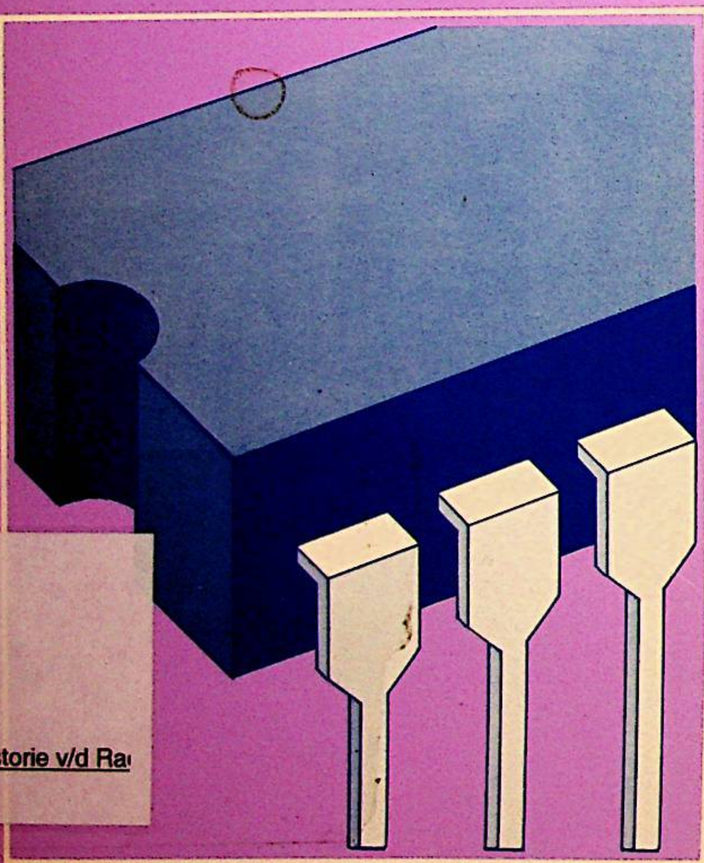
Modern Op-Amp Projects

R. A. PENFOLD



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PREFACE

Integrated circuit operational amplifiers have been much used in electronics projects for the home-constructor ever since the early types became available to the amateur user, and although these early devices may seem crude when compared to the more advanced devices that are currently available, they proved the extreme versatility and usefulness of this type of device. Whereas most integrated circuits are suitable for just one application, or perhaps a number of basically similar uses, operational amplifiers proved to be suitable for a very wide range of applications in many different areas of electronics.

Some of the early integrated circuit operational amplifier devices are still in common use today, and the standard 741C device, for example, is probably as popular as it has ever been. However, early types such as the 741C and 748C have limitations which make them unsuitable for use in certain applications, and the development of higher specification devices has further increased the usefulness of operational amplifiers. Hopefully this point is demonstrated by the wide range of constructional projects featured in this book which make use of the specialised operational amplifiers that are available today, including low noise, low distortion, ultra-high input impedance, high slew rate, and high output current types. Circuits using transconductance types are also included.

All the projects are easy to construct, and a stripboard layout is provided for each one so that even constructors of limited experience should be able to build any of the projects.

R. A. Penfold

CONTENTS

CHAPTER 1

OPERATIONAL AMPLIFIER BASICS

Many people are rather confused when they first start to deal with operational amplifiers due to the fact that an operational amplifier has two inputs. These are the inverting (–) input and the non-inverting (+) input, and what the device is actually amplifying is the voltage difference across the two inputs. In theory at any rate, the actual input voltages are of no importance, and it is only the difference between the input voltages that is of relevance. Thus, for example, having the inverting input at 1 volt and the non-inverting input at 1.1 volts is theoretically the same as having the inverting input at 10 volts and the non-inverting input at 10.1 volts. The input voltage differential is 0.1 volts in both cases.

The simple circuit of Figure 1 demonstrates the basic properties of an operational amplifier. If VR1 is adjusted to

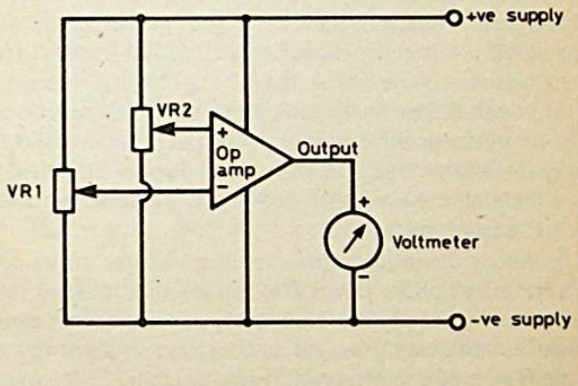

Fig. 1 Circuit to demonstrate the properties of an op amp

give an input potential of half the supply voltage to the inverting input of the operational amplifier, and VR2 is set for minimum voltage at the non-inverting input, the voltmeter will read zero volts at the output of the amplifier. In the case of a practical operational amplifier the output cannot actually go fully negative, and there will be at least a very small voltage present at the output of the device. In fact with most devices the output stage is such that the minimum output potential is in the region of one or two volts, but this is often of no real significance in practical applications.

If VR2 is steadily adjusted for increased voltage at the non-inverting input a point will be reached where the output starts to swing positive, and once this point has been reached only a very small increase in the voltage fed to the non-inverting input is needed in order to send the output fully positive. The reason for this is simply that at DC an operational amplifier has an extremely high voltage gain, a figure of around 100 to 106dB. (100000 to 200000 times) being quite typical. Thus a differential input voltage of only a fraction of a millivolt is sufficient to send the output fully positive or fully negative. In theory an operational amplifier is assumed to have an infinite DC voltage gain, but of course this cannot be achieved in practice. However, for most practical purposes it is safe to assume that operational amplifier devices do have infinite voltage gain. Note that the output goes positive if the non-inverting input is at the higher potential, and negative if the inverting input is the one which is at the higher voltage.

Although theoretically the output should be able to swing right up to the positive supply potential, no practical device can quite achieve this, and with many devices the output has a maximum potential of about one volt less than the positive supply voltage.

In theory the input impedance at each input of an operational amplifier is infinite, and no input current is consumed by the device. Of course, in practice this cannot be achieved, and many practical devices have an input impedance in the region of one Megohm. However, some FET input devices achieve extremely high input impedances and for practical purposes can safely be assumed to have an infinite input impedance at low frequencies. The input impedance inevitably falls somewhat at high frequencies due to stray capacitance at the inputs.

At first it might appear that an amplifier having differential inputs, infinite voltage gain, infinite input impedance, and the further assumed characteristic of zero output impedance, is of little practical value. Operational amplifiers were in fact originally designed specifically to perform mathematical operations in analogue computers, and it is from this that the name operational amplifier is derived. The characteristics of an operational amplifier do in fact make it an extremely versatile device, and it is likely that only a very small percentage of operational amplifiers are actually used in analogue computers these days.

Inverting Amplifier

Operational amplifiers are ideal for use where a DC or AC amplifier having specific levels of voltage gain and input impedance are required. There are two amplifying modes in which an operational amplifier can be used, and these are the inverting mode and the non-inverting mode. As these names suggest, the signal undergoes a 180 degree phase change through an inverting amplifier so that a positive going input produces a negative going output and vice versa. With a non-inverting amplifier the input signal produces a change in the output voltage that is of the same polarity as the input.

We will consider the inverting mode first, and Figure 2 shows the circuit diagram of a basic DC inverting amplifier. Note that there are dual power supplies with a central 0V earth rail, so that the output can be of either polarity with respect to earth. The non-inverting input is simply connected to the earth rail.

Ra and Rb form a negative feedback network which set the closed loop voltage gain and input impedance of the circuit. Note that the closed loop gain is the gain of the circuit as a whole, and not the voltage gain of the operational amplifier which is termed the "open loop" voltage gain. The closed loop gain must always be less than the open loop gain.

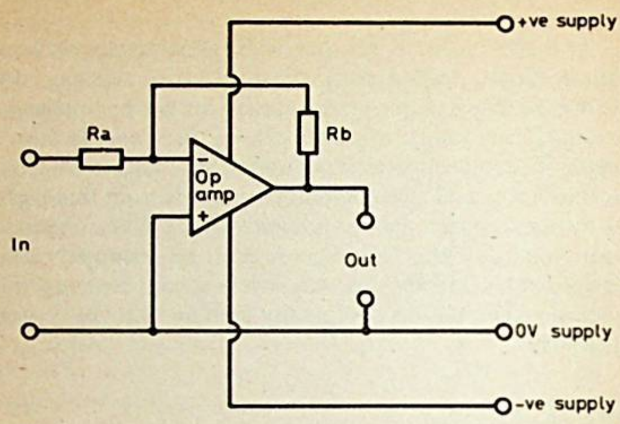

Fig. 2 The basic inverting amplifier configuration

The mathematics of this type of amplifier are very simple, with the input impedance being equal to the value given to Ra, and the voltage gain is equal to Rb divided by Ra. The required values for Ra and Rb are thus easily calculated with Ra being given the nearest preferred value to the required input impedance, and this figure is then multiplied by the required voltage gain in order to give the necessary value for Rb.

The way in which these two components set the input impedance and voltage gain is also very simple. Under quiescent conditions negative feedback through Rb will set the voltage at the inverting input and output of the operational amplifier at the same potential as the non-inverting input, or at the 0V supply rail potential in other words. This must be so, since a higher output voltage would unbalance the two input voltages due to the feedback through Rb, and would result in the output going negative until the two input voltages were balanced. Remember that in theory there is no voltage drop through Rb since no input current flows into the operational amplifier, and that any voltage difference across the inputs, no matter how small, is sufficient to send the output fully positive or fully negative. Obviously an excessively negative output would also unbalance the inputs and result in the output swinging positive in this case in order to balance the input voltages again. This balance is achieved with the output at the 0V rail potential.

If Ra is (say) one fifth of the value of Rb, and an input voltage of +1V is applied to the circuit, this input voltage unbalances the inputs of the operational amplifier and takes the inverting input positive with respect to the non-inverting input. The output swings negative in an attempt to rebalance the two inputs, and it must go 5 volts negative in order to achieve this. With the input at +1V and the output at –5V, by a simple potential divider action the required 0V is produced at the inverting input. The output must always go to five times the input voltage, and have the opposite polarity to the input signal in order to balance the input voltages of the operational amplifier.

If Rb is made higher in value there will be a larger voltage drop across this component and the output will need to produce proportionately higher voltages in order to balance the inputs. A lower value gives a reduced voltage drop, lower output voltage for a given input signal, and hence lower gain. Thus the closed loop voltage gain is controlled by Ra and Rb, and is equal to the ratio of Rb to Ra.

As the negative feedback action maintains the inverting input at the 0V earth rail potential, input signals produce a current flow in Ra that is the same as if Ra was actually connected to earth rather than to the inverting input. Thus the input impedance of the circuit is equal to the value given to Ra. Although no input current flows into the operational amplifier, there is a current flow through Ra, Rb, and the output circuitry of the operational amplifier, and an input current does flow into the amplifier circuit as a whole.

Non-Inverting Amplifier

The basic non-inverting amplifier configuration is shown in the circuit diagram of Figure 3. With this type of circuit the input signal is taken to the non-inverting input, and the latter is

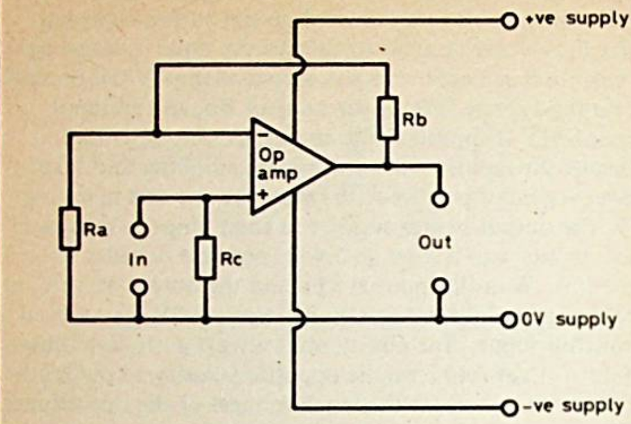

Fig. 3 The basic non-inverting amplifier configuration

biased to the 0V rail potential by a resistor (Rc in this case). The operational amplifier has an infinite input impedance, and the input impedance of the amplifier as a whole is therefore equal to the value of this bias resistor.

Ra and Rb again form a negative feedback network which set the closed loop voltage gain of the circuit. In this case though the voltage gain is equal to the sum of Ra and Rb divided by Ra, rather than simply Rb divided by Ra. For example, if Rb has a value which is five times larger than that of Ra, and an input of +1V is applied to the circuit, the output must assume a potential that gives +1V at the inverting input in order to balance the input voltages. The feedback is not maintaining the inverting input at the 0V rail potential as was the case with the inverting amplifier mode. The output needs to go six volts positive in order to give a voltage of 1V at the inverting input by the potential divider action across Rb and Ra, and the closed loop voltage gain of the circuit is therefore six times and not five times.

AC Amplifiers

Of course, the DC amplifier circuits of Figures 2 and 3 can also be used to amplify AC signals, and DC blocking capacitors can be added at the input and output if necessary. As we shall see later, it is possible to use a single supply rail for a circuit that only handles AC signals, and under certain circumstances it is possible to use a single supply for DC circuits.

An important point to bear in mind if an operational amplifier is being used as an AC amplifier is that although the DC voltage gain may be typically 100dB or more, and for most practical purposes can be regarded as being infinite, this gain reduces above frequencies of more than a few Hz. The graph of Figure 4 illustrates this point, and this actually shows the open loop voltage gain of the standard 741C operational amplifier at frequencies from one Hz to one Megahertz. As can be seen from Figure 4, even at a frequency of just ten Hz the gain has dropped significantly, and it continues to do so at a nominal rate of 6dB per octave with a voltage gain of only unity being reached at 1 Megahertz.

The 741C is obviously of little use as a radio frequency amplifier, and it cannot be used as a high gain audio amplifier since at the highest audio frequency of 20kHz the gain of the device is only about 50 times.

The reason for this severe rolling-off of the high frequency and audio frequency gain is that instability would otherwise occur. Like many operational amplifiers, the 741C has internal frequency compensation that enables it to be used at any closed loop gain above unity without instability occuring (provided stray positive feedback over the circuit is avoided). Some operational amplifiers require external frequency compensation and this is usually in the form of a single low value capacitor. The advantage of external frequency compensation is that it enables a greater bandwidth to be achieved provided the closed loop voltage gain of the circuit is greater than unity.

The 748C device is the externally compensated version of the 741C operational amplifier, and the graph of Figure 5 shows the open loop voltage gain of this device at frequencies from

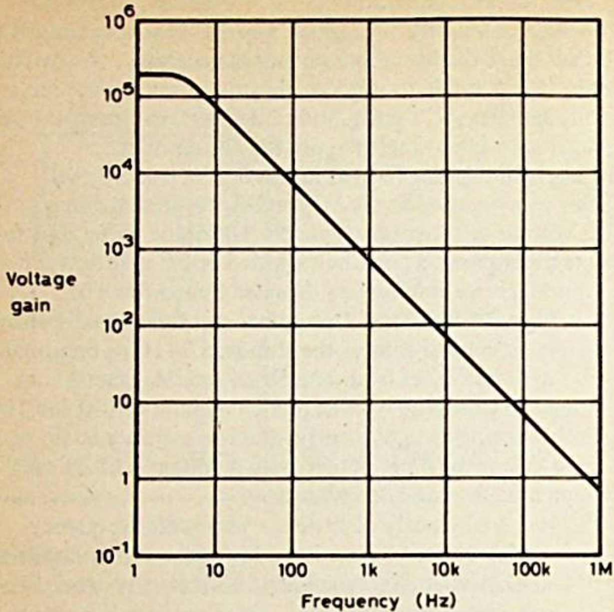

Fig. 4 Graph showing open loop gain versus frequency for the 741C device

1Hz to 1MHz using a 2pF compensation capacitor. At high frequencies the gain is over ten times higher than that provided by a 741C operational amplifier, but with a compensation capacitance of only 2pF the 748C must be used with a closed loop voltage gain of about one hundred or more or instability will arise.

Offset Null

In theory an operational amplifier circuit using either of the configurations described earlier has the inputs and output at the central 0V rail potential under quiescent conditions. In any practical circuit there are small errors in the various

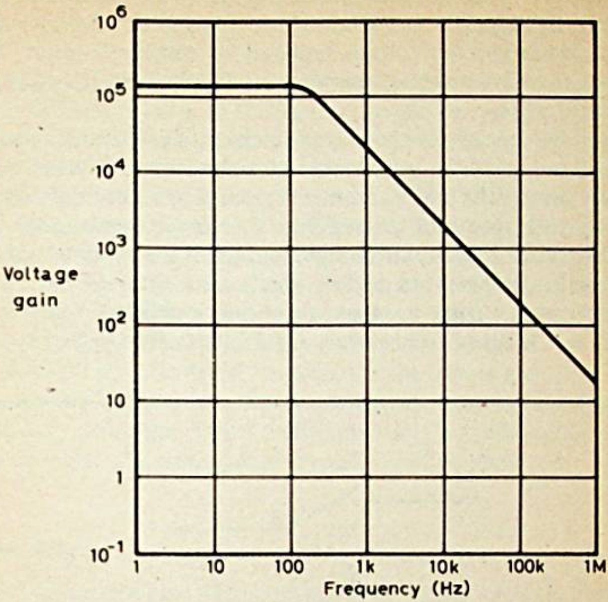

Fig. 5 The open loop gain of the 748C with a 2pF compensation capacitor

circuit voltages, and these occur due to the small input currents that flow into the operational amplifier, plus the fact that the differential input circuitry of a practical device is not going to be perfectly balanced. In other words, whereas the circuit should stabilise with the two inputs at precisely the same voltage, in practice the circuit will probably stabilise with a small potential (typically just two or three millivolts) across the two inputs. This can lead to the output drifting away from the 0V supply potential under quiescent conditions, especially if the amplifier has a high closed loop voltage gain since the error in the output voltage will be equal to the input offset voltage multiplied by the closed loop voltage gain of the circuit. With a closed loop gain of around one hundred to one thousand times the error in the output voltage would obviously be in the region of a few hundred millivolts, or even a few volts. A substantial biasing error such as this might not be of any consequence in AC amplifier applications, but it would be sufficient to render the amplifier unusable in most, if not all, DC applications.

The normal solution to this problem is to use an offset null control to enable the output voltage to be trimmed to the correct level. The 741C and many other devices use the offset null control circuit shown in Figure 6. Not all operational amplifiers use this circuit though, and there are a number of externally compensated devices which use a different configuration. There are also operational amplifiers available which are designed to have very small input offset voltages.

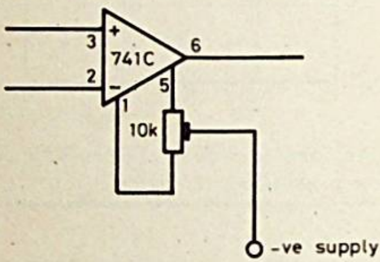

Fig. 6 An offset null control for a 741C (or equivalent) device

Latch-Up

As was stated earlier, in theory the actual input voltages to an operational amplifier are not important, it is the voltage difference across the inputs to which the device responds. With all practical operational amplifiers this is only true up to a point, and if one or both of the input voltages go outside certain limits the device ceases to function properly. This normally results in the output going fully negative or fully positive, and this is termed "latch-up". There are some devices that will function with the inputs virtually anywhere between the two supply rail potentials, but with most devices a malfunction occurs if one or both of the input voltages are very close to one or other of the supply rails. Thus, in our test circuit of Figure 1, if one of the potentiometers is set for practically zero volts at its wiper terminal it is quite likely that the proper voltage comparator action will not be obtained. The standard 741C device is one of those that perform well in this respect and is free from latch-up, but paradoxically some of the "improved" operational amplifiers that have become available in recent years seem to be inferior to the 741C in this respect.

With most operational amplifiers if the inputs are taken outside the limits of the supply potentials there is not just a likelyhood of the circuit malfunctioning, it is also quite possible that the operational amplifier will be damaged.

Slew Rate

In many applications the slew rate of an operational amplifier is of no real importance, but it is crucial in circuits where high frequencies and high signal levels are involved. The slew rate of an amplifier is the fastest rate of change that the output can achieve, and is normally specified in volts per microsecond. Slew rate is normally measured by applying a signal having an extremely high risetime and falltime to the input of an amplifier circuit using the device concerned, and using an oscilloscope to show the risetime and falltime of the signal at the output of the amplifier.

Many operational amplifiers have quite low slew rates, and the figure for the standard 741C device, for example, is only 0.5 volts per microsecond. This is a typical figure incidentally, and some examples of the 741C will inevitably be significantly slower than this.

This can result in an operational amplifier being unsuitable for an application where at first sight it might appear to be ideal. For example, the 741C might seem to be a good choice for use in the buffer amplifier at the output of an audio signal generator. However, if the signal generator has a maximum operating frequency of (say) 200kHz and a maximum output voltage of 10 volts peak to peak, with a triangular output waveform this would give a maximum slew rate of 4 volts per microsecond. For sinewave and squarewave outputs an even faster slew rate would be required, and the 741C would not be able to cope with such high combinations of signal level and operating frequency.

There are many devices having higher slew rates than the 741C, and as we shall see later, there are some very useful devices which have been designed to have very high slew rates.

Packages

One final point before moving on to the projects is that many operational amplifier integrated circuits are available in more than one package. For example, the 741C and 748C devices can be obtained in an eight leadout TO–99 metal encapsulation, an 8 pin DIL package, and a 14 pin DIL package. The 8 pin DIL version is by far the most common one, but I have been supplied with all three types over the years.

