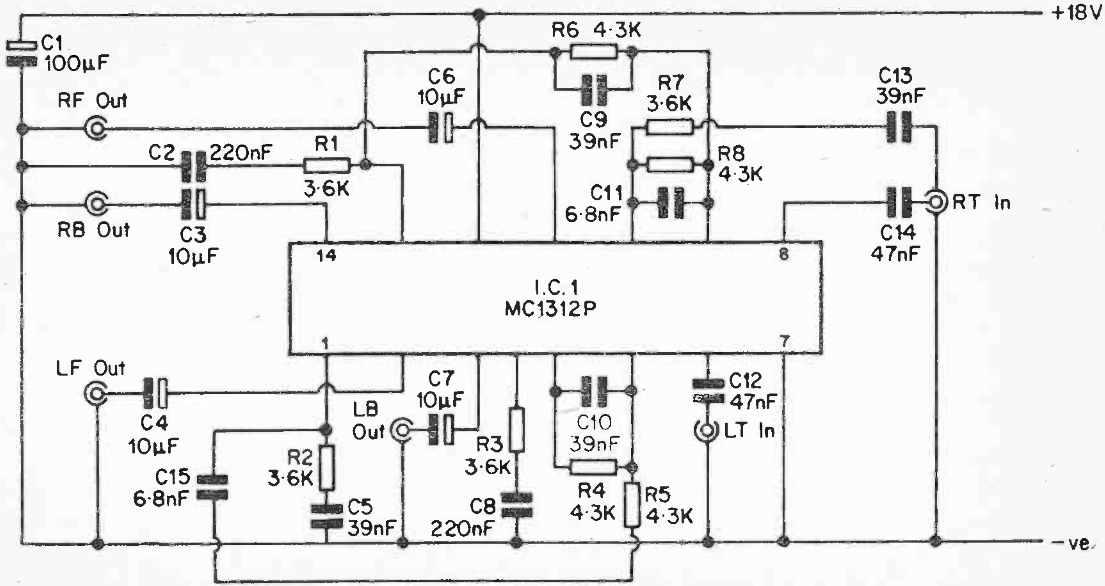

Fig. 7. The circuit diagram of the MC1312P(Q) SQ quadrophonic decoder.

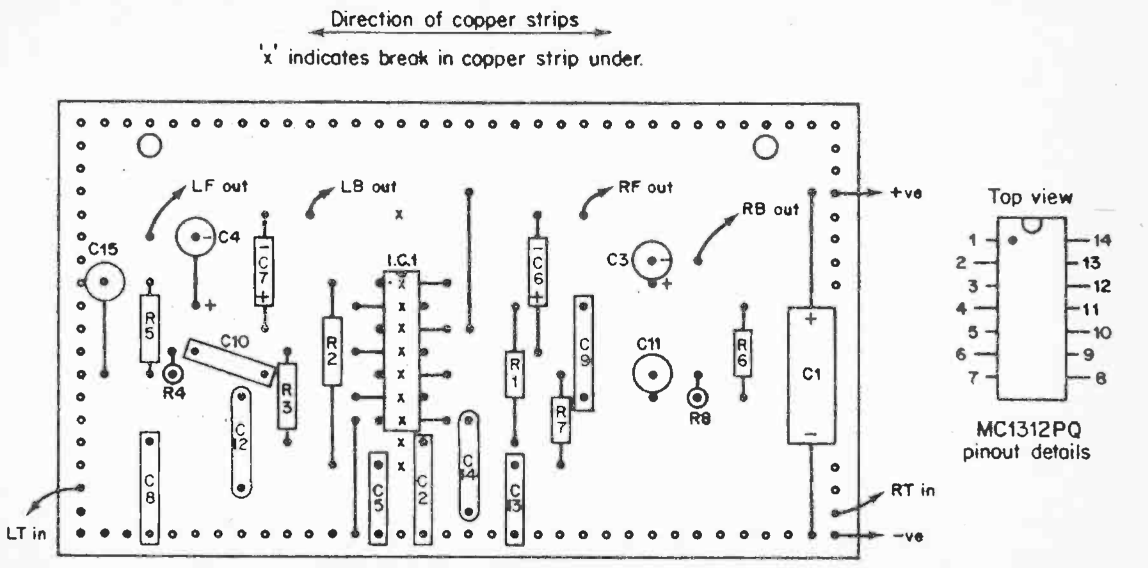

Fig. 8. The 0·1 in. pitch stripboard layout for the SQ decoder.

contained in a standard 14 pin Quad-in-line (QUIL) plastic package. It could easily be adapted to accept the MC1312P 14 pin DIL version which has the same basic pin-out arrangement.

With the SQ system the two left hand signals are encoded into a single signal, and the two right hand channels are similarly encoded, giving what are normally termed the left total and right total signals respectively. These terms are abbreviated to LT and RT respectively in Figs. 7 and 8. The decoded signals are normally fed to four speakers placed at the four corners of a rectangle, preferably some 8 to 16 feet between adjacent speakers, and with the listener at roughly the centre of the rectangle. The four signals are normally called the left front, right front, left back, and right back, and are abbreviated to LF, RF, LB, and RB respectively in Figs. 7 and 8. There are other speaker arrangements which give good results, and it may be worth experimenting a little here. Ideally the four speakers should all be of the same type, but quite good results will probably be obtained if the rear speakers are of slightly lower quality than the front pair. It is advisable to ensure that the speakers all have the same phasing, if possible.

The decoder requires a reasonably well smoothed supply of about 18 to 20 volts and the current consumption will be in the region of 10 to 20mA. The high input impedance of the unit enables it to be fed from a crystal or ceramic cartridge successfully without the need for any preamplifier, and the output can feed a high level input of the amplifier (or amplifiers if two stereo units are used). If a magnetic cartridge is to be used, either the decoder should be connected into the 'tape monitor' or some similar facility of the amplifier so that its RIAA preamplifier can be utilised ahead of the decoder, or a separate preamplifier (such as the one described in the previous section) must be utilized. The unit can also be fed from a tuner or a tape deck, and will act as a form of quadrophonic synthesiser when fed with an ordinary stereo signal.

SQ decoders often use a certain amount of front and rear channels blending in order to improve centre front to centre back separation. A normal 10% front channel blend and 40% rear channel blend can be achieved by adding a 47k resistor between pins 2 and 11 of I.C.1, and a 7.5k resistor across pins 3 and 14.

SQ encoded records are produced by C.B.S., and many SQ encoded records are produced by E.M.I. and are available in the U.K. through normal retailers of E.M.I. records.

Components MC1312P(Q) SQ DECODER

Resistors.	All miniature ¼ watt 5%
R1	3.6k
R2	3.6k
R3	3.6k
R4	4.3k
R5	4.3k
R6	4.3k
R7	3.6k
R8	4.3k

Capacitors.	
C1	100MFD 25vw
C2	220nF polycarbonate 5%
C3	10MFD 25vw
C4	10MFD 25vw
C5	39nF polycarbonate 5%
C6	10MFD 25vw
C7	10MFD 25vw
C8	220nF polycarbonate 5%
C9	39nF polycarbonate 5%
C10	39nF polycarbonate 5%
C11	6.8nF polystyrene 5%
C12	47nF type C280
C13	39nF polycarbonate 5%
C14	47nF type C280
C15	6.8nF polystyrene 5%

Integrated Circuit

I.C.1 MC1312P or MC1312PQ

CHAPTER 2

AUDIO POWER AMPLIFIERS

LM380N Audio Amplifier

The LM380N is a deservedly popular device which can provide a high quality output of up to about 2.5 watts, and requires minimal discrete circuitry. The normal version of this device is contained in a standard 14 pin DIL package, but an 8 pin version is available from a few sources, and is suitable for low power applications. The 14 pin version is the one almost invariably used in amateur electronic designs, and is the version which will be considered here.

Principal Ratings.

Supply voltage		8V. min.	22V. max.
Voltage gain	40 min.	50 typ.	60 max. (V/V)
Supply current (quiescent)		7mA. typ.	25mA. max.
Bandwidth at 2W into 8 ohms		100kHz.	
Output power into 8 ohms with 18 volt Sup. and suitable heatsinking		2.5 watts R.M.S. min.	
Input resistance		150k typ.	
T.H.D. at 1kHz		0.2% typ.	

Testbench Amplifier

The LM380N is a very versatile device which is suitable for most low to medium power applications. The circuit shown in Fig. 9 is for a testbench amplifier which is a very useful item of equipment to have in the electronics workshop. The circuit has many other possible applications however such as the audio amplifier section of a portable radio for example.

Like an operational amplifier, the LM380N has both inverting (–) and non-inverting (+) inputs, but unlike an operational amplifier it has an internal feedback network which sets the

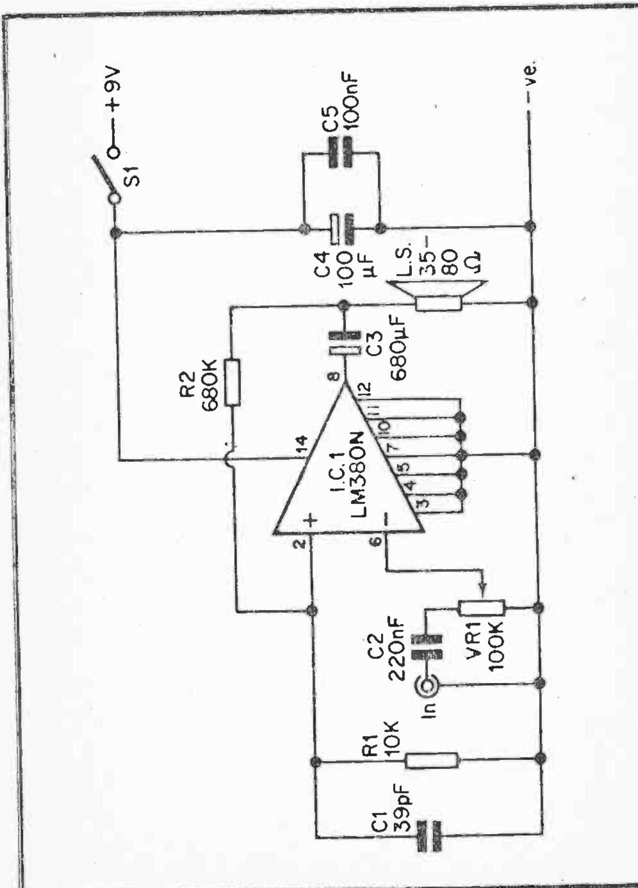

Fig. 9. The circuit diagram of the testbench amplifier.

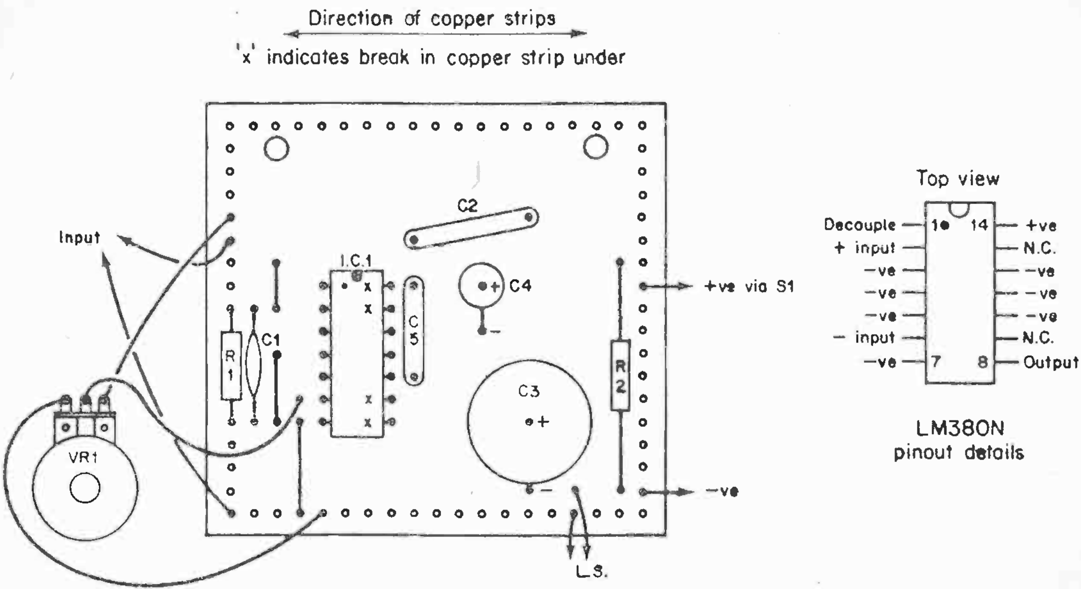

Fig. 10. Constructional details of the test bench amplifier.

closed loop gain at typically 50 times. The input signal can be applied to either input, and in this case it is the inverting input to which it is coupled. VR1 is the volume control, and no D.C. blocking capacitor is needed between its slider and the input of the LM380N. However, any D.C. bias on the input signal must not be fed to the input of I.C.1, and so D.C. blocking capacitor C2 is used at the input of the amplifier. C3 provides D.C. blocking at the output.

Although the LM380N is often regarded as a fixed gain device due to its internal negative feedback components, its gain can be decreased by an external negative feedback loop. It can also be increased by using a discrete positive feedback loop, and this is necessary in this instance since a testbench amplifier should ideally have high sensitivity.

The positive feedback components are R1 and R2, and these boost the gain to about 150 times. This gives an input sensitivity of about 15mV. for maximum output power. C1 reduces the amount of feedback at high frequencies, and aids good stability. Note that the loss through the feedback circuit of R1 and R2 must be greater than the normal gain of the LM380N, or oscillation will result, and so it is advisable not to substitute near values for either of these components.

C4 and C5 are supply decoupling components and S1 is the on/off switch.

The output power of the circuit depends upon the speaker impedance used. In this application a high output power is not really required, and a high impedance speaker will be satisfactory. The output power varies from about 180mW. with a 35 ohm load down to about 80mW. with an 80 ohm speaker. Higher output powers can be obtained by using a lower impedance speaker, with about 750mW. being available into the minimum recommended load impedance of 8 ohms. The LM380N has a class B output stage, and although the quiescent output current is only about 7mA., it will be much higher than this on volume peaks, reaching more than 100mA. if an 8 ohm load is used. The LM380N has thermal overload and output short circuit protection circuitry. In applications where an output power of a couple of watts or so are required, pins 3 to 5 and 10 to 12 of the device should be connected to a heatsink, which can conveniently be an area of p.c.b. copper laminate. With the output powers involved in this case there is no need for any heatsinking.

Construction

It is recommended that the unit be housed in a metal case, earthed to the negative supply rail so as to provide overall screening of the circuit. There is then no need to screen the internal input wiring of the amplifier from mains hum and other stray pick-up, or against stray feedback over the amplifier, as the input and output are out of phase and the feedback will not cause instability. Unless the amplifier is fed from a low impedance source it will be necessary for external connecting leads to be screened. The input impedance of the circuit is about 100k, and it will have little loading effect on normal audio signal sources.

Components TESTBENCH AMPLIFIER

Resistors.	Both miniature ¼ watt 5%
R1	10k
R2	680k
VR1	100k log carbon (can be ganged with on/off switch S1)
Capacitors.	
C1	39pF ceramic plate
C2	220nF type C280
C3	680MFD 10vw
C4	100MFD 10vw
C5	100nF type C280
Integrated Circuit.	
I.C.1	LM380N (14 pin package)

Switch.

S1	S.P.S.T. toggle type, or ganged with VR1

Miscellaneous.
Metal instrument case
Miniature 35 to 80 ohm impedance loudspeaker

TBA800 5 Watt Power Amplifier

The TBA800 is an inexpensive device which is capable of output powers up to 5 watts R.M.S. with reasonably good output quality. It is contained in a 12 pin quad-in-line package which is basically a standard 16 pin QUIL plastic package, but four of the pins have been replaced by a couple of heat tabs (as shown in the pinout diagram). Unless suitable heatsinking is applied to these heat tabs the device is restricted to a maximum output power of 2.5 watts R.M.S.

Principal Ratings

Supply voltage	5V. min.	30V max.
Quiescent current (24V. supply)	9mA. typ.	20mA. max.
Input sensitivity for 5 Watts R.M.S. into 15 ohms at 1kHz with 56 ohm feedback resistor	80mV. typ.	
Distortion from 50mW. to 2.5W. R.M.S. at 1kHz into 16 ohm load (24 volt supply)	0.5% typ.	

Simple Record Player Amplifier

Probably the most obvious application for the TBA800 device is as the basis of the amplifier for an inexpensive record player, where its low cost and appropriate output power and distortion levels make it a good choice. The circuit diagram of such an amplifier is shown in Fig. 11.

Pin 3 of I.C.1 is the input terminal and is biased by R2. The input sensitivity of the circuit is a little high for this application, while the input impedance is rather low. Series resistor R1 is therefore used at the input to both reduce sensitivity and provide some increase in the input impedance. This gives the amplifier as a whole an input impedance of 250k and a sensitivity of 350mV. R.M.S.; both figures being nominal.

