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ELECTRONICS DIGEST

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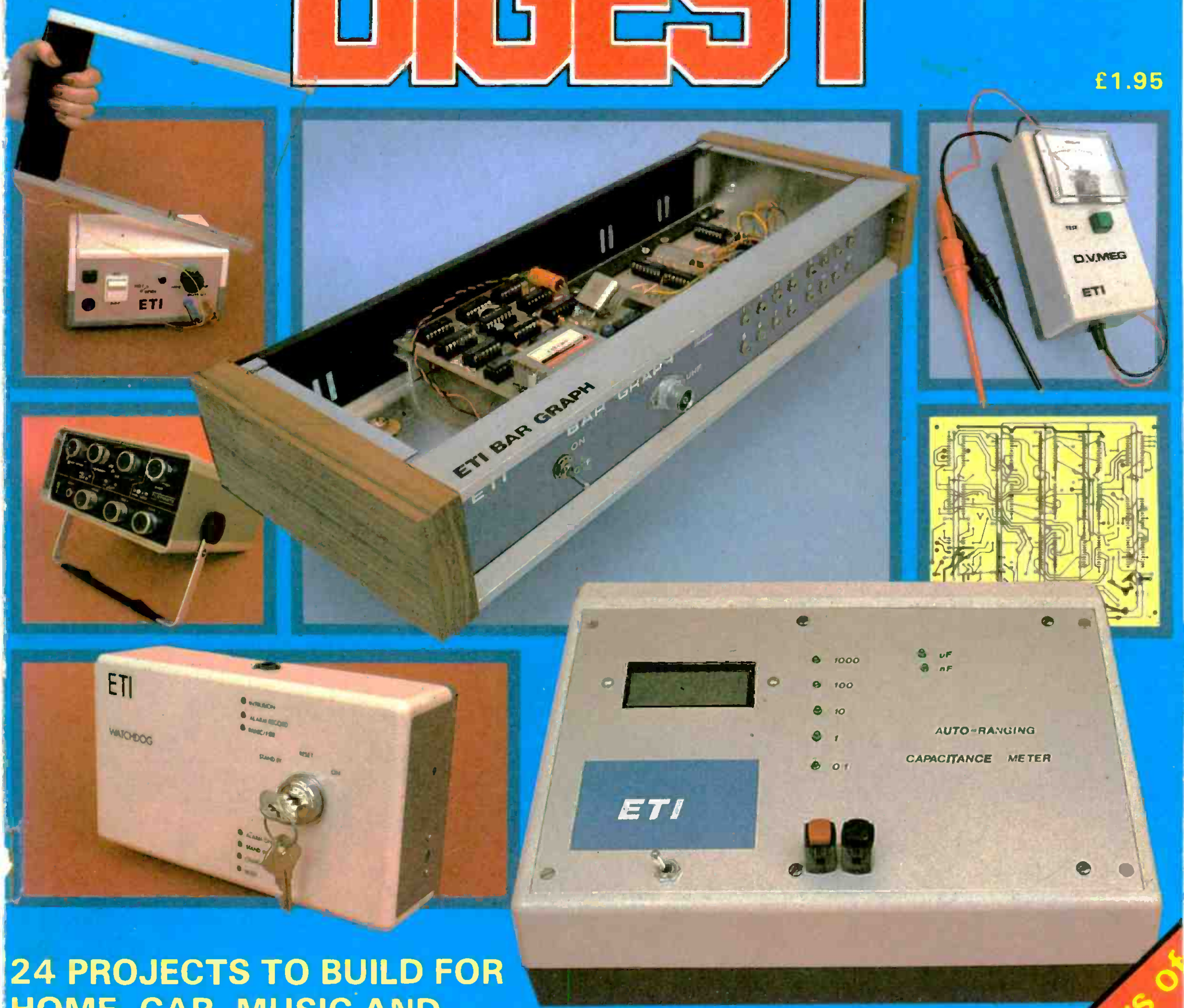

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ELECTRONICS DIGEST

Volume 4 No. 3

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INTRODUCTION

This issue of Electronics Digest has been compiled from projects published in Electronics Today International. We have put together a variety of projects including some of a serious nature such as the Watchdog Home Security System and the Car Alarm and some that are more geared towards the lighter side of electronics.

We also provide a PCB Service to make your life a little easier. We hope that you have fun constructing all these projects — and that electronics remains a source of pleasure for you for many years to come.

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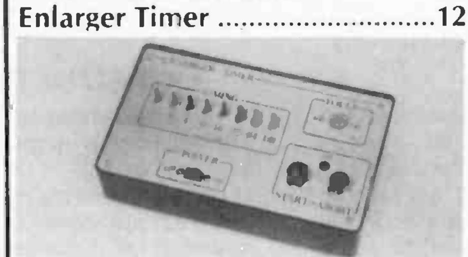
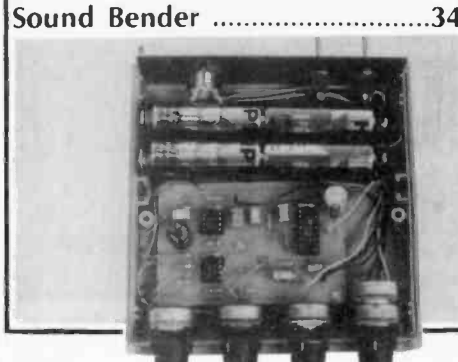
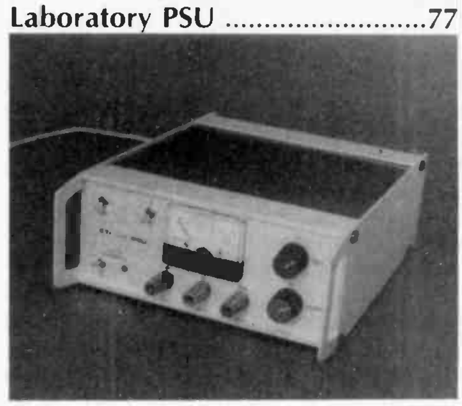






HOME SECURITY SYSTEM

Have you seen the price of meat these days? It's much too expensive to keep a watchdog. Our ever-vigilant electronic version will protect your valuables without creating ruinous butcher's bills. Design by Ray Marston. Development by Steve Ramsahadeo.

The ETI Watchdog acts as the heart of a top-quality home security system. When it is coupled to suitable input sensors (window and door-mounted security switches, pressure mats, thermostats, panic buttons, and so on), and to an output sound generator (alarm bell or siren), it gives excellent protection against burglars and front-door thugs, as well as acting as an automatic fire alarm unit.

The Watchdog is powered by a PP9-style Ni-Cd battery, which is intended to be permanently trickle-charged by a simple mains-powered charger unit. The Watchdog will operate for about two weeks with mains power removed. Particular care has been taken in the design of the entire home-security project to ensure that it has excellent reliability, with a high degree of immunity against false alarms induced by lightning strikes and radio interference. The system incorporates features such as timer-controlled auto-turn-off of the main (external) alarm, a built-in audio-visual alarm recorder, and automatic exit and entry delays on the main burglar alarm system.

Operating Modes

The Watchdog has three basic operating modes selected by a Yale-type keyswitch, these being 'standby', 'reset', and 'on'. When the unit is in the standby mode the basic burglar alarm circuitry is disabled but the fire alarm and panic facilities are fully armed: if a fire is detected by one or more of the thermostats, or if one or more of the panic buttons are momentarily closed, the external alarm bell or siren will be immediately activated by the Watchdog's built-in timer-controlled relay. Simultaneously, the internal audio-visual alarm recorder will latch on, giving a permanent indication that an alarm activation has occurred: after five minutes or so the external alarm will automatically turn off, but the internal alarm recorder will remain latched on until the key-switch is moved to the reset position.

When the Watchdog is set to the on mode the fire and panic alarm systems are immediately fully armed, as described above, but the burglar alarm system is only semi-armed for the first 50 s. This delay gives the owner time to check (from a built-in LED) that none of the security sensor switches are active and to then pass through armed doors and so on, without sounding the main alarm. At the end of this 50 s delay the burglar alarm system becomes fully armed, and the alarm then sounds and self-latches (for a timer-controlled period) if any sensors are subsequently activated. The burglar alarm system can be temporarily disabled at any time for another 50 s period by briefly operating a remote re-entry switch, which can take the form of either a key-switch or a concealed push-button switch, thus enabling authorised persons to re-enter the building without sounding the main alarm.

Sensors And Alarms

The Watchdog can be used with a variety of types of anti-burglary and fire-detecting input sensors, which can be coupled into the system via terminal strips that are built into the unit. The anti-burglary sensors can take the form of normally-open parallel-connected devices such as pressure mats, and normally-closed series-connected devices such as microswitches and magnetically-activated reed relays mounted on doors and windows. Fire protection can be obtained by wiring normally-open thermostats in parallel, and thug protection can be obtained by wiring normally-open panic (push-button) switches in parallel.

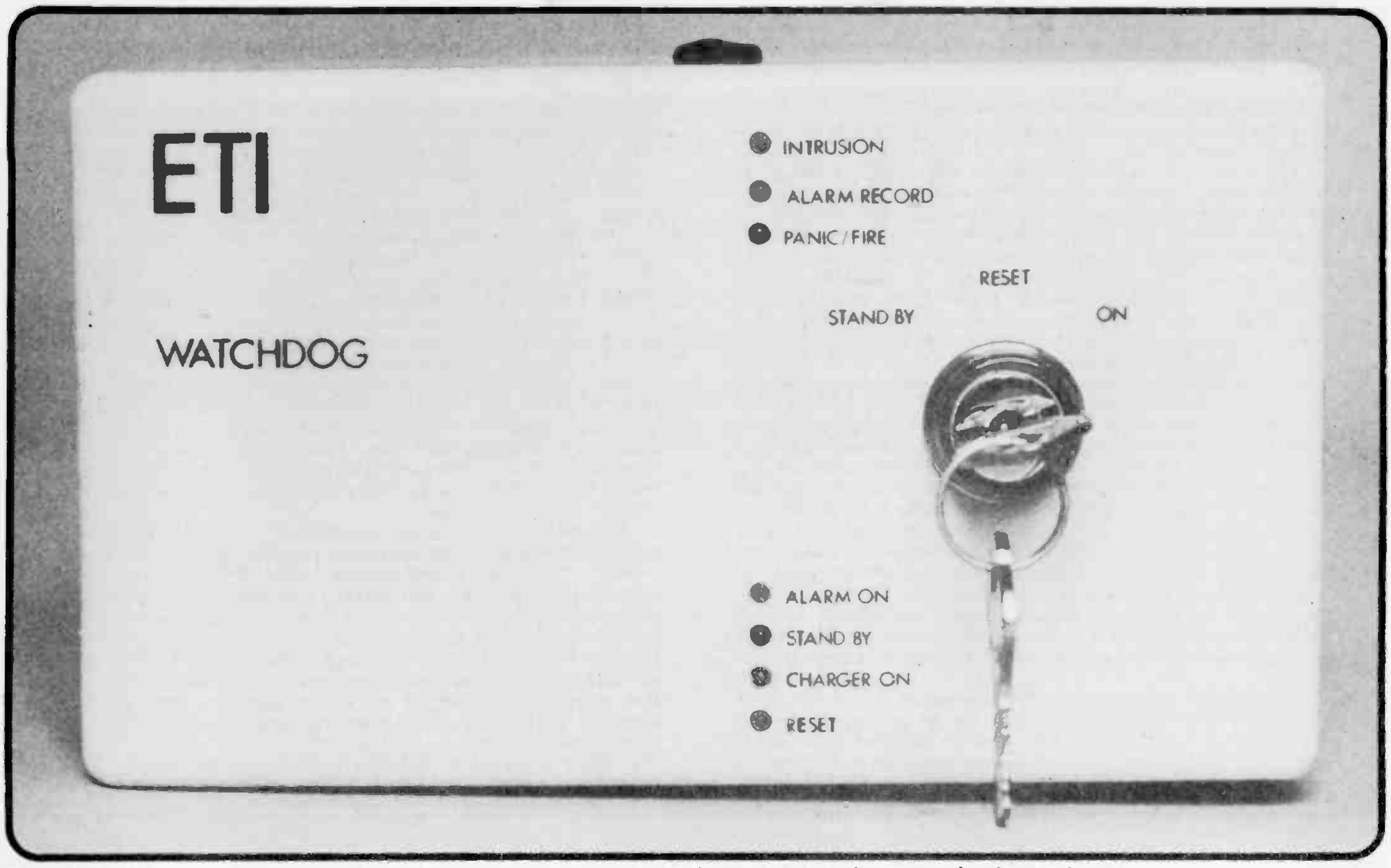
The external alarm can be any electro-mechanical or electronic siren or bell that is provided with its own power supply and draws an operating current of less than 5 A; the external alarm is activated by the contacts of relay RLA, which is built into the Watchdog unit.