The stripboard layouts in this book have been designed to suit 8 pin DIL versions of integrated circuits that are available in more than one encapsulation. TO–99 versions will also readily fit into these layouts, and no modifications would be necessary. 14 pin DIL devices will not fit into the layouts without some modifications, and it is advisable to avoid the 14 pin DIL versions (which is not difficult as these are very uncommon these days).

CHAPTER 2

OP-AMP PROJECTS

With the wide range of high performance operational amplifier integrated circuits available these days one could be forgiven for thinking that there are no applications where the standard 741C device is satisfactory, and that a higher specification device will always give improved results. In fact in many simple applications, such as where a low frequency oscillator or low gain audio amplifier is needed, there is no advantage in using a higher specification device.

Slide Timer

An example of a circuit of this type is the Slide Timer circuit of Figure 7. This circuit merely pulses a relay on for a period of about one second, at an interval that is continuously variable from about five to thirty seconds. A pair of normally open relay contacts are used to activate the automatic slide change mechanism of a slide projector, and thus provide slide changes at the required frequency.

In this circuit the operational amplifier is used in a very common oscillator configuration which actually uses the operational amplifier as a form of Schmitt trigger. R1 and R2 bias the non-inverting input of IC1 to half the supply voltage, but due to the inclusion of R3 this bias voltage will always be substantially increased or decreased, depending on whether the output of IC1 is high or low.

When power is first applied to the circuit C2 will have zero charge, and the inverting input will be at a lower voltage than the non-inverting input so that the output goes high. This takes the bias voltage at the non-inverting input to around 75% of the supply voltage. C2 charges from the output of IC1 via two routes, the main one being through the relatively low resistance provided by R5 and D1. The other charge path is through VR1 and R4, and this has little influence on the charge

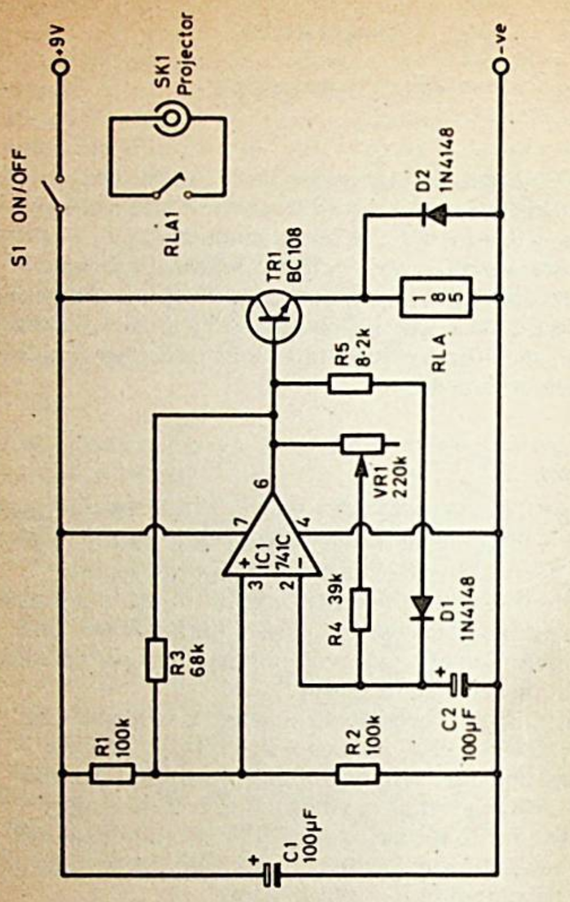

Fig. 7 The circuit diagram of the slide timer

rate due to the relatively high resistance values involved. It takes approximately one second for the charge on C2 to reach the same potential as that present at the non-inverting input of IC1, and the output of IC1 then starts to swing negative.

Due to the positive feedback through R3 this results in the voltage at the non-inverting input being reduced, the output going more negative, the non-inverting input voltage being further reduced, and so on, so that in practice the output almost instantly switches to the low state.

The voltage at the non-inverting input is then at only about 25% of the supply voltage, and C2 starts to discharge through R4, VR1, and the output circuitry of IC1. There is no discharge path through D1 and R5 since D1 blocks any significant current flow in this direction. After a period which can be varied from about five seconds with VR1 at minimum resistance to approximately thirty seconds with VR1 at maximum resistance, the charge on C2 falls below the voltage present at the non-inverting input. The output then starts to swing positive, and R3 again provides positive feedback so that the output quickly switches to the high state.

C2 then starts to charge up by way of R5 and D1 once again, and the whole cycle of events repeats itself indefinitely with a series of positive output pulses being produced by IC1.

These pulses are used to drive the relay, and Tr1 is an emitter follower buffer stage which enables sufficient output current to drive the relay to be obtained. Operational amplifiers are not power devices, and typically have a maximum output current of about 20mA.

There are obviously not very stringent requirements placed on the operational amplifier used in a circuit of this type since the device does not need to have a high slew rate, wide frequency response, ultra-high input impedance, low noise level, low distortion, or large output voltage swing in order to function properly in this circuit. The inexpensive 741C is as good as any other operational amplifier in this circuit.

Construction

Details of the 0.1in matrix stripboard component panel and wiring of the unit are shown in Figure 8. The unit is constructed in the normal way with a board of the correct size (13 strips by 19 holes in this case) first being cut out using a hacksaw, the breaks in the copper strips being made next, and then the two 3.3mm diameter mounting holes are drilled. The latter accept

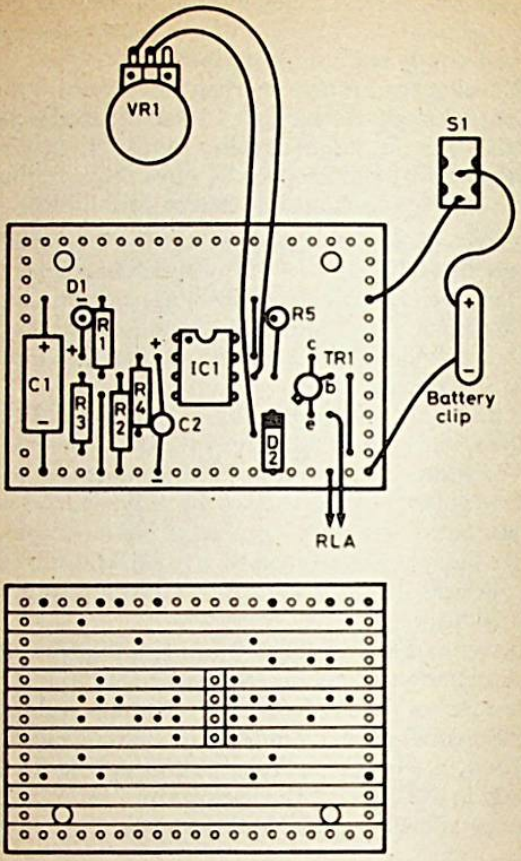

Fig. 8 Constructional details of the slide timer

either 6BA or M3 mounting bolts. The components and link wires are then soldered into place. Note that C2 should ideally be a tantalum capacitor and not an ordinary electrolytic type. Apart from the fact that a tantalum component will fit into the component layout much better, a tantalum component has greater reliability in this circuit than an electrolytic type. This is due to the rather high leakage levels and tolerances of electrolytic capacitors, especially higher value types.

The relay is not fitted on the component panel, but is mounted on the case at any convenient point. The method of mounting the relay must be varied to suit the particular component employed in the unit, but with virtually all modern relays it will simply be necessary to glue the component in place using a good quality general purpose adhesive.

The unit can be housed in virtually any small metal or plastic case, although some of the smallest types available are unsuitable as they have insufficient depth to accommodate the PP6 battery. Of course, if a physically large relay is used in the unit this will also necessitate the use of a larger case than would otherwise be required. The output socket is a 3.5mm jack, and this is connected to the socket on the projector via an ordinary twin lead (it does not have to be a screened type) terminated in the appropriate type of plug.

Components for Slide Timer (Figure 7)

Resistors, all 1/3 watt 5%

R1	100k	R2	100k
R3	68k	R4	39k
R5	8.2k		
VR1	220k lin carbon		

Capacitors

C1	100μF 10V electrolytic
C2	100μF 10V tantalum bead

Semiconductors

IC1	741C		
Tr1	BC108		
D1	1N4148	D2	1N4148

Switch

S1	SPST toggle type

Miscellaneous

Plastic or metal case

0.1in matrix stripboard
Control knob
Relay having a coil resistance of 185 ohms or more, an operating voltage of 6/12V, and at least one make contact of adequate rating.
3.5mm jack socket
PP6 battery and connector to suit (uses PP3 style connector)
Connecting cable, wire, solder, etc.

Audio Millivolt Meter

Of course, not all circuits are as undemanding as the Slide Timer circuit just described, and the Audio Millivolt Meter circuit of Figure 9 is an example of a circuit that uses high performance operational amplifier integrated circuits, and would give very poor results using 741C devices in place of those specified.

IC1 is simply used as a non-inverting unity voltage gain buffer stage which gives the instrument a high input impedance of about one Megohm, and this is necessary to ensure no significant loading on the equipment under test. Although it may not appear to be necessary for this device to have a particularly high level of performance, it should be borne in mind that the instrument has a frequency response that is virtually flat up to about 200kHz, and the least sensitive range of the unit is 1 volt RMS (the other two ranges provided being 100mV and 10mV fsd). The small signal bandwidth of the 741C device is adequate, but the slew rate of typically only 0.5 volts per microsecond is not, and a sinewave having an amplitude of 1 volt RMS and a frequency of 200kHz requires a slew rate of at least a few volts per microsecond.

High slew rate versions of the 741C are available, but MOSFET and JFET devices having high slew rates are less expensive and more readily available, and are a more practical choice. A TL081CP is specified for IC1, but there are alternatives such as the LF351 (which has a slew rate of 13 volts per microsecond) and the CA314OE (which has a slew rate of 9 volts per microsecond) which are equally suitable. All three devices have internal frequency compensation for voltage gains of unity or greater, incidentally.

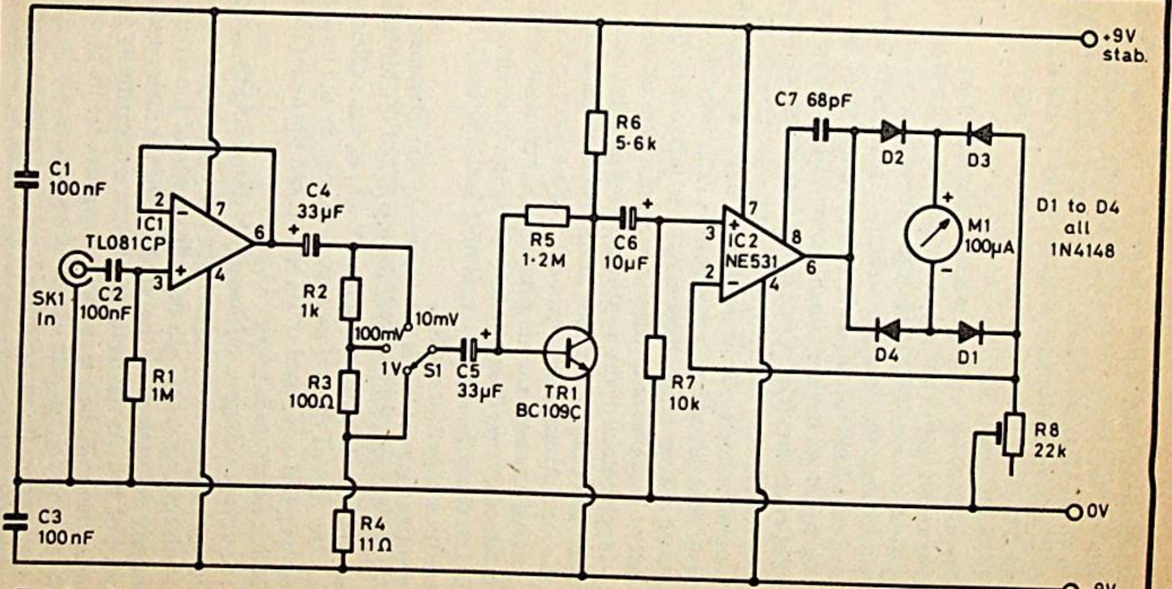

Fig. 9 The circuit diagram of the audio millivolt meter

C4 couples the output of IC1 to a simple attenuator which can be used to give a loss of 0dB, 20dB or 40dB, depending on the setting of S1. The latter is the range switch, and the basic sensitivity of the millivolt meter is 10mV RMS for full scale deflection so that the attenuator gives additional ranges of 100mV RMS and 1 volt RMS.

The output of the attenuator is coupled by C5 to a simple common emitter amplifier based on Tr1, and having a high voltage gain of around 40dB. This high voltage gain is necessary as the rectifier and meter circuitry requires an input of around 1 volt RMS for full scale deflection of the meter.

In order to obtain linear scaling it is essential to use an active rectifier circuit. This is due to the forward voltage drop across a semiconductor diode, with about 0.5 to 0.6 volts being required before a silicon rectifier will conduct reasonably well (germanium devices are rather better in this respect and have forward voltage drops of around 0.1 to 0.2 volts). The normal way of counteracting the non-linearity of semiconductor diodes in applications of this type is by using negative feedback with the diode or diodes included in the feedback path, and this is the method used here.

IC2 is used as a non-inverting amplifier and its non-inverting input is biased to the 0V rail by R7. C6 couples the output from Tr1 to the non-inverting input of IC2. C7 is the compensation capacitor for IC2.

The voltage gain of IC2 is determined by the resistance between the output and the inverting input, and the resistance between the inverting input and the 0V rail, as was explained earlier. The first of these resistances is formed by the bridge rectifier which is comprised of D1 to D4, and the meter (M1). The second resistance is provided by R8.

If the output voltage is sufficiently high to bring the bridge rectifier into conduction, the rectifier will contribute very little feedback resistance, and the voltage gain of the circuit is then largely determined by the resistance of the meter and the resistance of R8. As the meter will have a comparitively low resistance there will be a great deal of negative feedback and the closed loop voltage gain of IC2 will consequently be little more than unity. If, on the other hand, the output voltage is not high enough to bring the bridge rectifier into conduction, there will be a very high feedback resistance in this part of the circuit and IC2 will exhibit a very high closed loop voltage gain.

In practice this means that it is only necessary for the input signal to have a very low amplitude in order to send IC2's output about one volt or so positive or negative (depending on the polarity of the input signal). When the output achieves an amplitude of around 1 volt or so the bridge rectifier begins to conduct and the gain of the circuit drops back to little more than unity. The output needs to go about 1 volt above or below the 0V rail potential, rather than just 0.5 volts or so, since there is the voltage drop of two diodes to be overcome before the bridge rectifier conducts, and not just the voltage drop of a single diode.

With the meter driven to full scale deflection the waveform at the output of IC2 looks something like the waveform shown in Figure 10. The feedback effectively causes the output voltage of IC2 to be boosted by an amount equal to the voltage drop through the bridge rectifier so that this voltage drop is precisely counteracted. This gives the instrument linear scaling, and the existing scale of the meter can be used.

This is another example of a circuit where a high slew rate is required, and the high slew rate is necessary due to the output of IC2 having to very rapidly change from 1 volt or so positive to

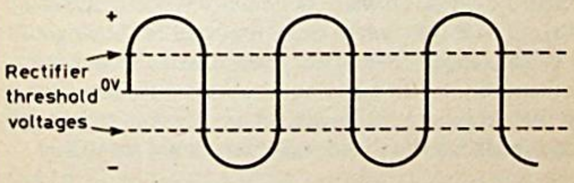

Fig. 10 The output voltage from IC2

1 volt or so negative, or vice versa, as the input signal crosses through zero volts. If the instrument is to function properly at frequencies of around one or two hundred kilohertz it is necessary for IC2 to have an extremely high slew rate indeed. The NE531 device specified for the IC2 position is specifically designed to have a high slew rate, and has an actual slew rate of 35 volts per microsecond with unity gain frequency compensation (a 100pF compensation capacitor). In this circuit a slightly lower compensation capacitance is utilized, and the slew rate is increased accordingly. The frequency response of the circuit does fall off slightly at frequencies of around 200kHz or so, but reasonable accuracy is provided at frequencies below this level. It will probably be found that C7 can be reduced substantially below the specified value of 68pF (probably to as little as 33pF) without the circuit becoming unstable, and there will then be very little fall off in performance even at frequencies in the region of 200kHz.

R8 controls the voltage gain of the output circuitry, and it also controls the current levels in the feedback circuitry of IC2. This enables the sensitivity of the circuit to be set at the appropriate level by adjusting R8.

The circuit is powered from dual 9 volt supplies, and the supply voltage is not too critical. Anything from about ± 6 volts to ± 10 volts should give good results. The supply must be stabilised though, since variations in the supply voltage will result in small but significant changes in the gain of Tr1. This would obviously affect the accuracy of the unit.

An interesting and important point regarding the NE531 device is that it does not have a particularly wide small signal bandwidth, and it is only when there is a large output signal level that it has a superior frequency response to most other operational amplifiers. In many applications this factor is an advantage since it makes the device no more prone to instability due to stray feedback than a normal operational amplifier.

Construction

A metal instrument case is probably the best type of case to use for a project of this type, but it may be necessary to use a larger type than might at first be expected since a reasonably large front panel is needed to accommodate the meter, and the accuracy of the unit does merit the use of a meter somewhat larger than the usual 60 x 45mm type should you wish to use a larger type. The only slight difficulty with the mechanical construction of the unit is the mounting of the meter which will require a large central cutout (38mm in diameter for most meters) in addition to the four smaller holes for the built-in fixing bolts. The main mounting hole can be cut using a fretsaw, a needle file, or a series of small holes can be drilled just inside the perimeter of the required cutout. With this last method it is necessary to punch out the metal within the ring of holes and then file out the hole to the correct size and a reasonably neat finish.

The 0.1in pitch stripboard layout for the millivolt meter is shown in Figure 11, and this is quite straight forward. The board has 16 copper strips by 30 holes and there are sixteen breaks in the strips, as can be seen from the diagram. A screened lead should be used to make the connections from the board to SK1, and the latter is a 3.5mm jack socket on the prototype, but any normal type of audio connector is suitable. Of course, the external lead at the input should be of the normal screened type used with audio test gear.

The circuit is designed for use with a mains power supply, and in order to minimise stray pick up of mains hum from the power supply the main circuitry should be kept reasonably well away from all the AC supply wiring. It might be found beneficial to screen the power supply from the main circuit. The mains tends to be poluted with a lot of RF interference these days, and it might be of benefit to use a mains RF filter to attenuate this interference since the higher frequencies of these signals makes them more easily coupled into the main circuit by stray coupling than is the 50 Hertz mains signal.

Note that attenuator resistors R2 to R4 are mounted on range switch S1 and are not fitted on the component panel. Refer to Figure 12 for wiring details of S1 and the three attenuator resistors.

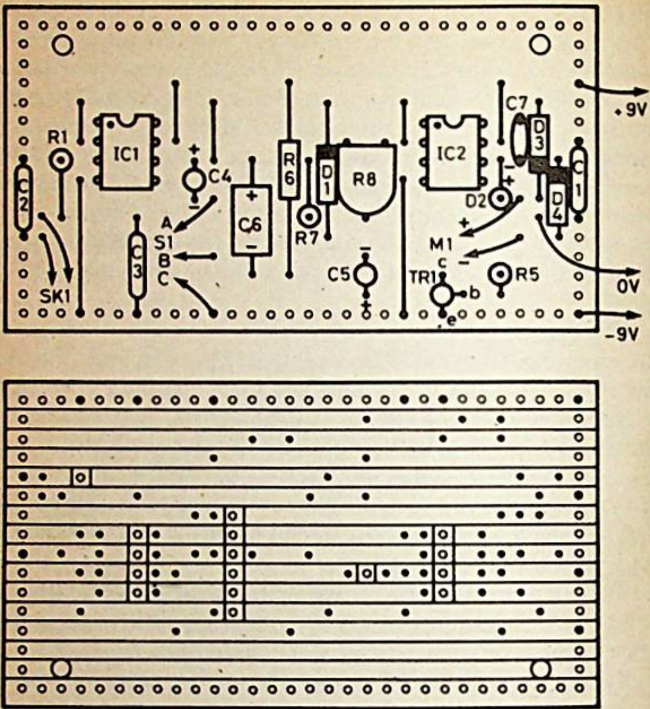

Fig. 11 Constructional details of the audio millivolt meter

Calibration

The unit is calibrated using a 1 volt RMS sinewave signal at a middle audio frequency of around 200 Hertz to 2kHz. Many audio frequency signal generators are capable of supplying a suitable signal, but many instruments will not have an accurately calibrated output level control, and may not be capable of

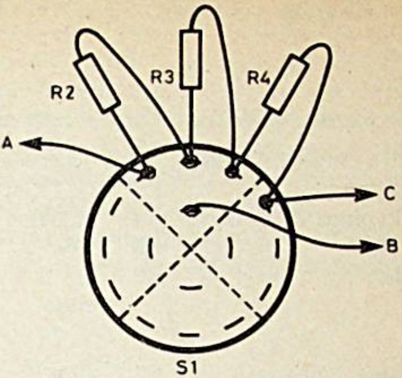

Fig. 12 The wiring to attenuator switch S1

giving an output level of 1 volt RMS with an accuracy of only around $\pm$ 2%. It is still possible to use a generator of this type to provide the calibration signal, but a multimeter set to a low AC volts range or an audio millivolt meter set to a suitable range must be used to monitor the output level of the generator so that the output can be accurately set at the required level.