C1 provides input D.C. blocking and VR1 is the volume control. Top cut is applied by C2 in conjunction with R1 when VR2 is adjusted for minimum resistance, but almost zero treble cut will be produced when VR2 is set at maximum resistance. Intermediate settings give intermediate degrees of treble attenuation. VR2 thus acts as a simple top cut type treble control.

Pin 12 is the output terminal of I.C.1 and C10 is the output D.C. blocking capacitor. C6 and C8 are compensation capacitors which help to prevent instability. However, the Zobel network consisting of C9 and R4 is also needed in order to ensure stability, and so is the high frequency filter capacitor (C3) which is connected at the input.

The voltage gain of the circuit (closed loop) is determined by an internal feedback resistor of I.C.1 and discrete shunt feedback resistor R3. Thus the sensitivity of the amplifier can be controlled to some extent by altering the value of R3. The voltage gain is approximately equal to 6,200 divided by the value of R3 in ohms, and is roughly 110 times in this case. C4 is merely a D.C. blocking capacitor.

C11 and C12 are supply decoupling capacitors, and additional smoothing and decoupling of the supply to the preamplifier stages of I.C.1 is given by C5. This gives the circuit good supply ripple rejection and a well smoothed supply is not essential for this circuit. C7 is a bootstrapping capacitor and helps to

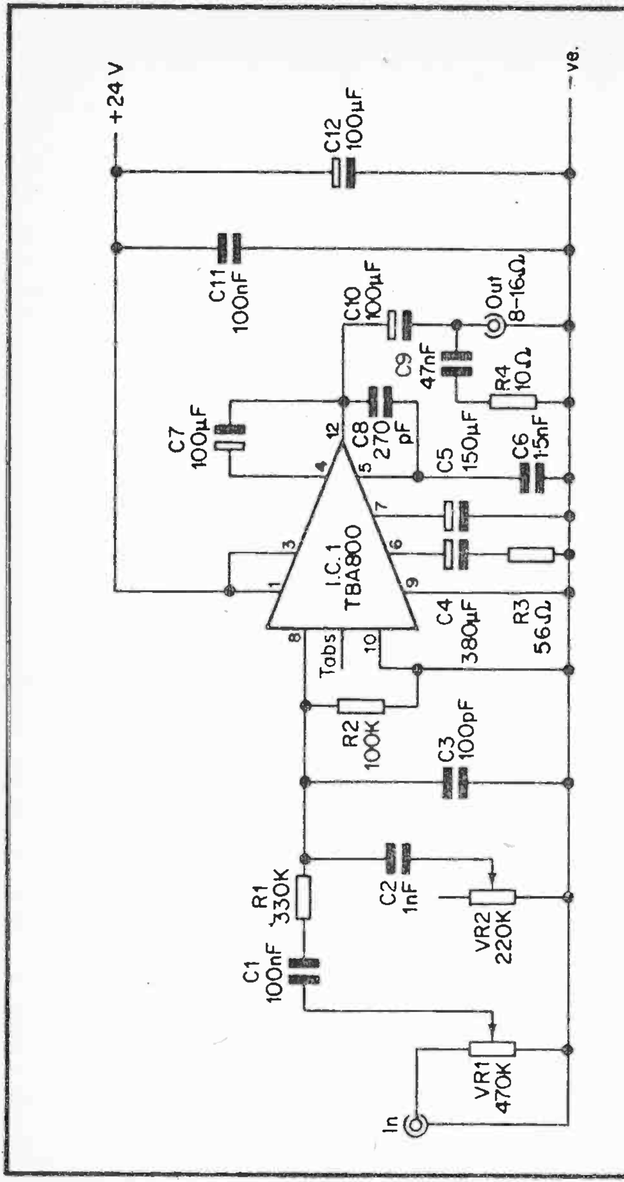

Fig. 11. The circuit diagram of the simple record player amplifier.

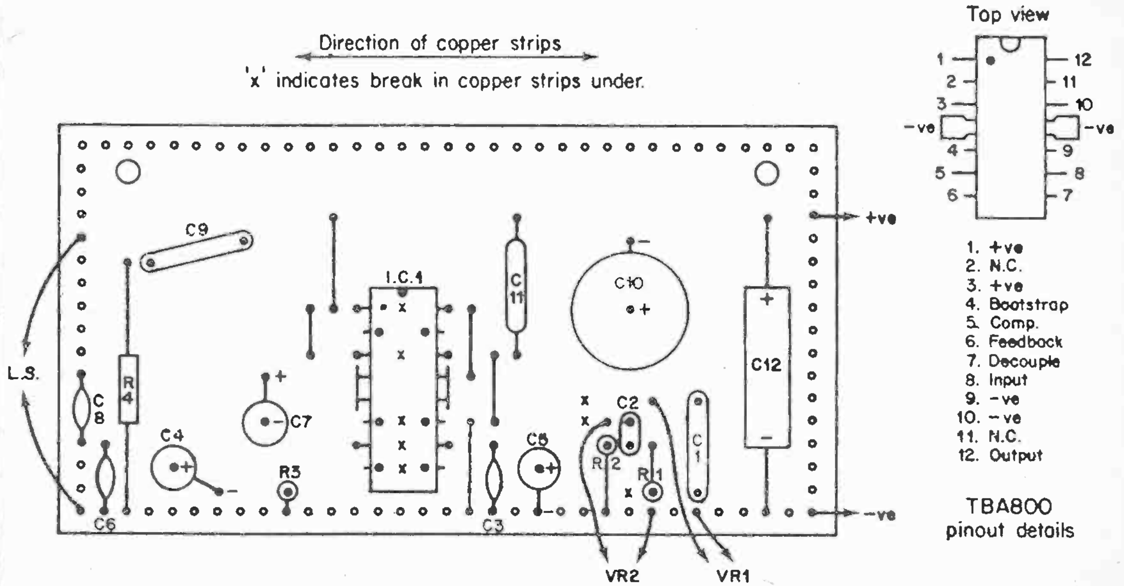

Fig. 12. The 0·1 in. stripboard layout for the record player amplifier.

optimise the maximum available output power for a given supply voltage.

Construction

This is quite straightforward, but as the TBA800 does not have output short circuit protection it is essential to ensure that no accidental short circuits are produced. Neither does it incorporate thermal shutdown protection circuitry, and so damage to the device could easily result if adequate heatsinking is not provided. This heatsinking can simply consist of two pieces of copper laminate board measuring about 43mm square, with one piece being soldered to each heat tab. The heat tabs must be carefully bent up vertically to facilitate this, and the connections to the heatfins must be made fairly quickly so that the I.C. is not damaged due to overheating.

The circuit requires a supply potential of 24 volts or so in order to give a maximum output power of 5 watts into a 15 or 16 ohm impedance load, and at high volume levels the current consumption will be in the region of 300 to 400mA. This really necessitates a stabilised supply since most unstabilised circuits capable of providing 24 volts at full load will give more than the maximum permissible 30 volt supply under quiescent and low volume conditions. With an 8 ohm load a supply voltage of about 18 volts or so will be sufficient to provide 5 watts R.M.S. of output, but the maximum supply drain will be in the region of 500mA. An 18 volt unstabilised supply should give a potential of less than 30 volts under quiescent conditions, and should therefore be suitable for use with the unit.

The amplifier has a fairly high input impedance and will therefore provide a satisfactory impedance match for a crystal or ceramic cartridge. The input leads should be screened types to avoid excessive pick-up of mains hum, and it is also advisable to earth the bodies of VR1 and VR2 for the same reason. The sensivitity of the unit is also suitable for use with most tuners and tape decks.

For stereo operation two amplifier boards must be constructed, of course, and the tone controls should be ganged. The convention with inexpensive audio equipment of this type is to have separate volume controls for each channel, thus dispensing with the need for a balance control. Slider type controls are then most convenient for use as the volume controls, but obviously ordinary rotary types can be used but are more difficult to adjust simultaneously and in co-ordinated manner.

Components SIMPLE RECORD PLAYER AMPLIFIER

Resistors.	All miniature ¼ watt 5%
R1	330k
R2	100k
R3	56 ohms
R4	10 ohms
VR1	470k log carbon
VR2	220k lin carbon

Capacitors.

C1	100nF type C280
C2	1nF Mylar p.c. mounting
C3	100pF ceramic plate
C4	330MFD 10vw
C5	150MFD 25vw
C6	1.5nF ceramic plate
C7	100MFD 15vw
C8	270pF ceramic plate
C9	47nF type C280
C10	1000MFD 16vw
C11	100nF type C280
C12	100MFD 40vw

Integrated Circuit.

I.C.1	TBA800

Miscellaneous.
Copper laminate board for heatsinks.

8 or 16 ohm impedance speaker capable of handling at least 5 watts R.M.S.
Power source.

SN76023 5 Watt Power Amplifier

The SN76023N device is contained in a 12 pin DIL package which is a modified 16 pin DIL, having the two centre pins of each row replaced by heat tabs. These tabs are fixed to a finned aluminium heatsink which is supplied ready fitted to the device, and is the only heatsinking that is required. There is also a version of the device, the SN76023ND which just has the heat tabs, and is not equipped with the heatsink.

Principal Ratings

Supply voltage	8V. min. 28V. max.
Output power for maximum T.D.H. of 1% with 8 ohm load and 24V. supply	5 watts R.M.S. typ.
Signal to noise ratio	-70dB. typ.
Quiescent supply current	Approx 10mA. typ.
Input sensitivity	30mV. R.M.S. for 5 watts R.M.S. output (adjustable)

Guitar Amplifier

An advantage of the SN76023N device over most of its rivals is that it is capable of providing quite low noise and distortion levels even when used to give high gain and (or) input impedance. This enables it to be used without the need for any other active circuitry in applications where one would normally expect a separate preamplifier stage to be employed ahead of the power amplifier. This guitar amplifier is an example of such a circuit. It achieves an input sensitivity of approx. 13mV. into 47k for 5 watts R.M.S. into 8 ohms, with all the amplification being provided by the SN76023N. Distortion is typically only a fraction of 1% at most output powers, reaching no more than about 2% at any output power up to 5 watts R.M.S. The signal noise ratio is about 60dB.

The full circuit diagram of the unit is shown in Fig. 13. The SN76023N is rather like an operational amplifier having a Class B output stage, as it has a high open loop voltage gain and both inverting (pin 12) and non-inverting (pin 1) inputs.

It is normal for the device to be used in the non-inverting mode, and this is indeed the case here. Negative feedback is applied to the inverting input in the usual op. amp. fashion, except that as only a single supply rail is used, a D.C. blocking capacitor must be included in series with the shunt feedback resistor. The feedback components are R3, R4, and C5. The closed loop voltage gain of the circuit is approximately equal to R4 divided by R3, or about 482 times with the specified values.

The input signal is applied to the non-inverting input by way of volume control VR1, and coupling capacitor C2. This input is biased to about half the supply rail potential by an internal potential divider circuit of the I.C. R1 and R2 couple this bias voltage to the non-inverting input, and C4 decouples the bias signal so as to give good supply ripple rejection. The inclusion of C5 in the circuit results in a gain of about unity at D.C., and so biasing the input to about half the supply voltage results in the required quiescent output potential of about this same level.

Due to the high gain of the circuit and the fact that its input and output are in phase, a number of components are needed in order to prevent instability. These are capacitors C1, C3, C6, C7, C10, and C11. C8 is simply a D.C. blocking output capacitor.

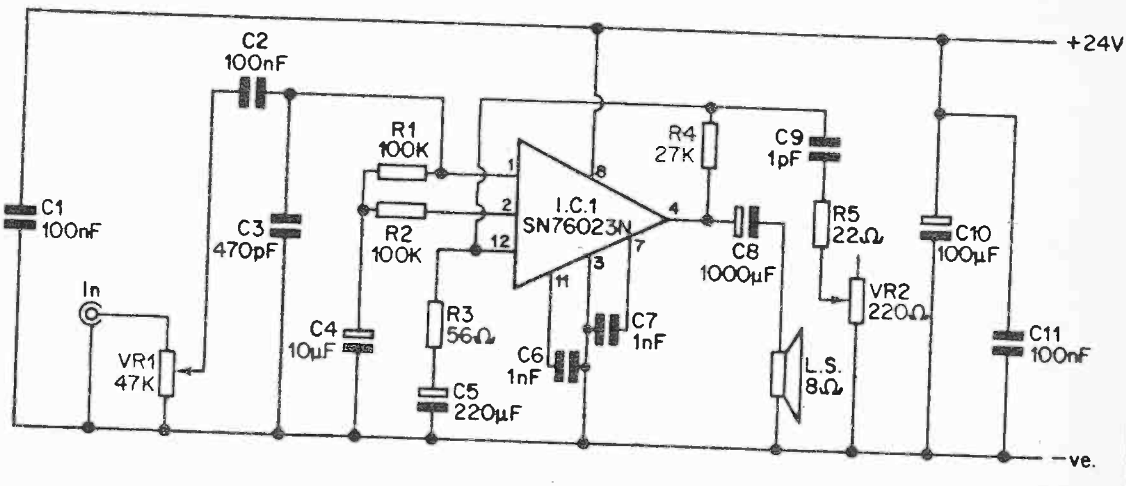

Fig. 13. The circuit diagram of the guitar amplifier.

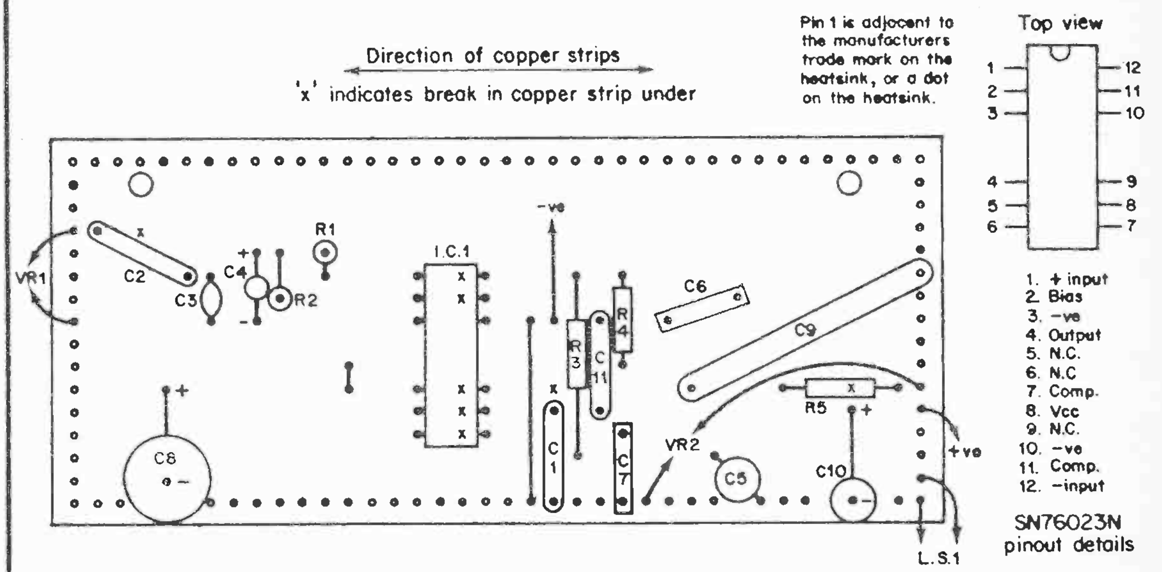

Fig. 14. Constructional details of the guitar amplifier component panel.

A simple tone control is included in the circuit, but this is not of the usual top cut variety. Just the opposite in fact, and it can be used to apply treble boost to the circuit. This can be used to give a more "brilliant" sound to a guitar. VR2 is the tone control potentiometer, and when adjusted for a fairly low resistance, C9 and VR2 have the effect of reducing the amount of feedback applied to the circuit, and thus boost the gain. However, as C9 has only a fairly low value, it only has a low impedance at high frequencies, and only treble frequencies are boosted. R5 limits the amount of boost applied to the circuit to a reasonable amount. With VR2 adjusted for maximum resistance there is very little treble boost applied to the signal, and the frequency response becomes virtually flat over the audio spectrum.

Construction

It is advisable not to solder in the I.C. until the other components and the link wires have been connected, since the heatsink of the device will otherwise obstruct and hinder the installation of some components. The components situated close to the I.C., in particular the vertically mounted ones, must have short leads so that they do not protrude too far above the board, and prevent the I.C. from being slotted into position.

A screened lead must be used to connect the board to VR1, and in fact all input wiring must be screened to prevent mains hum pick-up and stray feedback. The body of VR1 should be earthed to the negative supply rail. The SN76023N does not incorporate output short circuit protection, and so it should be ensured that there are no accidental short circuits on the output before power is applied to the circuit. It should be noted that the metal heatsink connects to the negative supply rail, and care must also be taken to ensure that there are no accidental short circuits to this.