Construction

The Watchdog system consists of two independent units. The main unit is intended to be wall-mounted and contains the main electronics of the system, together with the key-switch, LEDs, alarm-driving relay and the Ni-Cd battery. The second unit is a simple mains-powered Ni-Cd trickle-charger and can be mounted on a skirting board, close to a power socket, and coupled to the main unit by a twin lead.

Construction of the main unit should present very few problems, provided that the specified components are used and that the usual care is taken to observe semiconductor polarities and so on.

The Watchdog main unit and the trickle-charge unit.



The ETI Watchdog Security System is keyswitch operated and has plenty of LEDs to let you know exactly what's going on.

Start the construction by assembling all indicated components on the PCB, as shown by the overlay, and then test-fit the board into the specified case, together with the PP9-style Ni-Cd, taking care to ensure that clearance is available to give access to the wall-mounting knock-out holes moulded into each end of the case bottom.

Next, modify the top half of the case to accept the key-switch and the seven LEDs, etc, and the lower half of the case to accept the charger input socket and the PB-2720 transducer, then complete the circuit interwiring, noting that current-limiting resistors R11, R12 and R25 are wired directly in series with LEDs 3, 4 and 7 respectively.

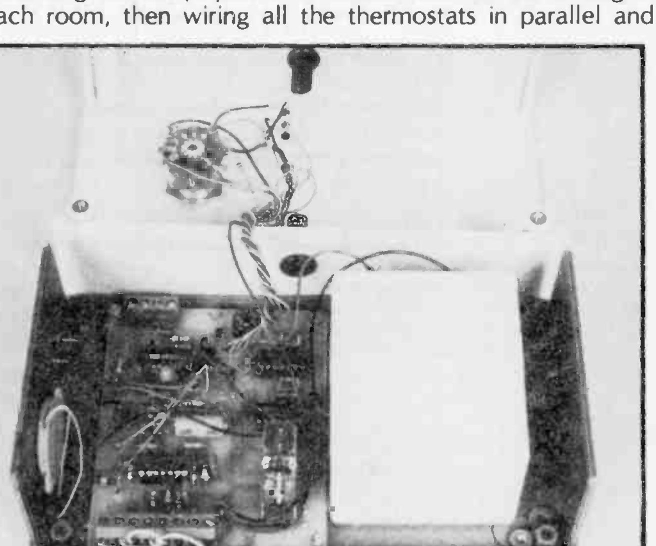
At this stage you can give the unit a brief functional check. Switch to the standby mode and check that the alarm can be activated by briefly shorting one of the N.O. thermostat or panic-button inputs. Check that the main (external) alarm turns off after a few minutes, but that the alarm recorder remains permanently latched on and produces a pulsed-tone audio signal. Check that the recorder can be turned off by switching to reset. Similarly, check that the burglar alarm circuitry functions as already described by switching to the on mode.

The charger circuit is so simple that its construction should present absolutely no problems at all. Note, however, that Q2 must be mounted on a small heatsink. When construction is complete switch the unit on, connect a DC current meter directly across the unit's output terminals, and check that a current indication of roughly 70 mA is given. Finally, connect the output of the charger to the charger socket on the main Watchdog unit, switch on, and check that LED7 (on the main unit) illuminates.

Installation And Use

The installation of a home security system is a fairly major undertaking, with many fine points to consider and individual decisions to be made regarding the degree of protection that is required and the types of sensors that are to be used. In the next few paragraphs we've outlined some very basic principles of installation; however, it's up to the individual reader to work out the details of his own sensor and alarm generator networks, and then couple these networks to the main Watchdog unit.

A complete home protection system should contain three distinct types of sensor networks, these being designed to give (1) fire protection, (2) anti-thug or panic protection, and (3) anti-burglary protection. Fire protection can be obtained by mounting normally-open thermostats close to the ceiling in each room, then wiring all the thermostats in parallel and



The main unit contains the alarm sensing circuitry and the PP9 Ni-Cd battery. Connections to the sensors depend on the application and none are shown on this prototype — but all the necessary wiring can enter the box through the two grommets in the lid.

HOW IT WORKS

The circuitry of the main unit can be broken down into two distinct sections, with the basic alarm circuitry plus alarm recorder to the right of key switch SW1a, and the burglary-detection circuitry to the left. The circuitry to the right of SW1a is permanently enabled and can be activated at any time by the panic and fire inputs: the burglary-detection circuitry is active only when the system is turned fully on by SW1.

The operation of the basic alarm circuitry and the alarm recorder is fairly simple. IC2a-IC2b is a long-period monostable (several minutes) and can be triggered by applying a high (logic 1) signal to R13 via the D2-D3 OR gate; the mono can thus be triggered by closing any of the fire-detecting thermostats or the panic buttons, or by a high output from IC1d. When the mono is triggered, the output of IC2b goes high for the duration of the monostable period and thus drives RLA (and the external alarm) on via Q1 for a preset period. Simultaneously, the output of IC2a goes low and causes special-purpose bistable IC2c-IC2d to self-latch into a state in which the output of IC2d goes low, driving LED6 (the visual alarm recorder) on and activating the IC3-IC4 audible alarm recorder circuitry. When the monostable turns off at the end of its timed period the external alarm also turns off, but the audio-visual alarm recorder remains active, giving a permanent indication that an alarm action has occurred. The monostable and the recorder can both be reset by briefly moving SW1 to the reset position.

The audible alarm recorder circuitry is quite simple. IC3a-IC3b form a low-frequency astable, which is gated on by a low input signal. IC3c-IC3d form a high-frequency astable (a couple of kilohertz) which is gated by the output of IC3b. The output of the high-frequency astable is fed to acoustic transducer TX1 via bridge-configured driver IC4. Thus, when the circuitry is activated by a low input signal, an audible pulsed-tone signal is generated by TX1.

The burglary-detection circuitry is designed around IC1 and is active only when SW1 is switched to the on position. Here, the N.O. and N.C. burglar-detecting security switches are wired in such a way that a high voltage is normally applied to R3, but this voltage goes low if any of the N.C. switches are opened or the N.O. switches are closed. The R3 voltage is inverted by both IC1a and IC1b, so that LED2 turns on and a high voltage is fed to one input of IC1c if an intrusion occurs; C1 and R5-C2 filter the signals from R3, to eliminate the effects of lightning-induced signals and transients. The other input of IC1c is controlled by the C3-R7 time-controlled network, which causes IC1c to be effectively disabled for a minute or so after SW1 is first moved to the on position or after the optional re-entry switch is momentarily closed. The output of IC1c is inverted by IC1d and fed to the D2 input of the D2-D3 OR gate, where it can control the action of the main alarm circuitry.

Thus, when the burglar alarm circuit is first switched on by SW1 the IC1a-IC1b section is fully enabled, so that LED2 will turn on if any of the security switches are incorrectly set, but IC1c-IC1d are disabled by the C3-R7 network, so that the main alarm will not activate under this condition. After a minute or so, however, the IC1c-IC1d section becomes fully enabled, so the main alarm will sound instantly if any subsequent intrusion is detected.

Note that the main Watchdog unit is powered by a single PP9-style Ni-Cd battery, which is intended to be permanently trickle-charged by an external mains-powered charger circuit. Also note that the circuit is provided with a total of seven LEDs, which give visual indications of the existing operating mode, the presence of sensor faults/actions, the presence of the charger current, and the record of an alarm action.

The charger circuit is very simple and is intended to apply a permanent trickle-charge current of roughly 70 mA to the Ni-Cd battery of the main Watchdog unit. Here, the mains voltage is stepped down, full wave rectified and smoothed by T1-D9-D10-C8, to provide roughly 13 V DC across C8. This voltage is used to power the constant-current generator that is designed around ZD1-R26-Q2-R27. ZD1 sets a standing voltage across emitter resistor R27, which thus determines the emitter current of Q2: since the emitter and collector currents of an active transistor are virtually identical, the collector of Q2 effectively acts as a constant-current source and is used to feed a trickle-charge current to the Watchdog Ni-Cd via D8. LED7 (in the main unit) illuminates when the charger is active.

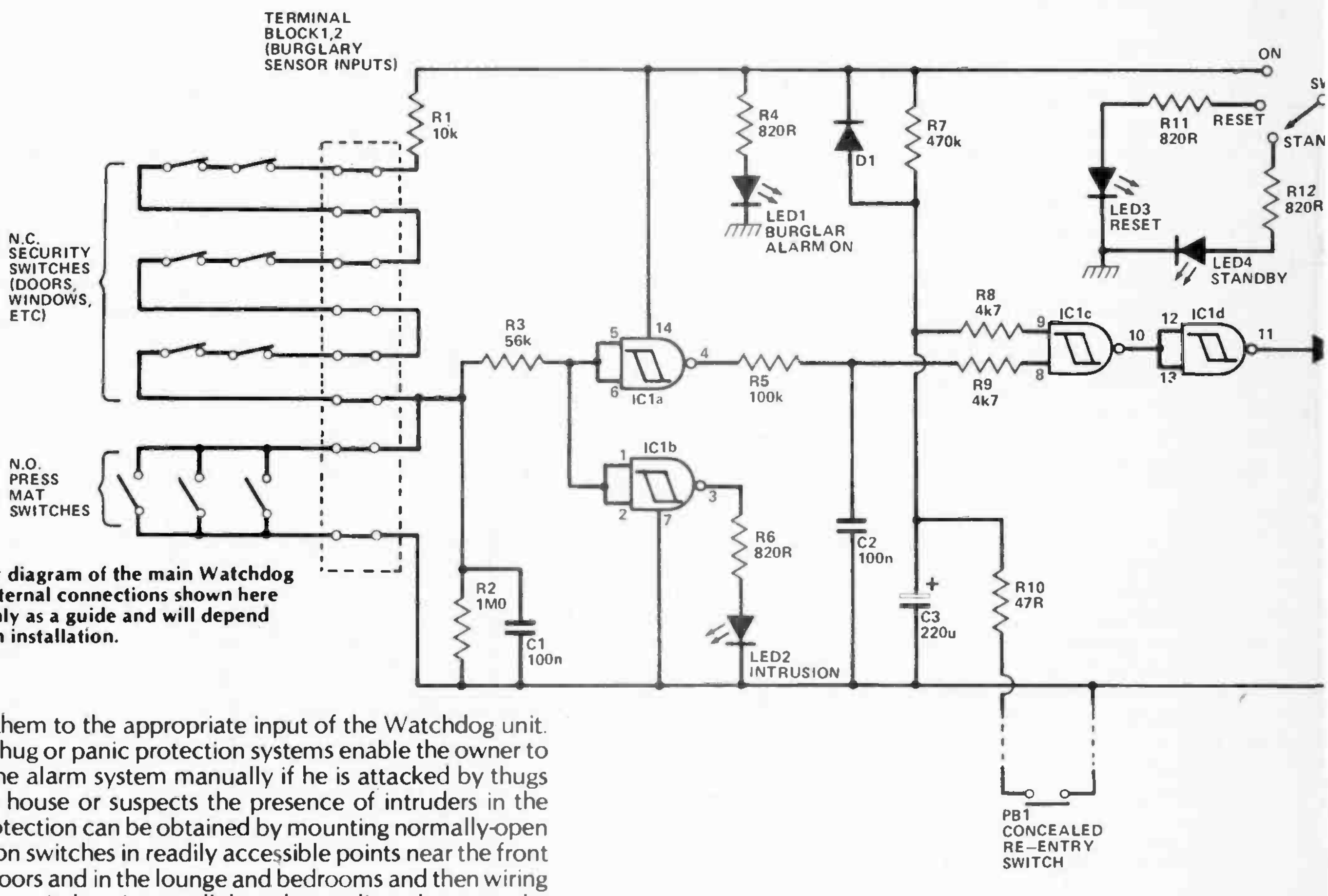


Fig.1 Circuit diagram of the main Watchdog unit. The external connections shown here are given only as a guide and will depend on your own installation.

coupling them to the appropriate input of the Watchdog unit.