Switch S1 to the 1 volt range, adjust R8 for minimum sensitivity (fully clockwise), switch the millivolt meter on, and connect the output of the generator to SK1, and adjust R8 for precisely full scale deflection of M1. The unit is then ready for use.

It is possible to calibrate the unit using a 100mV RMS or 10mV RMS signal if S1 is set to the appropriate range, but most multimeters will not be sufficiently accurate at these lower signal levels to enable the unit to be set up satisfactorily and this would necessitate the use of an audio voltmeter or a generator having an accurately calibrated output level control. Probably most constructors will not have access to either of these items of equipment.

It is also possible to calibrate the unit at less than full scale deflection, using (say) a 500mV RMS signal and adjusting R8 for half full scale deflection on M1. However, for optimum accuracy it is advisable to calibrate the unit at full scale deflection if possible.

Components for Audio Millivolt Meter (Figure 9)

Resistors, all 1/3 watt 5% unless noted otherwise

R1	1M	R2	1k 2% or better
R3	100 ohms 2% or better	R4	11 ohms 2% or better
R5	1.2M	R6	5.6k
R7	10k	R8	22k 0.1 watt horizontal preset

Capacitors

C1	100nF polyester	C2	100nF polyester
C3	100nF polyester	C4	33μF 10V tantalum
C5	33μF 10V tantalum	C6	10μF 25V electrolytic
C7	68pF ceramic plate		

Semiconductors

IC1	TL081CP	IC2	NE531
D1	1N4148	D2	1N4148
D3	1N4148	D4	1N4148
Tr1	BC109C		

Switch

S1 3 way 4 pole rotary (only one pole used)

Miscellaneous

Metal case
0.1in matrix stripboard
Control knob
100μA moving coil meter
3.5mm jack socket
±9 volt power source (see text)
Test leads, wire, solder, etc.

Simple AF Signal Generator

An AF signal generator of some kind is one of the most useful pieces of test equipment, and one of the first to be acquired by most electronics enthusiasts. The signal generator described here is simple and inexpensive to build but has quite good performance. It is based on a function generator integrated circuit and an operational amplifier. As we shall see later, the function generator device contains an operational amplifier and this is used in a logarithmic amplifier circuit.

The unit covers a frequency range of about 15Hz to 25kHz in three ranges which have the following (approximate) frequency spans.

Range 1 15Hz to 250Hz
Range 2 150Hz to 2.5kHz
Range 3 1.5kHz to 25kHz

Sine, square and triangular outputs are provided, and these have maximum peak to peak output voltages of 1.5 volts, 2 volts, and 4 volts respectively. By means of an output level control it is possible to set the output amplitude at anything from zero to the maximum figures quoted above. The output level does not vary significantly with variations in output frequency response measurements.

Operating Principle

The circuit is built around the 8038 function generator IC. This unit produces triangular and squarewave outputs direct from an oscillator, and then processes the triangular one to give a sinewave signal. The circuit used to convert the triangular waveform to a sinewave is an operational amplifier used as a logarithmic amplifier, and not as a normal linear amplifier. The basic idea is for the gain of the amplifier to fall slightly as the amplitude of the input signal is increased, so that a triangular waveform is rounded to give a sinewave of reasonable purity.

This method does not give a low enough level of distortion to permit the unit to be used to make distortion measurements on audio equipment, but the degree of purity is perfectly adequate for frequency response measurements and most other general audio testing. The actual total harmonic distortion in this case is in the region of 2%.

The circuit diagram shown in Figure 13 shows the way in which the transfer characteristic of the amplifier is shaped to have the required effect on a triangular input signal. As can be seen from this, there are a number of feedback paths from the output to the inverting input of the operational amplifier. With a low input signal level only R2 will provide a reasonably low impedance feedback path, as there will be too little voltage at the output of the operational amplifier to bias any of the diodes into conduction. It should be borne in mind that the inverting input is maintained at a fixed potential and the greater the output voltage swing, the greater the potential across R2 and the other feedback paths. At low input voltages then, R1 and R2 alone determine the closed loop gain of the circuit.

At higher output voltages D1 begins to conduct, and R3 is shunted across R2. This increases the amount of negative feedback and reduces the closed loop voltage gain of the circuit. If the output voltage is increased still further, D2 and

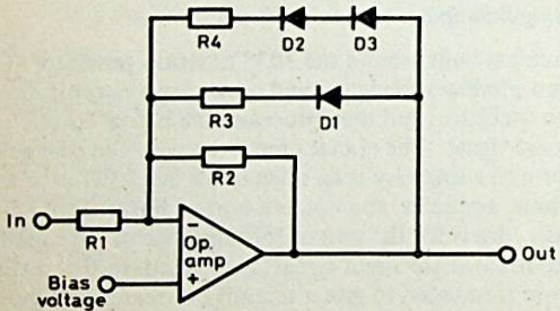

Fig. 13 An op-amp used as a logarithmic amplifier

D3 are biased into conduction so that R4 is shunted across R2 and R3, and the voltage gain of the circuit is further reduced.

In this way the required reduction in gain as the signal level rises is obtained, and in a practical circuit more sets of diodes in the feedback path might be required. Also, in a practical circuit each diode would have a parallel connected diode of opposite polarity so that the circuit would function properly with a negative going output.

The Circuit

The complete circuit diagram of the signal generator is given in Figure 14. There are three switched capacitors (C2 to C4) giving the unit its three ranges, and range switch S1 is used to select the desired timing capacitor.

VR1 and R1 form a potential divider circuit which provides a variable voltage that is used to control the charge and discharge currents of the timing capacitor. Raising and lowering these currents has the effect of increasing and reducing the operating frequency of the circuit. VR1 thus acts as the fine frequency control. R4 to R6 are used to control the relative charge and discharge currents, and by adjusting R5 it is possible to balance these so that the output has the correct 1 : 1 mark space ratio.

R2 and R3 are part of the sinewave conversion circuit, and are adjusted for optimum sinewave purity. R7 is a load resistor for the squarewave output of the 8038.

A signal generator should have a fairly low output impedance so that connecting the output to a normal load will not result in a significant drop in the output signal level. The outputs of the 8038 are at comparatively high impedances, and it is therefore necessary to include a buffer amplifier at the output.

This uses operational amplifier IC2 in the non-inverting mode. Its inverting input is connected direct to the output so that there is 100% negative feedback over the amplifier and it has unity voltage gain. IC2 is a CA313OT device which has a wide enough bandwidth and high enough slew rate to give good quality output waveforms, as well as giving a low output

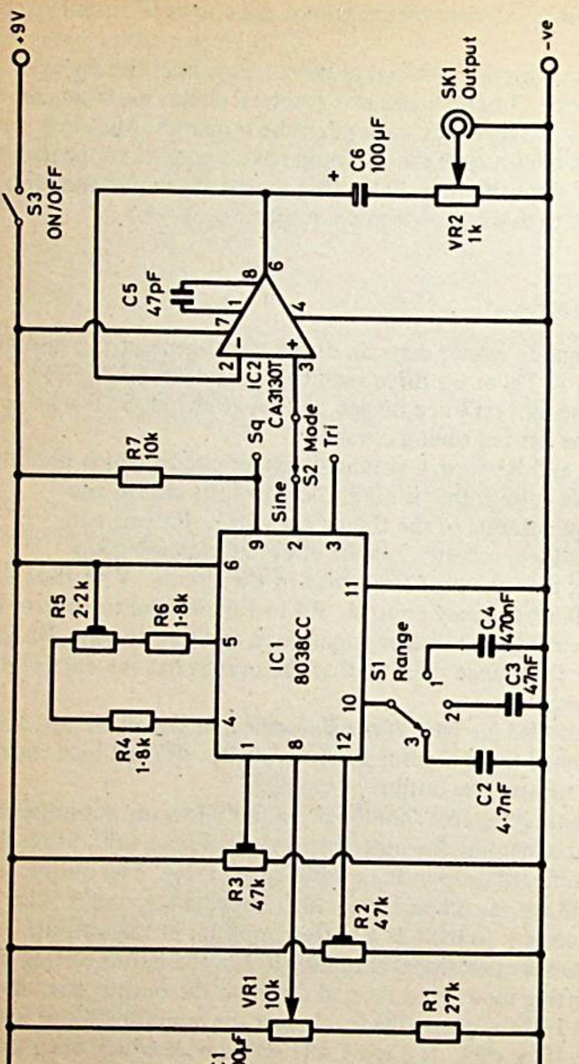

Fig. 14 The circuit diagram of the simple A.F. signal generator

impedance and presenting a high input impedance to IC1. The CA3130T is also capable of a high output voltage swing (virtually equal to the supply potential) so that there is no danger of the outputs being clipped and seriously distorted. The CA3130T requires a discrete compensation capacitor, and this is C5 in this circuit.

C6 couples the output of IC2 to the output level control, VR2. S3 is the on/off switch and C1 is a supply decoupling capacitor.

Although circuits which employ the 8038 IC usually have a supply potential of about 18 volts, this circuit was found to operate well using a 9 volt supply. The current consumption is about 9mA, and this can be provided by a small (PP3) size battery. However, if the unit is to be used a great deal it would probably be more economic to use a somewhat larger battery such as a PP9 or equivalent.

Construction

A metal instrument case makes an excellent housing for this project, and a type having approximate dimensions of 200 x 150 x 65mm will readily accommodate all the components and give ample panel space for the controls and output socket. If a smaller case is used it may well be found that the front panel becomes rather cramped.

Apart from the controls, output socket, battery and C2 to C4, the components are fitted onto a 0.1in matrix stripboard panel. This board is constructed in the usual manner and details of the board are provided in Figure 15. The board has 31 holes by 20 copper strips, and there are 15 breaks in the strips.

IC2 is a CMOS device and it should therefore be the last component to be connected. Also, it should be left in its protective packaging until it is to be fitted onto the board, and a soldering iron having an earthed bit should be used when connecting it.

Figure 16 shows the wiring to the controls and output socket. The identification letters used in this diagram correspond with those used in Figure 15. For example, point

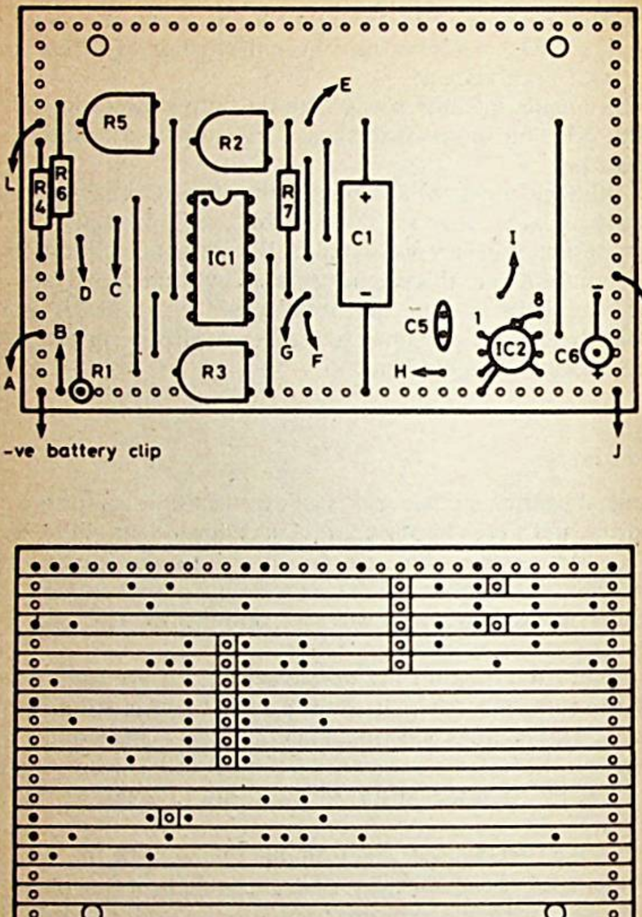

Fig. 15 Constructional details of the simple A.F. signal generator

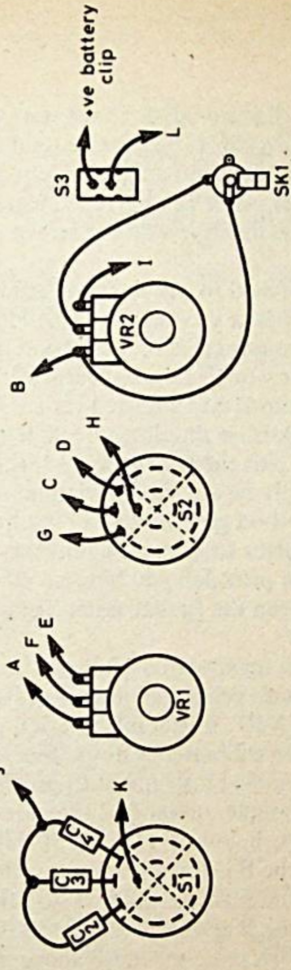

Fig. 16 The wiring to the controls and SK1

"A" in Figure 15 is joined to point "A" in Figure 16, the points marked "B" are joined, and so on.

Adjustment

Initially R2, R3 and R5 should all be set at a midway position. If an oscilloscope is available, this can be used to display the squarewave output of the unit, and R5 can be adjusted for a 1 : 1 mark space ratio. This can be done most accurately with just one or two cycles displayed on the screen of the oscilloscope.

With the unit switched to the sinewave mode, R2 and R3 are then adjusted for the best waveshape. An alternative method of adjusting the three presets is to feed the output to an amplifier/loudspeaker, or to a crystal earpiece. With the unit set to the sinewave mode and adjusted for an output frequency of around 100 to 500Hz it should be possible to hear the fundamental signal, plus the harmonics at higher frequencies. The presets can simply be adjusted to minimise the harmonics. I found that this method gave very good results. In fact it would seem to be better to adjust the unit this way instead of using an oscilloscope provided you have no difficulty in differentiating between the fundamental signal and the harmonics.

In order to obtain maximum usefulness from the unit it is necessary to add a scale calibrated in output frequency around the control knob of VR1. If suitable test equipment against which the unit can be calibrated is not available, an alternative method is to use a musical instrument to provide calibration frequencies. For example, on range 2 the unit could be calibrated using the D below middle C (150Hz), the G below middle C (200Hz), the B below middle C (250Hz), the D above middle C (300Hz), the F above middle C (350Hz), the G above middle C (400Hz), the B above middle C (500Hz), the second F above middle C (700Hz), the second B above middle C (1kHz), and the B above this (2kHz). The frequencies quoted above are not precise, but are very close to the frequencies of the specified notes in each case. It should be possible to estimate intermediate calibration points with adequate accuracy.

When using a musical instrument to calibrate the unit the output of the generator must be fed to an amplifier/speaker or a crystal earpiece. The generator can then be carefully adjusted to produce the same pitch as the calibration note.

Components for Simple AF signal Generator (Figure 14)

Resistors, all 1/3 watt 5% except presets

R1	27k	R2	47k 0.1 watt horizontal preset
R3	47k 0.1 watt horizontal preset	R4	1.8k
R5	2.2k 0.1 watt horizontal preset	R6	1.8k
R7	10k		
VR1	10k lin carbon	VR2	1k lin carbon

Capacitors

C1	100μF 10V electrolytic	C2	4.7nF plastic foil 5% or better
C3	47nF plastic foil 5% or better	C4	470nF plastic foil 5% or better
C5	47pF ceramic plate	C6	100μF 10V electrolytic

Semiconductors

IC1	ICL8038CC	IC2	CA3130T

Switches

S1	3 way 4 pole rotary (only 1 pole used)	S2	3 way 4 pole rotary (only 1 pole used)
S3	SPST miniature toggle type		

Miscellaneous

Metal instrument case
0.1in matrix stripboard
3.5mm jack socket (SK1)
PP3 battery and connector to suit (see text)
Four control knobs
Wire, solder, etc.

Simple Mixer

Few cassette decks and recorders have any form of built-in mixing, and although there are normally microphone and high level inputs (for use with a tuner, second recorder, etc.), it is normally a case of using one set of inputs or the other. This is of no consequence when making copies of tapes, records, or when making a straight forward recording from a tuner. The lack of any form of mixing facility is a major drawback for more creative forms of recording, such as making a tape to accompany a slide or cine show.

Usually all that is needed when making a tape recording of this type is a simple mixer, such as the one described here, which can mix a microphone signal with a high level signal, and provides sufficient output to drive a high level input of the cassette deck or recorder. Normally background music from (say) a second recorder will be fed into the high level input of the mixer, and a commentary is added to this via a microphone connected to the microphone input of the mixer. Many simple mixer designs cannot be used in this way as they do not have sufficient gain to bring the weak microphone signal up to a similar level to the music signal, but in this design there is up to 40dB of gain from the microphone input to the output (there is up to unity voltage gain from the high level input to the output).

The voltage gain available at the microphone input together with the input impedance of 47k make the unit suitable for use with a high impedance dynamic microphone or an electret type having a built-in step-up transformer. The unit is not recommended for use with crystal or low impedance dynamic (cassette type) microphones.

The Circuit

The complete circuit diagram of the Simple Mixer is provided in Figure 17, and this is based on a TL072CP dual operational amplifier. This device is a BIFET type which has a JFET input stage on the same chip as the remaining (mostly bipolar transistor based) circuitry. BIFET devices are ideal for use in

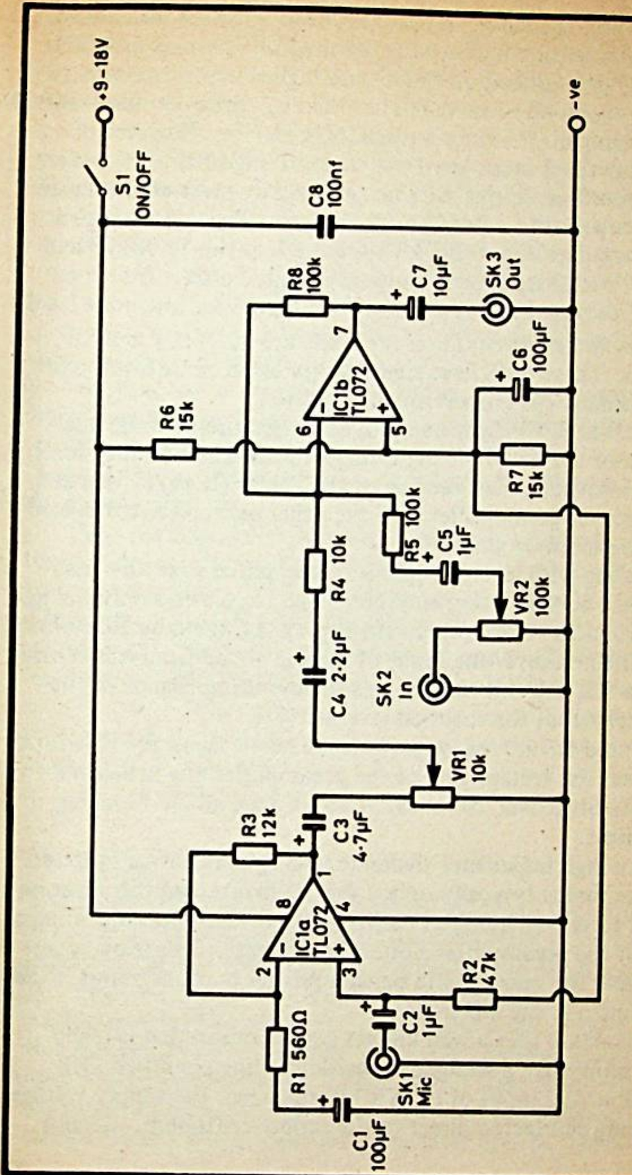

Fig. 17 The circuit diagram of the simple mixer

audio circuits as they have low levels of noise and distortion. They also usually have a wider bandwidth (about 3 or 4 MHz at unity gain instead of 1MHz) and higher slew rate about 10V per microsecond or more (rather than 0.5 volts per microsecond) when compared to the standard 741C device. However, it should perhaps be pointed out that although BIFET types are low noise devices, this does not necessarily mean that they are vastly superior to a 741C in this respect. The 741C is also a low noise device, and BIFET types such as the TL081CP and LF351 would seem to be only marginally better. There are BIFET devices which are designed to have very low noise levels, and the two most common types are the TL071CP and the LF356. These both have significantly lower noise levels than the standard 741C operational amplifier.

The TL072CP device employed in this circuit is the dual version of the TL071CP, and this gives the circuit a high level of performance. One section of the device (IC1a) is used as a microphone preamplifier, and the other section (IC1b) is used as a simple mixer stage.

Dealing with the microphone preamplifier first; this uses IC1a as a non-inverting amplifier. The circuit uses only a single supply, and a centre-tap on the supply is formed by R6, R7 and C6. The non-inverting input of IC1a is biased from this centre-tap via R2, and this resistor sets the input impedance of the preamplifier at the required level of 47k.

R1 and R3 are the negative feedback network for IC1a, and these set the voltage gain of the preamplifier at a little over 26dB (a little over 20 times). C1 to C3 are all DC blocking capacitors.

As a high impedance dynamic microphone has an output voltage that is typically only a few millivolts, and this must be raised to several hundred millivolts in this instance, the voltage gain of the preamplifier alone is inadequate. This is overcome by giving the microphone signal a further boost of about 20dB (10 times) in the mixer stage.

The mixer uses a well known configuration that is really little more than a straight forward inverting amplifier. The non-inverting input of IC1b is biased to half the supply voltage by being connected direct to the supply centre-tap. R5 and

R8 form the negative feedback network as far as the high level input is concerned, and set the voltage gain of IC1b at unity. C5 provides DC blocking at this input and VR2 is the level control for this input.