For an output power of 5 watts R.M.S. a supply potential of 24 volts and an 8 ohm speaker are required. 8 ohms is the minimum recommended load impedance. A high load impedance can be used safely, but will give reduced output power. With a supply of about 26 to 27 volts the output power will be increased to approximately 6 watts R.M.S. At maximum output the current drain will be something in the region of 500mA. As the I.C. is used close to its maximum permissible supply potential of 28 volts, a stabilised supply is really required. Battery operation is not really feasible due to the relatively high voltage and current demands of the unit.

It is possible that slight instability will be evident when the amplifier is tried out, this manifesting itself as a very noticeable roughness on loud and high frequency signals. This is most likely to occur when the unit is used in the same cabinet as the speaker with a short lead connecting the two. This can be cured by connecting a small choke in the non-earthy speaker lead, preferably close to the circuit board. The exact value of the choke is not critical, and only needs to be low (a few microhenries rather than millihenries), but it must have a low resistance and be capable of handling a current of an amp. or so. A suitable component could simply consist of any small piece of ferrite rod with a winding consisting of about 30 to 40 turns of 18 s.w.g. enamelled copper wire.

Components GUITAR AMPLIFIER

Resistors.	All miniature ¼ watt 5%
R1	100k
R2	100k
R3	56 ohms
R4	27k
R5	22 ohms
VR1	47k log carbon
VR2	220 ohms wirewound

Capacitors.	
C1	100nF type C280
C2	100nF type C280
C3	470pF ceramic plate

C4	10MFD 15v w
C5	220 MFD 16vw
C6	1nF polycarbonate or ceramic
C7	1nF polycarbonate or ceramic
C8	1000MFD 16vw
C9	1 MFD type C280
C10	100 MFD 40vw
C11	100nF type C280

Integrated Circuit.
I.C.1 SN76023N

Miscellaneous.
Case.
8 ohm impedance loudspeaker capable of handling at least 6 watts R.M.S.

TBA820 Low Voltage Power Amplifier

The TBA820 is designed primarily for use in battery powered equipment where its low quiescent current consumption and ability to operate well at low supply voltages are great advantages. The device is contained in a standard 14 pin QUIL plastic package.

Principal Ratings

Supply voltage	3V. min. 16V. max.
Quiescent current drain (9V. supply)	4mA. typ.
T.H.D. for 500mW. R.M.S. into an 8 ohm load at 1kHz with 9v. supply	0.4% typ. with 120 ohm feedback resistor (0.8% typ. with 33 ohm resistor).

Voltage gain (closed loop)	Variable up to about 56dB.
Maximum output power (12 volt supply and 8 ohm load)	2 watts R.M.S.

Intercom Amplifier

A device designed for use in the output stages of small battery operated equipment obviously has a vast range of possible applications. It is particularly suited to this application as it can provide most of the very high voltage gain that is needed, and only a simple single transistor preamplifier stage is required in order to provide the necessary small additional voltage gain.

The circuit diagram of the intercom amplifier is shown in Fig. 15.

The voltage gain of the TBA820 (closed loop) is determined by an internal feedback resistor, and a discrete feedback resistor which is the shunt element of this network. R5 is the discrete resistor in this case, and the selected value of 10 ohms gives about the maximum usable gain that can be obtained from the device. The voltage gain is approximately equal to the value of R5 divided into 6,000, or about 600 times in this case. C4 is merely a D.C. blocking capacitor.

R4 is the input bias resistor, and this can have any value up to about 100k or so, depending upon the required input impedance. C3 provides input D.C. blocking, and C8 has the same function at the output of the amplifier. C7 and R6 are bootstrapping components, and these help to give the highest possible output power for a given supply voltage and speaker impedance. C5 and C6 are frequency compensation components and are needed to prevent instability.

The preamplifier stage is based on Tr1 which is used in the common base mode. This mode of operation gives a suitably

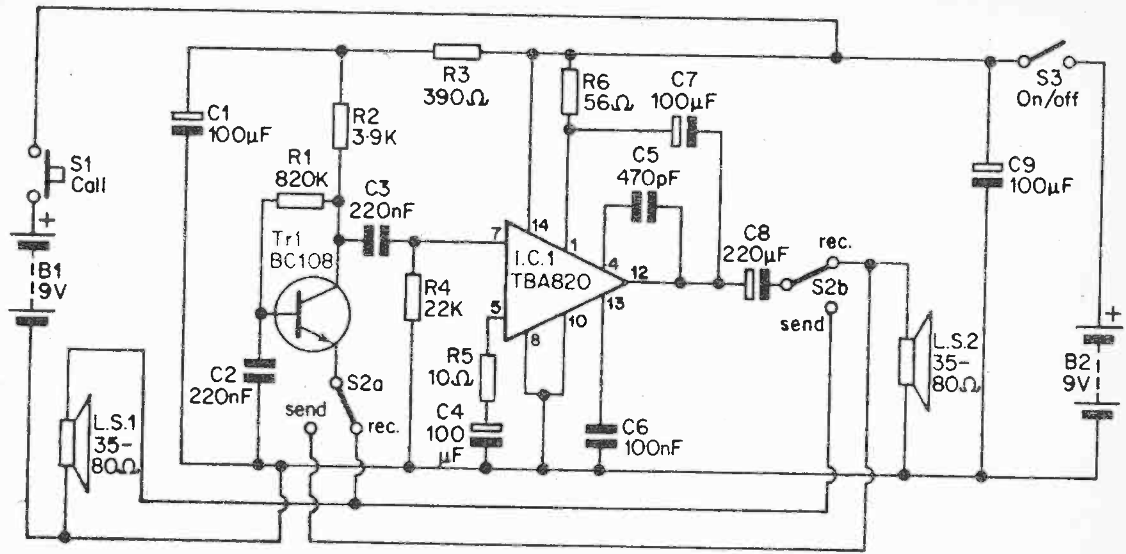

Fig. 15. The circuit diagram of the intercom unit.

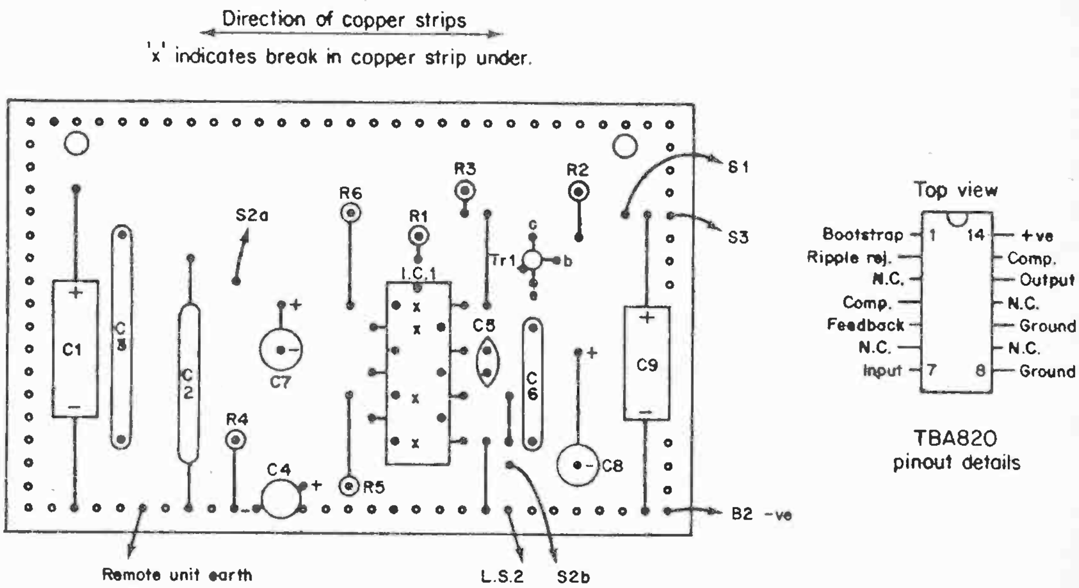

Fig. 16. Details of the 0·1 in. matrix stripboard layout for the intercom.

low input impedance to match the microphone and gives the moderate amount of extra voltage gain that is required. As is normal practice with intercom designs, the speakers double as microphones in the interest of economy.

S2 determines which speaker is connected to the amplifier's input (and actually acts as the microphone) and which one is fed from the output of the amplifier. S2 should really be a biased switch, biased to the 'receive' position. For the remote unit to call the master station it is then only necessary for the user to close S1 (to connect power from BY1 to the amplifier and thus switch the unit on), and then talk into the microphone to attract the attention or the person at the master station. An ordinary D.P.D.T. switch can be used, but it will then be necessary for operator at the master station to always set S2 manually to the 'receive' position after use. Otherwise it will be impossible for the remote station to call the master one. S3 is the ordinary on/off switch situated at the main station.

C1, R3 and C9 are supply decoupling components. A capacitor of about 50mfd. in value can be connected between the negative supply rail and pin 2 of the TBA820 in order to improve the ripple rejection of the device, but this was not found to be necessary in this particular application.

Construction

The unit will work well with a connecting cable of up to about 15 metres in length, and due to the low impedance of the signals carried in the cable it is unlikely that it will be necessary to use a screened connecting lead. The cable must be a three core type, and thin three core mains lead is the obvious choice.

Components INTERCOM AMPLIFIER

Resistors.	All miniature ¼ watt 5%
R1	820k
R2	3.9k
R3	390 ohms
R4	22k
R5	10 ohms
R6	56 ohms

Capacitors.	
C1	100MFD 10vw
C2	220nF type C280
C3	220nF type C280
C4	100MFD 10vw
C5	470pF ceramic plate
C6	100nF type C280
C7	100MFD 10vw
C8	220MFD 10vw
C9	100MFD 10vw

Semiconductors.	
I.C.1	TBA820
Tr1	BC108

Switches.	
S1	push to make non-locking push button type
S2	D.P.D.T. toggle or push button type, biased one way
S3	S.P.S.T. toggle, etc.

Miscellaneous.
Two miniature high impedance speakers

CHAPTER 3

TIMER DEVICES

NE555V General Purpose Timer I.C.

The NE555V (and its many equivalents) can be used as a monostable to provide output pulses from microseconds to hours in duration, or as an astable it can have an operating frequency of as much as 1MHz or as little as one cycle per hour or so. It is an extremely versatile device and is one of the most frequently used I.C.s, in amateur designs. It is contained in a standard 8 pin DIL plastic package.

Principal Ratings

Supply voltage	5V. min. 16V. max.
Supply current (5V. supply)	3mA. typ.
Supply current (15V. supply)	10mA. typ.
Maximum output current (sorce or sink)	200mA.
Power dissipation	600mW. max.
Temperature stability	0.005% per degree C.

Enlarger Timer

When used in the monostable mode the NE555V makes an excellent basis for a simple but very accurate electronic timer, such as the enlarger timer circuit of Fig. 17.

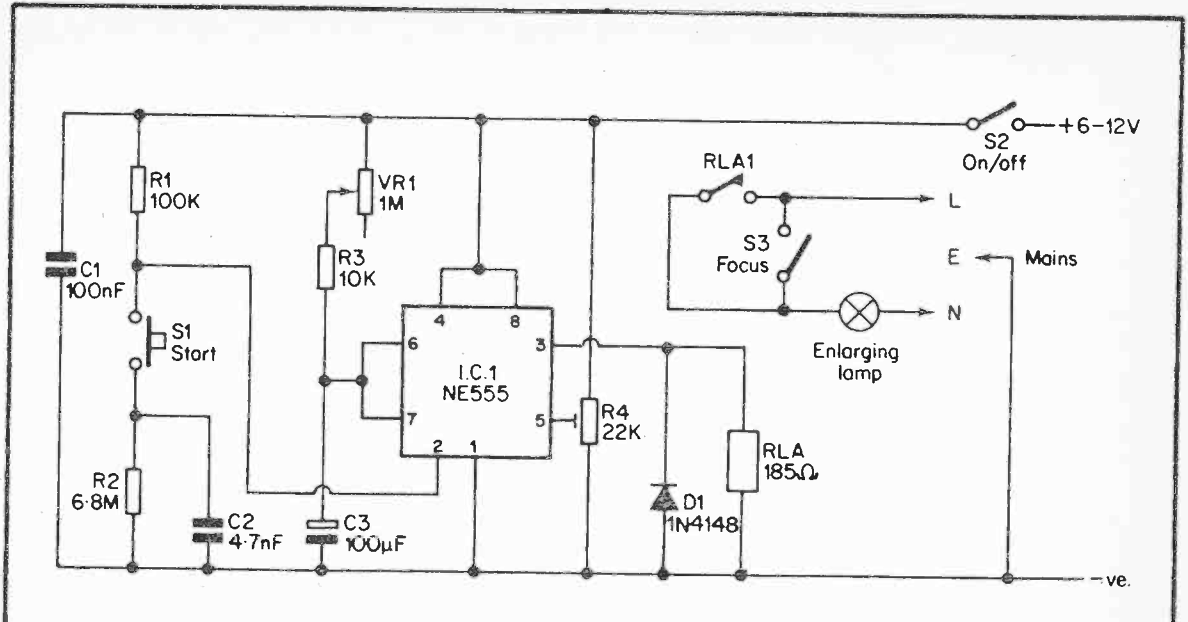

Fig. 17. The circuit diagram of the enlarger timer.

When S2 is closed and power is initially applied to the circuit, R1 will hold the trigger terminal of I.C.1 (pin 2) at virtually the full positive supply potential, and so the circuit is not triggered, and the output at pin 2 assumes a very low voltage. Therefore no power is applied to the relay, and no

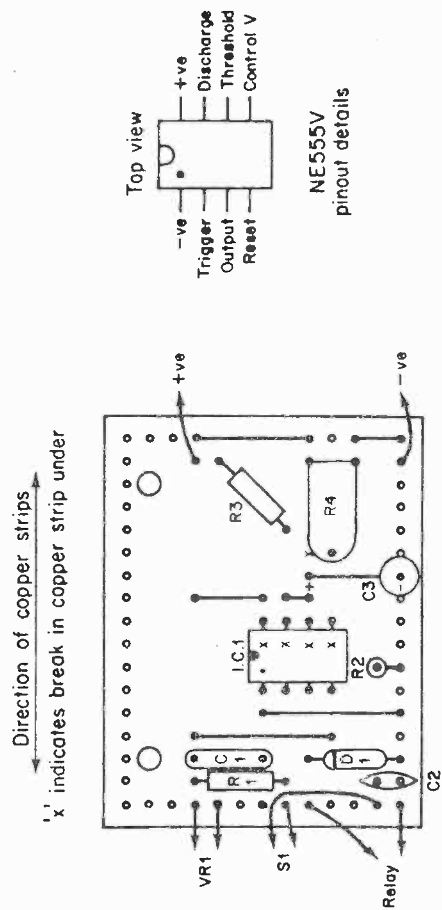

Fig. 18. The stripboard layout for the enlarger timer unit.

power is applied to the enlarging lamp through the pair of normally open relay contacts (RLA1).

If push button switch S1 is now closed, the voltage at I.C.1 pin 2 will momentarily be taken to the negative supply rail voltage by C2, although it will quickly return to its previous level as C2 becomes fully charged through R1. This brief negative pulse triggers the circuit and sends the output to a potential of several volts. This activates the relay, the relay contacts close, and the enlarging lamp is turned on. Before the circuit was triggered, C3 was held in a discharged state by an internal transistor of I.C.1. Once the circuit is triggered, this transistor is switched off and C3 begins to charge up via VR1 and R3.

When the charge on C3 equals approximately two thirds of the supply voltage, this is sensed by an internal comparator of the 555 which has one input taken to pin 6 of the device. The comparator then sets the circuit back to its original state with the relay and enlarger lamp switched off, and C3 discharged through the internal transistor of the 555.

The time taken for C3 to reach this two-thirds charged state is determined by the time constant of VR1 plus R3 and C3. The output pulse duration is approximately equal to 1.1 CR seconds (C can conveniently be in mfds. and R in Meg. ohms, rather than farads and ohms). This gives the circuit a nominal timing range of approximately 1.1 to 111 seconds with VR1 varied from minimum to maximum resistance. However, the tolerances of the timing components are likely to be quite high, VR1 probably having a tolerance of ± 20%, and that of C3 being no better than this, and quite probably worse. It is therefore not possible to accurately predict the time constant produced by practical components in this case.

This makes it necessary to have some means of trimming the timing range to the appropriate time span, so that errors in the timing component values can be compensated for. The purpose of R4 is to permit such trimming, and this component shunts the internal potential divider of I.C.1 which sets the discharge threshold level for C3. By taking R4 slider up towards the top of its track, the threshold voltage is increased, and so is the time taken for C3 to charge to this voltage. If R4 slider is taken towards the bottom end of its track the threshold voltage is reduced, and so is the time taken for C3 to charge to the threshold potential. In this way R4 can either increase or decrease the delay times provided by the unit, and thus compensate for any deficiency of excess in the timing component values. In practice it is adjusted to give a timing range of 1 to 100 seconds (approx.).