Anti-thug or panic protection systems enable the owner to activate the alarm system manually if he is attacked by thugs inside the house or suspects the presence of intruders in the house. Protection can be obtained by mounting normally-open push-button switches in readily accessible points near the front and rear doors and in the lounge and bedrooms and then wiring all of the switches in parallel and coupling them to the Watchdog unit.

BUYLINES

The piezo-electric transducer can be obtained from Ambit International. Electrovalue are stockists for the PCB terminal blocks. The 6 V relay, DC socket and key-switch are available from Watford Electronics.

All the other semiconductor devices used in this project should be available from advertisers in this issue eg Technomatic, Greenweld etc. The Samos 002 case can be obtained from West Hyde Developments.

Special-purpose security sensors, panic buttons and alarms are available from specialist security companies such as Strathand Security, 44 St Andrew's Square, Glasgow G1 5PL.

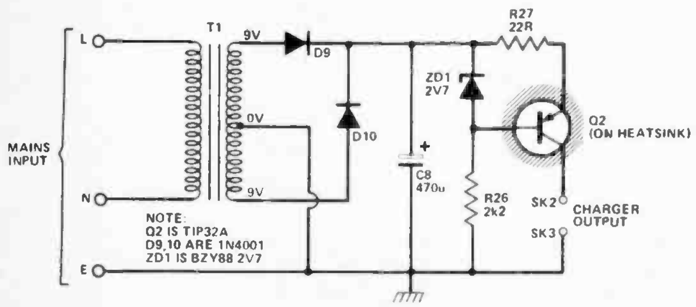
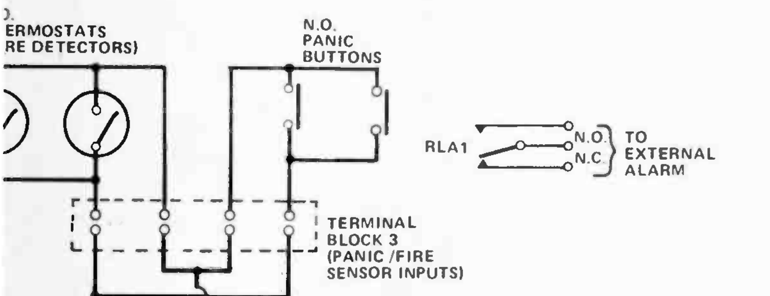


Fig.2 (Above) Circuit diagram of the trickle-charger.



Burglary protection can be obtained either by using a 'perimeter defence' system that detects the intruder as soon as he enters the building through a protected door or window, or by using a 'spot defence' system that detects him only after entry has been made (as he opens internal doors or treads on concealed pressure mats), or by using a combination of the two systems.

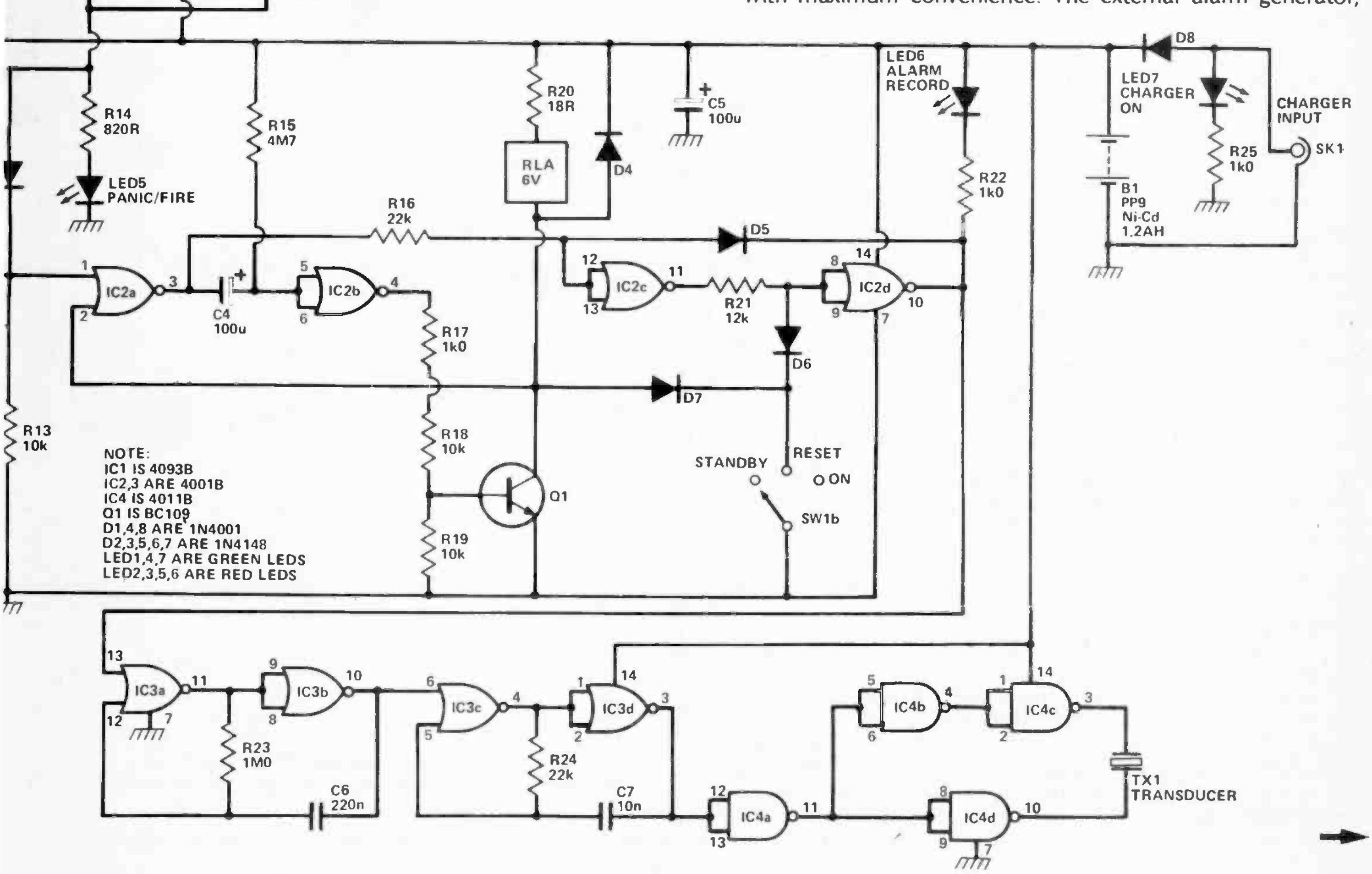
Selected doors and windows are easily protected with a reed-relay/magnet combination: the magnet is installed in the door or the opening window, opposite a reed-relay installed in the frame, so that the relay is normally closed but opens when the door/window is opened. All the relays are then wired in series and connected to the appropriate input of the Watchdog unit, so that the alarm activates when any of the protected doors/windows are opened.

Pressure mats come in a variety of sizes and are easily hidden under rugs and carpets. They are usually normally-open devices, so any number can be wired in parallel and fed to the appropriate input of Watchdog.

Modus Operandi

When planning the installation, the house-owner must try to think like a burglar. Normally, the burglar enters a house from an easy access point that is obscured from the view of the neighbours, ie a back door or window. Often, he breaks in using tools 'borrowed' from the owner's shed or garage, so these two places should be included in the owner's defence system. Invariably, the burglar's first action on entering the property is to secure a rapid escape route, ie to open the back door. He then starts hunting for stealable goodies.

Once you've designed and installed your own sensor systems you can couple them up to the main Watchdog unit, which should be mounted in a central part of the house, such as on a landing or at the foot of the stairs, so that it can be operated with maximum convenience. The external alarm generator,



complete with its own power supply arrangement, can then be coupled to the output of the unit so that it activates when relay RLA closes.

If you decide to mount a re-entry switch on the front door of the house (so that you can enter the building without activating the alarm), take care to conceal its wiring. If required, a number of re-entry switches can be wired in parallel so that, for example, the system can be temporarily disabled from either the front door or the main bedroom.

The alarm system is very simple to use. The panic and fire alarm side of the circuit is permanently enabled and can be operated at any time. The anti-burglar section is enabled only when the main key switch is set to the on position. If LED2 lights at the moment of turn-on it means that part of the burglary sensor system is either open or closed when it should not be, possibly due to an open door or a chair resting on a pressure mat, for example. The fault must be rectified before the system is put to full use.

If you leave the house or pass through a protected area after turning the system on, remember to use the re-entry facility before returning to the unit, or you'll sound the alarm and annoy the neighbours.

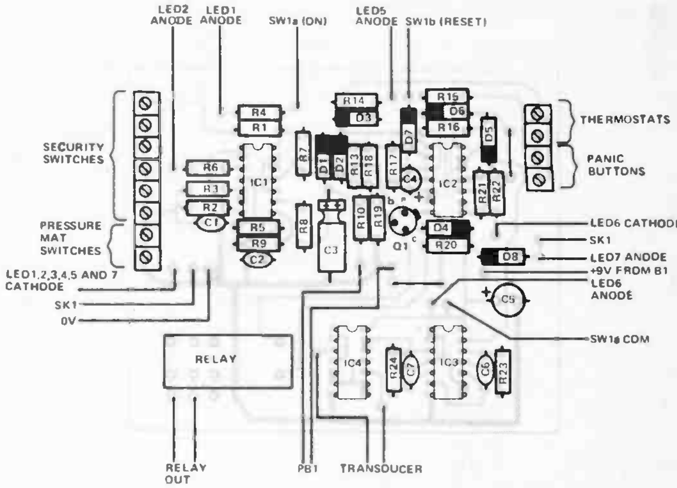
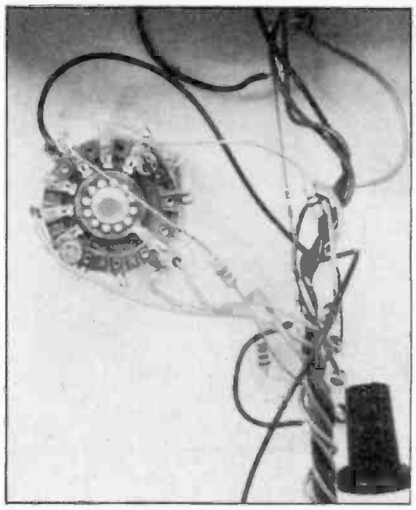
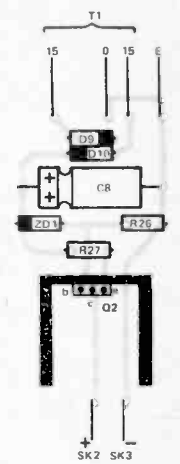


Fig.3 Overlay for the main board. Note that R11,12 and 25 are mounted off-board.



Details of the keyswitch wiring. The off-board resistors can also be seen.