The output from the microphone preamplifier is fed to microphone level control VR1, and from here the signal is coupled by C4 to a second input feedback resistor. This is R4, and with R8 sets the voltage gain of IC1b, as far as the microphone signal is concerned, at 20dB. IC1b is used in what is often termed the "summing mode", and this term is derived from the fact that the output of the amplifier must assume a voltage that gives a current flow through R8 which matches the sum of the currents through R4 and R5.

The negative feedback action stabilises the potential at the inverting input of IC1b at half the supply voltage, and the virtual earth that is formed here isolates the two inputs from one another so that there is no interaction between the two level controls.

Only negligible levels of distortion are produced by the circuit at audio frequencies, and the signal to noise ratio of the circuit (assuming an output level of around 500mV to 1V RMS) is well over –60dB (unweighted). An output level of up to about 2 volts RMS can be accommodated before clipping occurs, and this can be raised to over 5 volts RMS if an 18 volt supply (such as two PP3 size batteries in series) is used instead of a 9 volt supply. The current consumption of the circuit is only about 4.5mA.

Construction

The 0.1in pitch stripboard layout and details of the mixer wiring are shown in Figure 18. The component panel has 13 copper strips by 24 holes and is constructed in the normal way. IC1 does not require any special handling precautions. Provided a metal case (earthed to the negative supply rail) is used for this project it is not necessary to use any screened leads on the interior of the unit. There is little risk of stray feedback causing instability, and the case will screen the sensitive circuitry from mains hum and other interference. Of course, external cables which connect to the input and output sockets should be the usual audio screened type.

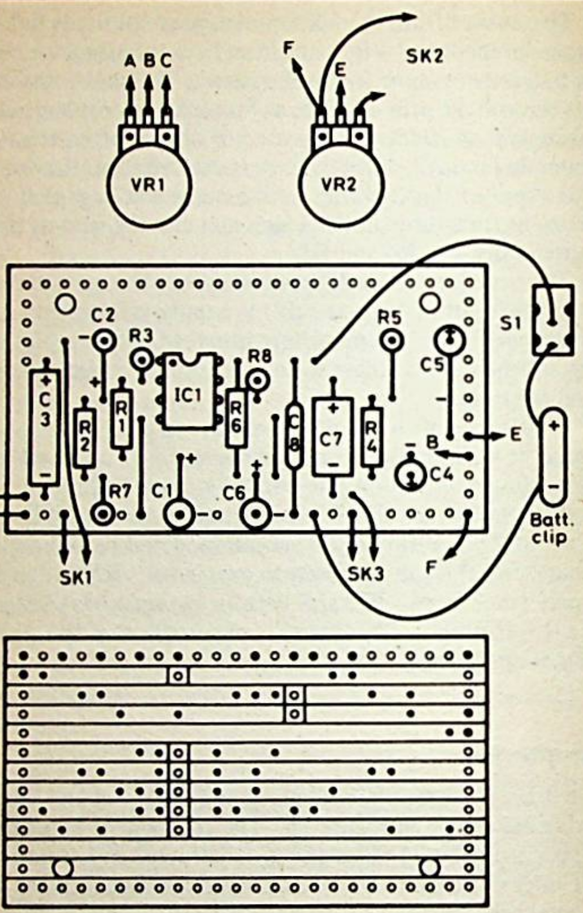

Fig. 18 Constructional details of the simple mixer

Components for Simple Mixer (Figure 17)

Resistors, all 1/3 watt 5%

R1	560 ohms	R2	47k
R3	12k	R4	10k
R5	100k	R6	15k
R7	15k	R8	100k
VR1	10k log carbon	VR2	100k log carbon

Capacitors

C1	100μF 10V electrolytic	C2	1μF 63V electrolytic
C3	4.7μF 63V electrolytic	C4	2.2μF 63V electrolytic
C5	1μF 63V electrolytic	C6	100μF 10V electrolytic
C7	10μF 25V electrolytic	C8	100nF polyester

Semiconductor

IC1 TL072CP

Switch

S1 SPST miniature toggle type

Miscellaneous

Case
0.1in matrix stripboard
Three standard jack sockets (SK1 to SK3)
Two control knobs
PP3 battery and connector to suit (see text)
Wire, solder, etc.

Microphone Preamplifier

Low impedance dynamic microphones (the type used with most cassette recorders) are quite inexpensive and mostly have an output of reasonably good quality. A major drawback of this type of microphone though, is the extremely low output level which is usually in the region of a few hundred microvolts RMS. This makes it difficult to achieve a really good signal to noise ratio when using this type of microphone due to the high voltage gain needed to bring the signal level up to a usable level for most applications. Also, unlike a tape preamplifier (which handles a similar input level) a low impedance dynamic microphone preamplifier does not use any treble cut, or any form of equalisation for that matter. A microphone preamplifier does not, therefore, have the advantage of a considerable subjective improvement in signal to noise ratio that is obtained using equalisation.

This microphone preamplifier has a voltage gain of up to almost 70dB (nearly 3000 times), and any low impedance dynamic microphone should be capable of producing an output of 1 volt RMS or more from the unit. This is sufficient to enable such a microphone to be used with virtually any power amplifier, mixer, or other item of equipment which does not have an input suitable for direct use with a low impedance dynamic microphone. The signal to noise ratio (referenced to a 1 volt RMS output level) is better than –60dB, which is very good when the high sensitivity of the circuit is taken into account.

The circuit can easily be modified for use with a high impedance dynamic microphone, or an electret type having an integral step-up transformer, and as described later, this merely entails changing the value of two resistors and two capacitors. This version of the circuit has less gain and the signal to noise ratio is consequently improved by over 10dB.

The Circuit

The circuit diagram of the microphone preamplifier is shown in Figure 19, and as can be seen from this, two stages of amplification are used. The first uses IC1 in the inverting mode, and the second uses IC2 in the non-inverting mode.

In order to obtain a good signal to noise ratio it is necessary for the device used in the IC1 position to have a very low noise level, and this is achieved by using the NE5534A operational amplifier in this stage of the unit. This is rather an expensive device when compared with most other operational amplifiers, but in audio applications it gives an improvement in signal to noise ratio of around 10 to 20dB when compared with the less expensive alternatives. The NE5534A is a bipolar device incidentally, and does not have a JFET input stage.

R2, R3 and C2 form a centre-tap on the single supply and this is used to bias the non-inverting input of IC1. R1 and R4 form the negative feedback network and these set the input

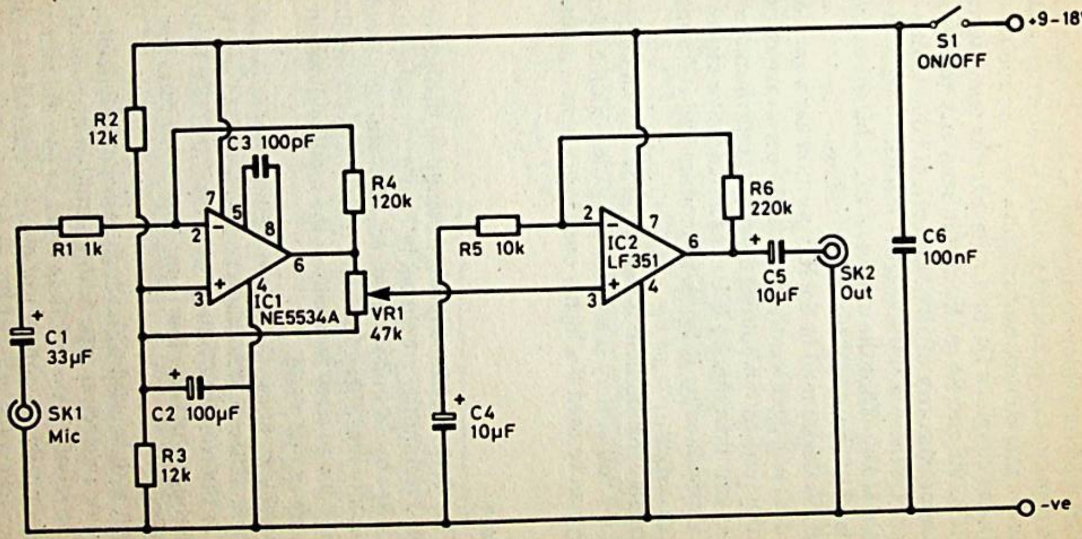

Fig. 19 The circuit diagram of the microphone preamplifier

impedance of the circuit at about 1k and the closed loop gain of IC1 at a little over 40dB (100 times). The NE5534A does not have internal frequency compensation, and discrete compensation capacitor C3 is needed in order to prevent instability.

VR1 is the gain control, and in addition to the audio output signal this also couples a DC bias voltage to the non-inverting input of IC2. R5 and R6 set the audio frequency voltage gain of IC2 at approximately 27dB (23 times), but due to the inclusion of C4 there is 100% negative feedback and only unity voltage gain at DC. Thus the output of IC2 is biased to the required level of about half the supply voltage.

The LF351 device used in the IC2 position has a low noise level and does not significantly degrade the performance of the circuit in this respect. Of course, the noise performance of IC2 is less important than that of IC1 since any noise produced by the latter is amplified by IC2. Noise produced by IC2 appears direct at the output.

The current consumption of the circuit is only about 5 to 6mA, and a 9 volt battery is suitable as a power source. An 18 volt supply can be used if a higher overload margin is required.

Construction

Details of the 0.1in matrix stripboard component panel for the Microphone Preamplifier are shown in Figure 20, and a board having 12 strips by 24 holes is required. Construction should not produce any real difficulties, and there is little possibility of instability due to stray feedback as the input and output of the circuit are out-of-phase.

It is advisable to house the unit in a metal case to provide screening, and a diecast aluminium box is ideal. The very high sensitivity of the circuit obviously makes it vulnerable to stray pick-up, although the fairly low input impedance eases the problem slightly.

If the unit is to be used with a high impedance dynamic microphone or an electret type having a built-in transformer, R1 should be raised to about 22k in order to give the unit a higher input impedance, and C1 can then be reduced to 1.5μF in value. Increasing the value of R1 might excessively reduce

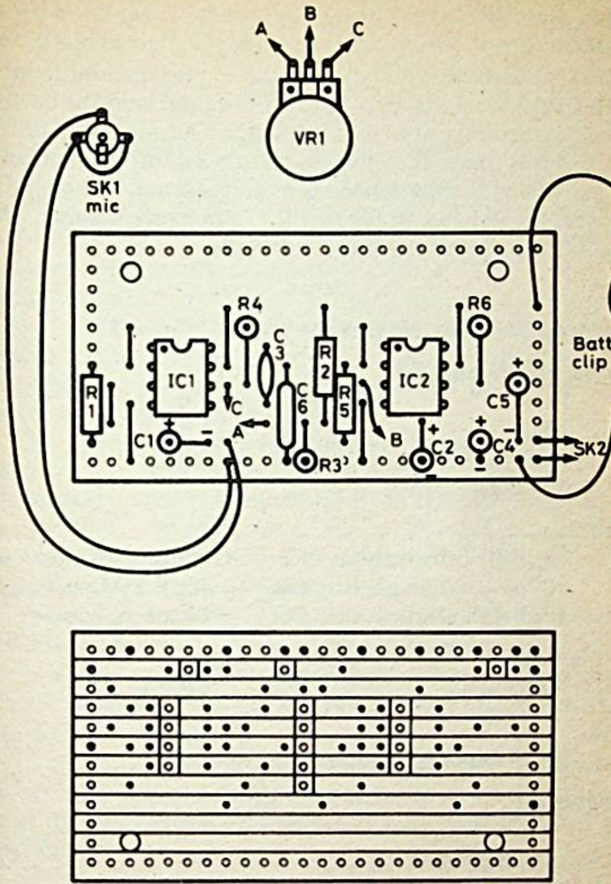

Fig. 20 Constructional details of the microphone preamplifier

the voltage gain of the input stage, and R4 can be increased to about 470k if a boost in gain is required. Due to the lower voltage gain of IC1 (even with R4 at around 470k) C3 must be increased to about 1nF in order to maintain good stability.

Note that with no microphone plugged into the unit there is 100% negative feedback over IC1, and a danger that the device will oscillate strongly at a high frequency. A simple way of avoiding this is to use the break socket on SK1 to short circuit the input when the microphone is unplugged, and this method of connection is shown in Figure 20. 3.5mm jack sockets normally have a break contact incidentally.

Components for Microphone Preamplifier (Figure 19)

Resistors, all 1/3 watt 5%

R1	1k	R2	12k
R3	12k	R4	120k
R5	10k	R6	220k
VR1	47k log carbon		

Capacitors

C1	33µF 10V electrolytic	C2	100µF 10V electrolytic
C3	100pF ceramic plate	C4	10µF 25V electrolytic
C5	10µF 25V electrolytic	C6	100nF polyester

Semiconductors

IC1	NE5534A	IC2	LF351

Switch

S1 SPST miniature toggle type

Miscellaneous

Case
0.1in matrix stripboard
Two 3.5mm jack sockets (SK1 and SK2)
PP3 battery and connector to suit
Control knob
Wire, solder, etc.

Magnetic Pick-Up Preamp

A weakness of many hi-fi amplifiers that are a few years or more in age is a relatively poor signal to noise ratio when using a magnetic cartridge. One reason for this is simply that audio semiconductors have steadily improved over the years, and many amplifiers of more than a few years of age have circuitry which uses components that are not a match for the best available today. Another problem that can arise with equipment of more than a few years old is breakthrough of noise from the mains supply, and this is caused by the high level of noise on the mains supply these days, a level that causes problems with many amplifiers which were designed at a time when far lower noise levels were present. A third reason for a relatively low signal to noise ratio is that many budget amplifiers use inexpensive components that give far from "state of the art" performance.

One way of obtaining improved performance under these circumstances is to use an external magnetic pick-up preamplifier, and feed the output from this into a high level input ("tape", "tuner", "aux." etc.). Provided a high quality preamplifier, such as the one described here, is used, a very large improvement in performance can be obtained.

This design has a signal to noise ratio that is better than –80dB (unweighted and referenced to an output level of 700mV RMS). The distortion at 1kHz with an output level of 700mV RMS is only about 0.004%, and the circuit produces no significant loss of quality.

Of course, the preamplifier does not have to be used as an add-on unit for an existing hi-fi amplifier, and it should also be of use to someone who wishes to construct their own hi-fi amplifier.

The Circuit

Refer to Figure 21 for the complete circuit diagram of the Magnetic Pick-Up Preamplifier. The high performance of the circuit is achieved by using a ZN424E low noise, low distortion operational amplifier as the basis of the unit. This device has

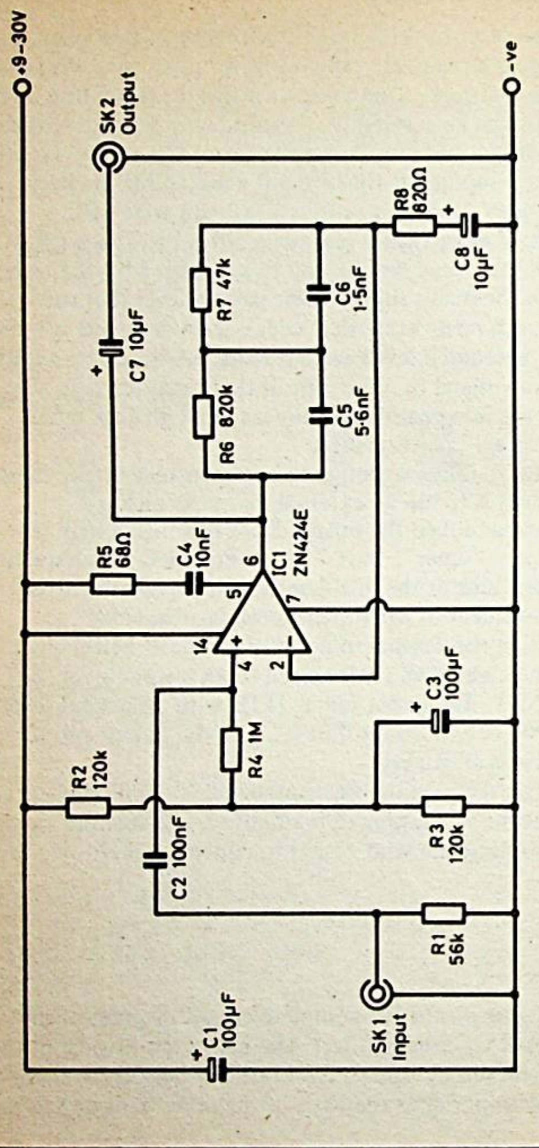

Fig. 21 The circuit diagram of the magnetic pick-up preamplifier

a lower open loop voltage gain than most operational amplifiers, the typical figure being 86dB (20000 times), but it also has the unusual feature of a low level of open loop distortion. The total harmonic distortion is only about 1.5% at audio frequencies and with an output voltage of 700mV RMS.

Although 1.5% may not seem very low, it must be borne in mind that this is the open loop figure, and the use of negative feedback not only reduces the voltage gain of an amplifier, it also reduces distortion (provided the amplifier is not overloaded). In this circuit, at middle audio frequencies, negative feedback reduces the gain of the amplifier by nearly 52dB (400 times), and the distortion is reduced by the same amount. Thus the total harmonic distortion is only 0.004%, and is too small to be of any significance.

IC1 is used in the non-inverting mode, with R4 biasing the non-inverting input from a centre-tap on the supply lines formed by R2, R3 and C3. C2 couples the input signal to the non-inverting input of IC1, and R1 is used to reduce the input impedance of the circuit to a suitable level (about 50k). Some pick-ups require an input impedance of 100k, and in such cases R1 should be increased to 120k in value.

A magnetic cartridge preamplifier does not have a flat frequency response, but must provide treble cut and bass boost. The maximum cut and boost is about 20dB, and this is needed to counteract the treble boost and bass cut used during the recording process, so that overall a flat frequency response is obtained. The treble boost and cut are used to provide noise reduction, and the bass cut and boost are needed to prevent excessive groove modulations.

The required tailoring of the frequency response is obtained by using frequency selective negative feedback. At middle audio frequencies the impedance of C5 is low in relation to that of R6, and R6 is effectively short circuited. The impedance of C6 on the other hand, is very high in comparison to that of R7. At middle audio frequencies the voltage gain of the amplifier is largely determined by R7 and R8. The specified values give a voltage gain of about 35dB (57 times) at these frequencies.

At lower frequencies the impedance of C5 becomes higher so that the required bass boost is produced. R6 limits the maximum boost, and at very low frequencies the impedance of C8 starts to rise significantly and the response of the circuit is rolled off. This sub-audio roll off helps to avoid problems with strong very low frequency signals caused by record warps etc.

At high audio frequencies C6 shunts R7 and this gives the treble cut. A resistor could be added in series with C6 to limit the amount of attenuation above the upper limit of the audio spectrum, but this would not give any audible difference in performance, and not doing so reduces the possibility of problems with RF breakthrough.

The current consumption of the circuit is only about 3mA, and a 9 volt supply is sufficient, although a higher supply potential (30 volts maximum) gives an improved overload margin.

Construction

The 0.1in matrix stripboard for the Magnetic Pick-Up Preamplifier is shown in Figure 22. There should be no problems with the construction of the unit, and although the input and output of the amplifier are in-phase, the fairly low gain at high frequencies prevents any problems with instability due to stray feedback.

Once again this is a project which should be housed in a metal case such as a diecast aluminium type to provide screening, and the unit is very vulnerable to stray pick-up of mains hum. This is due to the bass boost which gives very high sensitivity at 50Hz, and the fairly high input impedance of the circuit. If a mains power supply is used for the unit this must be kept as far away from the preamplifier wiring as possible, and it would be advisable to add a screen in the case between the preamplifier and power supply sections. The lead connecting SK1 to the component panel should be a screened type.

Although the unit has been described here as a mono unit, if a stereo version is required it is simply necessary to build two preamplifier boards, one for each stereo channel. These

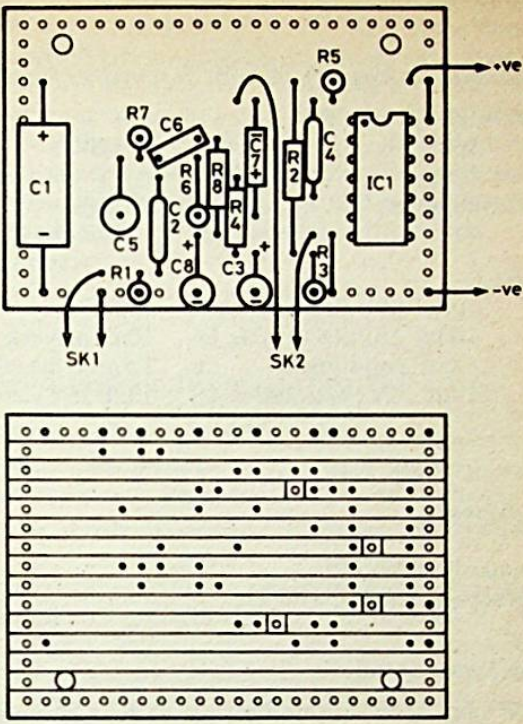

Fig. 22 Constructional details of the magnetic pick-up preamplifier

can have a common case and power source. The same is also true of the Microphone Preamplifier and Simple Mixer units described earlier, incidentally.

It should be found that the output of the unit is sufficient to drive a high level input of an amplifier, but it is possible that with some combinations of amplifier and pick-up there will be insufficient output. In such cases R8 can be reduced slightly in value (say to 470 ohms) in order to increase the gain of the circuit. It is not advisable to make R8 much less than about 470 ohms in value.