It should be noted that if pin 2 is held at a low voltage, the output will remain at a high voltage at the end of the timing period until pin 2 is returned to a suitably high voltage. This could cause problems on short timing intervals if S1 was simply to be connected between pin 2 and the negative supply, since the minimum timing period is only one second, and the user might depress S1 for longer than one second, thus increasing the timing period. This is why C2 has been included between S1 and the negative rail. This ensures that no more than a very brief trigger pulse of about two mS. is applied to pin 2 regardless of how long S1 is depressed. R2 ensures that C2 is discharged reasonably quickly once S1 is opened, so that the circuit is ready to operate again when S1 is closed once again. However, the time constant of R2 and C2 is made sufficiently long to prevent any problems with accidental retriggering of the circuit when S1 is released due to the inevitable contact bounce.

S3 is a 'focus' switch, and this can be used to switch on the enlarger lamp independently from the timer circuit, so that the enlarger can be focussed.

Construction

The circuit can be powered from a 9 volt battery of fairly high capacity such as a PP9 size, but the quiescent current consumption is likely to be 10mA. or so, rising to over 50mA. when the relay is activated. This would give rather a short battery life. Since the unit is connected to the mains supply anyway, a simple unstabilised nominal 9 volt powered supply is a more suitable source.

R4 is adjusted to give the correct timing range by trial and error. VR1 is given a scale calibrated in seconds, and again, trial and error is required in order to find each calibration point.

Components **ENLARGER TIMER**

Resistors	All miniature ¼ watt 5% except R4
R1	100k
R2	6.8 Meg
R3	10k
R4	22k 0.1 watt horizontal preset
VR1	1 Meg lin carbon

Capacitors.	
C1	100nF type C280
C2	4.7nF ceramic
C3	100MFD 10vw (preferably a tantalum type, see text)

Semiconductors.	
I.C.1	NE555V or equivalent
D1	1N4148

Switches.	
S1	Push to make non-locking push button type
S2	S.P.S.T. toggle type
S3	S.P.S.T. toggle type

Miscellaneous.
Relay having nominal 6/12 volt coil and coil resistance of about 185 ohms or more. A least one normally open contact of adequate rating for load.

NE556V Dual Timer I.C.

The NE556V contains two timer circuits of the same type as the NE555V within a standard 14 pin DIL plastic package. The timers share common supply connections but are otherwise independent of each other. The performance parameters are the same for each section of the NE556V as they were for the NE555V device covered in the previous section.

Pulsed Tone Generator

As mentioned in the previous section, timer circuits of the 555 type can be employed in either the monostable (single shot) or astable (free running) mode. The enlarger timer is an example of the device used in the monostable mode, and this circuit uses the two timers of the NE556V device in the astable mode to produce a pulsed tone alarm signal. A pulsed tone is superior to a straight forward continuous tone in that its intermittent nature makes it less monotonous, and thus also more noticeable.

The circuit of the pulsed tone generator is given in Fig. 19. The audio tone is generated by I.C.1b and is fed to a medium impedance loudspeaker via coupling capacitor C4. The output stage of a 555 type timer circuit is more than adequate to drive a medium impedance speaker, but it is recommended that a speaker impedance of less than 25 ohms should not be used. A high impedance speaker can be used, but the output power will be significantly reduced if this is done.

When used in the astable mode the trigger terminal is connected to the threshold terminal, and a resistor is usually connected between the threshold and discharge pinouts. This is in fact the configuration used here. When power is applied to the circuit, C3 will be discharged and so the trigger terminal will be at zero volts. This triggers the circuit, the discharge transistor of the I.C. is switched off, and C3 begins to charge via R3 and R4. The output goes high during this period.

When the charge on C3 reaches two thirds of the supply potential, this is sensed by the threshold terminal and the circuit is reset to its original condition. The output then goes low and the discharge transistor begins to discharge C3 through R4. When the charge on C3 reaches approximately one third of the supply voltage, the potential at the trigger terminal is sufficiently low to retrigger the circuit. This sends

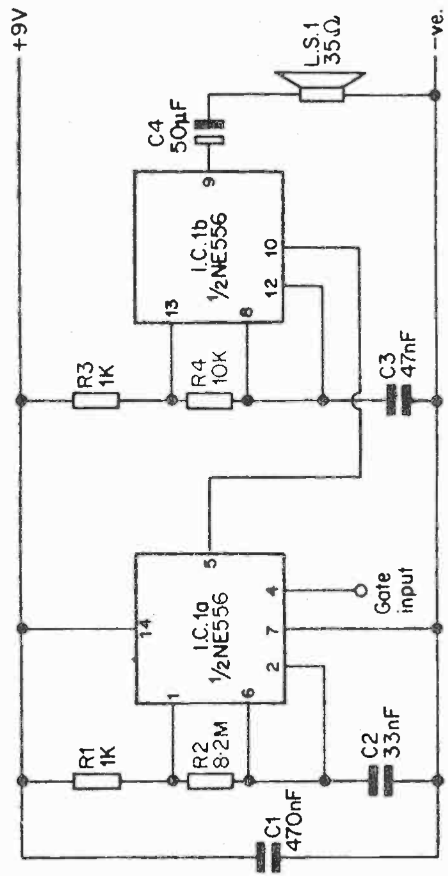

Fig. 19. The circuit diagram of the pulsed tone generator.

the output high once again and C3 begins to charge up once again. Of course, when it achieves a charge level of two thirds of the supply voltage the circuit is reset once again, and so the circuit continuously oscillates in this manner. The time during which the output is high is controlled by the time constant of R3 plus R4 and C3, while the low output period is controlled by the time constant of R4 and C3. The high output time must obviously always be longer than the low

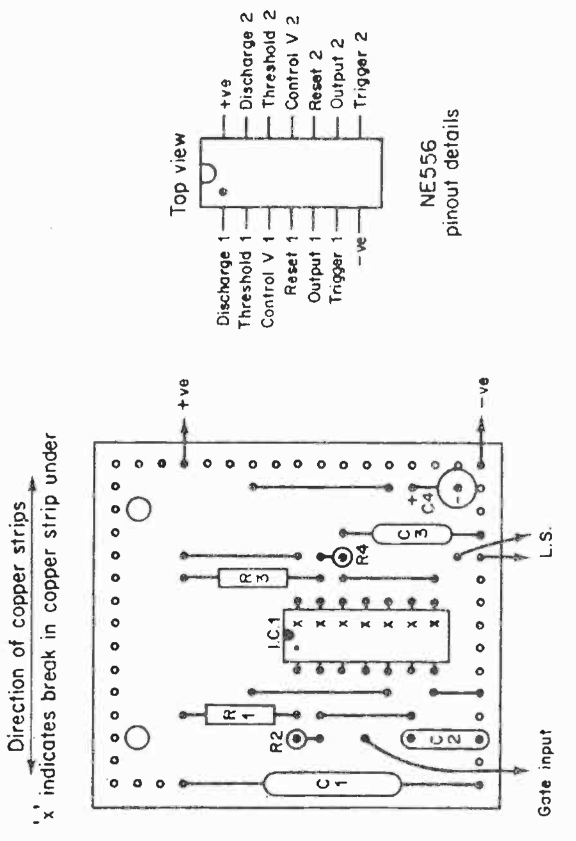

Fig. 20. Details of the 0·1 in. stripboard layout for the pulsed tone generator.

output time when using this basic circuit., but by making R3 much lower in value than R4 an output having a mark/space ratio of nearly one to one is obtained.

The frequency of operation is approximately equal to 1 divided by 1.1CR where C is the value of the timing capacitor in mfds. and R is the total effective timing resistance in Meg. ohms (i.e. 2 x R4 + (1 x R3)). This gives a nominal output frequency of 921Hz with the specified values.

I.C.1a is used as a low frequency astable circuit having an operating frequency of only about 1.7Hz. Its output is coupled to the reset terminal of I.C.1b. If the reset terminal is left floating or taken to the positive supply, the circuit will function normally. However, if it is taken below about 0.5 volts the circuit is forced to assume its initial stage and oscillation is blocked.

Therefore, when I.C.1a output is high, the tone oscillator functions normally, but when this output goes low the tone oscillator ceases to produce an output signal. In this way the audio output signal is rapidly switched on and off to produce a pulsed output signal.

The entire circuit can be controlled electronically by applying a control signal to pin 4 (the reset terminal) of I.C.1a. If this terminal is taken below about 0.5 volts the low frequency astable will be blocked, and its output will go low. This in turn blocks the tone generator astable and no output is produced. If pin 4 is taken above about 0.5 volts (or left floating) the circuit will function normally. Note however, that when the circuit is switched off using pin 4, the circuit still draws a supply current of something in the region of 14mA. This rises to about 40 to 50mA. (average) when the circuit is running normally.

Components   **PULSED TONE GENERATOR**

Resistors.	All miniature ¼ watt 5%
R1	1k
R2	8.2 Meg
R3	1k
R4	10k
Capacitors.	
C1	470nF type C280
C2	33nF type C280
C3	47nF type C280
C4	50MFD 10vw
Semiconductor.	
I.C.1	NE556V or equivalent
Loudspeaker.	
L.S.1	miniature 35 ohm impedance speaker (see text)

The ZN1034E Precision Long Timer

Simple timers such as the NE555V can provide good results at timing periods up to a few minutes in duration, but they tend to give comparatively poor results over longer timing intervals.

This is due to the high values of the timing components needed in order to give long delays, making it necessary to use electrolytic capacitor. These components usually have quite high leakage currents, and when employed in conjunction with a high value timing resistor this leads to inconsistent and unreliable results, or even complete failure of the circuit.

Timers such as the ZN1034E use an alternative approach to generating long time delays, and one which gives far superior results, although admittedly at the cost of increased circuit complexity. The basic arrangement of this type of circuit is to have a stable C – R oscillator feeding a digital counter and logic control circuit. When the circuit is triggered, the counter circuit begins to count the oscillator pulses and the output of the circuit goes high. After a certain number of pulses have been received, the counter and logic control circuits reset the circuit to its original state. The ZN1034E contains a 12 stage binary counter, and the timing interval ends after 4095 clock cycles. Whereas the output time from a 555 timer I.C. is nominally 1.1CR, the ZN1034E produces an output pulse in the region of 2500 to 7500 CR, and this enables output times of hours or even days to be obtained using ordinary components.

Supply voltage	4.75V. min.	5.25V. max.
Supply voltage if internal regulator is used (together with suitable series resistor)	6V. min.	450V. max.
Supply current	5mA. typ., must not exceed 50mA.	
Timing capacitor	3300pf min.	
Timing resistor	5k min.	10Meg. max.
Output current	25mA. max.	
Temperature stability (internal cal.)	0.01% per degree C.	

Switch Off Delay Timer

This simple timer can be used to switch on an item of electrical equipment for some predetermined time. It could, for example, be used to switch off a bedside radio or bedroom light automatically, after the preset time has elapsed. This time can be adjusted from 5 minutes to 30 minutes in five minute increments. By adding extra timing resistors and using a 12 way switch this can be increased to a range of 5 minutes to one hour, if desired.

The circuit of the timer is shown in Fig. 21. The ZN1034E has an internal 5 volt shunt regulator with its positive terminal at pin 5. This must be fed from the main supply voltage via a suitable current limiting resistor, and this resistor must have a value which is computed to give a current flow of 7mA. plus the required maximum output current. The supply current must not be allowed to exceed 50mA. In this case the output current is only a few mA. and a value of 220 ohms gives an adequate current with a 9 volt battery or mains power supply (R9). Pin 4 is the positive supply connection and is fed from the regulated supply. It can be fed from an external 5 volt regulated supply in applications where this is available, and the internal regulator is then simply ignored.

The device has both Q and $\bar{Q}$ (not Q) outputs, the former being normally low and going high during the output pulse, and the latter having the opposite states. In this case it is necessary for the load to be switched on during the output pulse, and so the Q output is used to drive the base of Tr1 via current limiting resistor R8. Normally the Q output is low and so Tr1 is switched off, and so is the relay which forms its collector load. When the circuit is triggered, Tr1 and the relay are switched on as the Q output goes high, and they remain on until the end of the timing period. So does the load which is controlled by the normally open relay contacts, RLA2.

The trigger terminal (pin 1) is actually tied to the negative supply rail so that the circuit is triggered as soon as power is applied to the circuit. This is achieved by momentarily depressing S1. Relay contacts RLA1 close as soon as power is connected to the circuit, so that the unit continues to receive power when S1 is released, but it will be automatically switched off at the end of the timing period, along with the load.

D1 is the normal protective diode, and it suppresses the high back E.M.F. which would otherwise be generated across the relay coil as it de-energised, and which could possibly damage Tr1 and I.C.1 if not suppressed.

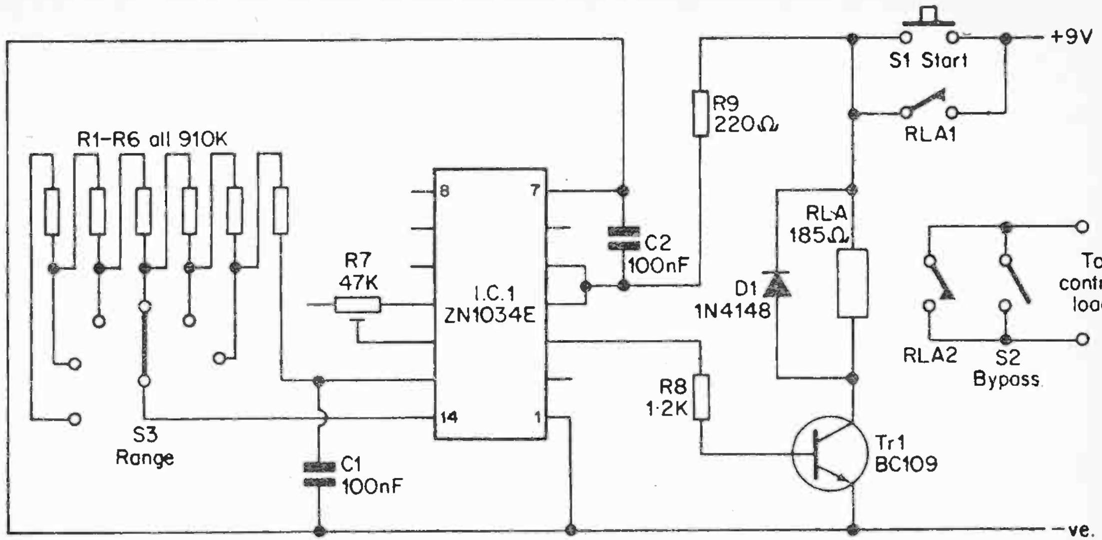

Fig. 21. The circuit diagram of the switch off delay timer.

C1 is the timing capacitor, and R1 to R6 form the timing resistance. As S3 is adjusted in an anticlockwise direction it increases the number of timing resistors in circuit. In the fully clockwise direction only one resistor is in circuit, and

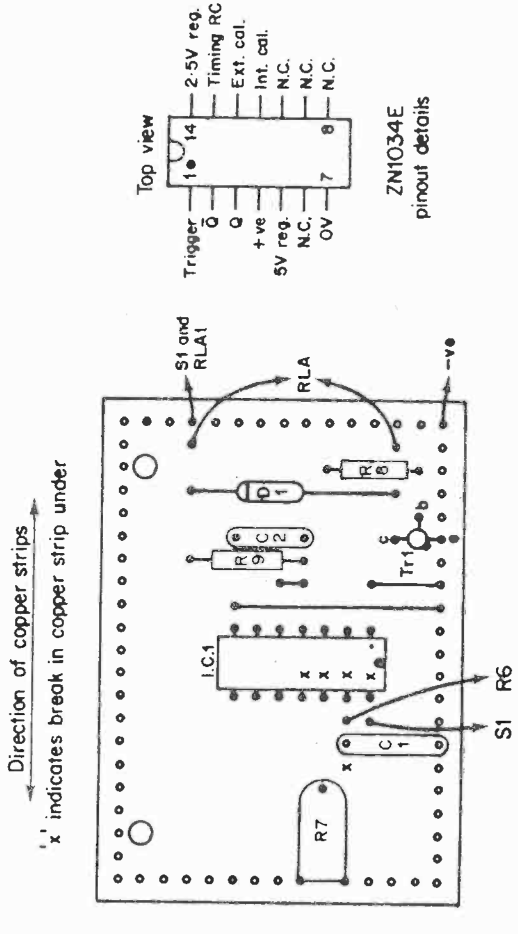

Fig. 22. The 0·1 in. matrix stripboard layout for the switch off delay timer.

this gives an output pulse of 5 minutes in duration. In the next position two resistors are connected into circuit, and so the output pulse duration is doubled to 10 minutes. The next position gives three resistors in circuit and a delay of 15 minutes, the one after that connects four resistors to produce a pulse of 20 minutes, and so on. Six resistors are used in all giving a maximum delay of 30 minutes, but by using a switch having more positions it would be possible to include more resistors and increase the maximum delay up to about 1 hour.