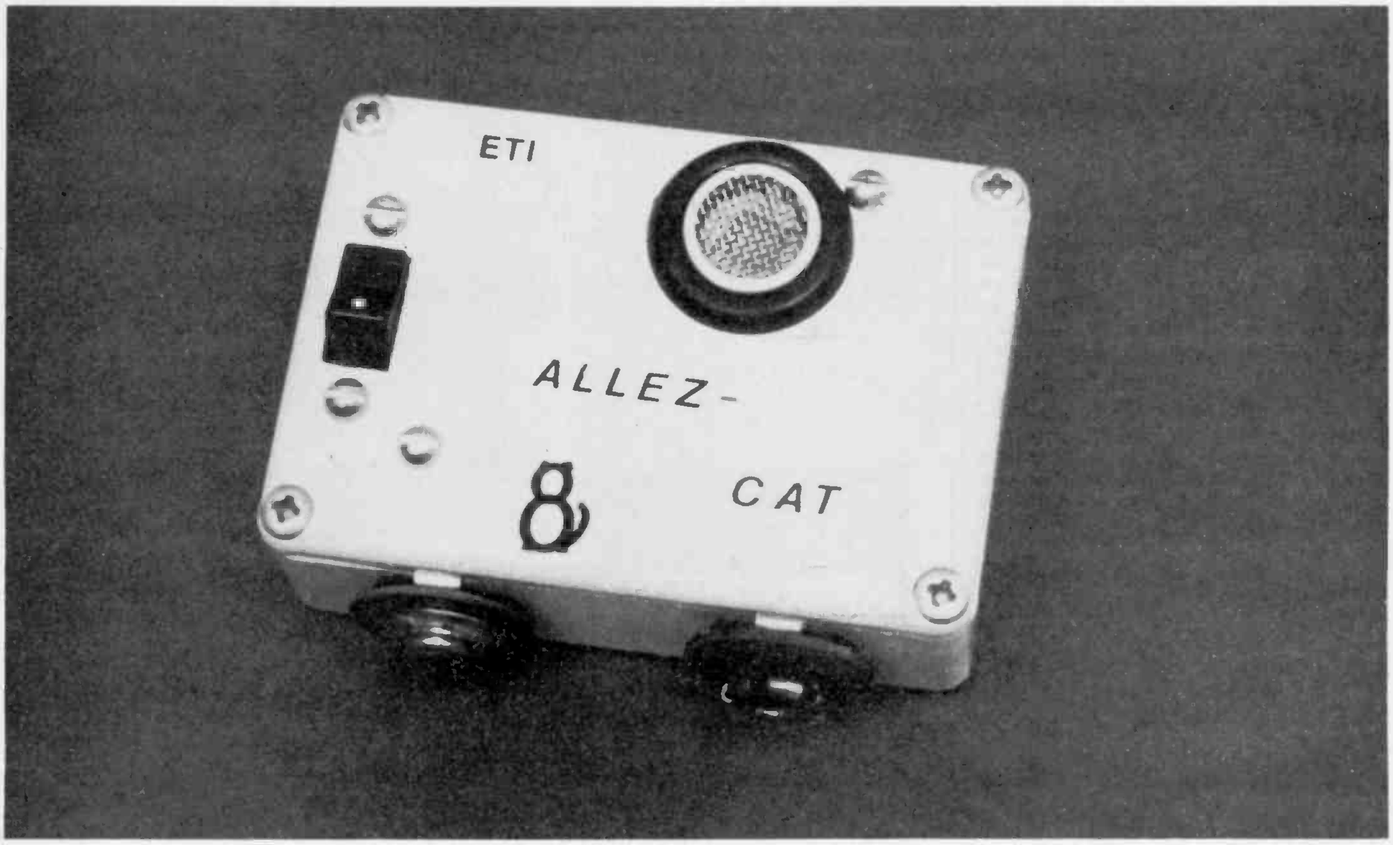
PARTS LIST

Resistors (all ¼ W, 5%)	
R1,13,18,19	10k
R2,23	1M0
R3	56k
R4,6,11,12,14	820R
R5	100k
R7	470k
R8,9	4k7
R10	47R
R15	4M7
R16,24	22k
R17,22,25	1k0
R20	18R
R21	12k
R26	2k2
R27	22R
Capacitors	
C1,2	100n ceramic
C3	220u 16 V axial electrolytic
C4	100u 10 V tantalum
C5	100u 16 V electrolytic (PCB type)
C6	220n polycarbonate
C7	10n polycarbonate
C8	470 40 V axial electrolytic
Semiconductors	
IC1	4093B
IC2,3	4001B
IC4	4011B
Q1	BC109
Q2	TIP32A
D1,4,8,9,10	1N4001
D2,3,5,6,7	1N4148
ZD1	BZY88 2V7
LED1,4,7	0.125" green LED
LED2,3,5,6	0.125" red LED
Miscellaneous	
T1	9-0-9 @ 75 mA
SW1	two-pole six-way wafer key switch
SK1	DC socket and plug
SK2,3	4 mm sockets (and plugs)
TX1	piezo-electric transducer
RLA	6 V DPCO, PCB-mounting

PCB-mounting terminal blocks; Verocase (order code 202-21031G); case for charger unit (order code Samos 002).

PEST CONTROL

If I could talk to the animals. . . I'd tell 'em to go away. Get the message across with this harmless little gadget. Design and development by Phil Walker.



Is your garden the main attraction for all the local neighbourhood cats, dogs, hedgehogs, rats, voles, aardvarks and other small furry creatures? If so, this device should protect your seedlings, kiddies' sandpit or dustbin from their attentions until you can make a more permanent arrangement. It is harmless to the animals themselves, being merely annoying rather than painful.

Our first design involved frequency-shifting a tape of Barry Manilow up into the ultrasonic range so that only animals could hear it, but it was felt that this would lay us open to prosecution by the RSPCA. Consequently a second approach was adopted.

Many people have noticed that, when they operate the ultrasonic remote controls on their TVs or hi-fis, their household pets will protest loudly and leave the vicinity rapidly. We have designed this little device to capitalise on the phenomenon, in reponse to our readers requests for a gadget capable of protecting precious plants without the need for violence!

The 'Allez-Cat' consists of two basic parts: the first is an oscillator tuned to 40 kHz while the second is a voltage doubler and pulse generator. The pulses are about 10 milliseconds long and occur two to three times a second. This is done to reduce battery drain and increase the annoyance factor (for a cat, dog, hedgehog . . .). The voltage doubling action increases the available output power for a given battery voltage.

The whole device (apart from the battery) fits into a small plastic box and can use 6 V or 9 V batteries such as a PP1 or PP9, as convenient. If it is to be used out of doors, some form of protection from rain or other water should be provided as the transducer isn't waterproof.

Construction

The device is constructed in a small plastic box which is modified to clip onto a PP9 or similar battery using the battery connectors. The assembly of the PCB should offer no problems provided that care is taken with the polarity of capacitors and diodes. Do not fit the transducer to the board at this stage.

Mount the switch on the lid of the box using two short bolts. The large grommet should fit fairly loosely into the large hole in the lid and the transducer may be pushed into its centre using reasonable care. (Push in from the rear and avoid pressing the mesh front). Attach some thin single strand wire to its pins, and put the lid on one side for the time being.

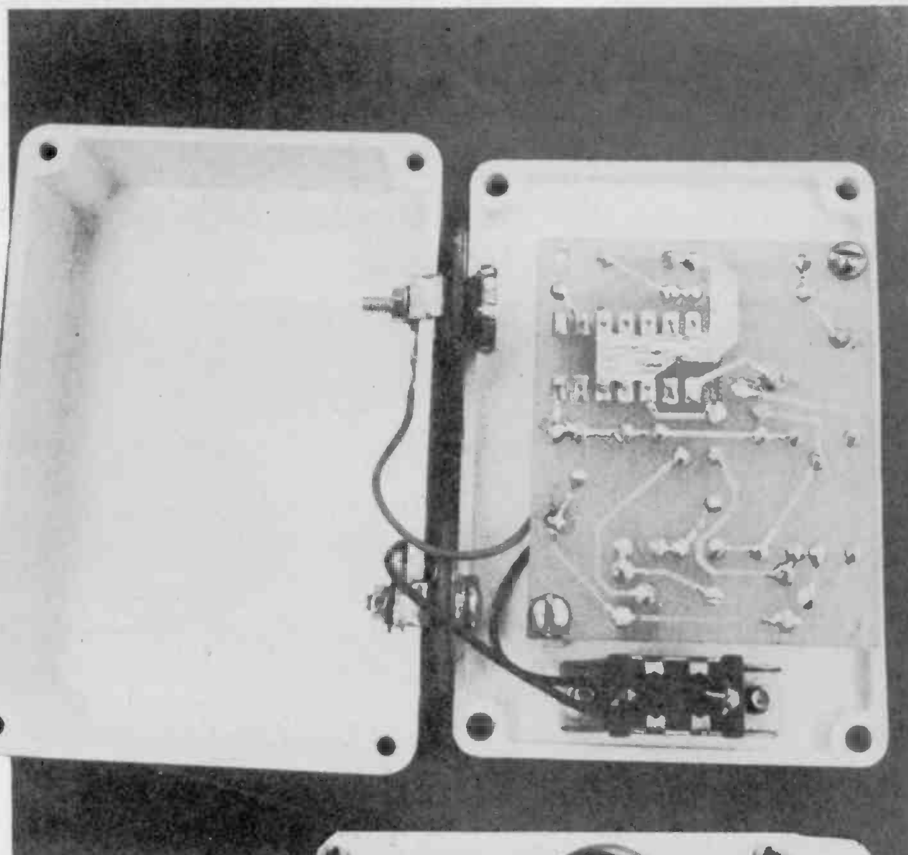
Assemble the battery connectors to the side of the box as shown in Fig. 1, remembering to attach wires about 2" long to the solder tags first. The small grommet is cut in half (ie into two discs), each half being sandwiched between the battery connector and the case. This gives a steadier mounting, as the rear of the connectors is bowed slightly. Attach the wire from one connector to its proper hole on the

PCB. Then attach another short length of wire to the other supply hole on the PCB.

Bolt the PCB to the lid of the box using the spacers and ensure that the wires to the transducer are soldered to the proper holes on the board. Connect the two remaining wires to the switch and assemble the two halves of the box with self-tapping screws (these are provided with the box). You are now ready to control the migratory habits of the animal kingdom.

HOW IT WORKS

IC1-R6-R7-R8-C4 together with TX1 form a power oscillator which drives TX1 at its resonant frequency of 40 kHz for maximum efficiency. Q1,Q2 and their associated components form a standard multivibrator oscillator with a low natural frequency. R5 and Q3 form an electronic switch which applies the negative spikes occurring at the junction of C3, R3 and D2 to IC1. This means that the power supply to IC1 consists of short spikes of up to twice the battery voltage. The duration of the spikes depends on the current drawn by IC1, since this current flow alters the mark/space ratio of the multivibrator from a square wave to a series of widely spaced pulses. The repetition rate of these pulses is determined by C2 and its associated timing resistors.



The Allez-cat PCB easily fits into a small Vero potting box, with the battery connectors bolted through the side of the case (above). The whole project then clips directly onto a PP9 battery (left).

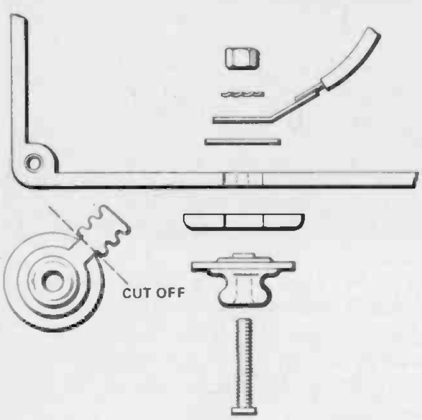


Fig.1. (right) Constructional details of the battery terminals.

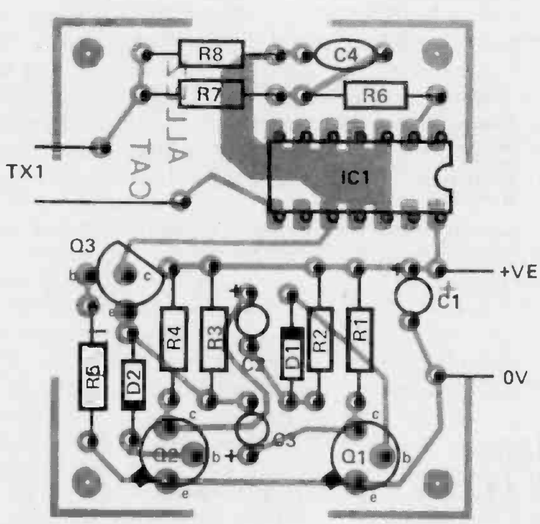


Fig.2. (below) Component overlay for the Allez-cat pest control project.

PARTS LIST

Resistors (all ¼W, 5%)	
R1,6	1k0
R2,3,4,5	10k
R7	100R
R8	10R
Capacitors	
C1,2,3	100u tantalum
C4	10n polycarbonate
Semiconductors	
IC1	LM380
Q1,2	BC107
Q3	BC182L
D1,2	1N4148
Miscellaneous	
SW1	miniature rocker or slide switch
TX1	40 kHz ultrasonic transmitter

PCB (see Buylines); PP9 battery; pair of PP9 battery connectors; two solder tags; fixing hardware; grommets; case (Vero 75-1469L).

BUYLINES

No problems with any of the parts for this project, with the possible exception of the ultrasonic transmitter. The catch is that most mail order companies sell these in matching pairs, but for this application we don't need the receiver. A quick chat with Watford Electronics established that they are prepared to oblige with single transducers, so direct your enquires thence. PCBs, as ever, are available from us at the price listed on page 81.