Components for Magnetic Pick-Up Preamplifier (Figure 21)

Resistors, all 1/3 watt 5%

R1	56k	R2	120k
R3	120k	R4	1M
R5	68 ohms	R6	820k
R7	47k	R8	820 ohms

Capacitors

C1	100µF 40V electrolytic	C2	100nF polyester
C3	100µF 16V electrolytic	C4	10nF polyester
C5	5.6nF polystyrene	C6	1.5nF polycarbonate
C7	10µF 25V electrolytic	C8	10µF 25V electrolytic

Semiconductor

IC1	ZN424E

Miscellaneous

Case
0.1in matrix stripboard
Power source, sockets, wire, etc.

Active Tone Controls

Passive tone control circuits have the obvious advantage of needing no power supply, and a further point in favour of this type of circuit is a slight reduction in expense when compared to a comparable active tone control circuit. However, it can often be awkward to incorporate a passive tone control circuit into equipment due to the losses that occur through this type of circuit. When set for a flat response there is typically a 20dB loss through a passive tone control network, and when adjusted to give bass or treble boost the circuit is in fact just giving reduced losses at these frequencies, rather than a genuine boost to the processed signal.

An active tone control circuit is really just a passive tone control circuit connected in the feedback path of an amplifier so that frequency selective negative feedback is obtained, and the required tailoring of the amplifier's frequency response is produced. Using this type of circuit it is possible to produce a design that has a nominal voltage gain of unity, and when adjusted to give boost or cut, does actually give voltage gain or attenuation over the appropriate frequency band. The advantage of a circuit of this type is that it can be added into a signal path without altering the overall gain of the system. Thus, if you have a preamplifier and a power amplifier that are compatible with one another, adding an active tone control of this type between the two is unlikely to introduce any matching problems.

The Circuit

Figure 23 shows the circuit diagram of the Active Tone Controls, and this is a conventional arrangement having bass control VR1 and treble control VR2. With the sliders of VR1 and VR2 fully towards the left there is maximum feedback, and hence full bass and treble cut. With the wipers at the opposite ends of their tracks there is minimum feedback and therefore maximum bass and treble boost.

The controls have no significant effect at middle audio frequencies (around 800Hz), and provide maximum boost and attenuation figures of about 12dB. The full amount of cut and boost is only available at the extremes of the audio frequency range, and 12dB is about the most that would ever be needed in practice.

IC1 is used in the inverting mode, and the non-inverting input of this device is therefore simply biased to half the supply potential by R1 and R2. C2 decouples any noise which might otherwise be fed to the non-inverting input from the supply lines via R1 and R2, or picked-up due to stray coupling.

The levels of noise and distortion produced by the circuit are negligible, even with the controls set for maximum boost (which still results in the circuit having only a very low level of voltage gain). When one or both of the controls are set for a cut in the response of the circuit, at some frequencies IC1 has a closed loop gain of less than unity. With some internally compensated operational amplifiers a closed loop gain of less

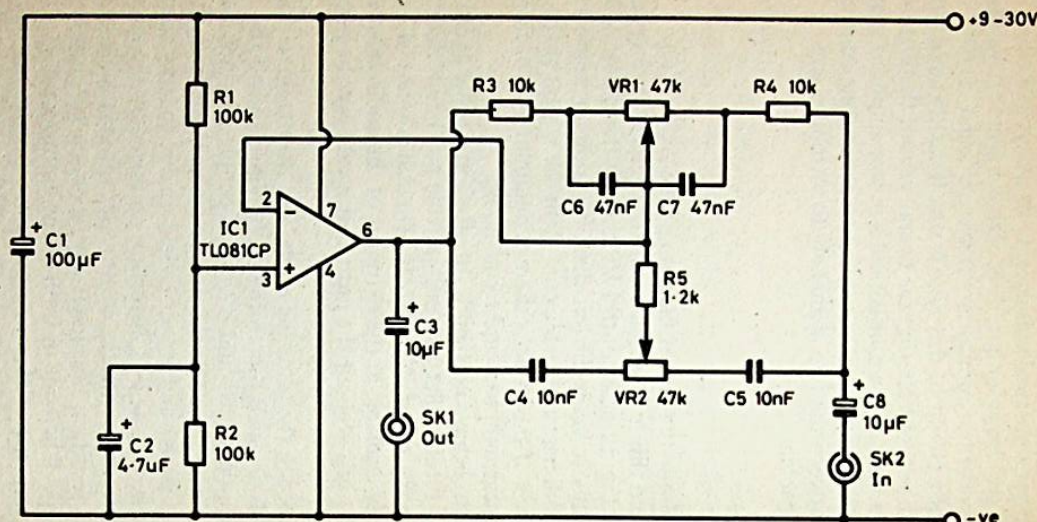

Fig. 23 The circuit diagram of the active tone controls

than unity can cause instability, and the internal compensation is only for closed loop voltage gains of unity or greater. Several TL081CP devices were tried in the circuit and no problems with instability were experienced. The circuit also works well using a 741C device, and in practice is is unlikely that any noticeable fall off in performance would ever be apparent using this device, although the noise and distortion levels are slightly higher than those obtained using the TL081CP.

Construction

The tone controls can be built as a self contained unit, but it is more likely that they will be used as part of a larger project such as a hi-fi amplifier or receiver. In either case the 0.1in matrix stripboard layout of Figure 24 can be used, and this is based on a board having 15 copper strips by 26 holes.

If a stereo version of the unit is required it will be necessary to make up two component panels, and the two bass control potentiometers would of course be a dual gang 47k potentiometer, as would the two treble control potentiometers. The circuit will work well using any supply voltage between about 9 and 30 volts, and it is not necessary to have a particularly well smoothed supply.

Components for Active Tone Controls (Figure 23)

Resistors, all 1/3 watt 5%

R1	100k	R2	100k
R3	10k	R4	10k
R5	1.2k	VR1	47k lin carbon
VR2	47k lin carbon		

Capacitors

C1	100μF 40V electrolytic	C2	4.7μF 63V electrolytic
C3	10μF 25V electrolytic	C4	10nF polyester
C5	10nF polyester	C6	47nF polyester
C7	47nF polyester	C8	10μF 25V electrolytic

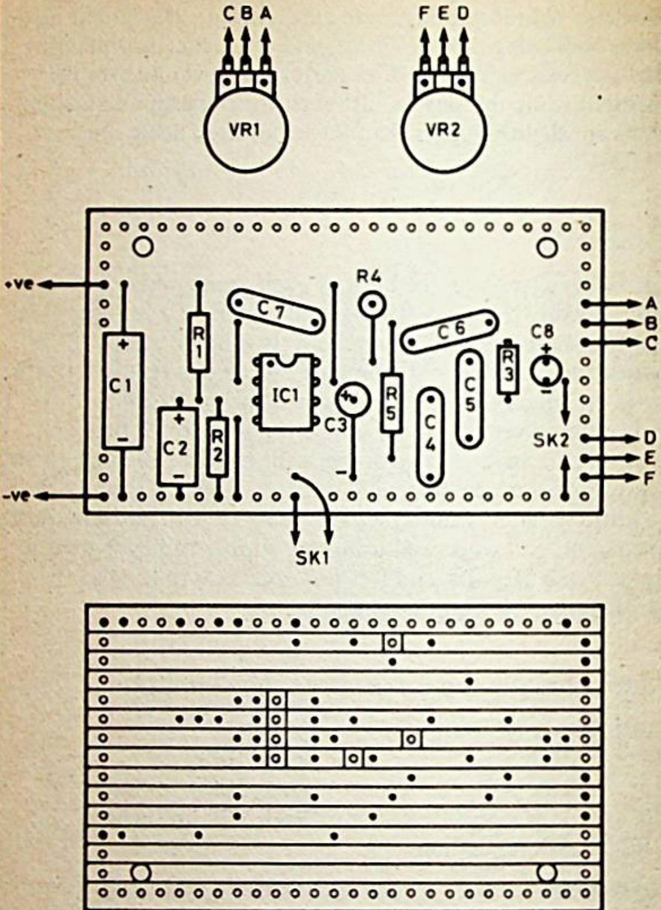

Fig.24 Constructional details of the active tone controls

Semiconductor
IC1 TL081CP

Miscellaneous
0.1in matrix stripboard
Two control knobs
Power source
Wire, solder, etc.

Microphone Compressor

Operational amplifiers are ideal for use in voltage controlled amplifiers (VCAs) since a change in resistance in the feedback network is converted to a change in voltage gain. This circuit uses a VCA which is based on an operational amplifier and a JFET; the latter being used as a voltage controlled resistor.

The circuit is designed for use with a high impedance dynamic microphone or an electret microphone having a built-in step-up transformer. The circuit is a form of automatic volume control and gives a virtually constant output level over a wide range of input signal levels. Units of this type are useful in virtually any application where a microphone is used, since without any form of automatic gain control there are likely to be times when the microphone signal is inadequate to give good results, and others when it becomes excessive and causes overloading.

The Circuit

Figure 25 shows the circuit diagram of the Microphone Compressor unit, and IC1 is used in the VCA.

Tr1 is the voltage controlled resistor, and under quiescent conditions this has its gate terminal biased to the negative supply rail by R11. This gives a gate-to-source voltage of zero so that Tr1 is switched on and exhibits a drain-to-source resistance of only a few hundred ohms. A JFET device, incidentally, is normally switched on and requires a reverse gate bias in order to bias it off. This is not like an ordinary bipolar transistor which is normally switched off, and requires

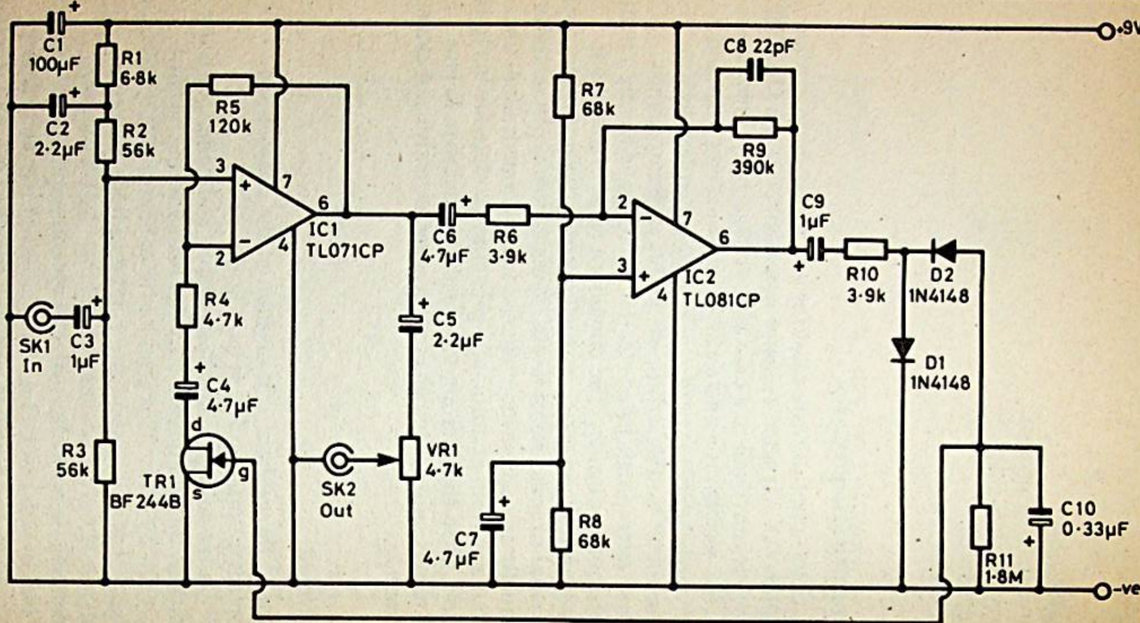

Fig. 25 The circuit diagram of the microphone compressor

a forward base bias in order to bring the device into conduction. It is also important to bear in mind that the input impedance of a JFET device is very high; being typically a few thousand Megohms, and is not just a few kilohms which would be typical for a bipolar transistor.

IC1 is used in the non-inverting mode, and its non-inverting input is biased by R1 to R3. R1 and C2 prevent noise from the supply lines from being coupled to the non-inverting input via the bias circuit. The closed loop voltage gain of IC1 at audio frequencies is determined by a negative feedback loop which has R5 as one section and the series resistance of R4 and Tr1's drain-to-source as the other section. With Tr1 switched on the closed loop gain is approximately 26dB (20 times) or so, but if Tr1 is switched off its drain-to-source resistance increases dramatically (to around a thousand Megohms). The voltage gain of the input stage would then only be about unity. Intermediate bias levels give intermediate levels of gain, of course.

Some of IC1's output is coupled by C5 to output level control VR1, and from here the signal is taken to the output socket. The remainder of IC1's output is coupled by C6 to an inverting amplifier having its voltage gain set at about 40dB (100 times) by R6 and R9. C8 gives increased negative feedback at high frequencies an thus reduces the high frequency gain of the circuit. This was found to be necessary in order to prevent the circuit from becoming slightly unstable.

The output of IC2 is coupled by C9 and R10 to a simple rectification and smoothing circuit based on D1, D2 and C10. This circuit produces a negative bias in the presence of a suitably strong input signal, and this reverse bias is applied to the gate of Tr1.

With only a low input level the bias produced is too small to have any significant effect on the circuit, but above a certain threshold level Tr1 starts to switch off and the closed loop gain of IC1 is reduced. The further the input level is raised above this threshold, the lower the closed loop gain of Tr1 becomes, until the gain of IC1 has dropped to about unity, and the circuit is then saturated. Provided the input level is kept above the threshold level, but below the saturation level, the gain of the circuit automatically adjusts itself to give a virtually constant output level.

The input threshold level is less than one millivolt RMS, and above 20mV or so is needed at the input before the circuit saturates. The output from a high impedance dynamic microphone or an electret type having a built-in step-up transformer is normally within these limits.

The attack time of the circuit is set by the values of R10 and C10. This time has been made very short so that the circuit responds almost instantly to any large rise in input level, and thus prevents overloading of the item of equipment which is fed with the output signal of the unit. The decay time is set by the values of R11 and C10, and is much longer. However, this time is still only a fraction of a second, and the unit quickly responds to any sudden fall in signal level. It is essential for the decay time to be comparitively long since too short a time constant here would result in severe distortion. Some constructors might prefer a longer decay time, and this is really a matter of personal preference, and may also depend to some extent on the exact use to which the unit will be put. In order to extend the decay time it is merely necessary to increase the value of R11.

The circuit has a current consumption of only about 4mA and a PP3 battery is a suitable power source. Any other source capable of providing 9 volts at about 4mA or so is also suitable, and it is not essential for the supply to be especially well smoothed. The maximum output level from the unit is about 20mV RMS, but this can be attenuated using VR1 to give any level below 20mV RMS.

Construction

Details of the 0.1in matrix stripboard for the Microphone Compressor unit are shown in Figure 26, and a board having 39 holes by 15 copper strips is needed. There are a number of link wires and breaks in the copper strips, and care should be taken not to omit any of these.

This is another project where the high sensitivity to stray pick up of mains hum and other electrical noise make it

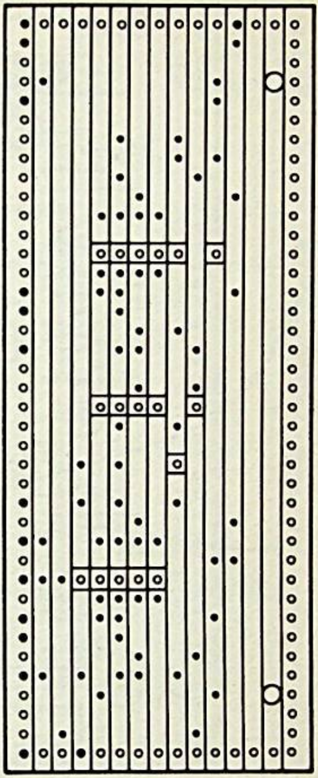

Fig. 26 Constructional details of the microphone compressor

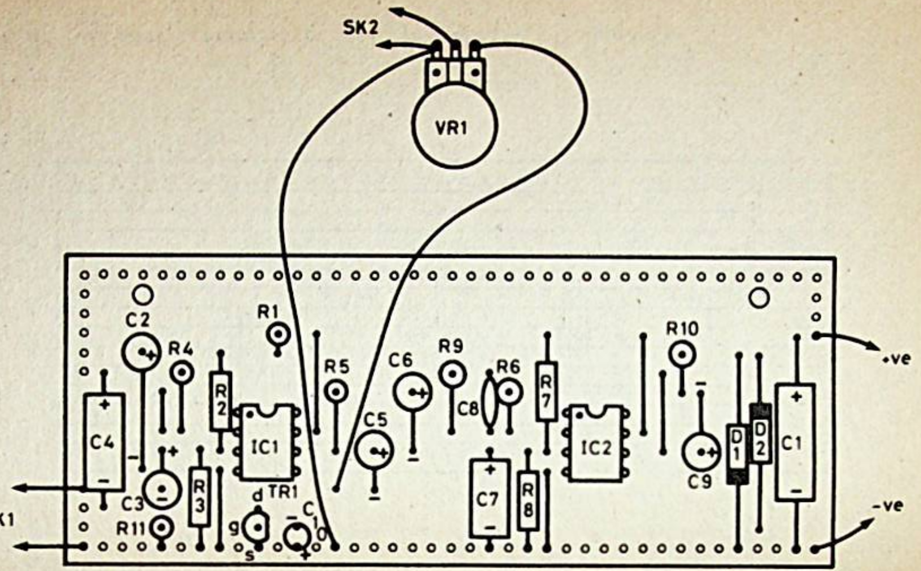

Fig. 26a Constructional details of the microphone compressor

advisable to use a metal case so that the circuitry is screened.

Components for Microphone Compressor (Figure 25)

Resistors, all 1/3 watt 5% (10% over 1M)

R1	6.8k	R2	56k
R3	56k	R4	4.7k
R5	120k	R6	3.9k
R7	68k	R8	68k
R9	390k	R10	3.9k
R11	1.8M		
VR1	4.7k log carbon		

Capacitors

C1	100μF 10V electrolytic	C2	2.2μF 63V electrolytic
C3	1μF 63V electrolytic	C4	4.7μF 63V electrolytic
C5	2.2μF 63V electrolytic	C6	4.7μF 63V electrolytic
C7	4.7μF 63V electrolytic	C8	22pF ceramic plate
C9	1μF 63V electrolytic	C10	0.33μF 35V tantalum

Semiconductors

IC1	TL071CP	IC2	TL081CP
Tr1	BF244B		
D1	1N4148	D2	1N4148

Miscellaneous

Case
0.1in matrix stripboard
Two standard jack sockets (SK1 and SK2)
Control knob
PP3 battery and connector to suit
Wire, solder, etc.

Central Image Canceller

This simple circuit is fed with a stereo signal, and by means of a phasing process cancels out the centre of the stereo image. The output is a mono signal. Circuits of this type are used to "eliminate" the soloist (who will normally appear at the centre of the sound stage) so that a record or tape can be used to provide a backing for the user.

Another use for this type of circuit is in quasi-quadraphonic equipment. Here the output of the unit (which often has a comparatively high level of ambience signals) is fed to an amplifier which drives a loudspeaker placed at the rear of the room, or is fed to a stereo amplifier which drives speakers at the rear of the room and on opposite sides. Despite the simplicity of these arrangements they can give very good results if fed with a suitable input signal.

The Circuit

The circuit must mix the two stereo channels, but they must be out-of-phase so that the signals which form the central stereo image will be cancelled out. These signals appear in-phase in both channels incidentally.

An operational amplifier is the obvious basis for a circuit of this type since the inverting and non-inverting inputs make it easy to obtain the required antiphase mixing. The circuit diagram of the Central Image Canceller appears in Figure 27.

R3 biases the non-inverting input of IC1 from a centre-tap formed on the supply lines by R1, R4 and C3. R2 and VR1 form a negative feedback network which sets the closed loop voltage gain of IC1 at unity for an input applied to SK1, with VR1 being adjusted to give this level of gain. Of course, the signal is inverted between SK1 and the output at SK3.

Signals applied to SK2 are coupled to the non-inverting input of IC1 by way of DC blocking capacitor C4 and attenuator resistor R5. Assuming that an input signal having a fairly low source impedance is fed to SK1, VR1 and R2 set the voltage gain from the non-inverting input to the output at 6dB (two times). However, R3 and R5 form a 6dB attenuator so that there is unity voltage gain from SK2 to the output at SK3. There is no signal inversion between SK2 and SK3.

In practice, due to component tolerances and the unpredictability of the source impedances of the signals fed to the inputs there might be slightly more or less than unity gain fro SK2 to SK3. This is of no great consequence since VR1 can still be adjusted to give equal sensitivity at SK1 and SK2 so that an identical signal appearing at both inputs is precisely phased out

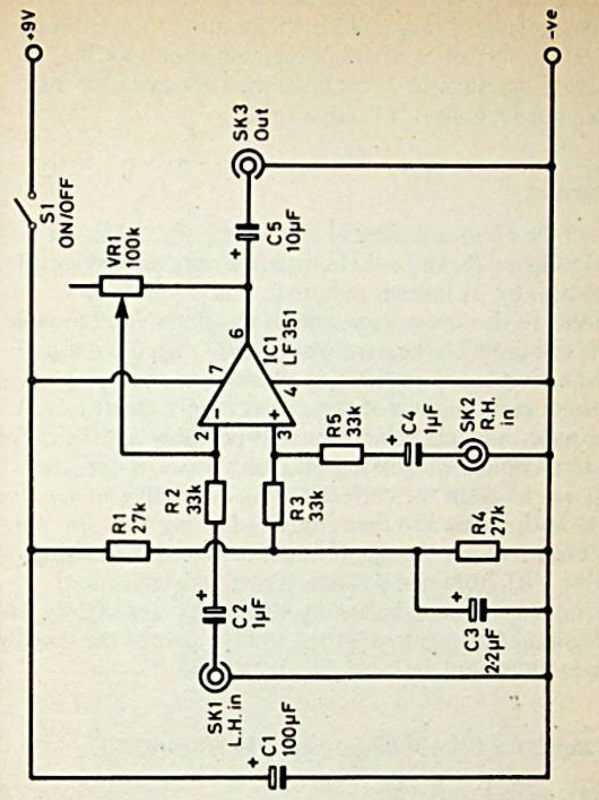

Fig. 27 The circuit diagram of the central image canceller

and does not appear at the output. In fact, even if the inputs are at slightly different levels it will still be possible to adjust VR1 to give a very high degree of cancellation.