With pins 11 and 12 of I.C.1 connected together, an internal 100k calibration resistor is connected into circuit, and the output pulse length is approximately 2736 CR seconds. However, in many applications, including this one, it is necessary to have some means of trimming the output times accurately to the correct figures.

This could be achieved by having a series of presets for the timing resistors, so that each range could be individually adjusted, but this would be rather time consuming. A better method in many respects is to have some means of adjusting the output pulse length other than by altering the timing component values.

This can be achieved by connecting a 47k preset between pins 11 and 12, as in this case. This gives a calibration resistance which is adjustable from 100k to 147k, and enables the output pulse to be varied from approximately 2376 CR seconds to about 4095 CR seconds. By empirical means R7 is adjusted to give an acceptably high degree of accuracy on one range, and then the other ranges will automatically have a similar degree of accuracy (provided the timing resistors are close tolerance types as specified in the components list).

It is possible to obtain a wider adjustment range by ignoring the internal calibration resistor and connecting an external one between pins 12 and 7. A 220k preset and a 47k resistor in series will give a range of very approximately 2500 CR seconds to 7500 CR seconds, but the temperature stability of the circuit will be degraded by the use of external calibration components.

Construction

Current consumption of the unit is something in the region of 50mA., and this could be supplied by a high capacity 9 volt battery, but a simple mains supply unit is probably a more economical alternative in most cases. In this event it would be preferable to connect S1 and RLA1 in the mains supply to the unit, as the mains power supply would otherwise be left running continuously. The timer circuit would then be fed direct from the output of the 9 volt power supply unit.

Components SWITCH OFF DELAY TIMER

Resistors.	All miniature ¼ or ½ watt, tolerance as indicated below, except R7
R1 to R6	910k 2% or better (6 off)
R7	47k 0.1 watt horizontal preset
R8	1.2k 5%
R9	220 ohms 5%
Capacitors.	
C1	100nF type C280
C2	100nF type C280
Semiconductors.	
I.C.1	ZN1034E
Tr1	BC109
D1	1N4148
Switches.	
S1	Push to make non-locking push button type
S2	S.P.S.T. toggle switch
S3	6 way 2 pole rotary switch (only one pole used)

Miscellaneous.
Relay having at least 2 normally open contacts, and nominal 6/12 volt coil of 185 ohms or more in resistance.

4047 CMOS Monostable/Astable

There are a number of timer I.C.s included in the various families of logic I.C.s, and the 4047 device is one of these. It is a versatile device which can be used in a number of monostable and astable modes, and has features in common with other CMOS devices, which is the logic family it comes from. It is contained in a standard 14 pin DIL package (other packages are used, but only the 14 pin DIL version is available from normal amateur retail sources).

Principal Ratings

Supply voltage (devices having 'A' suffix)	3V. min.	15V. max.
Supply voltage ('B' suffix devices)	3V. min.	20V. max.
Fanout	50 min.	
Input impedance	1,000,000, Meg. typ.	
Quiescent supply current (10V. supply)	0.2 microamps typ.	
Input capacitance	5pf typ.	
Package dissipation	500mW. max	

Christmas Tree Lights Flasher

The 4047 is basically a low power astable multivibrator, but it also incorporates a large amount of logic circuitry which enables it to operate as a negative edge triggered monostable, a positive edge triggered monostable, or a retriggerable monostable. It can also be used as a true or complement gating astable, as well as a straight forward free running type. It must be admitted that these circuit types can be formed from CMOS gates, and at a slightly lower cost than using the 4047, but the latter has the advantage of greater stability with variations in supply voltage, and gives good predictability of the output frequency (or pulse length). The device can be used in a wide range of simple and more complex equipment, and the example given here illustrates the use of the device as a simple astable in a Christmas tree lights flasher project.

The circuit diagram of the unit is shown in Fig. 23. The device is designed to switch the lights on for approximately 1.5 seconds, and then off for about 0.5 seconds, in a continuous sequence. This is superior to the more normal method of simply fitting a flashing bulb into the chain of lights, as it gives improved regularity and reliability.

In the free running astable mode the various inputs to the device are not fed from any external source, and are merely connected to the appropriate supply rail. R1 and C2 are the timing components, and in the monostable mode the output pulse is approximately equal to 4.4CR seconds, and this is the length of one cycle when the 4047 is used in the astable mode. The output frequency in Hz is the reciprocal of this, or about 0.5Hz with the specified values.

This is the output frequency at the normal antiphase Q and $\overline{Q}$ (not Q) outputs, but these are obtained from the astable via a divide by two circuit, and so the astable actually operates at double this frequency. A buffered but otherwise direct output from the astable circuitry is available at pin 13 of the 4047.

In this case the Q output at pin 10 and the astable output at pin 13 are used, and the $\overline{Q}$ is simply ignored. The two used outputs drive the base of Tr1 via current limiting resistors R2 and R3. A relay is used as the collector load for Tr1, and when Tr1 and the relay are switched on, a pair of normally open relay contacts (RLA1) will close and switch the lights on. If either pin 10 or pin 13 is high, the relay will be activated and the lights will be turned on since Tr1 will

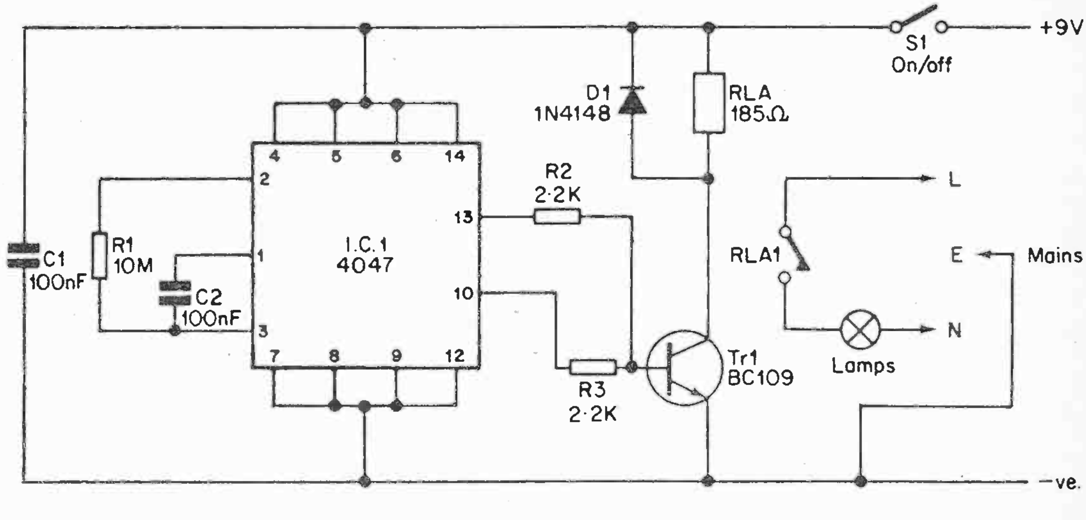

Fig. 23. The circuit diagram of the Christmas tree lights flasher.

receive a strong base current from the high output(s).

Pin 10 will go high for about 1 second and then low for the same period of time, causing the lights to be switched on for one second and then off for one second. However, the output at pin 13 is at double the pin 10 frequency, and pin 13 will therefore be high for half the time that pin 10 is in the low

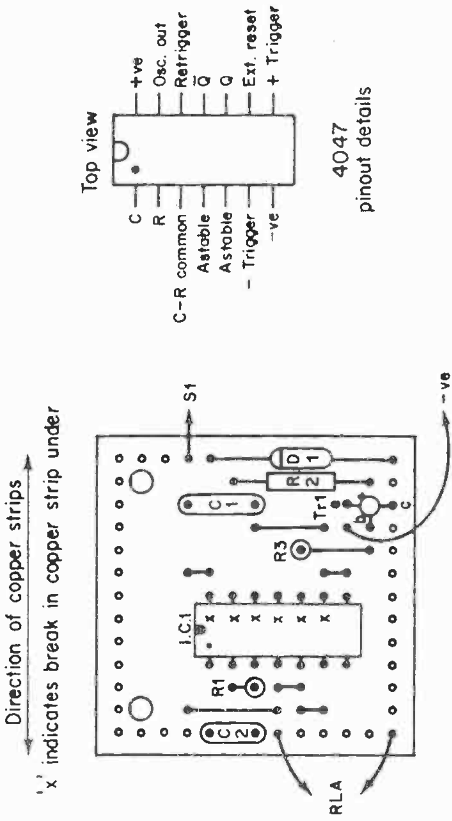

Fig. 24. The 0·1 in. matrix stripboard layout for the Christmas tree lights flasher.

state. This gives a modified sequence of 1.5 seconds on, and 0.5 seconds off, which in practice seems to give a better effect than a straightforward 1 to 1 mark space ratio.

S1 is the on/off switch and C1 is a supply decoupling capacitor. D1 is the normal protective diode. The current consumption of the 4047 device is negligible, except for the output current of a few mA. which flows into Tr1 base when one or both of the used outputs are high. Most of the supply current is that consumed by the relay, which gives an average current consumption in the region of 37mA. (50mA. peak). This could be supplied by a large 9 volt battery such as a PP9 type, but a mains power supply would be more economical in the medium–long term. S1 would then switch the mains input to the power pack rather than the 9 volt D.C. output, which would be taken direct to the flasher circuit.

Construction

As I.C.1 is a CMOS device, it can be damaged by high static voltages. It will probably be supplied in some form of protective packaging and it should be left in this until it is to be soldered into circuit. It should then be handled as little as possible and a soldering iron having an earthed bit should be used when connecting it. Alternatively a socket or Soldercon pins can be used.

This circuit is connected to the mains, in common with other such circuits described in this book and elsewhere, it should not be undertaken by inexperienced constructors.

Components — CHRISTMAS TREE LIGHTS FLASHER

Resistors.	All miniature ¼ watt 5% (10% over 1 Meg)
R1	10 Meg
R2	2.2k
R3	2.2k

Capacitors.

C1	100nF type C280
C2	100nF type C280

Semiconductors.

I.C.1	CMOS 4047 device
D1	1N4148
Tr1	BC109

Switch.

S1	S.P.S.T. toggle type

Miscellaneous.

Case.

Relay having nominal 6/12 coil of 185 ohms or more in resistance, and at least one normally open contact of adequate rating.

Chapter 4

OPERATIONAL AMPLIFIERS

The Standard 741C Device

Operational amplifier I.C.s are widely used in amateur electronic designs, and the 741C industrial standard device is the one most frequently employed. It is available in 8 and 14 pin DIL packages and in an eight pin TO-99 package. The 8 pin DIL plastic package is the version which is now normally sold by component retailers.

Principal Ratings

Supply voltage	± 18V. max.
Input resistance	300k min. 2 Meg. typ.
Voltage gain	20,000 typ. 200,000 max.
Input capacitance	1.4pf typ.
Supply current (± 15V. supply)	1.7mA typ. 2.8mA. max.
Input offset voltage	2mV. typ. 6mV. max.
Output voltage swing (± 15V. supply, 10k load)	±12V. min. ±13V typ.
Common mode rejection ratio	70dB. min. 90dB. typ.
Slew rate	0.5V. per micro sec. typ.

Fuzz Effect Unit

This circuit, which is shown in Fig. 25, is designed to produce the popular 'fuzz' type effect, and it also demonstrates how the gain of an amplifier employing an op. amp. I.C. is determined by the values of two discrete resistors.

The circuit is basically a standard op. amp. non-inverting amplifier. This type of circuit is really intended for the amplification of D.C. signals rather than A.C. signals, and this is the reason that dual power supplies are used with a central earth rail. The non-inverting amplifier is biased to the earth potential by bias resistor R2.

The gain of an operational amplifier is very high, being typically 200,000 times in this case. When employed as a practical amplifier it is normal for negative feedback to be applied to the circuit to reduce the gain of the circuit as a whole to a more suitable level for the particular application involved. The gain of the I.C. is known as the open loop gain, and the gain of the circuit as a whole is termed the closed loop gain.

An operational amplifier is a form of differential amplifier, and it is so called because it is the voltage difference across the two inputs that is amplified. If the inverting input is at a higher potential than the non-inverting one, a negative output swing is produced. Reversing the comparative input states causes a positive swing at the output.

If we assume that the slider of VR1 is at the top of its track, the output of the amplifier will connect direct to the inverting input. The output and inverting input will be stabilised by a negative feedback action at the same potential as the non-inverting input. This must be so since an excess in the inverting input voltage would take this input higher in potential than the non-inverting one, causing the output to swing negative to balance the input voltages. Similarly, a defficiency in the inverting input's potential would send the output positive, raising the voltage at the inverting input to once again match that at the non-inverting one. The circuit is then simply a unity gain buffer amplifier.

If the slider of VR1 is taken down its track, the circuit will still stabilise under quiescent conditions with the output at the same voltage as the non-inverting input, since the negative feedback action will still apply, and will correct any drift in the potential at the inverting input. However, due to the potential divider action across BR1 and R1, the output voltage has to change slightly more than it did previously in order to maintain a balance at the two inputs. In a theoretically perfect circuit the op. amp. would have infinite voltage gain and input impedance, zero output impedance, and the inputs would have to be at precisely the same voltages in order to balance the circuit. In a practical circuit none of these criteria are truly achieved, and this leads to the output drifting away from the earth potential in order to balance the inputs. The further VR1 slider is taken down its track, the more the output will have to drift away from earth potential in order to balance the inputs.

This drift in the output is known as an "offset" voltage, and in a D.C. amplifier it has to be eliminated by an offset null control. For the 741C this consists of a 10k preset connected between pins 1 and 5 of the device, with the slider connecting to the negative rail. This can be adjusted to establish the quiescent output voltage at earth potential. Such a control is not necessary in this case since it is alternating signals that are being handled, and slight drift at the output is of no practical consequence provided input and output D.C. blocking capacitors are included in the circuit. These are C1 and C2 respectively.

In the same way that taking VR1 slider further down its track causes the offset voltage of the circuit to be amplified, it will also cause any input signal to receive increased amplification as well. A rise in the potential at the non-inverting input for example, will cause the output to swing positive by an amount just large enough to balance the two inputs. The further VR1 slider is taken down its track, the greater the output voltage swing required in order to compensate for a given change in input voltage. Thus VR1 controls the closed loop gain of the circuit.

The voltage gain of the circuit is actually equal to (R1 + R2) divided by R1, where R1 is the resistance between the inverting input and earth, and R2 is the resistance between the inverting input and the output.

Of course, the purpose of a fuzz unit is not to provide amplification, but to distort the input signal to produce high

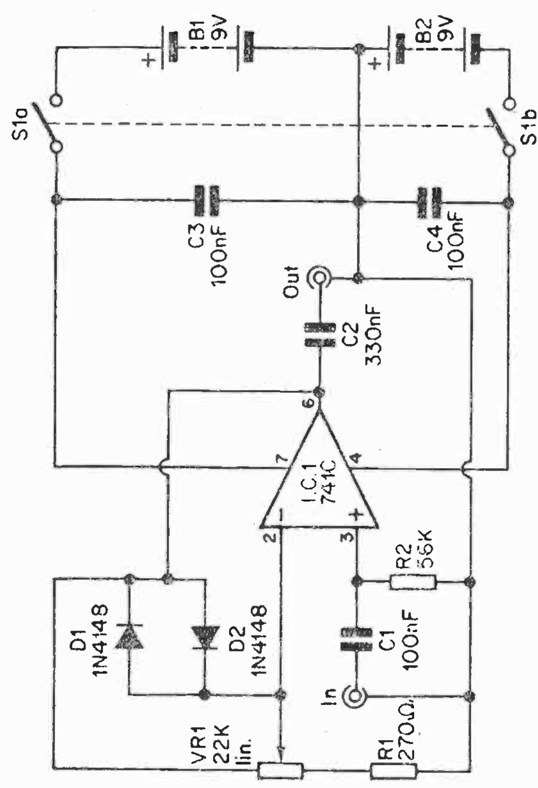

Fig. 25. The circuit diagram of the fuzz effect unit.

frequency harmonics, and give the "fuzz" type sound to the processed signal. In this circuit the distortion is produced by including D1 and D2 in the feedback circuit. With output voltages of less than about 1 volt peak to peak these diodes will not become conductive and will have no significant effect. At signal voltages above this level they conduct heavily and

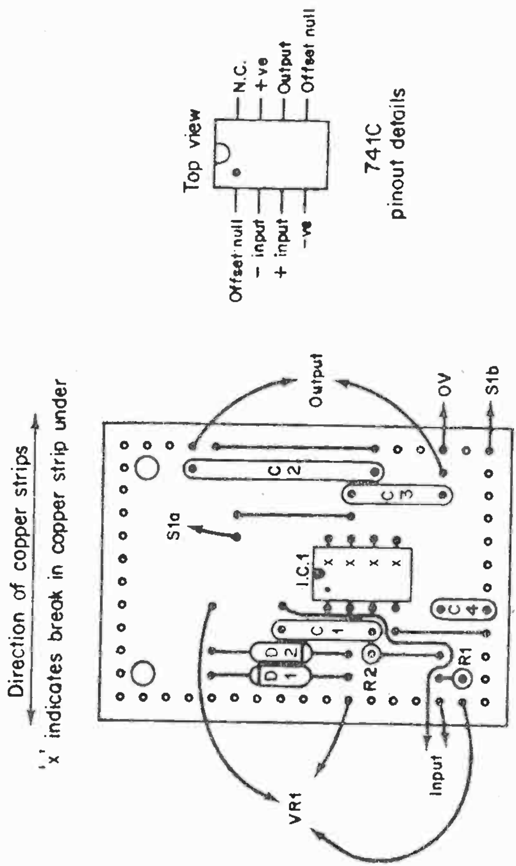

Fig. 26. Details of the 0·1 in. matrix stripboard panel for the fuzz effect unit.

effectively short circuit the output to the inverting input, reducing the closed loop gain to only about unity.