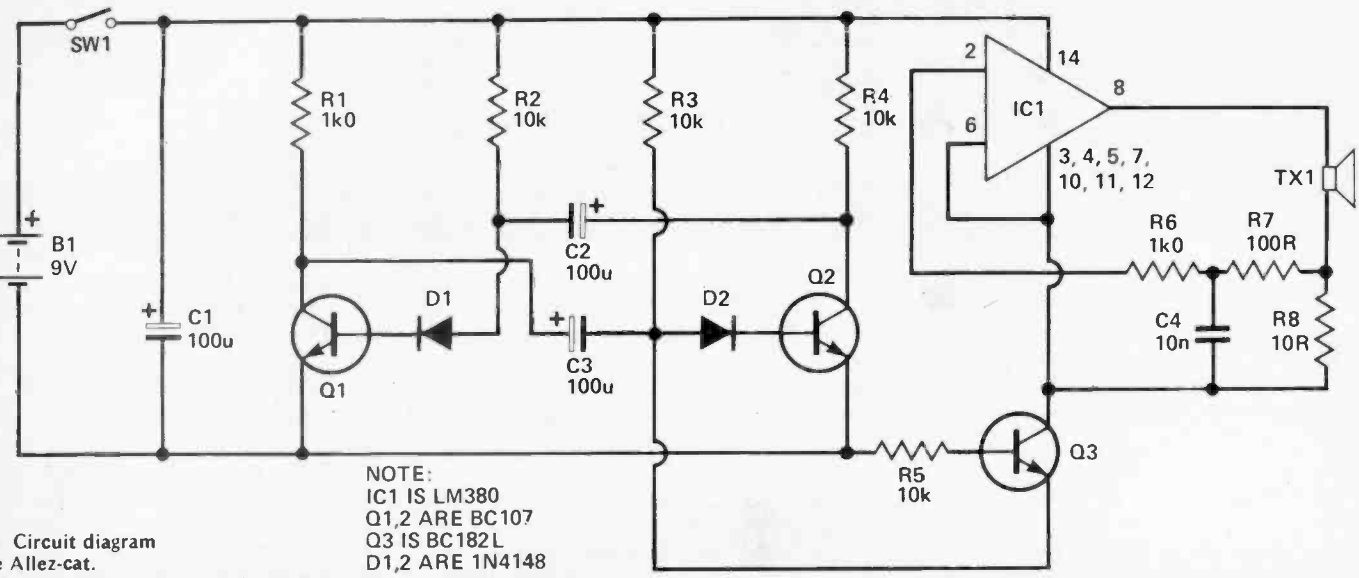
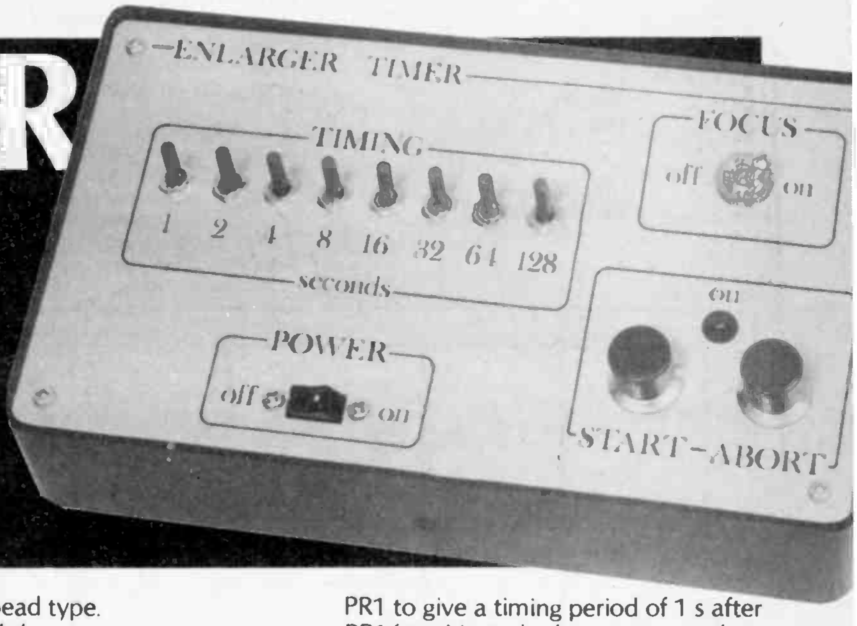


Fig.3. Circuit diagram of the Allez-cat.

ENLARGER TIMER



At last — an enlarger timer designed for the photographer by a photographer. Design and development by Tony Alston.

Most enlarger timers use either potentiometers or rotary switches for time period selection; however, both these methods suffer from some drawbacks. First the potentiometers, although inexpensive, are generally difficult to line up with a timing mark especially under darkroom conditions, and also tend to wear out after long usage. The rotary switch overcomes the alignment problem but the timing period range is normally limited to the number of click stop positions — mostly 12 ways, maximum 18 ways. The ETI Enlarger Timer does not suffer from either of these problems; it uses only eight toggle switches in conjunction with the 2240 programmable timer IC to offer a wide range of accurate and easily selected timing periods ranging from one second to 4 minutes 15 seconds in one-second steps.

This flexibility is due to the programmable eight bit counter, oscillator and control flip/flop featured within the timer IC. Having set the time base to 1 s using PR1, R4 and C2, each single switch (SW1-8) will give the basic timing periods of 1,2,4,8,16,32,64 and 128 s; by switching in more than one switch, any combination of timing periods can be achieved as previously mentioned.

Construction

As can be seen from the photographs, all switches and push-buttons are mounted on the front panel together with the LED indicator. Suitable mains input and output sockets such as Bulgin miniature mains type are mounted on the rear case panel, input nearest the transformer, output (for the enlarger) by the PCB. The PCB design will enable an easier and neater assembly, but make sure to orientate components D1-D5, Q1, IC1, C2, C3 and LED1 as shown in the overlay diagram. Note that C2 must be a tantalum bead type.

Once all the components are mounted on the PCB and the switches, LED1, sockets and transformer are wired to the board, make sure that the panel assembly does not foul the transformer or the relay when fitted to the case.

Setting Up

This couldn't be easier; having checked all connections, connect the timer to the mains and the enlarger to the unit. First put switches SW1-8 and SW10 in the off positions, put SW9 (on/off switch) in the on position, operate focus switch SW10 and the enlarger lamp will light. Switch SW10 to the off position. Adjustment to the timing range can now be made; switch SW1 only (1 s switch) to on and adjust PR1 to give a timing period of 1 s after PB1 (start) is pushed — a stopwatch or digital watch is ideal for this.

Using The Timer

Switch on SW9 (on/off switch) and power will be applied to the circuitry. SW10 can be used for focusing the enlarger; cancel SW10 once this is done. Select the timing period required using a combination of SW1-8, push PB1 and the enlarger lamp timing cycle will commence; after this period the timer will stop/reset. LED1 will be on during timing period as a visual indication. If cancellation of a timing period is needed press PB2 which will abort and reset the timer. If any interference from RLA/2 is experienced, fit a 100n 600 V capacitor as marked on the circuit and overlay diagrams (C4).

HOW IT WORKS

The heart of the ETI Enlarger Timer is the 2240 programmable timer IC which features a time base oscillator, programmable eight bit counter and a control flip-flop that can be used in monostable or astable mode. Here it is used in the monostable mode.

On application of a positive pulse to pin 11 (trigger) via PB1 and R1, the timing cycle is started. The trigger input activates the time base oscillator, enables the counter section and sets the counter outputs low from their normally high state. This switches on Q1 and activates RLA for the time duration as set by the SW1-8 combination. The timing sequence is completed when a positive pulse is applied to pin 10 (reset) via R3 from the output bus, disabling the time base and counter sections and returning the counter outputs to a high state.

The duration of the timing cycle T_o is given as:

$$T_o = nT = nRC \text{ seconds}$$

(R in ohms, C in farads)

where T (= RC) is the time base period as set by the timing components at pin 13 (PR1, R4 and C2) and n is an integer in the range of $1 \leq n \leq 255$ as determined by the combination of counter outputs (pins 1-8) via SW1-8 to the output bus. The time base as set by PR1, R4 and C2 is 1 s.

The binary-counter outputs are the open collector type and can be shorted together to the common pull-up resistor R6. Thus the time delays associated with each counter input can be added together; for example, if pin 6 is connected by SW6 to the output bus the duration of the timing cycle, T_o, is 32T. (T is 1 s as previously stated). Similarly, if pins 1,5, and 6 are all connected to the output bus via their appropriate switches SW1, SW5 and SW6 the total time delay is 49T (1 + 16 + 32). In this manner the timing cycle can be programmed to be from 1 s to 255 s (four minutes 15 s) in 1 s steps by proper choice of switches SW1-8.

The enlarger lamp is powered from the AC outlet socket and receives its current via the RLA/2 contacts for the duration of the selected timing period. An LED is incorporated as a visual indicator; it is switched on by RLA/1 and remains on for the timing period. Manual cancellation is provided for by PB2 which applies a positive pulse to pin 10; this can be used at any point in the timing period. SW10, the focusing switch, overrides the RLA/2 contact regardless the output state of IC1 thus enabling the enlarger to be focused.

The power supply consists of T1, D2-D5 and C3 which provides smoothing.

PARTS LIST

Resistors (all ¼ W, 5%)

R1,2,6	10k
R3	47k
R4	33k
R5	22k
R7	1k5

Potentiometer

PR1	22k miniature horizontal preset

Capacitors

C1	10n disc ceramic
C2	22u 16 V tantalum
C3	1000u 25 V axial electrolytic
C4	100n 600 V mixed dielectric

Semiconductors

IC1	uA2240CP
Q1	2N3702
D1	1N4148
D2-5	1N4001
LED1	0.2" red LED

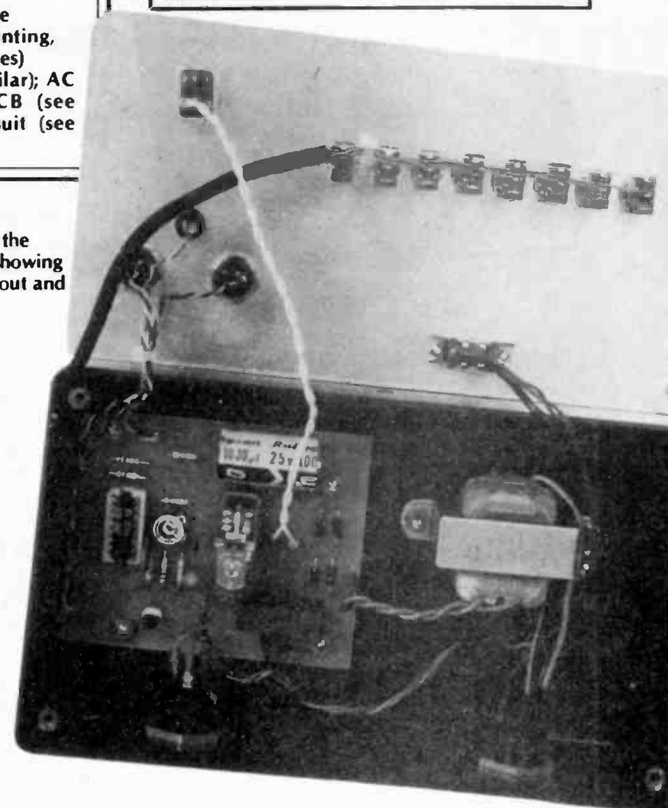
Miscellaneous

PB1,2	momentary action push-button
SW1-8	SPST miniature toggle
SW9	SPST 240 V 3 A miniature rocker
SW10	SPST 240 V 3 A toggle
RLA	12 V DPDT PCB-mounting, 205R coil (see Buylines)

Transformer (12 V, 250 mA or similar); AC outlet socket (Bulgin type); PCB (see Buylines); mains lead; case to suit (see Buylines).

BUYLINES

Most of the components used in this project can be easily obtained for component retailers or mail order firms. IC1 is available from Technomatic and the relay is from Watford Electronics. The case we used is from Tandy, order no. 270-627; and if you don't want to make your own PCB from the foil pattern at the back of the magazine, then take a look at the advert for our PCB Service on page 91.



(Left) Inside the prototype, showing the neat layout and tidy wiring.