With a simple tone applied to the inputs it is possible to obtain a very large amount of attenuation indeed, and 80dB can be easily achieved. However, with a complex input signal containing a wide range of frequencies it is not possible to obtain quite such a high level of attenuation due to slight phase shifts which occur in the circuit, and which vary slightly over the audio frequency range. With VR1 adjusted for optimum results it is nevertheless possible to obtain about 60dB of attenuation overall, and this is more than adequate for the intended applications of the circuit.

Construction

Details of the component panel and wiring of the unit are shown in Figure 28, and a 0.1in matrix stripboard having 13 copper strips by 21 holes is required. The layout is not critical due to the low voltage gain of the circuit, and because the unit will probably be used with a fairly high input signal level the circuit is not particularly sensitive to stray pick-up. The current consumption of the circuit is only about 1.5mA, and the most practical power source is probably a PP3 size 9 volt battery which will give a great many hours of operation.

VR1 can be given the correct setting by feeding an identical signal to both inputs and then simply adjusting VR1 for minimum output. In use though, it might be found that slightly offsetting VR1 from this position gives improved central cancelling since channel balancing is likely to vary slightly from one programme source to another, and the part of the signal you wish to attenuate might be slightly off-centre.

Components for Central Image Canceller (Figure 27)

Resistors, all 1/3 watt 5%

R1	27k	R2	33k
R3	33k	R4	27k
R5	33k		
VR1	100k lin carbon		

Capacitors

C1	100μF 10V electrolytic	C2	1μF 63V electrolytic
C3	2.2μF 63V electrolytic	C4	1μF 63V electrolytic
C5	10μF 25V electrolytic		

Semiconductor

IC1	LF351

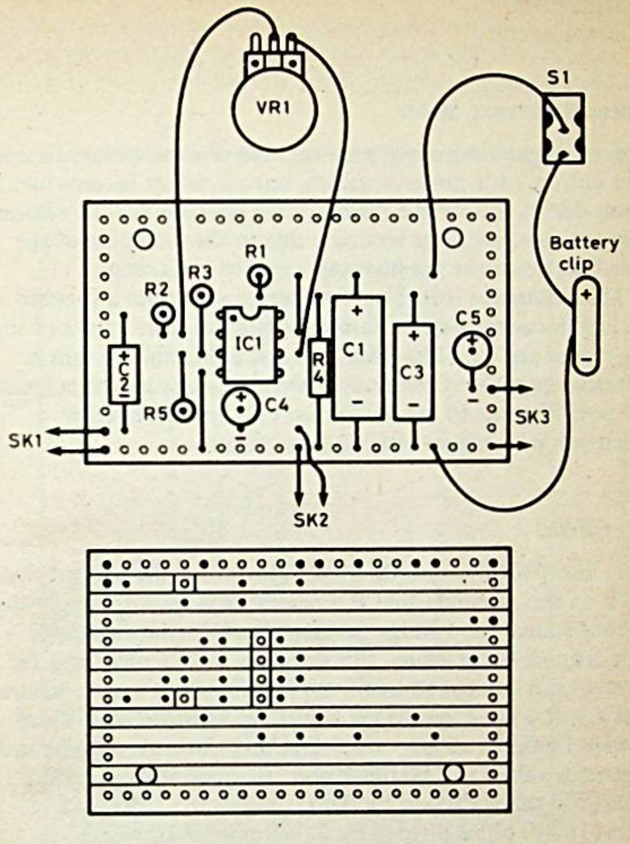

Fig. 28 Constructional details of the central image canceller

Switch
S1 SPST miniature toggle type

Miscellaneous
Case
0.1in matrix stripboard

PP3 size battery and connector to suit
Control knob
Wire, solder, etc.

Linear Resistance Meter

Most analogue multimeters are capable of measuring resistance over quite a wide range of values, but are rather inconvenient in use due to the reverse reading scale which is also non-linear. This can also give poor accuracy due to the cramping of the scale that occurs at the high value end of each range.

This resistance meter has five ranges and it has a forward reading linear scale on each range. The full scale values of the five ranges are 1k, 10k, 100k, 1M and 10M, and the unit is therefore capable of reasonably accurate measurements from a few tens of ohms to ten Megohms. Few circuits employ resistors having values outside these limits.

The Circuit

Most linear scale resistance meters, including the present design, work on the principle that if a resistance is fed from a constant current source the voltage developed across that resistance is proportional to its value. For example, if a 1k resistor is fed from a 1mA current source from Ohm's Law it can be calculated that 1 volt will be developed across the resistor (1000 ohms divided by 0.001 amps = 1 volt). Using the same current and resistance values of 100 ohms and 10k gives voltages of 0.1 volts (100 ohms divided by 0.001 amps = 0.1 volts) and 10 volts (10000 ohms divided by 0.001 amps = 10 volts).

Thus the voltage developed across the resistor is indeed proportional to its value, and a voltmeter used to measure this voltage can in fact be calibrated in resistance, and will have the desired forward reading linear scale. One slight complication is that the voltmeter must not take a significant current or this will alter the current fed to the test resistor and impair linearity. It is therefore necessary to use a high impedance voltmeter circuit.

The full circuit diagram of the Linear Resistance Meter is given in Figure 29, and the constant current generator is based on IC1a and Tr1. R1, D1 and D2 form a simple voltage regulator circuit which feeds a potential of just over 1.2 volts to the non-inverting input of IC1a. Tr1 is driven from the output of IC1a, and there is 100% negative feedback from the

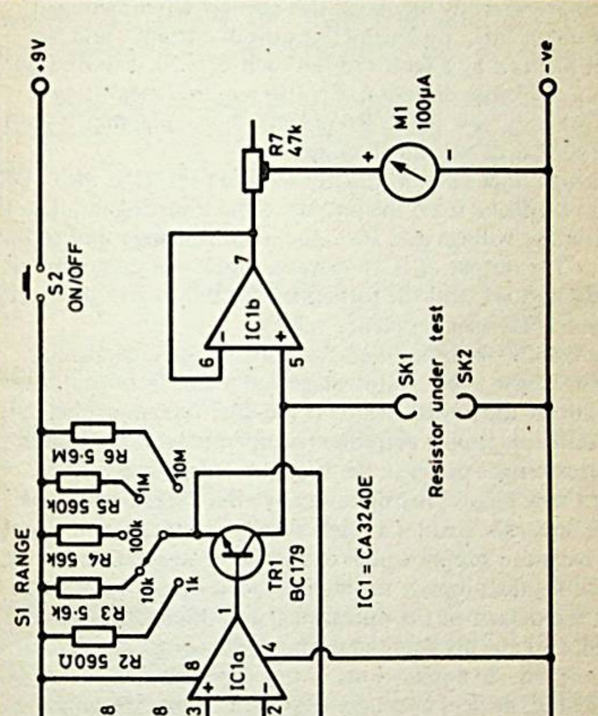

Fig.29 The circuit diagram of the linear resistance meter

emitter of Tr1 to the inverting input of IC1a so that Tr1's emitter is stabilised at the same potential as IC1a's non-inverting input. In other words it is stabilised a little over 1.2 volts below the positive supply rail potential. S1 gives five switched emitter resistances for Tr1, and therefore five switch emitter currents. As the emitter and collector currents of a high gain transistor such as the BC179 device used in the Tr1 position are virtually identical, this also gives five switched collector currents. By having five output currents, and the current reduced by a factor of ten each time S1 is moved one step in a clockwise direction, the five required measuring ranges are obtained. R2 to R6 must be close tolerance types to ensure good accuracy on all ranges.

The high impedance voltmeter section uses IC1b with 100% negative feedback from the output to the inverting input so that there is unity voltage gain from the non-inverting input to the output. The output of IC1b drives a simple voltmeter circuit using R7 and M1, and the former is adjusted to give the circuit the correct full scale resistance values.

The CA324OE device used for IC1 is a dual operational amplifier having a MOS input stage and a class A output stage. These enable the device (which is the dual version of the CA314OE operational amplifier) to operate with the inputs and outputs right down to the negative supply rail voltage. This is a very helpful feature in many circuits, including the present one, as it enables a single supply rail to be used where a dual balanced supply would otherwise be needed. In many applications the negative supply is needed simply in order to permit the output of the operational amplifier to reach the 0 volt rail, and not because the output will ever go negative of the 0 volt rail. In applications of this type using the CA314OE or CA324OE devices normally enables the negative supply to be dispensed with. Incidentally, the CMOS output stage of the CA313OT device also enables the output to swing right down to the negative supply potential.

As the CA3240E has a MOS input stage for each section the input impedance is very high (about 1.5 million Megohms!) and obviously no significant input current flows into the device. This, together with the high quality of the constant current source, and the practically non-existant distortion through IC1b due to the high feedback level, gives the circuit excellent linearity.

With no resistor connected across SK1 and SK2 M1 will be taken beyond full scale deflection and overloaded by about 100 or 200%. This is unlikely to damage the meter, but to be on the safe side a push-to-test on/off switch is used. Thus power is only applied to the circuit when a test resistor is connected to the unit, and prolonged meter overloads are thus avoided.

A small (PP3 size) 9 volt battery is a suitable power source for this project which has a current consumption of around 5mA, and does not require a stabilised supply.

Construction

Refer to Figure 30 for details of the component panel and wiring of the resistance meter. A 0.1in matrix stripboard having 17 holes by 14 copper strips is required, and construction of the board is quite straight forward apart from the fact that IC1 is a MOS device and therefore needs the normal MOS handling precautions. Use an 8 pin DIL IC socket for this device (or Soldercon pins) and do not plug it into circuit until the unit is in other respects complete. Leave IC1 in its protective packaging until it is to be fitted into place, and touch the pins of the device as little as possible.

In order to minimise the amount of wiring from S1 to the component panel R2 to R6 are mounted on S1, as shown in Figure 30. S1 is a 12 way type with an adjustable end stop (set for 5 way operation).

As far as the mechanical construction of the unit is concerned the only slightly awkward part of this is the mounting of the meter, and as this was covered in an earlier project (the Audio Millivoltmeter) it will not be described again here.

The unit is calibrated by first setting R7 at maximum resistance (fully anticlockwise), switching S1 to the "100k" range, and connecting a 100k 1% resistor across the test terminals. Switch the unit on by operating S2, and then carefully adjust R7 for precisely full scale deflection of the

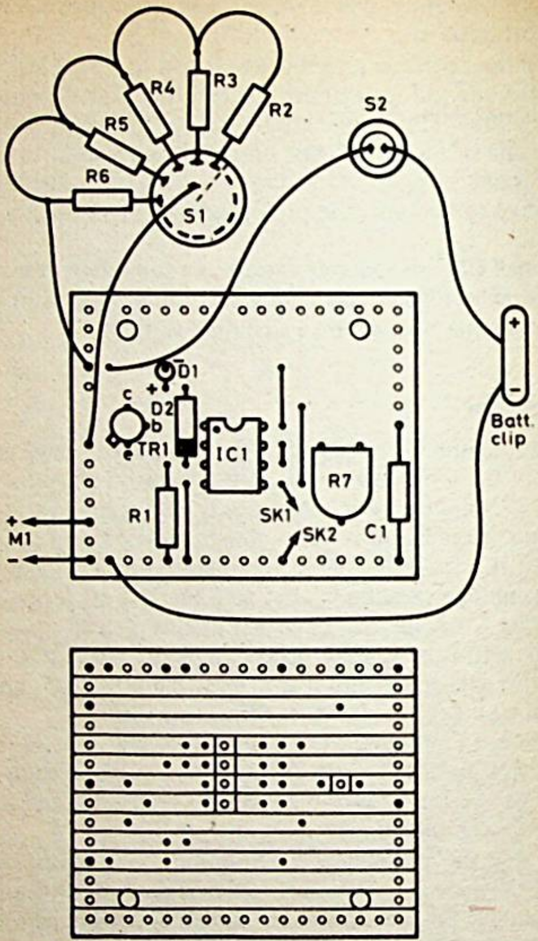

Fig. 30 Constructional details of the linear resistance meter

meter. The unit is then ready for use.

It is not essential to alter the numbering of the meter's scale as the 0 to 100 is easily translated into resistance values on any of the ranges covered by the unit.

Components for Linear Resistance Meter (Figure 29)

Resistors, ½ watt 2% or better except where noted otherwise

R1	1.8k 1/3 watt 5%	R2	560 ohms
R3	5.6k	R4	56k
R5	560k	R6	5.6M
R7	47k 0.1 watt horizontal preset		

Capacitor

C1	100nF polyester

Semiconductors

IC1	CA3240E		
Tr1	BC179		
D1	1N4148	D2	1N4148

Switches

S1	5 way 1 pole rotary (see text)	S2	Push to make – release to break type

Miscellaneous

Case
0.1in matrix stripboard
100μA moving coil meter (M1)
PP3 size battery and connector to suit
Two 1mm wander sockets (SK1 and SK2)
Control knob
Wire, solder, etc.

Scratch Filter

When playing an old or well used record results are often marred by the continuous "crackling" sound of what has become known as "surface noise". This noise can be greatly reduced by using a filter which gives a fairly rapid roll off at frequencies above a few kilohertz, since surface noise consists almost entirely of high frequency components. There is obviously some loss of the wanted signal, but subjectively results are often greatly improved.

This stereo scratch filter design is based on an LM3900N integrated circuit which contains four Norton or current differencing amplifiers. These differ from an ordinary operational amplifier in that the output responds to the relative input currents rather than the relative input voltages. Although the LM3900N has from time to time been described as a "quad 741", it is not a quad version of the 741, and Norton amplifiers cannot be directly substituted for conventional (voltage differencing) operational amplifiers.

The LM3900N is primarily intended for applications such as waveform generators, audio amplifiers, etc., where only a single supply rail is used. The amplifiers in the device give low levels of distortion and noise, and this filter has a signal to noise ratio (unweighted) of over 70dB relative to an output level of 775mV RMS (2 volts peak to peak). The –6dB point of the filter is at about 5 to 6kHz, and above this there is a nominal attenuation rate of 12dB per octave up to the limit of the audio frequency range. Above about 20kHz the attenuation of the filter remains almost constant and may actually reduce slightly, and this is caused by stray capacitance and phase shifts. However, the performance of the filter above the audio frequency range is of no practical importance, and the attenuation of about 20dB provided at high audio frequencies plus the virtually flat response at frequencies up to about 4kHz ensures good results.

The Circuit

Figure 31 shows the circuit diagram for one channel of the filter. The other channel is essentially the same, but uses the other two sections of the LM3900N of course. The pin numbers in brackets show the equivalent pin numbers for the other channel.

IC1a is used as a simple inverting buffer stage having unity voltage gain, and this ensures that the active filter circuit is fed from a low source impedance and functions correctly. The

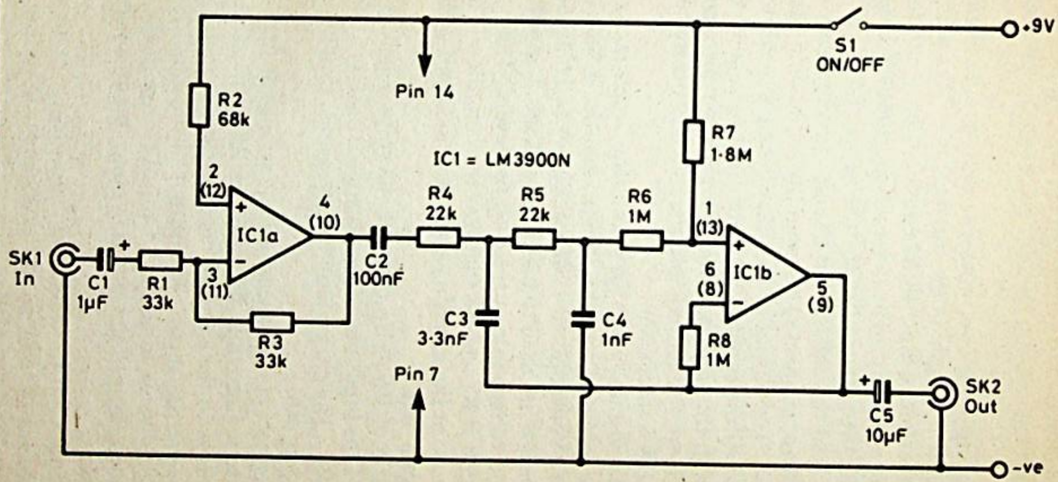

Fig. 31 The circuit diagram of the scratch filter

method of biasing a Norton amplifier and setting the input impedance and closed loop voltage gain at the required level is quite different to the system used for an ordinary operational amplifier.

In the inverting mode two resistors are used to set the input impedance and voltage gain in the usual way, and R1 plus R3 therefore set the voltage gain of the circuit at unity and the input impedance at unity. The non-inverting input is simply fed with a bias current from the positive supply rail via a single resistor, and this resistor has a value approximately double that of the resistor connected between the output and the inverting input. The output of the amplifier then needs to assume about half the supply voltage in order to balance the input currents to the amplifier (which is achieved automatically by a negative feedback action), and the required quiescent output potential is obtained.

IC1b is used as a unity voltage gain buffer amplifier having a high input impedance, and this buffer amplifier is then used as the basis of a conventional 12dB per octave active filter. Three components are used to bias IC1b and set its input impedance and voltage gain at the required levels. These are R6 to R8, and it is R7 and R8 that bias the amplifier. This is just the same as for the inverting amplifier mode and will not be covered again here.

It is the ratio of R6 to R8 that sets the voltage gain of the amplifier, and the input impedance is approximately equal to the value of R6. The closed loop voltage gain is simply equal to R8 divided by R6.

This configuration might seem a little strange at first, but is quite easy to understand. With R6 and R8 at the same value, any input voltage to the amplifier will alter the current flowing into the non-inverting input. The output of the amplifier is automatically adjusted to give an identical change in the current flowing through R8 into the inverting input due to a negative feedback action.

With R6 and R8 at the same value, any change in input voltage will be matched by an identical change in output voltage so that the input currents remain balanced, and the closed loop voltage gain of the circuit is thus unity. If R8 had a value (say) ten times higher than that of R6, then the output would change by an amount ten times greater than the input signal in order to balance the input currents, and a closed loop voltage gain of ten times would be achieved. If R8 was to be made lower in value than R6 the voltage gain of the circuit would be less than unity, and this is something that cannot be achieved using a conventional operational amplifier in the non-inverting mode.

The filter components are R4, R5, C3 and C4, and these set the –6dB point at the required frequency of approximately 5.5kHz.

The current consumption of the circuit is about 6mA and a PP3 size 9 volt battery is a suitable power source, although it would probably be better to use a larger type such as a PP9 if the unit is to be given a large amount of use. The circuit is sensitive to noise on the supply lines due to the coupling from the supply lines to the non-inverting inputs via R2 and R7. If a mains power supply is used it must either be a very low noise type, or the bias currents must be obtained via decoupling networks and not direct from the positive supply.

Construction

The Scratch Filter is constructed using a 0.1in matrix stripboard having 18 copper strips by 26 holes using the component layout and wiring scheme shown in Figure 32. Resistors and capacitors used in the right hand channel have an "a" suffix, while those that have no suffix are for the left hand channel. Since the LM3900N has four amplifiers and only two are needed per channel, only one LM3900N is, of course, sufficient for a stereo version of the unit.

Phono sockets were used at the input and output of the prototype Scratch Filter, but it is probably best to use sockets of a type that match those fitted to the equipment with which the filter will be employed. The best way to use the filter is to connect it into the "tape monitor" facility of the amplifier, or some similar arrangement that enables the filter to be connected between the preamplifier and power amplifier stages, and easily bypassed when not required. It is not a good idea to

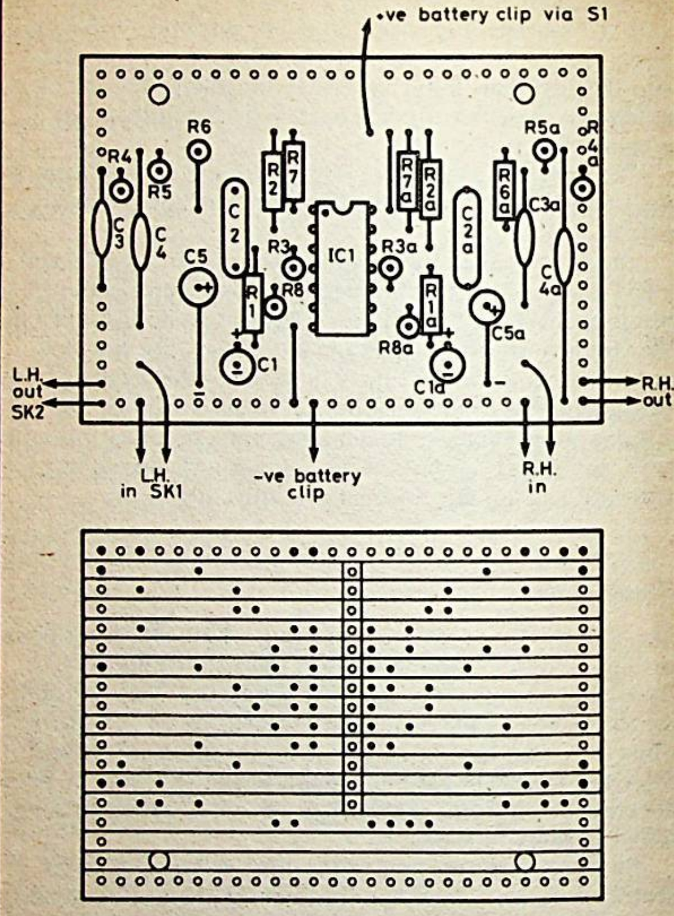

Fig. 32 Constructional details of the scratch filter

use the circuit to process a low level signal, such as the direct output from a magnetic pick-up, since this would give a comparitively poor signal to noise ratio.