With VR1 set for maximum gain, which is about 81 times, this results in signals being greatly amplified on output levels of up to 1 volt peak to peak, but receiving only unity gain above this level. This produces a form of clipping and gives the required distortion. Using VR1 to reduce the amount of gain results in reduced "fuzz", as only fairly strong input signals will then generate a high enough output voltage swing to initiate clipping. VR1 thus controls the depth of the "fuzz".

Construction

It is advisable to house the unit in a metal case to provide screening from sources of electrical interference, and screened input and output leads should be used for the same reason. Results will probably be satisfactory with the unit connected between the instrument and the amplifier, or between a pre-amplifier and power amplifier. No bypass switching is included in the circuit since the fuzz can be eliminated by adjusting VR1 slider to the top of its track so that the circuit acts as a simple unity gain buffer stage. The input impedance of the circuit is approximately equal to the value given to R2, or a little over 50k with the specified value. This can be varied to suit individual requirements, but it must not be made too high in value as this would lead to stability problems.

Components. FUZZ EFFECT UNIT

Resistors.	All fixed values are miniature ¼ watt 5%
R1	270 ohms
R2	56k
VR1	22k lin carbon

Capacitors.	
C1	100nF type C280

C2	330nF type C280
C3	100nF type C280
C4	100nF type C280

Semiconductors.

I.C.1	741C
D1	1N4148
D2	1N4148

Switch.

S1	D.P.S.T. toggle type

Miscellaneous.
Case
Two 9 volt batteries (PP3, etc.)

CA3140 MOSFET Op. Amp.

Although the 741C operational amplifier is a very versatile and useful device, it falls well short of theoretical perfection in many respects, making it unsuitable for certain applications or at least adding complications to the circuitry where it can just about be pressed into service. This has led to a number of op. amp. I.C.s being produced which improve upon the 741C's performance in some way. Some such devices just concentrate on one particular aspect of performance, such as the low noise and high slew rate devices that are available. Others are designed to give an overall improvement on the 741C's performance, and the CA3140 is an example of such a device. It is available in a standard 8 pin DIL package using the same pinout configuration as the 741C, and in most applications it can be used as a direct replacement for the 741C, often giving improved performance. This 8 pin DIL version has an 'E' suffix. A 'T' suffix version is also available, and this uses the standard op. amp. 8 pin TO-99 encapsulation and pinout arrangement. There is also an 'S' suffix version which has a TO-99 encapsulation with the lead-out wires performed into the 8 pin DIL format.

Principal Ratings.

Supply voltage	± 18V. max.
Large signal voltage gain	100dB. typ.
Supply current (±15V. supply)	4mA. typ.
Common Mode rejection ratio	90dB. typ.
Input resistance	1,500,000 Meg typ.
Differential mode input voltage	± 8V. max.
Common mode D.C. input voltage	V+ plus 8V. max. V– minus 0.5V. min.
Input offset voltage	5mV. typ.
Input terminal current	1mA. max.
Slew rate	9V/microsec. typ.
Gain/bandwidth product	4.5MHz typ.

Six Range Resistance Meter

The most obvious advantage of the CA3140 over the 741C device is the much higher input impedance of the former, which is achieved by the use of a PMOS f.e.t. input stage. This is followed by a bipolar transistor amplifier and output stage which are fabricated on the same chip as the input stage. There are other advantages to the CA3140, such as its increased bandwidth and slew rate, and also its higher output voltage swing. Further, it has the unusual property of being able to operate with its inputs at any voltages between the supply rails, or even slightly outside these limits.

These features enable the CA3140 to operate satisfactorily in the six range linear-scale resistance meter circuit of Fig. 27, whereas an ordinary 741C device would fail completely if used in this manner.

The basic principle of this type of circuit is very simple, and merely consists of applying a constant current to the resistor under test, and then measuring the voltage developed across the test component. If, for example, a constant current of 1mA. is applied to the test resistor and the resultant voltage is measured by a voltmeter having a f.s.d. sensitivity of 1 volt, from Ohms Law it is apparent that a 1k resistor will give a voltage of 1 volt and thus f.s.d. of the meter. A 500 ohm resistor would only give half a volt and half f.s.d. of the meter, a 100 ohm resistor would develop just 0.1 volts and 10% of f.s.d. and so on. In other words there is a linear relationship between the meter reading and the test resistor value, making the basic configuration suitable for use as the basis of a linear scale forward reading resistance meter. This is generally more accurate and convenient to use than the non-linear reverse reading resistance scales used on most multimeters.

In the circuit of Fig. 27 Tr1 is used as a conventional single transistor constant current source, and the series of six emitter resistors (R2 to R7) enables six output currents to be obtained. S1 selects the appropriate output current, and acts as the range switch. On the 100 ohm range the output current is about 56mA., so that about 5.6 volts is developed across a 100 ohm test resistor, and f.s.d. of the voltmeter circuit is produced. The latter uses a conventional micro-amp. meter (M1) and series resistor (R8) giving a f.s.d. sensitivity of about 5.6 volts, preceded by a simple unity gain buffer stage incorporating I.C.1.

The buffer stage is needed to ensure that no significant current is drawn by the voltmeter circuit. This is essential because a test current of only about 0.56 micro-amps is fed to the test resistor, and a current of some 100 micro-amps is needed to give f.s.d. of the voltmeter circuit if the buffer amplifier is omitted. The 1,500,000 Meg. input impedance

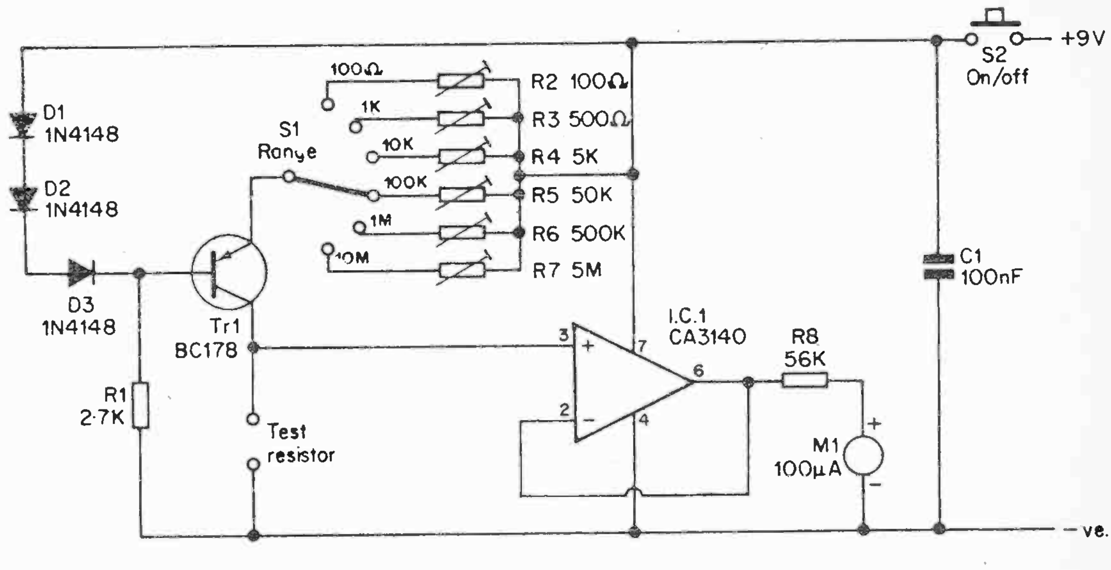

Fig. 27. The circuit diagram of the six range resistance meter.

of the CA3140 ensures that no significant current is drawn by the voltmeter circuitry.

In this circuit the ultra-high input impedance of the CA3140 is an obvious advantage over the 741C, but in this respect the input impedance of the 741C (when used with 100% negative

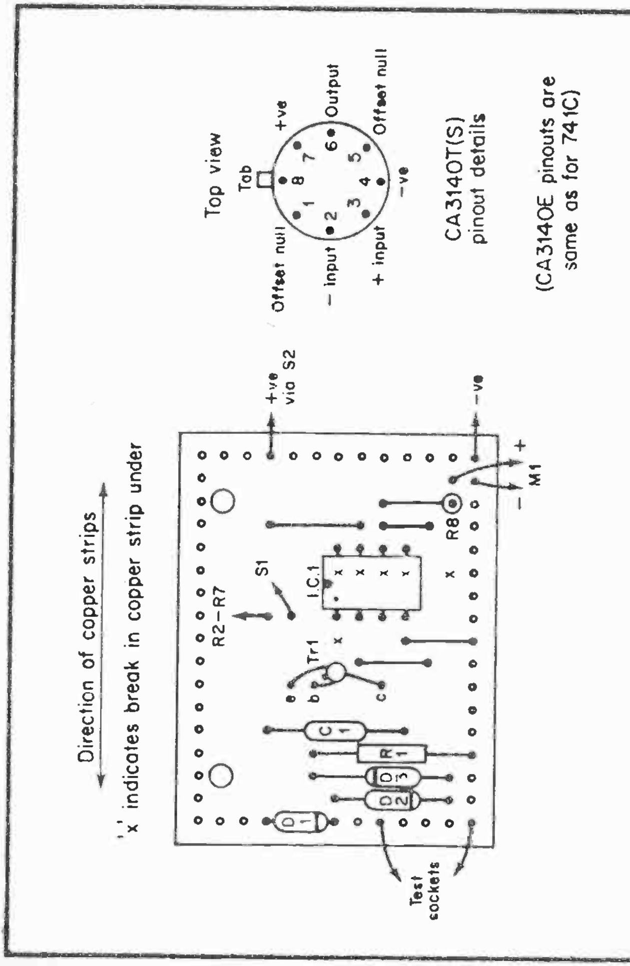

Fig. 28. The layout of the 0·1 in. pitch stripboard panel for the six range resistance meter.

feedback as in this case) might be just about adequate. The main advantage of the CA3140 over the 741C is its ability to operate with input voltages down to and just below the negative supply rail voltage, and the ability of its output terminal to swing down to virtually the negative supply potential. This enables the circuit to operate successfully using a single supply rail, whereas the 741C would need dual supplies due to its inability to operate with input voltage below about 2 volts, and because its minimum output voltage is also about 2 volts.

The emitter resistors of Tr1 are all presets so that they can be adjusted to give precisely the correct f.s.d. sensitivity on each range. S2 is the on/off switch, and this is a non-locking push button type. This should not be closed until the test resistor is connected into circuit since the meter will be driven beyond f.s.d. if this should happen to be done. The same thing will occur if the test resistor has a value which is higher than the selected f.s.d. sensitivity of the unit. However, the meter will not be damaged as a maximum overload of only about 25 to 30% can occur.

Construction

In order to minimise the amount of interconnections between the component panel and the controls, the six preset resistors are mounted on S1.

When calibrating the unit start with all the presets adjusted for maximum resistance. Then switch to the 100 ohm range, connect a 100 ohm close tolerance (2% or better) resistor across the test terminals, and adjust R2 for precise f.s.d. of the meter. R3 to R7 are then similarly adjusted using 1k, 10k, 100k, 1 Meg., and 10 Meg. calibration resistors of low tolerance.

Components. SIX RANGE RESISTANCE METER

Resistors.	Fixed values are miniature ¼ watt 5%, presets are 0.25 watt vertical types.
R1	2.7k
R2	100 ohms preset
R3	500 ohms preset
R4	5k preset
R5	50k preset
R6	500k preset
R7	5 Meg preset
R8	56k

Capacitor.	
C1	100nF type C280

Semiconductors	
I.C.1	CA3140
Tr1	BC178
D1	1N4148
D2	1N4148
D3	1N4148

Switches.	
S1	6 way 2 pole rotary type (only one pole used).
S2	Push to make non-locking type.

CA3130 CMOS OP. AMP.

The CA3130 is a high performance op. amp. which has a PMOS input stage, bipolar main amplifier stage, and CMOS output stage all fabricated on the same chip. This gives the device an ultra-high input impedance, high slew rate, wide bandwidth, and a peak to peak output voltage swing only a few mV. less than the supply potential under light loading conditions. One drawback of the device is the fact that the CMOS output stage results in a maximum supply voltage rating of only 16V. (or ± 8V.), although this is of no real consequence in most applications. The CMOS output stage consumes no significant current when the output is fully positive or negative, giving a quiescent current consumption of only a few hundred microamps under these conditions. The device is contained in a standard 8 pin T0-99 can and either has ordinary straight leadout wires (given a 'T' suffix after the type number) or the leads preformed into an 8 pin DIL configuration (given an 'S' suffix after the type number).

Principal Ratings.

Supply voltage	5V. min.	16V. max.
Input offset voltage	8mV. typ.	
Large signal voltage gain	100dB. typ.	
Input resistance	1,500,000 Meg. typ.	
Input capacitance	4.3pF typ.	
Input terminal current	1mA. max.	
Common mode D.C. input voltage	V+ to V– minus 0.5V.	
Differential mode input voltage	± 8V. max.	
Common mode rejection ratio	90dB. typ.	
Gain bandwith product (no comp. cap.)	15MHz typ.	
Gain bandwidth product (47pF comp. cap.)	4MHz typ.	
Slew rate (no comp. cap.)	30V./μS. typ.	

Slew rate (56pF comp. cap.) 10V./μS. typ.

1MHz Crystal Calibration Generator.

The CA3130 is one of the most versatile devices currently available. It can, for example, be used perfectly well in a number of radio frequency applications, even though apart from a few specialist devices, op. amps. are not usually suitable for use in this area of electronics. Without any compensation capacitor being used the CA3130 has a gain bandwidth product of 15MHz, and a slew rate of 30V./μS., and it is this that enables it to be used in some R.F. applications such as the crystal calibrator circuit of Fig. 29.

Here the device is used as a low gain non-inverting amplifier, with the non-inverting input being biased by R1 and R2. R3 and R4 are the feedback resistors which determine the closed loop gain of the amplifier and C3 provides D.C. blocking.C4 is the output D.C. blocking capacitor.

Positive feedback is applied over the circuit by C2 and X1. This feedback will only be at the series resonant frequency of the crystal where it has a very low impedance. At other frequencies it has a very high impedance and will provide insufficient feedback to be of significance. Since the circuit has a gain of more than unity (about 2.5) and at the series resonant frequency of X1 there will be only minute losses through the positive feedback circuit, oscillation is produced at the series resonant frequency of X1 which is 1MHz. C2 can be trimmed to bring the frequency of operation to precisely the correct figure by compensating for stray circuit capacitances which would otherwise introduce errors.

The circuit has more gain than is needed to just sustain oscillation, and so the circuit oscillates quite violently, producing harmonics of the fundamental at the output. These harmonics are simply signals at frequencies which are multiples of the fundamental frequency (2MHz, 3MHz, 4MHz, etc.). In a signal approximating a squarewave the even harmonics are rather weak (in a theoretically perfect squarewave they are not present at all), and so the bias resistors, R1 and R2, have been given values that give a non-symmetrical output waveform. This minimises any variations in the strengths of the harmonics apart from the gradual fall off in amplitude with increasing frequency, although there will inevitably be some harmonics which are comparatively weak but still quite usable. Of course, it is the harmonics that are used as the calibration signals when checking the calibration of a short wave receiver, or marking the tuning dial of a newly constructed set, the 1 MHz fundamental being outside the S.W. bands.

Construction

X1 is not mounted on the board, but is fitted into a chassis mounting crystal holder of the appropriate style. Neither is C2 mounted on the board, as trimmers of the same nominal capacitance can differ considerably in physical shape and size. C2 is soldered to one tag of the crystal holder.