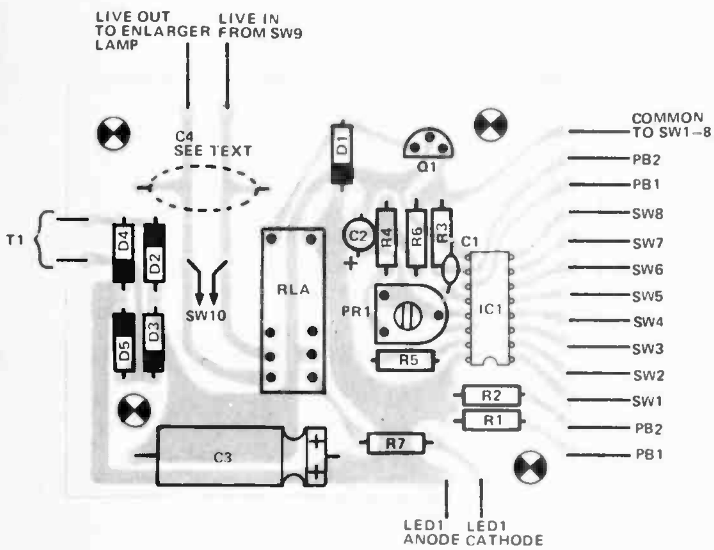


Fig. 1 Component overlay of the ETI Enlarger Timer.

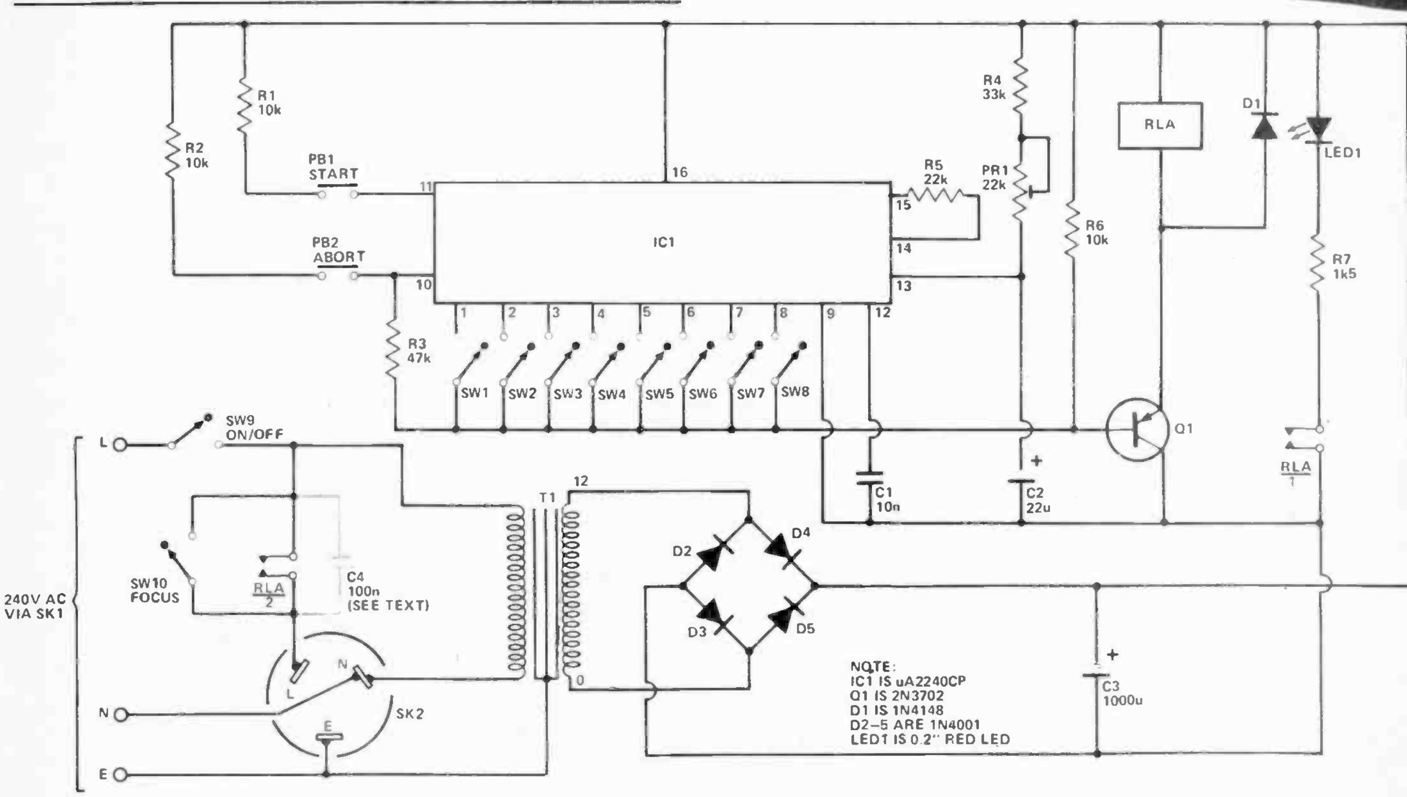


Fig. 2 Complete circuit diagram of the timer. C4 may be necessary to suppress switching noise from the relay contacts.

PHONE BELL SHIFTER

If you want to customise your phone without incurring a visit from the British Telecom heavies clutching their wirecutters, then this is the project for you. Design by Ray Marston. Development by Steve Ramsahadeo.

A major problem with the present-day British telephone system is that all bell-type phones generate virtually identical ringing sounds, thereby creating all sorts of problems. In the home, for example, while watching TV you'll often hear a phone ring and instantly leap to your feet to answer the call, only to find that the ringing sound comes from a phone in the TV programme. Similarly, when you are relaxing in the garden on a summer's day, or when you have wandered into a neighbouring office at work, you'll often hear a distant phone ring and will not know if the call is intended for you or a neighbour.

The ETI Phone Bell Shifter project is designed to overcome these identification problems by giving your phone its own distinctive sound. The project consists of a sophisticated five IC circuit; its input is provided by an inexpensive transducer, acoustically coupled to the phone body to detect the ringing of the phone bells, and its output is taken to an external 15R speaker from a built-in alarm-tone generator. A distinctive pulsed-tone alarm sound is generated which is synchronous with the detected phone bell signal. The speaker is connected to the output of the unit using twin flex, and can either be placed near the phone (so that the unit acts as a phone identifier) or can be placed at a remote location (so that the unit acts as a phone bell extender).

A unique feature of our Phone Bell Shifter project is that it uses a micropower signal-sampling technique to detect the presence of a ringing signal and then automatically switch the unit into the full-power mode. Because of this technique, the unit draws a quiescent current of a mere 7 uA, thus giving up to two years of continuous operation from a single PP9 supply battery.

The Phone Bell Shifter project is provided with two preset pots plus a normal volume control, and is simple to set up and use. The acoustic pickup transducer (which forms the input of the unit) is simply placed underneath the telephone body (below the bells) and the sensitivity of the unit is then initially adjusted by one preset to give the desired trigger action when the phone rings: once this pot has been initially set, it requires no subsequent adjustment.

Construction

All of the major electronics of this project (except the input transducer and the output speaker) are built into a single Verobox, which also houses the PP9 supply battery, and construction should present very few problems.

A number of high value resistors are used in the circuit, so extra care should be taken to ensure that moisture and contaminated grease are not allowed to shunt down the effective values of these components; when PCB construction is complete and the circuit has been tested, the board should be given a coating of varnish to prevent the ingress of moisture.

When the PCB construction is complete, fit the PCB and the PP9 battery into the recommended Verobox and complete the interwiring to RV1, SW1 and the input and output sockets. Fit the specified speaker in a second Verobox and then connect it to the input of the main unit.

To set the unit up initially, simply adjust sensitivity control PR1 so that the unit's alarm activates when the phone bells ring, and then adjust PR2 so that the alarm volume does not fall to zero when the main volume control is turned fully down.

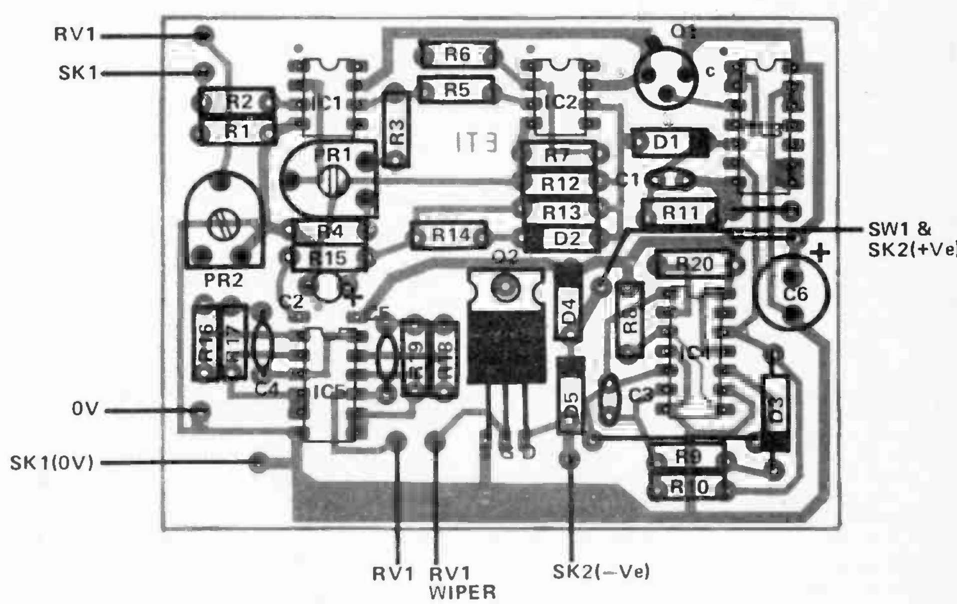
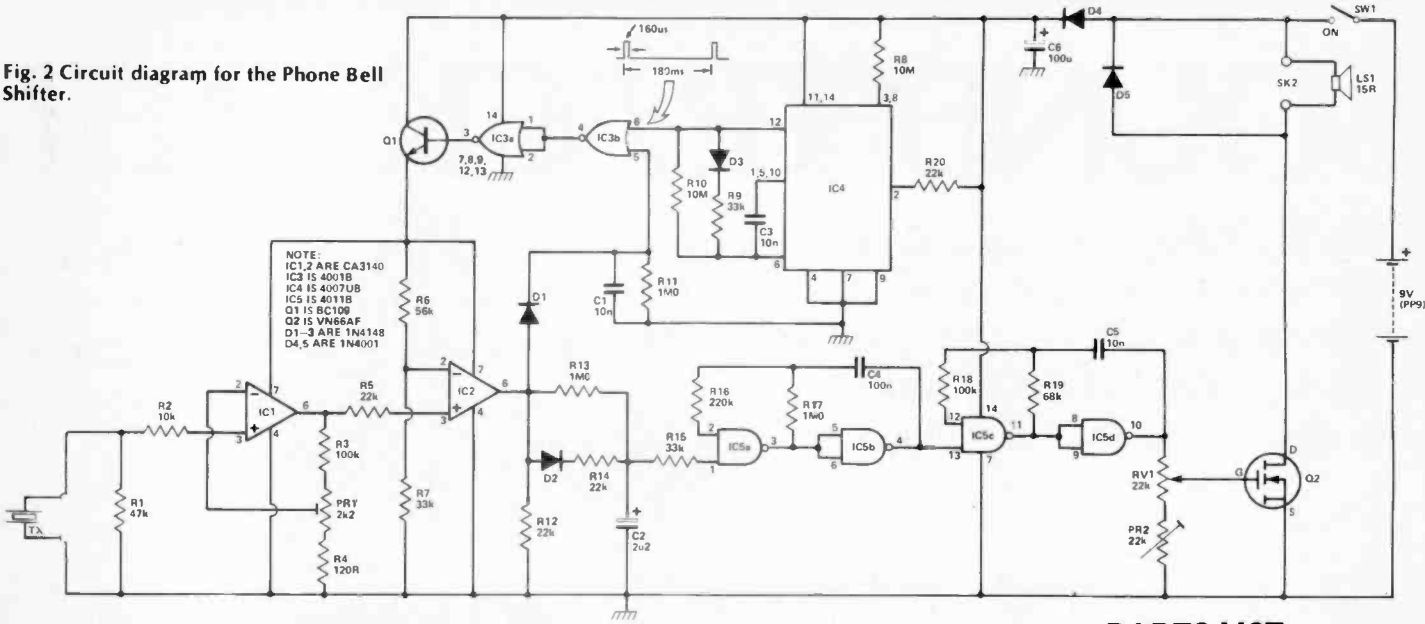


Fig. 1 Component overlay.