Components for Scratch Filter (one channel) (Figure 31)

Resistors, all 1/3 watt 5% (10% over 1M)

R1	33k	R2	68k
R3	33k	R4	22k
R5	22k	R6	1M
R7	1.8M	R8	1M

Capacitors

C1	1µF 63V electrolytic	C2	100nF polyester
C3	3.3nF ceramic plate 10%	C4	1nF ceramic plate 10%
C5	10µF 25V electrolytic		

Semiconductor

IC1 LM3900N

Switch

S1 SPST miniature toggle type

Miscellaneous

Case
0.1in matrix stripboard panel
PP3 size battery and connector to suit
Input and output sockets (e.g. four phono sockets)
Wire, solder, etc.

Note that for a stereo unit two of each resistor and capacitor are required (one for use in each stereo channel).

Battery Monitor

This simple device can be used to monitor a 12 volt car or boat battery and give warning if the battery voltage starts to fall to an unacceptable level. The unit has four LED indicators, and these switch on if the supply voltage falls below a certain threshold level, with a different threshold voltage being used for each LED. The approximate threshold potentials are 10, 11, 12 and 13 volts, but these can easily be changed, as explained later.

The circuit utilizes a quad comparator, and strictly speaking this device not not a quad operational amplifier. The difference between a comparator and an operational amplifier is very small, and they are largely interchangeable. The device used in this circuit is the MC3302P, and this has four identical comparators which have common positive and negative supply pins. Like an operational amplifier, there are two inputs (inverting and non-inverting) plus an output for each section of the device.

The comparators only really differ from an operational amplifier in that the output terminal connects to the open collector of a common emitter (NPN) output transistor. In normal use the output transistor is used to supply current to a load of some kind if the inverting input is at a higher voltage than the non-inverting one, and cut off power to the load if the comparative input states are reversed. Of course, the fact that an operational amplifier does not have an open collector output does not preclude its use as a voltage comparator, and if a comparator is given a discrete output load of some kind (a resistor of a few kilohms in value is sufficient) it functions as an operational amplifier.

The Circuit

Figure 33 shows the full circuit diagram of the Battery Monitor, and this consists of four virtually identical stages. The only difference between the stages is the voltage of the zener diode used in each, and this voltage is chosen to give the desired threshold voltage.

If we consider the stage which utilizes IC1a, the load for the output transistor of IC1a is LED indicator D2 and its series current limiting resistor R4. There is no form of output current limiting built into the output stage of each comparator, and discrete components are needed to ensure that the maximum permissible output current of 20mA is not exceeded.

The inverting input of IC1a is fed from the supply lines via the potential divider formed by R1 and R2. Obviously the

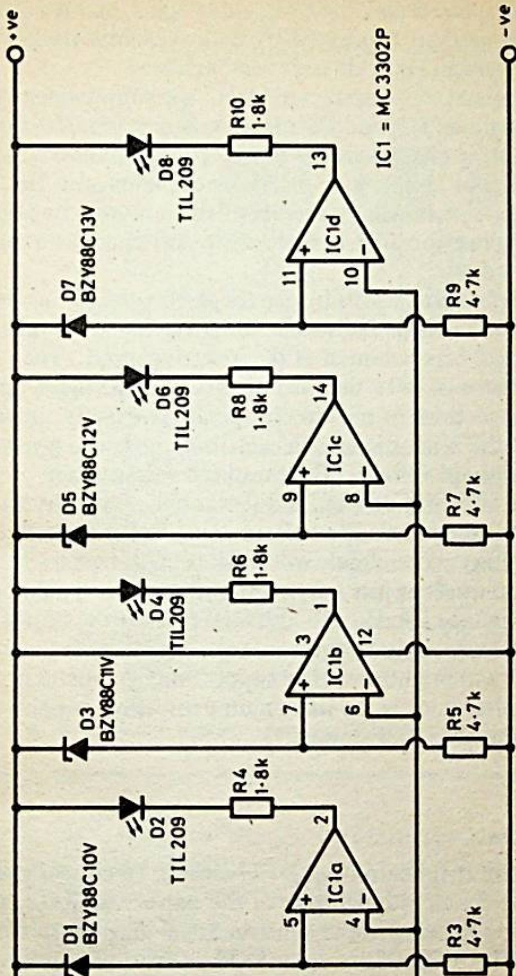

Fig. 33 The circuit diagram of the battery monitor

voltage fed to the inverting input will vary with changes in the supply voltage, but it will always be quite small, and is unlikely to become more than 100mV (0.1 volts). This bias voltage is fed to the inverting input of each comparator.

IC1a's non-inverting input is fed from the supply lines by way of zener diode D1, and D1's load resistor is R3. Under normal operating conditions the supply voltage should be a couple of volts or more above the avalanche voltage of D1, and about two volts or so will be present at the non-inverting input. The output transistor of IC1a is therefore switched off and D2 is not switched on.

If the supply voltage falls below about 10 volts, D1 ceases to conduct and the voltage fed to the non-inverting input falls below the small bias potential at the inverting input. The output transistor of IC1a then switches on and D2 lights up.

The other sections of the circuit operate in exactly the same manner, but the zener diode is in each case chosen to give a different threshold voltage. The threshold voltages can obviously be altered if desired, and it is simply necessary to choose a zener diode having a voltage equal to the desired threshold voltage. The circuit will work using zeners having operating potentials of just a few volts, although it would be advisable to reduce R4, R6, R8 and R10 to 1k if the unit is to be used with low supply voltages in order to give a reasonable LED current and brightness. The supply voltage must not exceed 28 volts which is the maximum permissible supply voltage for the MC3302P device.

Construction

A 0.1in matrix stripboard having 19 holes by 14 copper strips is used as the constructional basis of the Battery Monitor, as can be seen by refering to the constructional diagram shown in Figure 34. The MC3302P is not a MOS device and requires no special handling precautions. In fact electrical construction of the unit is perfectly straight forward and should not give the constructor any real problems.

Mechanical construction must obviously be varied to suit the circumstances under which the unit will be used. This is

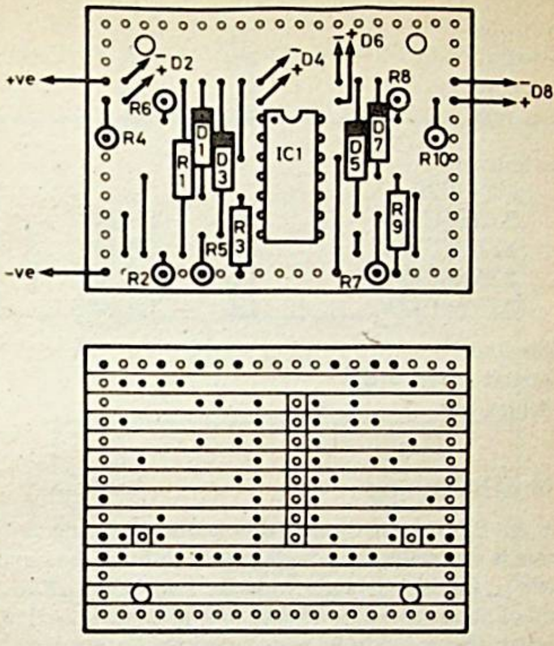

Fig. 34 Constructional details of the battery monitor

really just a matter of using a little initiative, and is again something that should not give any real difficulties.

The current consumption of the unit with all the LEDs switched off is only about 1.5 to 2mA, but this obviously rises considerably when one or more of the LEDs are switched on. In fact the increase is about 6mA per LED.

Components for Battery Monitor (Figure 33)

Resistors, all 1/3 watt 5%

R1	15k	R2	100 ohms
R3	4.7k	R4	1.8k
R5	4.7k	R6	1.8k
R7	4.7k	R8	1.8k
R9	4.7k	R10	1.8k

Semiconductors

IC1	MC3302P		
D1	BZY88C10V	D2	TIL209
D3	BZY88C11V	D4	TIL209
D5	BZY88C12V	D6	TIL209
D7	BZY88C13V	D8	TIL209

Miscellaneous

0.1in matrix stripboard
Wire, solder, etc.

Auto Swell Pedal

This circuit is a volume control that is operated via a switch, and there is normally unity voltage gain through the circuit. If, however, the switch is placed in the "down" position, over a period of about two seconds the unit provides a level of attenuation that steadily increases until the output signal is completely cut off. By returning the switch to the "up" position the output signal is "faded" back up to its original level over a period of about two seconds.

The circuit is probably most useful as a foot pedal which enables an instrument to be "faded out" or "faded in" without the need to operate any controls by hand. Circuits of this type can also be used in mixers to enable a signal to be easily brought up and down in level.

The Circuit

This circuit is based on a CA3080E operational transconductance amplifier, and this type of device is substantially different to a conventional operational amplifier. Rather than the differential input voltage giving an output voltage swing of a certain level, the differential input voltage gives a certain output current from the device. What makes this type of operational amplifier so useful is the addition of an amplifier bias input, and the current fed to this input controls the conductance (gain) of the device.

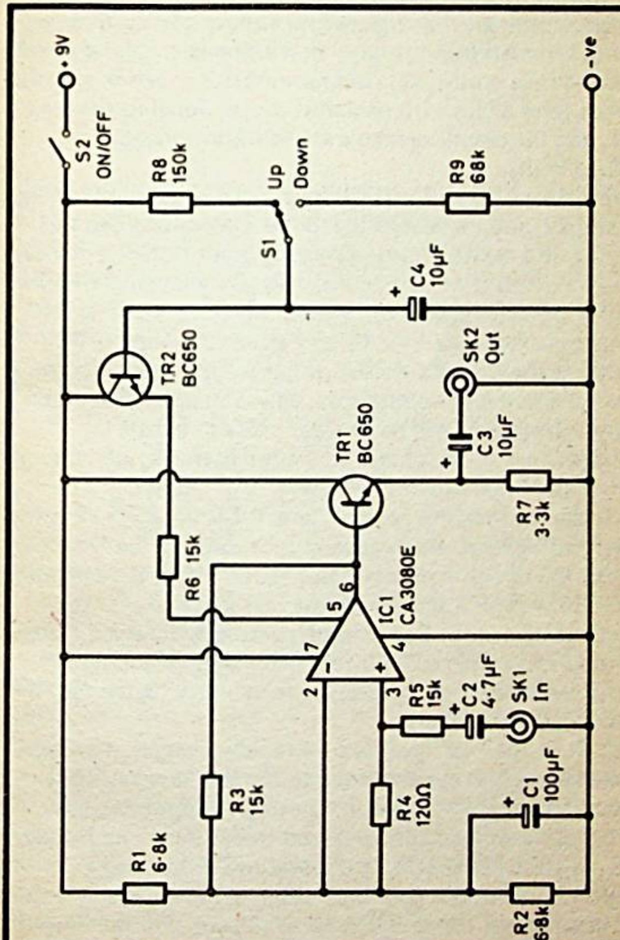

Fig. 35 The circuit diagram of the auto swell pedal

If a resistor is connected across the output of the amplifier, the differential input voltage will produce a certain current through this resistor, and a certain voltage will therefore be developed across the load resistor. In other words the device acts as a simple voltage amplifier. If a small bias current is applied to the bias input, the output current and voltage for a given differential input voltage will be quite small. Using a large bias current gives a large output voltage and current for a given differential input voltage, and a simple current controlled amplifier is thus produced. Adding a resistor in series with the bias input gives a bias current that is proportional to the bias voltage, and the circuit operates as a voltage controlled amplifier (VCA).

In this circuit R1, R2 and C1 form a centre-tap on the single 9 volt supply, and the inverting input is biased direct to this centre-tap. R4 biases the non-inverting input to the centre-tap, and a low value is used here in order to give a low noise level. The input signal is applied to the non-inverting input via DC blocking capacitor capacitor C2 and attenuator resistor R5. The latter is needed because the circuit would otherwise give a substantial amount of voltage gain, and in practice it is unlikely that any voltage gain will be needed. R5 also boosts the input impedance of the circuit to a more useful figure of about 15k (it would otherwise only be about 120 ohms!).

R3 is the load resistor for IC1, and Tr1 is used as an emitter follower buffer stage which ensures that the load fed from the output of the circuit does not significantly affect the operation of IC1. R6 is used in series with the bias input so that the required voltage controlled amplifier action is obtained, and Tr2 is used as an emitter follower buffer stage which boosts the input impedance to the bias input to a very high level (several Megohms in fact).

With S1 in the "up" position C4 rapidly charges at switch-on and a strong bias current is fed to the bias input of IC1. The circuit values have to be chosen to give approximately unity voltage gain through the circuit under these conditions.

If S1 is placed in the "down" position C4 discharges through R9, and over a period of a couple of seconds or so the bias current fed to IC1 reduces to zero, and the output signal is steadily reduced to zero as well. If S1 is returned to the "up" position the bias current fed to IC1 is gradually returned to its original level as C4 charges up, and so the output signal is also restored to its original level.

The current consumption of the circuit is only about 3mA, and a small (PP3 size) 9 volt battery is probably the most practical power source for the unit. Of course, a small mains power supply unit could be used if preferred, and the circuit is not sensitive to noise on the supply lines.

Construction

The 0.1in matrix stripboard component layout, plus the wiring etc. of the unit, are shown in Figure 36. A stripboard having 14 copper strips by 26 holes is needed. The CA3080E is a bipolar device and not a MOSFET type incidentally. Construction of the electronics is quite straight forward and the normal techniques are used. Standard (¼in or 6.35mm) jack sockets are used at the input and output of the prototype, but any preferred type of audio connector can be used.

If the unit is constructed as a pedal unit the circuit can be housed in a case having a built-in pedal and switch (which is used in the S1 position). Alternatively an ordinary case can be used and S1 can be a heavy duty push button type which must be of the latching type or it will be necessary to continuously operate the switch in order to keep the unit in the "up" or "down" position (depending on which way round the switch is connected).

Of course, if the unit is built into a mixer or other piece of equipment, or it is built as a add-on unit for a piece of equipment such as this, S1 can be an ordinary miniature toggle type.

The input signal should be no more than about 5 volts peak to peak or clipping and a serious degradation of quality will occur. The "fade up" and "fade down" times of the unit are those which the author considered to be best, but both times can easily be modified. The "fade up" time is proportional to the value of R8, and the "fade down" time is proportional to the value of R9.

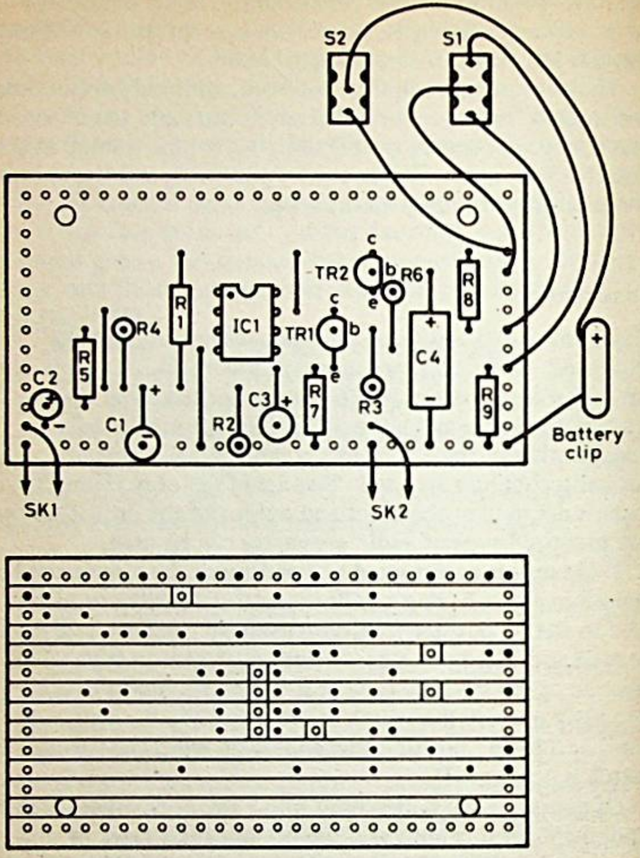

Fig. 36 Constructional details of the auto swell pedal

If only a fairly small amount of attenuation is needed at maximum "fade down", the track of a 10k potentiometer can be wired across the supply lines, and the lower end of R9 is then connected to the wiper terminal of this potentiometer instead of being connected to the negative supply rail. The maximum attenuation provided by the unit can then be adjusted from its normal high level right down to zero attenuation by means of this additional control.

Components for Auto Swell Pedal (Figure 35)

Resistors, all 1/3 watt 5%

R1	6.8k	R2	6.8k
R3	15k	R4	120 ohms
R5	15k	R6	15k
R7	3.3k	R8	150k
R9	68k		

Capacitors

C1	100μF 10V electrolytic	C2	4.7μF 63V electrolytic
C3	10μF 25V electrolytic	C4	10μF 25V electrolytic

Semiconductors

IC1	CA3080E		
Tr1	BC650	Tr2	BC650

Switches

S1	See text	S2	SPST miniature toggle type

Miscellaneous

Case
0.1in matrix stripboard
PP3 battery and connector to suit
Two standard jack sockets (SK1 and SK2)
Wire, solder, etc.

Volume Expander

Many types of music, particularly various types of classical music, have a very wide dynamic range. The dynamic range is simply the difference between the maximum and minimum sound levels, incidentally, and can be over 70dB (although in most cases it is not quite this high).

This usually results in a certain amount of compression being used in one form or another when music having a high dynamic range is recorded or transmitted, especially in the case of a cassette recording where the available dynamic range of the recording system is likely to be very limited. Without some form of compression to restrict the dynamic range of the signal either loud passages of music overload the recording or transmission medium, or quite passages fall below the noise level (or a combination of the two).

Music that has undergone even a fairly modest amount of carefully applied compression can be rather lacking in impact, and an improvement can often be obtained by using a certain amount of expansion during playback or reception in order to boost the effective dynamic range of the signal. In order to give the desired effect an expander is required that will not affect low level signals, but produces a small boost in volume at middle dynamic levels, steadily rising to an increase of about 10 or 12dB at the highest dynamic levels. In this way the lowest volume levels (and the background noise) remain unaltered, but the highest volume levels are substantially boosted. Of course, the noise is also boosted, but this is not noticeable as the wanted signal is strong enough to mask the noise.

Thus the use of a volume expander gives an apparent increase in signal to noise ratio, and can give more impact to music which requires a wide dynamic range. A volume expander cannot be expected to precisely counteract any compression used during the recording process or during transmission, but it can nevertheless produce a substantial subjective improvement when used with many programme sources.

The Circuit

An operational transconductance amplifier makes a good basis for a volume expander since it is essential that the expansion is applied gradually, and is not virtually switched on and off. A voltage controlled amplifier using an operational transconductance amplifier tends to naturally give a suitable control characteristic.

The circuit diagram of the Volume Expander is shown in Figure 37, and the operational transconductance amplifier used

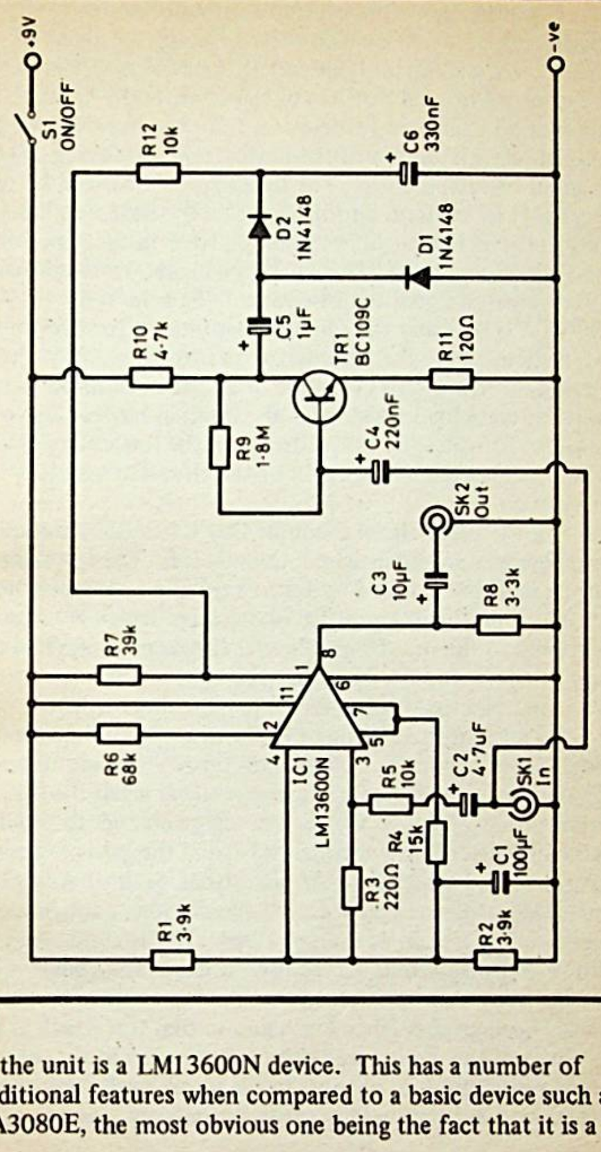

Fig. 37 The circuit diagram of the volume expander

in the unit is a LM13600N device. This has a number of additional features when compared to a basic device such as the CA3080E, the most obvious one being the fact that it is a dual device. However, it is economically competitive with alternative forms of VCA which offer a comparable level of performance if, as here, only one section of the device is used.