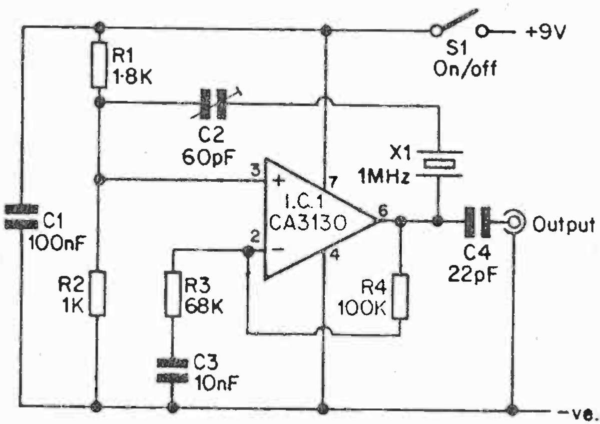

Fig. 29. The circuit diagram of the 1MHz calibration generator.

The unit can be set up against a standard frequency transmission such as the 5MHz Rugby signal. With the standard frequency transmission tuned in on the receiver, and the output from the calibrator loosely coupled to the receiver, a beat note should be produced from the two signals. It is quite likely that this will be at a low audio or sub-audio frequency, and in the case of the latter a phasing type effect will be produced and will enable the signal to be detected aurally. C2 is merely adjusted to give the lowest possible beat note, and careful adjustment should

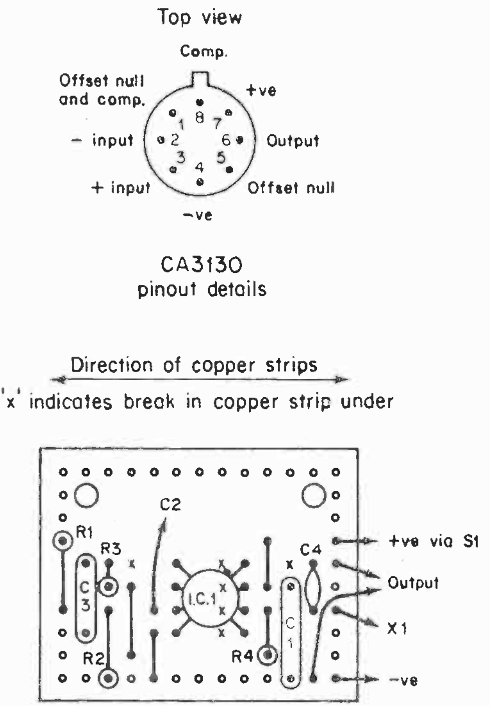

Fig. 30. The 0·1 in. matrix stripboard layout for the 1 MHz calibration generator.

produce an accuracy of 1Hz or better. It is not really essential to calibrate the unit in this way, since its accuracy will be more than adequate for normal requirements even if this adjustment procedure is omitted.

It will probably not be necessary to make a direct connection between the input of the receiver and the calibrator's output socket.

An adequate coupling should be obtained simply by connecting a short insulated lead to the receiver's aerial terminal and then placing this close to the calibrator. At higher harmonics it may be necessary to also connect a short lead to the output socket of the calibrator, and twist this together with the aerial lead. Do not use a coupling that is very much tighter than is necessary to give strong and readily identifiable calibration signals, as this would lead to signals breaking through on spurious responses of the receiver, and could give misleading results.

The prototype calibrator gives strong harmonics to beyond 30MHz, and the unit thus covers the entire S.W. frequency spectrum.

Components. **1MHZ CRYSTAL CALIBRATION GENERATOR**

Resistors.	All miniature ¼ watt 5%
R1	1.8k
R2	1k
R3	68k
R4	100k

Capacitors.	
C1	100nF type C280
C2	10/60pF ceramic trimmer (or similar)
C3	10nF type C280
C4	22pF ceramic plate

Semiconductor.

I.C.1	CA3130

Switch.

S1	S.P.S.T. miniature toggle type

Crystal.

X1	1MHz HC-6U series resonant (or other calibration frequency in 100kHz to 5MHz range).

Miscellaneous.

Case

The LM3900 Quad Norton Amplifier

Although the LM3900N is often described as a quad operational amplifier device, it does not contain four op. amps. of the conventional voltage difference type. The amplifiers respond to the difference in the current flowing into their two inputs, rather than to the voltage difference across them. This type of amplifier is now usually termed a 'Norton' amplifier. The LM3900N contains four of these amplifiers in a standard 14 pin DIL plastic package. The amplifiers share common supply connections.

Principal Ratings.

Supply voltage	4V. min.	36V. max.
Supply current	6.2mA. typ.	
Open loop voltage gain	70dB. typ.	
Gain bandwidth product	2.5MHz typ.	
Slew rate	0.5V/µSec. typ.	

Supply rejection ratio	70dB. typ.
Output resistance	8k typ.
Output voltage swing	V+ minus 1V. typ.
Power dissipation	570mW. max.
Operating temperature	0 deg. C. min. 70 deg. C. max.

Three Channel Mixer.

The LM3900N is primarily intended for use in audio and other low frequency A.C. applications rather than in D.C. applications and therefore it is normal for only a single supply to be used with this device. The device has possible applications in preamplifiers. signal generators, and a great many other types of low frequency circuit. The example given here is the three channel audio mixer circuit of Fig. 31. This has an input for a high impedance dynamic microphone, and two guitar inputs. The input sensitivity of the former is approximately 5mV. for 500mV. at the output, and the guitar inputs have a sensitivity of about 13mV. for an output of 500mV. All three inputs have an input impedance of 47k (nominal).

The circuit consists basically of four amplifiers, one section of the LM3900N being utilized in each. Each amplifier uses the same inverting amplifier configuration which requires just three resistors plus input and output D.C. blocking capacitors in addition to the Norton amplifier. The method of operation is very straight forward.

If we take the mic. preamplifier stage which is based on I.C.1a, for example, C2 is the input coupling capacitor and the value of input resistor R2 sets the input impedance of the circuit at approximately the value selected for this position. The closed loop voltage gain of the circuit is equal to the value of the inverting input bias resistor (R3) divided by the input resistor (R2), or a little over 20 times with the specified value. R1 is the bias resistor for the non-inverting input, and must have a value of roughly double that of the other bias resistor in order to bias the output of the amplifier to the usual level of about half the supply potential. C3 is the output D.C. blocking capacitor.

The biasing of the amplifier is stabilised by a negative feedback action. R1 provides a bias current to the non-inverting input and so the output begins to go positive, but only by an amount which is just sufficient to produce a current flow through R3 that equals that through R1. The input currents are then balanced, and there is no reason for the output to swing further positive. R1 must obviously have a value of about double R3 for the circuit to balance with the output at about half the supply voltage.

A positive voltage applied to the input of the circuit will cause a current to flow through R2 into the inverting input, and unbalance the circuit. This results in the output going negative, but only by an amount which reduces the current flow through R3 to a level which once again balances the two inputs. In other words, the increased input current through R2 will be matched by an identical drop in current through R3, so that the circuit remains balanced. Since R2 has a value which is only about one twentieth of that of R3, in order to achieve this balance the negative output voltage swing must be a little over twenty times the positive input swing. Thus the gain of the circuit is equal to R3 divided by R2.

What is termed a "virtual earth" is formed at the inverting input, since the voltage here does not alter with changes in the input voltage. The current flow through R2 is therefore determined solely by the input voltage and value of R2, and the input impedance of the amplifier must be equal to R2.

The other two preamplifiers are basically the same as the one described above, but the component values are altered slightly as less voltage gain is required.

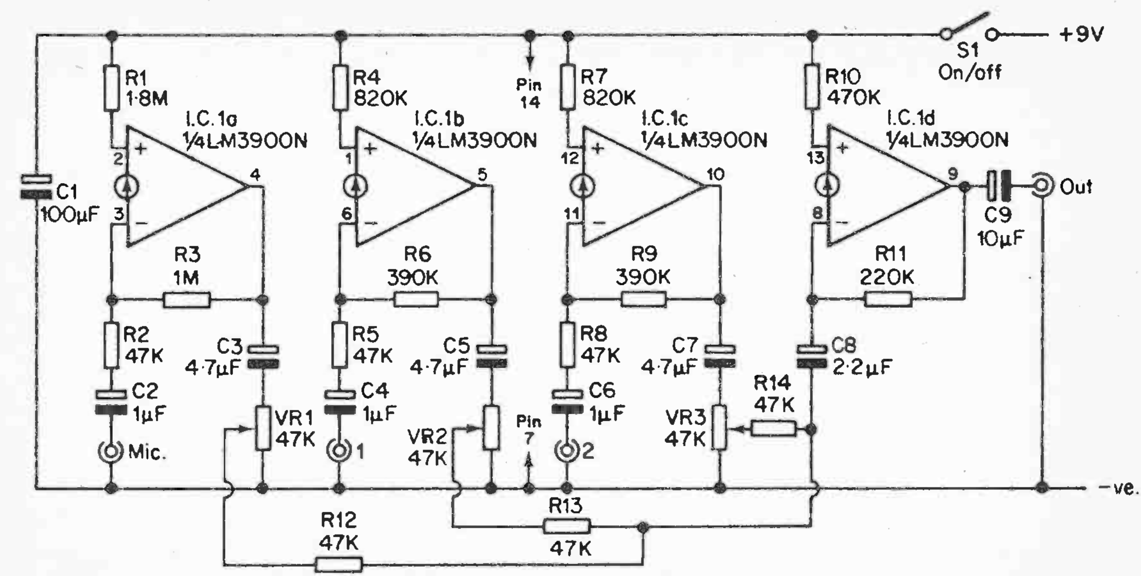

Fig. 31. The circuit diagram of the three channel mixer.

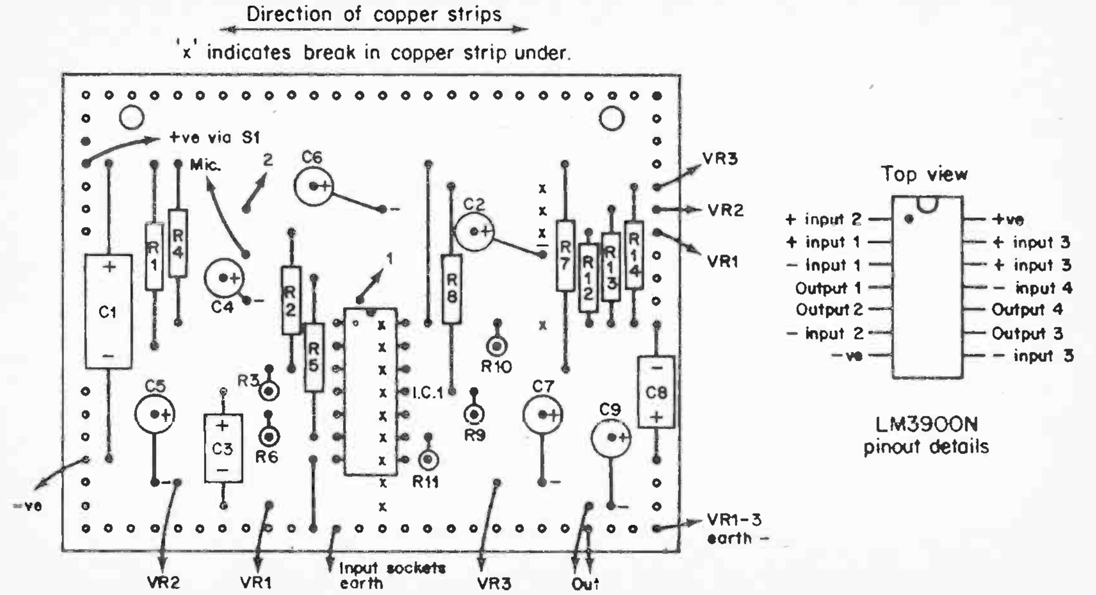

Fig. 32. The 0·1 in. stripboard layout for the three channel mixer.

I.C.1d is used as the mixing amplifier, and this has three series input resistors, one being fed from the output of each pre-amplifier via a potentiometer (VR1 to VR3), the potentiometers being used to adjust the individual channel levels to the required amplitudes. I.C.1d acts as a form of electronic adder circuit. If one input goes positive, the output will swing negative in order to balance the two input currents of the device. If all three inputs go positive by an equal amount, the current flow into the inverting input will increase by three times as much as if one input only was affected. This requires the output to swing sufficiently positive in order to compensate for all three input currents. The circuit thus gives the required adding or mixing action with the three signals being combined at the output socket.

Construction.

It is recommended that the project should be housed in a metal case, earthed to the negative supply rail, to provide screening from sources of electrical interference. None of the connecting leads within the case need to be screened types, but in order to ensure good stability the input leads should be kept well away from the output lead. External connecting cables should be screened types.

A mains power supply can be used, but it must not be situated close to the mixer unit as this could result in a high level of mains hum. Also, the supply must have a very well smoothed output as any noise on the supply lines will be coupled to the non-inverting inputs of the Norton amplifiers via their bias resistors.

Components. **THREE CHANNEL MIXER**

Resistors.	All are miniature ¼ watt 5% (10% over 1 Meg)
R1	1.8 Meg
R2	47k
R3	1 Meg

R4	820k
R5	47k
R6	390k
R7	820k
R8	47k
R9	390k
R10	470k
R11	220k
R12	47k
R13	47k
R14	47k

Capacitors.

C1	100MFD 10vw
C2	1MFD 10vw
C3	4.7MFD 10vw
C4	1MFD 10vw
C5	4.7MFD 10vw
C6	1MFD 10vw
C7	4.7MFD 10vw
C8	2.2MFD 10vw
C9	10MFD 10vw

Potentiometers.

VR1 to VR3	47k log. slider potentiometers (3 off)

Semiconductor.

I.C.1	LM3900N

Switch.

S1	S.P.S.T. miniature toggle type.

Miscellaneous.
Case.
Control knobs for VR1 to VR3
9 volt battery (PP3 etc.)

CHAPTER 5

MISCELLANEOUS CIRCUITS

The 723C Voltage Regulator

The 723C is virtually a standard voltage regulator I.C. which can be used in a wide variety of operating modes, and gives a very high level of performance. Despite its versatility and high performance, it is one of the least expensive I.C.s currently available. The device is available in both T0-99 (10 pin) and 14 pin DIL plastic packages, but the 14 pin version is the one supplied by most component retailers.

Principal Ratings.

Input voltage	40V. max.	9.5V. min.
Input-output voltage difference	38V. max.	3V. min.
Output current	150mA. max.	
Power dissipation	800mW. max.	
Operating temperature range	0 – 70 degrees C.	
Reference voltage	7.15V. typ. 6.8V. min. 7.5V. max.	
Supply current (zero output current)	2.3mA. typ.	4mA. max.
Output voltage range	2 to 37 volts	

Bench Power Supply.

The 723C is suitable for either fixed or variable voltage applications, and the example given in the circuit diagram of Fig. 33 is for a simple bench power supply having an output voltage which is adjustable from approximately 4 to 13 volts. The maximum continuous output current is 500mA., and current limit circuitry ensures that this figure is not significantly exceeded, even if the output is short circuited. The output voltage only drops by about 25mV. between zero output current and maximum load, and the output noise level is quite low at only about one millivolt or so.

The mains supply is connected to the primary winding of mains transformer T1 by way of on/off switch S1. LP1 is merely the on/off indicator. T1 provides the necessary voltage step down and isolation from the mains, and its centre-tapped secondary winding feeds a push-pull type full-wave rectifier using D1 and D2. The rough D.C. output from the rectifier is given a large amount of smoothing by C1. This gives an unloaded output voltage of about 25 volts, but this falls to only about 16 volts at full load.

I.C.1 basically consists of an operational amplifier, a Darlington pair emitter follower output stage, and a highly stable reference voltage source. It also contains a transistor which can be used to provide output current limiting, and a Zener diode which is only utilized in negative voltage regulator applications (this diode is absent on the T0-99 version of the device).

The reference voltage output is at pin 6 of the device, and this is fed to the non-inverting input of the op. amp. In this case the reference voltage is connected to the non-inverting input via a potential divider which is comprised of R1 and R2. This is necessary because the minimum output voltage of the circuit is equal to the reference voltage, and the reference voltage is 7.15V. (nominal). A somewhat lower minimum output voltage is desirable, and so the potential divider is used to effectively reduce the reference voltage and minimum output potential of the device to about 4 volts. This enables it to be used with TTL and other low voltage circuits which the unit would not otherwise be able to cater for.

The output stage of the 723C is not able to handle the currents and powers involved here, and so the output of the device is used to drive a discrete emitter follower output stage using power transistor Tr1. The output from Tr1 emitter is fed to the output socket via R3, which is part of the current limit circuit.