Fig. 2 Circuit diagram for the Phone Bell Shifter.



HOW IT WORKS

First ignore the effects of the signal-sampling circuitry and assume that Q1 is replaced by a short-circuit, so that supply power is continuously applied to IC1 and IC2.

The acoustic pickup device used in the circuit is a PB-2720 transducer, which is placed under the phone body, below the bells. This transducer has a poor low frequency response, but is quite sensitive to the high frequency (3 to 5 kHz) overtones of the phone bell. The output of the transducer is applied directly to the input of non-inverting variable-gain amplifier IC1.

Although the input of IC1 is DC-grounded by R1-R2, the CA3140 op-amp used in this position is able to respond to input signals all the way down to 0 V: consequently, the output of this op-amp (pin 6) corresponds to an amplified but positively half-wave rectified version of the input signal. This signal is fed to the input of non-inverting voltage comparator IC2, which is reference-biased at about 3 V by R6-R7.

Thus, the overall action of the IC1-IC2 circuit is such that the output of IC2 is normally low, but changes into a series of 3-5 kHz square waves in the presence of bell-ring input signals. These square wave signals are processed by the R13-D2-R14-C2 dual-time-constant integrating network, so that a high DC voltage (6 V) is developed across C2 in the presence of a true ring signal, and this voltage is used to activate the IC5 alarm-tone generator circuitry. Note that, because of the integrating action of D2-R14-C2, the input signal must be sustained for greater than 50 ms to initiate a high switching action in IC5, so the circuit has excellent immunity to transient signals caused by physically banging the telephone body or PB-2720 transducer. Similarly, because of the actions of R13-C2, the IC5 alarm circuit does not turn off until several hundred milliseconds after the bell-ring input has been removed, and thus gives a single continuous action from a double (ring-ring) input signal.

The IC5 oscillator circuitry is quite simple. IC5a-IC5b is a low-frequency gated astable (about 6 Hz), with its output fed to the input of 1 kHz gated astable IC5c-IC5d. The output of the 1 kHz astable is fed to the input of VFET Q2 via volume control pot RV1, and Q2 uses the external 15R speaker as its drain load. In the absence of an input signal (from C2) the two astables and Q2 are cut off and the circuit consumes virtually zero standby current: in the presence of a high signal from C2 the low frequency astable activates and pulses the 1 kHz astable on and off at a 6 Hz rate, thereby producing a pulsed tone in the 15R speaker.

Note that the supply to the major sections of the Phone Bell Shifter circuitry is decoupled from speaker/Q2 transients by D4 and C6.

In this circuit, IC5 and Q2 consume virtually zero quiescent current, but the IC1-IC2 ring detector stages would, if continuously powered from a 9 V supply, consume a quiescent current of about 4 mA and would thus flatten a PP9 battery in less than two days of continuous running. To overcome this problem and vastly extend the battery life, we use a signal-sampling technique in which a micropower oscillator network (IC4-IC3-Q1) is used to feed pulses of supply power to the IC1-IC2 ring detector circuit and simultaneously check the output of the detector for signs of such a signal; if a signal is detected, the circuitry then applies full power to IC1-IC2, so that the signal can be fully processed. The sampling pulses are only 160 us wide and are repeated every 180 ms; consequently the mean power consumption of the detector circuit is reduced by a factor of 1125, to a mere 3.5 uA. The micropower oscillator network also consumes a running current of 3.5 uA so the total quiescent current of the entire Phone Bell Shifter circuit works out at 7 uA, thus giving up to two years of continuous running from a single PP9 supply battery.

IC4 is a special-purpose micropower oscillator circuit, designed around a 4007UB dual complementary pair plus inverter CMOS chip: a full description of this oscillator was given in the October edition of Designer's Notebook. The output pulses of IC4 are fed to one input of the IC3a-IC3b OR gate and then passed on to emitter follower Q1, which consequently connects supply power to the IC1-IC2 ring detector circuitry for 160 us once every 180 ms. Simultaneously, the output of IC2 is peak-detected (inspected) during the sampling period and the resulting signals (if any) are stored in C1 and fed to the second input of the IC3 OR gate. Consequently, if a bell-ring signal is absent during the sampling period, the output of IC2 will be low and zero voltage will appear across C1, so another sample pulse will be applied 180 ms later. If, on the other hand, a bell-ring signal is present during the sampling period, the output of IC2 will switch high and store a high voltage on C1, thereby taking the second input of the IC3 OR gate high. This causes Q1 to turn fully on, so that the supply is semi-permanently connected to IC1-IC2 and the input signals may be fully processed.

PARTS LIST

Resistors (all ¼ W, 5%)

R1	47k
R2	10k
R3,18	100k
R4	120R
R5,12,14,20	22k
R6	56k
R7,9,15	33k
R8,10	10M
R11,13,17	1M0
R16	220k
R19	68k

Potentiometers

RV1	22k logarithmic
PR1	2k2 miniature horizontal preset
PR2	22k miniature horizontal preset

Capacitors

C1,3,5	10n ceramic
C2	2u2 16 V tantalum
C4	100n ceramic
C6	100u 16 V PCB electrolytic

Semiconductors

IC1,2	CA3140
IC3	4001B
IC4	4007B
IC5	4011B
Q1	BC109
Q2	VN66AF
D1,2,3	1N4148
D4,5	1N4001

Miscellaneous

SW1	SPST miniature toggle
LS1	15R 2 W, 5" x 3" elliptical speaker
SK1,2	phono socket
TX1	PB-2720

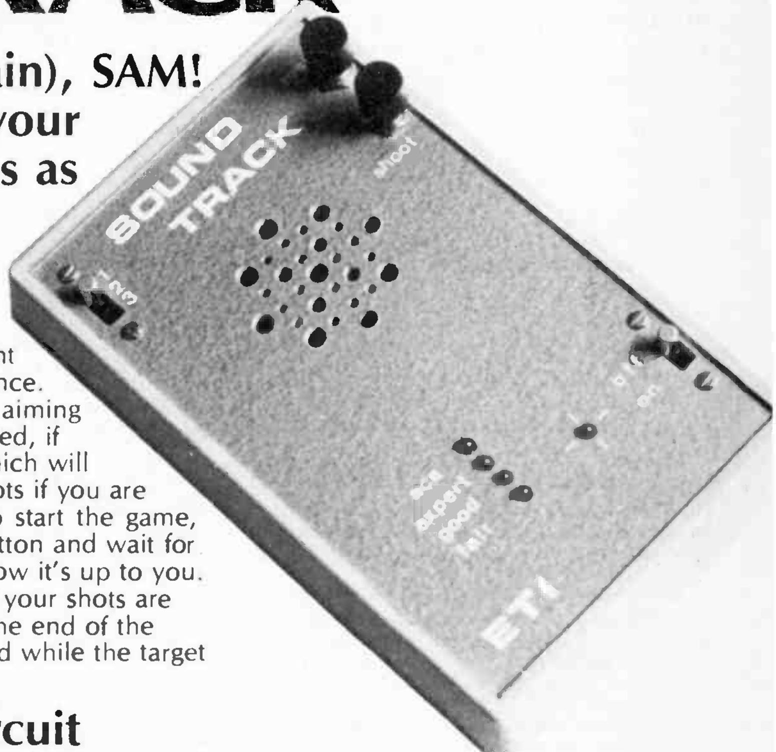
Case for electronics (Verobox order no. 202-21391A; case for speaker (order no. 202-21030K).

BUYLINES

LS1 might prove a little difficult to obtain from your local component emporium, but Watford Electronics have agreed to supply the speaker on request. The VMOS transistor (Q2) is available from Electrovalue, and Ambit International are stockists for the PB-2720. The PCB can be obtained from our PCB Service — see page 91 for details.

SOUND TRACK

Play it again (and again, and again), SAM! When you feel like working off your aggressions, try to zap the nasties as they fly past. Design and development by Phil Walker.



The ETI Sound Track is an 'arcade' game you can carry in your pocket. It requires no special displays as all the cues are sounds. The object of the game is to intercept all 15 of the attackers with your own armament. In order to do this you have to judge the best moment to fire from the simulated sound of the attacker. It is made more realistic by the fact that both volume and frequency changes due to Doppler shift are included. As the game progresses the speed of the attack increases to prevent you getting too used to one pace. Also there are three levels of skill which determine how difficult it is to hit the attackers at all.

At the end of the game, if enough of the attackers have been intercepted, an LED will light up to give an assessment of your performance. As an option, an aiming control can be fitted, if space permits, which will allow multiple shots if you are quick enough. To start the game, press the reset button and wait for the first attack. Now it's up to you. Bear in mind that your shots are effective only at the end of the shooting noise and while the target light is on.

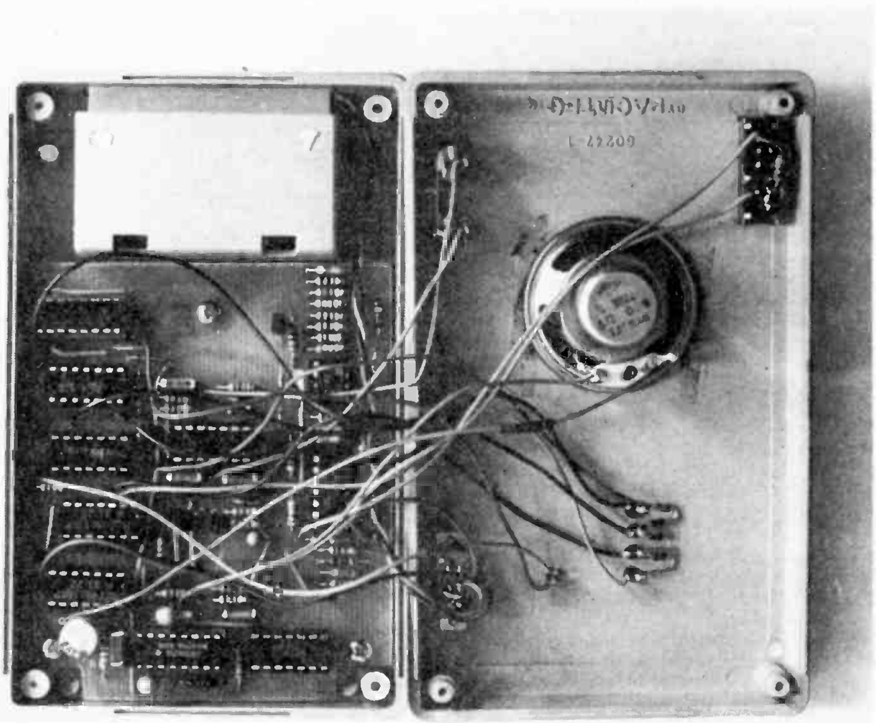
The Circuit

The circuit for this project uses standard op-amps, CMOS counters and gates and a special sound effects IC. This allows us to make fairly realistic sounds to simulate an object flying past, some sort of weapon being fired and an explosion if a successful interception has been made. In order to make the completed project hand-held, the PCB is fairly crowded but quite a lot has been put onto it.

The heart of the system is a voltage controlled oscillator operating at a frequency of less than 0.2 Hz. This provides two outputs; one is an asymmetrical triangle wave which controls the attack sound effect and simulates the position of the target while the other output is a logic signal to drive the score counters. The VCO frequency is modified by the attack counter such that the attacks proceed more rapidly as the game progresses.

The fire control section of the circuit produces two signals. The first of these is a long pulse which causes the shooting sound to be made by the sound generator. The second, immediately after, is a short pulse which enables the hit detector. If at the same time the ramp from the VCO is within the limits of the window discriminator in the hit detector, then a HIT will be registered and the HIT counter updated. At the same time the sound generator will be switched to provide an explosion effect.

The sound select logic and analogue control switching (in the absence of any other demand) will



assume an attack sequence and configure the sound generator to give a mixture of white noise and a tone. As the ramp voltage from the VCO falls, simulationg an attack, the volume will increase to a maximum and then decrease again. Simultaneously, as the volume reaches its peak, the pitch of the tone will decrease rapidly and stablise at a lower level to simulate Doppler shift. While the ramp voltage returns to its starting level the sound generator is inhibited.