The other additional features of the LM13600N are the inclusion of a Darlington Pair emitter follower stage for use at the output of each section of the device, and linearizing diodes at the input of each section. The linearizing diodes can be fed with a small bias current, and this enables the device to handle a significantly higher signal level (about 10dB in fact) before serious distortion occurs. This enables a higher overload margin and (or) increased signal to noise ratio to be achieved.

The VCA is basically the same as that used in the previous circuit, although here the amplifier is an inverting type rather than a non-inverting type (which is of no practical importance). R8 is the discrete load resistor for the internal buffer stage of IC1, and R6 provides the bias current for the linearizing diodes. R7 sets the quiescent voltage gain of the circuit at a little under unity.

Some of the input signal is coupled by C4 to the input of a common emitter amplifier which utilizes Tr1. The amplified output from Tr1 is coupled by C5 to a rectifier and smoothing circuit which produces a positive bias voltage which is roughly proportional to the input signal level. This signal is applied to the amplifier bias input of IC1 via R12.

If only a low input signal level is present there will be little or no current flow through R12 and into the amplifier bias input of IC1, and the voltage gain through the circuit will not be significantly changed. At higher signal levels the bias potential produced across C6 becomes large enough to produce a significant current flow through R12, and the gain of the circuit is boosted somewhat. At high signal levels of around 500mV RMS or so the bias voltage becomes high enough to produce a boost in gain of around 12dB. This is about the maximum expansion that can be used without the signal processing becoming obvious.

C6 has been given a fairly low value so that the attack and decay times of the circuit are both fairly short, and the unit responds very rapidly to changes in the input level. However, the attack and decay times are made sufficiently long to prevent significant distortion from being produced.

The current consumption of the circuit is only about 6mA, and a small 9 volt battery such as a PP3 makes an economic power source.

Construction

The 0.1in matrix stripboard panel for the Volume Expander is detailed in Figure 38, and a board having 19 copper strips by 27 holes is required. SK1 and SK2 are phono types on the protype, but the unit will probably be most easily wired into a system if these are types which match those used on the equipment with which the expander is employed.

If a stereo expander is required it will be necessary to make up two boards, one to process each stereo channel. Alternatively, a longer piece of stripboard could be used, and the otherwise unused section of IC1 could then be used as the basis of the additional expander circuit. The supply centre tap provided by R1, R2 and C1 could be used for the additional expander circuit, but all other components (apart from IC1 of course) would need to be duplicated in the extra expander circuit. The semiconductor leadout and pinout diagram at the end of this book gives details of the pin connections to the second section of the LM13600N device.

The circuit should give good results if fed with the output of a tuner, cassette or tape deck, or a record deck via a suitable preamplifier. Some items of equipment might be found to have too little output to drive the circuit properly, and it will then be necessary to reduce the value of R11, or to even replace it with a shorting link. It is just possible that some signal sources will provide too strong a signal so that the expansion commences at a fairly low level, and in such cases R11 can be increased in value. If preferred, R11 could be replaced by a 1k linear potentiometer which would be adjusted for (subjectively) the best results.

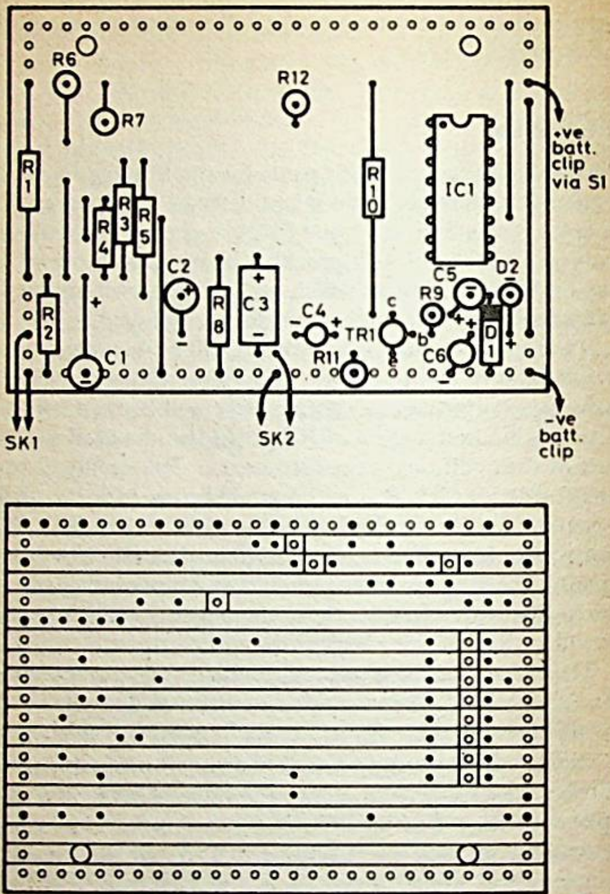

Fig. 38 Constructional details of the volume expander

Components for Volume Expander (Figure 37)

Resistors, all 1/3 watt 5% (10% over 1M)

R1	3.9k	R2	3.9k
R3	220 ohms	R4	15k
R5	10k	R6	68k
R7	39k	R8	3.3k
R9	1.8M	R10	4.7k
R11	120 ohms	R12	10k

Capacitors

C1	100μF 10V electrolytic	C2	4.7μF 63V electrolytic
C3	10μF 25V electrolytic	C4	220nF 35V tantalum
C5	1μF 63V electrolytic	C6	330nF 35V tantalum

Semiconductors

IC1	LM13600N		
Tr1	BC109C		
D1	1N4148	D2	1N4148

Switch

S1 SPST miniature toggle type

Miscellaneous

Case
0.1in matrix stripboard
Two phono sockets (SK1 and SK2)
PP3 battery and connector to suit
Wire, solder, etc.

CW Filter

Although the IF filtering of short wave receivers has tended to improve somewhat over recent years, few receivers have a very narrow bandwidth available, and this results in more adjacent channel interference during CW (Morse) reception than is really necessary. A simple way of obtaining improved results with a receiver that does not have a built-in CW filter is to use an add-on audio filter having a narrow bandwidth. If the receiver has a good filter for SSB reception, using an audio filter plus the SSB filter during CW reception should in fact provide excellent results.

The CW filter described here is simply connected between the headphone or loudspeaker socket of the receiver, and either the headphones or an external loudspeaker (having an impedance of 8 ohms or more). As the circuit provides unity gain at pass frequencies and a low impedance output, there should be no problems with a mismatching when the filter is in use.

The frequency response of the circuit peaks at approximately 800 Hertz, and the –6dB bandwidth is about 300 Hertz or so. The –20dB points occur at about 350 Hertz and 2kHz. This is sufficient to normally give a substantial reduction in adjacent channel interference, but the response is not so narrow and peaky that using the receiver with the filter in circuit becomes difficult, with the wanted signal tending to drift out of the passband and become lost.

The Circuit

The complete circuit diagram of the CW filter is shown in Figure 39.

R1 and R2 form a simple attenuator which is needed to counteract the gain of the circuitry that follows, and thus give unity voltage gain overall. The rest of the circuit is a conventional operational amplifier bandpass filter having the circuit values chosen for the desired centre frequency and a high Q value (to give the required fairly sharp, peaky response).

An ordinary operational amplifier would not give good results in this circuit due to the limited current drive capability of typically just a few milliamps. A few high current operational amplifiers are available, but even these can usually only provide a few tens of milliamps, and cannot provide a very high output power. This circuit is based on a TDA2006 device which is primarily intended for use as an audio power amplifier having an output power of up to about 10 watts, and the output overload protection circuitry within the device limits the output current to about 3 amps!

Although intended for use as an audio power amplifier, the TDA2006 is a very versatile device because it can really be regarded as an operational amplifier having a high power class

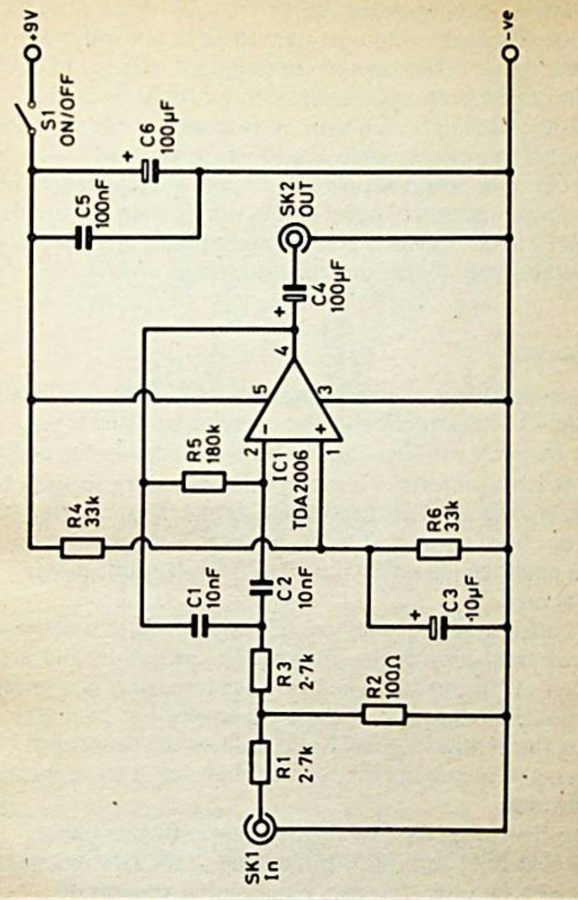

Fig. 39 The circuit diagram of the C.W. filter

B output stage. The TDA2006 has inverting and non-inverting inputs with a high input impedance (5 Megohms) at each of these, plus a high open loop voltage gain over the audio frequency range. Unlike many audio ICs which are basically operational amplifiers, the TDA2006 does not have any internal biasing components, and it can therefore be used, as it is here, in standard operational amplifier circuit configurations.

The lowest recommended supply voltage for the TDA2006 is 12 volts, but it seems to operate well using a 9 volt battery as the power source. The current consumption is around 20mA or so, and this can rise substantially if the circuit is used at high volume with an 8 ohm load. It is therefore advisable to use a fairly large battery such as a PP9 type to power the circuit. A mains power supply can be used if preferred and the circuit is not sensitive to noise on the supply lines. A supply voltage of 12 or 15 volts is perfectly acceptable provided C6 has a suitably high maximum voltage rating.

Construction

The component layout and wiring of the CW Filter are illustrated in Figure 40. Construction of the component panel is very straight forward, and there are no breaks needed in the copper strips. IC1 has preformed leads and it is necessary to splay the leadouts slightly in order to fit them onto a 0.1in matrix. IC1 should not need a heatsink if the circuit is battery powered, but this might be necessary if a 12 or 15 volt mains power supply is used.

If an operating frequency other than 800 Hertz is desired, the centre frequency of the filter can be changed by altering the values of C1 and C2. The operating frequency is inversely proportional to the value of these capacitors, but note that reducing the centre frequency also reduces the bandwidth of the filter, and increasing the centre frequency gives an increase in bandwidth.

When first using the filter it may appear to give a large reduction in gain, and this is simply due to the fact that at most frequencies the filter does give a substantial amount of attenuation. The CW note must be fairly accurately tuned to the centre of the filter's response in order to obtain a strong output, and after a little experience with the unit this will be found to be quite easy.

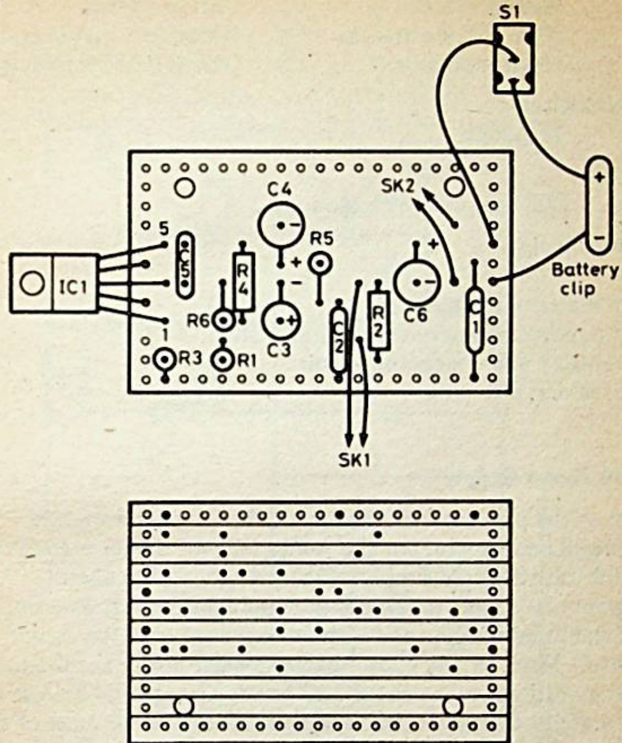

Fig. 40 Constructional details of the C.W. filter

Components for CW Filter (Figure 39)

Resistors, all 1/3 watt 5%

R1	2.7k	R2	100 ohms
R3	2.7k	R4	33k
R5	180k	R6	33k

Capacitors

C1	10nF polyester	C2	10nF polyester
C3	10μF 25V electrolytic	C4	100μF 10V electrolytic
C5	100nF polyester	C6	100μF 10V electrolytic

Semiconductor

IC1 TDA2006

Switch

S1 SPST miniature toggle type

Miscellaneous

Case
0.1in matrix stripboard
Two standard jack sockets (SK1 and SK2)
PP9 battery and connector to suit
Wire, solder, etc.

Mains Power Supply

Most of the projects described in this book require a single, non-regulated supply, but the Audio Millivolt Meter requires a dual 9 volt stabilised supply. Figure 41 shows the circuit diagram of a simple mains power supply unit which gives dual 9 volt stabilised outputs and is suitable for use with the Audio Millivolt Meter or any circuit having similar power requirements. With modification the circuit can be used to provide a single 9 volt stabilised supply suitable as a power source for most of the other projects in this book.

The circuit is quite conventional with the mains supply being connected to the primary winding of step-down and isolation transformer T1 by way of on/off switch S1. The output of the centre tapped secondary winding of T1 is fed to two separate push-pull rectifier and smoothing circuits with a positive supply being developed across C1, and a negative supply being produced across C2. Both supplies are quite well smoothed but vary in potential from about 18 volts under no load to 12 volts under full load.

A simple regulator circuit is therefore used to reduce each output voltage to the appropriate level, and provide stabilisation. R1, D5 and R3 form a simple zener shunt

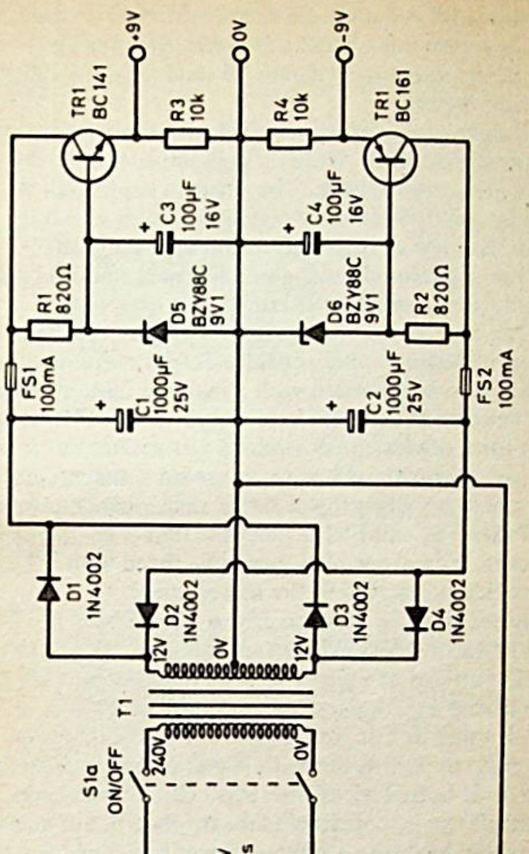

Fig. 41 The circuit of a simple 9V-0V-9V mains power supply

regulator which gives a very well smoothed and stabilised output potential of nominally 9.1 volts. This is fed to the base of emitter follower buffer transistor Tr1, and this gives an output voltage of about 8.5 volts at its emitter terminal (due to the voltage drop of about 0.6 volts across the base –

emitter terminals of Tr1). This is obviously a little below the required level of 9 volts, but this error would not normally be of any practical importance. If preferred, an output of 9.4 volts can be obtained by changing D5 (and D6) to a BZY88C10V 10 volt zener diode.

R3 is simply a load resistor for Tr1, and ensures that the circuit operates correctly if only a high impedance load is connected across the output. The negative supply rail is stabilised by a complementary regulator circuit which is essentially the same as that used to stabilise the positive supply. The regulator circuits give additional smoothing and there is only a very small noise content on each output of the supply. The emitter follower buffer stages give the circuit a low output impedance which enables output currents of up to about 100mA to be provided with a minimal drop in the output voltage. The circuit does not have output current limiting or any other form of electronic overload protection, but it is unlikely that overloads will occur in use since the circuit is designed to be part of a project rather than a bench power supply. Fuses FS1 and FS2 protect the unit if an overload should occur. The mains plug should be fitted with a 3 amp fuse to provide protection in the mains wiring.

For output currents of up to 50mA T1 can be a 12–0–12 volt type having a secondary current rating of 100mA or more. For output currents of up to 100mA T1 should be a 12–0–12 volt type having a secondary current rating of 200mA or 250mA. Tr1 and Tr2 do not require any heatsinking for currents of up to 50mA, but with higher output currents it is advisable to fit both devices with small clip-on heat sinks.

The circuit can be constructed on stripboard, but a neater solution would be to use a printed circuit board which would accommodate all the components with the only exception of the on/off switch. The 0V supply rail must be connected to the mains earth for safety reasons, and the metal case (or any exposed metal-work if a non-metalic type is used) must also be earthed for safety reasons. The circuit must be housed in a case having a screw-on lid or cover so that dangerous mains wiring cannot be exposed simply by unclipping the lid or a cover. It is also a good idea to insulate any exposed mains connections so that there is no risk of anyone receiving an electric shock from the unit when the circuitry is exposed.

Components for 9V–0V–9V Mains Power Supply (Figure 41)

Resistors, all 1/3 watt 5%

R1	820 ohms	R2	820 ohms
R3	10k	R4	10k

Capacitors

C1	1000μF 25V electrolytic
C2	1000μF 25V electrolytic
C3	100μF 16V electrolytic
C4	100μF 16V electrolytic

Semiconductors

Tr1	BC141	Tr2	BC161
D1	1N4002	D2	1N4002
D3	1N4002	D4	1N4002
D5	BZY88C9V1	D6	BZY88C9V1

Miscellaneous

Case

PCB or stripboard

T1	240V/12V–0V–12V Transformer (see text for current rating)		
FS1	100mA Fuse	FS2	100mA Fuse
S1	Mains switch		

Wire, solder, sockets, etc.

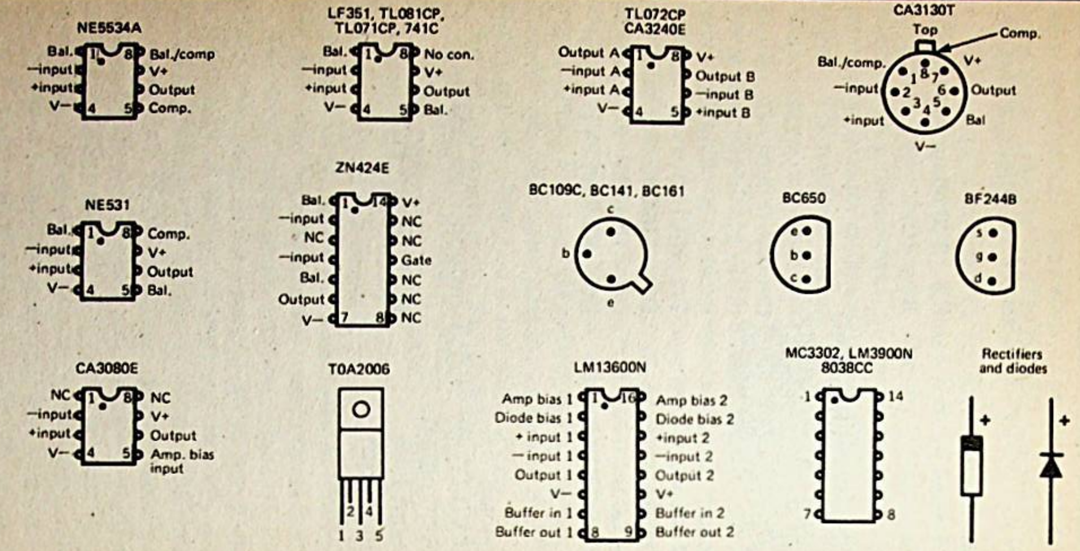

Fig. 42 I.C. pinouts (top views) and transistor leadouts (base views)

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