Pin 4 is the inverting input of the op. amp. and connected to VR1 slider. With VR1 slider at the top end of its track, a negative feedback action will stabilise the op. amp. with its two input voltages balanced. In other words the output potential will be equal to the reference voltage of about 4 volts, and will be stabilised at this voltage by the negative feedback action. Taking VR1 slider further down its track still results in the two inputs of the op. amp. being balanced and stabilised at the same level, but due to the potential divider action across VR1 it becomes necessary for the output voltage to increase in order to maintain this balance. Thus, the further VR1 slider is taken down its track, the higher the output voltage becomes, and so the output voltage can be controlled by means of VR1. R4 limits the maximum output voltage at approximately the required level of 13 volts (in practice it will be possible to obtain a maximum voltage of a little more than 13 volts, but the regulation efficiency will be seriously degraded above this level).

Pins 2 and 3 connect to the base and emitter (respectively) of the current limit transistor, and these two terminals are wired across R3 so that the transistor is forward biased by the voltage produced across R3 when an output current flows However, the current limit transistor is a silicon device and will not begin to conduct until it receives a forward bias of about 0.6 volts, and this does not occur until an output current of about 500mA. flows. Therefore, the current limit transistor does not normally have any effect on the circuit. On output currents of 500mA. or more, the transistor will conduct heavily, and it will divert current from the output of the op. amp. (to which its collector connects) through the load to

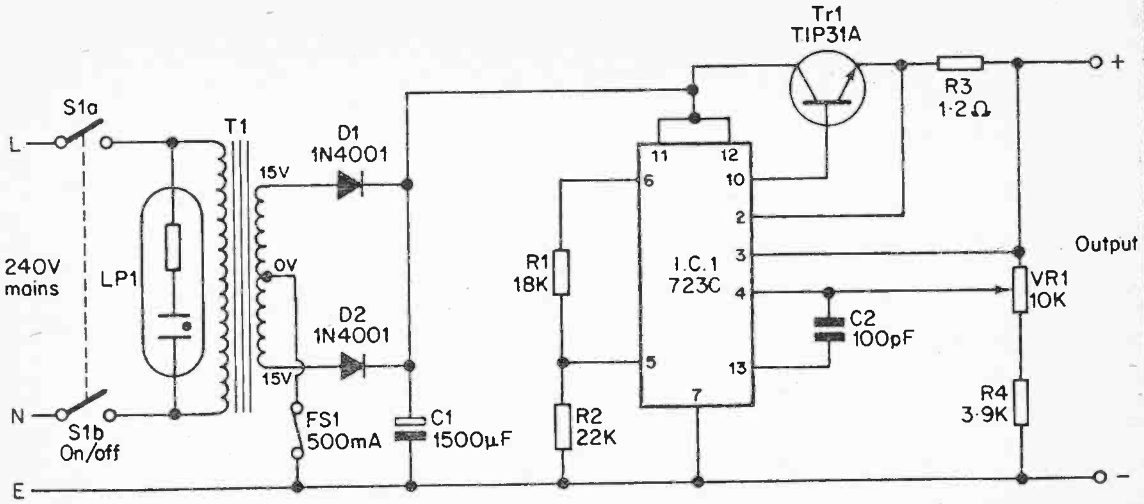

Fig. 33. The circuit diagram of the bench power supply.

earth. This reduces the drive to the output stage and causes the output voltage to fall. The more the loading of the output is increased, the more heavily the current limit transistor conducts, and the lower the output voltage becomes. This prevents the output current from more than slightly exceeding

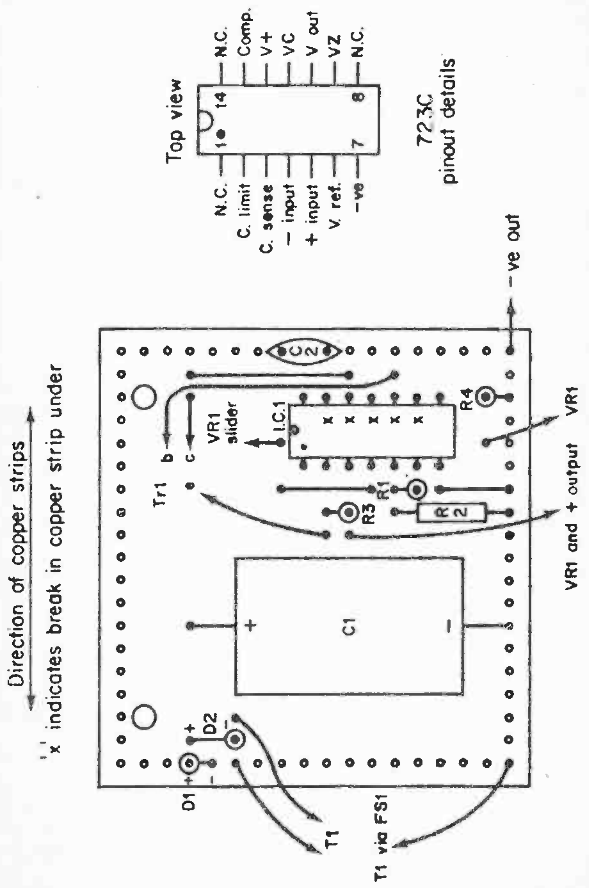

Fig. 34. Details of the 0·1 in. matrix stripboard layout for the bench power supply.

500mA., and greatly reduces the risk of the unit being damaged by accidental short circuits or overloads.

C2 is the compensation capacitor for the op. amp. If a really low output noise level is required, a capacitor of about 5 to 10mfd. can be shunted across R2. However, for normal requirements the noise level of the unit will be perfectly satisfactory without this capacitor. The transient response of the circuit can be improved by adding a 0.1MFD. plastic foil capacitor across the output. This should be mounted on the output sockets.

Construction.

The 0.1in. matrix stripboard layout for the bench power supply is shown in Fig. 34. The board has 18 copper strips by 19 holes and only accommodates the small components plus C1. T1 and the fuseholder are mounted on the case, which should preferably be of metal construction. Tr1 can then be mounted on the case, which will act as a suitable heatsink. Alternatively Tr1 can be mounted on a medium or large size commercial heatsink, and in either case it must have its metal heat pad (which connects to its collector terminal) insulated from the heatsink and (or) case. The case and any other exposed metal parts must be properly earthed.

A voltmeter can be used to monitor the output voltage so that the required output voltage can be set using VR1, but a less expensive alternative is simply to give the control knob of VR1 a scale calibrated in terms of output voltage.

Components. **BENCH POWER SUPPLY**

Resistors.	All miniature ¼ watt 5%
R1	18k
R2	22k
R3	1.2 ohms
R4	3.9k
VR1	10k lin carbon

Capacitors.

C1	1500MFD 25vw
C2	100pF ceramic plate

Semiconductors.

I.C.1	723C 14 pin DIL package
Tr1	TIP31A
D1	1N4001
D2	1N4001

Switch.

S1	Double pole rotary mains on/off switch

Transformer.

T1	Standard mains primary, 15 - 0 - 15 volt secondary rated at 500mA. or more

Miscellaneous.
Case
Output sockets.
Mains neon panel indicator (must have integral series resistor for mains operation).
Control knob.
Insulation set and heatsink for Tr1
500mA fuse and fuseholder to suit

The ZN414 T.R.F. Radio I.C.

The ZN414 is mainly intended for use as the basis of a T.R.F. (tuned radio frequency) medium wave or medium/long wave radio. The device contains an input buffer stage, three stages of R.F. amplification, and a transistor detector/A.G.C. (automatic gain control) stage. It provides a level of performance which is superior to that of virtually any discrete T.R.F. radio design, although it does not quite equal the performance of most superhets as far as selectivity, gain, and A.G.C. performance are concerned. However, compared to even a very basic superhet design a SN414 based receiver is extremely easy to construct and adjust, and is also comparatively inexpensive. The ZN414 is contained in a standard T0-18 lead encapsulation, like BC108, BC109, and similar transistors!

Principal Ratings.

Supply voltage	1.2V. min.	1.6V. max.
Supply current	0.3mA. typ.	0.5mA. under strong signal conditions.
Power gain	72dB. typ.	
A.G.C. range	20dB. typ.	
Output	30mV. approx.	
Input resistance	4 Meg. typ.	
Operating temperature range	0 – 70 degrees C.	
Useful operating frequency range	150kHz – 3MHz	
Audio T.H.D.	less than 2%.	

Miniature M.W. Radio

Probably the most obvious application for the ZN414 is in a miniature M.W. radio using a simple circuit such as the one shown in Fig. 35.

The ZN414 requires a discrete bias resistor to be connected between the output and input of the device, and R1 is the bias resistor in this case. It biases the input via the ferrite aerial coil L1, as this prevents the bias resistor from shunting the input impedance of I.C.1. This input impedance must be fairly high as the tuned circuit couples straight into the ZN414, no low impedance coupling winding being used. A low input impedance would result in a very wide bandwidth and very poor results. C1 couples the earthy end of the tuned circuit to the negative supply rail. VC1 is the tuning capacitor.

A discrete load resistor and R.F. filter capacitor are needed at the output of the device, and these are R2 and C2 respectively. Note that I.C.1 has no direct connection to the positive supply; all the positive supply current being obtained via R2. When a strong signal is received the current drain of I.C.1 increases, causing a reduction in the voltage at pin 1. This effective reduction in the supply voltage to the device causes a loss of gain which gives a simple A.G.C. action.

The audio output from I.C.1 is coupled to a straightforward single stage audio amplifier which utilises Tr1 in the common emitter mode. Unbypassed emitter resistor R5 reduces the gain of this stage somewhat, but this is necessary because clipping of the output signal could otherwise occur. The output signal level is about 1 volt or so peak to peak, and is sufficient to give good volume from a crystal earpiece. It is unlikely that good results will be obtained using any type of magnetic earpiece or headphones.

Current consumption of the receiver is only about 0.45 to 0.65mA. from a 1.5 volt supply, which can be a small torch cell such as an HP16 or HP7. A 9 volt battery such as a PP3 can only be used to power the unit if a suitable series voltage dropper resistor is used, together with a decoupling capacitor across the normal supply rails of the receiver. Suitable values are 15k and 10MFD. respectively.

Construction

The unit should be mounted in a plastic case, or one constructed from some other non-metallic substance. Metal cases are not suitable as they would screen the ferrite aerial and prevent any significant signal pick-up. For similar reasons the ferrite aerial should not be positioned very close to a large metal component, such as a metal framed tuning capacitor or a large metal tuning knob.

The unit should work perfectly well using any ordinary medium

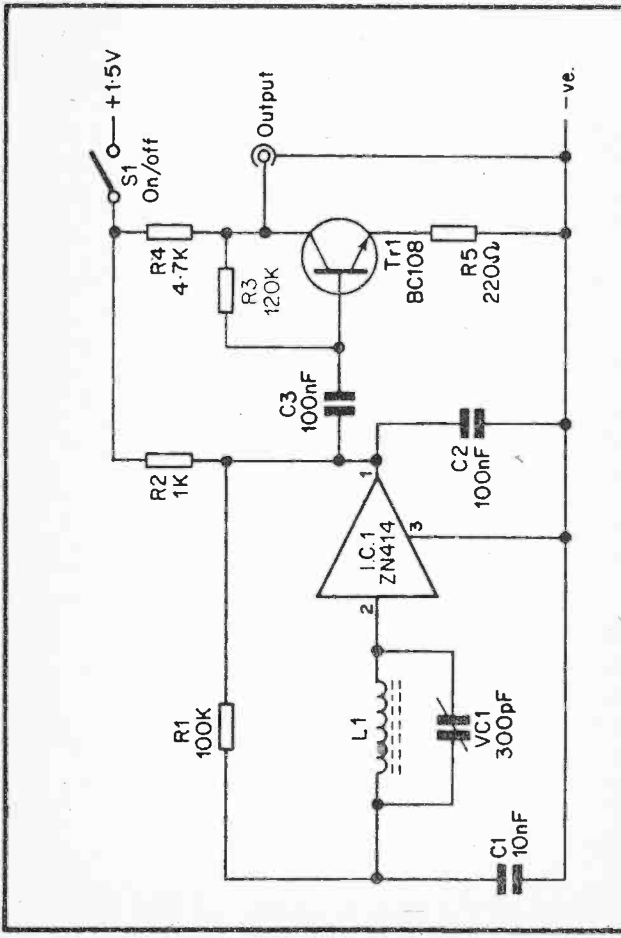

Fig. 35. The circuit diagram of the miniature M.W. radio.

wave ferrite aerial and tuning capacitor of appropriate value.

Any coupling windings are either removed or simply ignored, and any tappings on the main winding are also ignored. The prototype was tested using an Ambit International medium wave aerial coil type MWC2 and a Jackson 300pF solid dielectric tuning capacitor. The ferrite rod employed measured 9.5mm (3/8 in.) by about 77mm (3 in.). The aerial coil is moved up and down the ferrite rod to find a position that permits full coverage of the medium wave band using VC1, and then it is securely taped in that position. This is the only

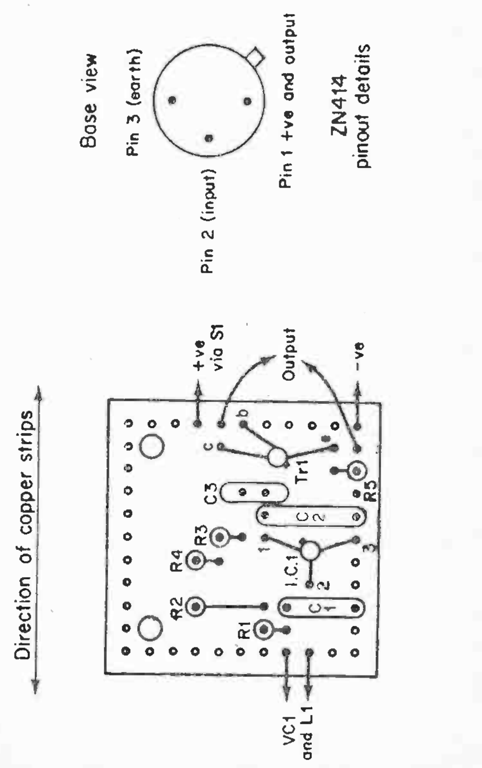

Fig. 36. The 0·1 in. matrix stripboard layout for the miniature M.W. radio.

alignment the finished receiver requires.

It may be rather difficult to connect the unit to a 1.5 volt battery since these do not have the usual radio battery type press stud connectors, of course. Neither is any form of connector available for this type of cell (for just a single cell anyway). Simply soldering the connections to the battery seems to be satisfactory in practice, although the connections must be made quickly to prevent the battery from becoming too hot. An alternative is to use an elastic band to bind a bared lead to each battery terminal.

Components. **MINIATURE MEDIUM WAVE RADIO**

Resistors.	All miniature ¼ watt 5%
R1	100k
R2	1k
R3	120k
R4	4.7k
R5	220 ohms
Capacitors.	
C1	10nF type C280
C2	100nF type C280
C3	100nF type C280
VC1	300pF Jackson solid dielectric (see text)
Semiconductors.	
I.C.1	ZN414
Tr1	BC108
Switch.	
S1	S.P.S.T. subminiature toggle type

Miscellaneous.
Case of non-metallic construction
Medium wave ferrite aerial (see text)
1.5 volt battery (HP16 or HP7)
Control knob.
3.5mm jack socket, and crystal earpiece with 3.5mm jack plug.

The 2N5777 Photo-Darlington.

The 2N5777 is one of the most simple I.C.s, containing the equivalent of just two transistors connected as a Darlington pair. The first transistor is a photosensitive type, and its output is amplified by the second transistor. The result is a photosensitive device of high sensitivity and quite fast response; and at a cost which is roughly comparable to that of other inexpensive photocells such as the ORP12 cadmium sulphide cell and the OCP71 germanium photo-transistor.

Principal Ratings.

Maximum permissible collector – emitter potential	25V.
Maximum permissible collector current	250mA.
Maximum permissible power dissipation	200mW.
Peak sensitivity wavelength	0.85 microns.
Dark current (12V. supply, 25 degrees C.)	less than 0.1μA.
Current gain (VCE = 5V, IC = 0.5mA.)	2,500 min.
Maximum permissible collector – base potential	25V.
Emitter – base breakdown voltage	8V. min.
Rise time	75μS. typ.
Fall time	45μS. typ.

Single IC Projects

- There are now a vast range of IC's available on the amateur market, the majority of which are not necessarily designed for use in a single application and can offer unlimited possibilities.

- All the projects contained in this book are simple to construct and are based on a single IC. A few projects employ one or two transistors in addition to an IC but in most cases the IC is the only active device used.

- A strip board layout is provided for each project, together with any special constructional points and setting up information, making this book suitable for beginners as well as more advanced constructors.

- The book is divided into five chapters: Chapter 1 covers Low Level Audio Circuits; Chapter 2 covers Audio Power Amplifiers; Chapter 3 covers Timer Devices; Chapter 4 covers Operational Amplifiers and Chapter 5 covers Miscellaneous Circuits.

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