If either shooting or explosion effects are demanded, these will take precedence over the attack sound. The explosion is produced by envelope-shaping the white noise source in the chip while the shooting sound is given by an audio frequency VCO, frequency-modulated by a much lower frequency triangle wave.

The display given by the LEDs is to give some indication of the number of successful interceptions made in a game. The first LED will light when eight out of the 15 attacks have been stopped. The next will light at 12, then 14, and finally 15. There is one other LED which flashes each time a HIT is possible, but note that the shoot button usually has to be pressed before it lights.

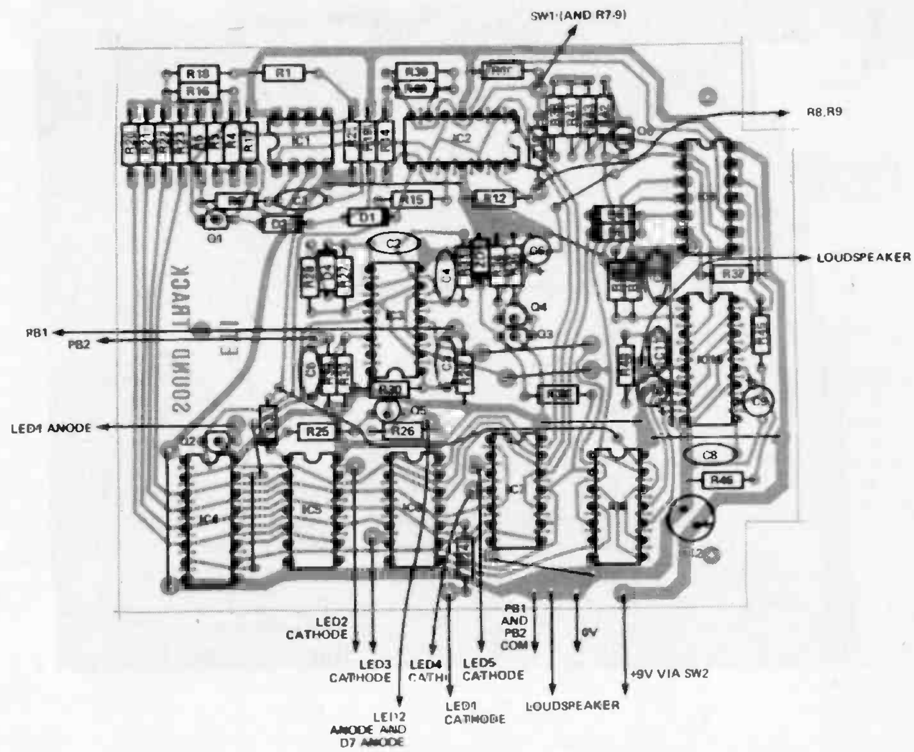


Fig. 1 Component overlay for the Sound track hand-held 'arcade' game. Note that some components are mounted off-board; see the photographs.

Construction

No major problems should be encountered in making this project; care must be taken when soldering the board as there are many places where tracks run between IC pins. Make sure that all the links are in place and that diodes, ICs and polarised capacitors are the right way round. Low profile IC sockets may be used but the case we used may then be a little tight.

SW1, R6 and R7 were mounted so that they fitted beside the battery compartment on one side while PB1 and PB2 went the other side. The LEDS are mounted on the front of the box so that they poke through the panel; use a little glue to hold them in place. Some interconnection work and components have to be put on to these (D7-10) and this should be kept as close tothe panel as possible. If there is room, fit RV1 and R7 but this will only be possible if a very small potentionmeter is available or a different box is used.

All interwiring should be carried out using thin flexible wire and kept as short as practicable. When fixing the loudspeaker check first that it will fit in the desired position and adjust fixing pillars etc. to ensure this. It is intended that it fits with part of the cone overlapping the battery compartment so a little shaving with a sharp knife may be required. When the speaker

PARTS LIST

Resistors (all ¼W, 5%)

R1,2,4,12,39	100k
R3,5,6,22,29, 30 R31,32,33,35	1M0
R7,15	220k
R8,46	6k8
R9,10,42	15k
R11,41	82k
R21,27	2M2
R14,23	470k
R16	180k
R17,19,37	27k
R18	150k
R20	4M7
R25	47k
R24,26,38	1k0
R28,48	10k
R34,36	22k
R40	3k3
R43	4k7
R44	1k5
R45	68k
R47	12k

R13 is not used

Potentiometer

RV1	100k linear

Capacitors

C1,8	100n polycarbonate
C2	220n polycarbonate
C3,5	4n7 ceramic
C4,11	1n0 ceramic
C6	2u2 16 V tantalum
C7	10n ceramic
C8	68n polycarbonate
C9,10	10u 10 V PCB electrolytic
C12	100u 10 V PCB electrolytic

Semiconductors

IC1	TL082
IC2	TL084
IC3	4093B
IC4	4520B
IC5	4012B
IC6	4023B
IC7,8	4011B
IC9	4066B
IC10	SN76495 (see Buylines)
Q1,2,3,4,6	BC182L
Q5	BC212L
D1-10	1N4148
ZD1	5V6 400 mW zener
LED1,3,4,5	Green miniature LED
LED2	Red miniature LED

Miscellaneous

SW1	miniature 3-position slide switch
PB1,2	miniature push-to-make push-button
SW2	miniature slide switch (on/off)

Case (Pac-tec type HP); PP3 battery and connector; knob for pot; loudspeaker (1" dia, 8 ohm) — see Buylines; PCB (see Buylines).

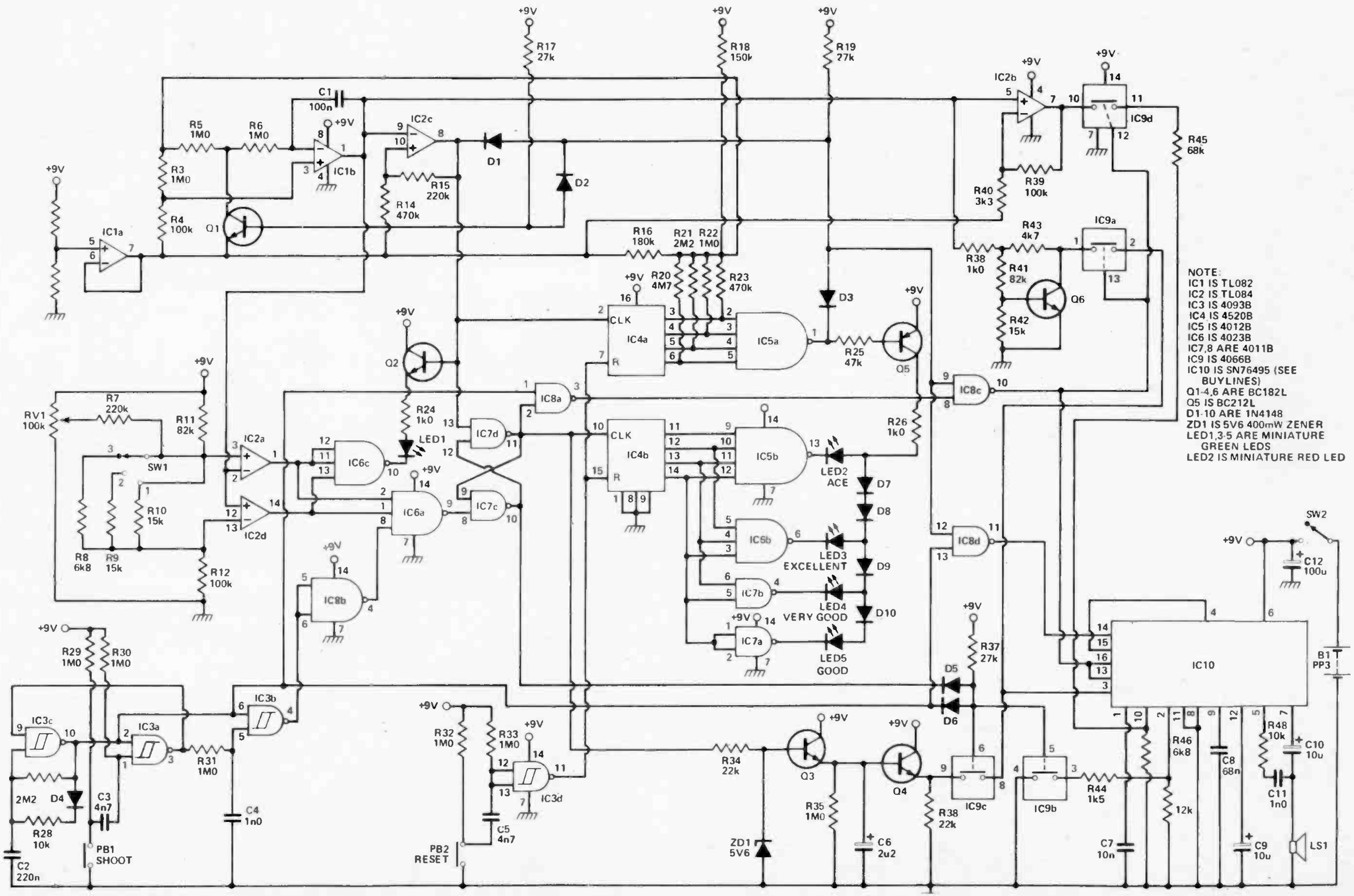


Fig. 2 Circuit diagram for the Sound Track.

position is known, drill a series of holes in the panel and glue it into position.

The wiring may now be completed and the box assembled to finish the project. Fit a PP3 battery to the connectors and it should be ready.

HOW IT WORKS

IC1b buffers the voltage at the junction of R1 and R2 to give a reference at half the supply. IC1a and IC2a form a very low frequency voltage controlled oscillator. R20-23 make a simple D-to-A converter which varies VCO frequency by a small amount as the game progresses. The timing for the whole game is derived from the VCO and provision is made by D1-3 to stop the circuit oscillating when the required 15 attacks have been counted by IC5a

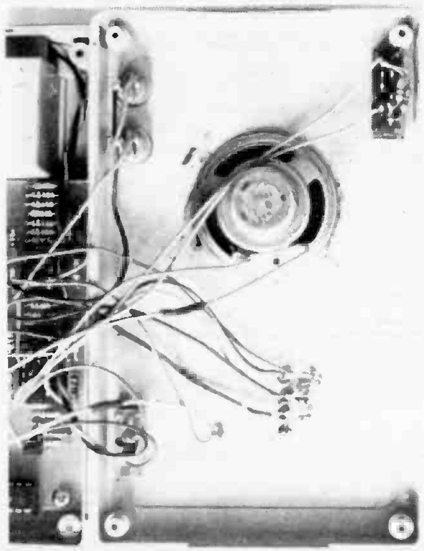
IC2a and IC2d form a window comparator whose position and width can be varied by RV1 and SW1. IC3a and IC3c are connected as a monostable and are triggered by PB1 being closed. C3 ensures that the period of the monostable is not affected by further closures of PB1. When the monostable time ends, IC3b is enabled for a short time determined by R31 and C4. This signal is inverted by IC8b and is applied with the outputs from the window comparator circuit to IC6a. If all the inputs to this IC are high at the same time this signifies a "HIT" and th eoutput of IC6a will go low. This action causes the latch formed by IC7c and IC7d to be set with IC7c output high. The resulting low on IC4b clock input increments that counter, increasing the score, while further counting on the same attack run is prevented by the latch action in IC7c/d. IC5b,6h,7b and 7d decode the outputs from IC4b to give a suitable display on the LEDs when Q5 is enabled by IC5a at the end of the game. IC3d is used to debounce the reset switch PB2 and the circuit at its input ensures that only a short pulse is available at its output.

The analogue control signals for the sound generator chip IC10 are produced in three parts and switched into circuit when required by IC9. The control signals for IC9 and IC10 are derived by IC8a, 8c,8d, and an AND gate made up of D6, D5, and R37.

The analogue control signals are produced individually. The Doppler style fall in frequency as each attack progresses is produced by IC2b. This device has a fairly high gain and at the start of the attack its output is driven to the positive rail by the ramp output from the VCO. As the ramp voltage falls past the reference voltage the output of IC2b will change from positive to negative quickly (but not instantaneously). If there is no other sound required at the time this output will modulate the oscillator in IC10 via R45.

Another effect required to simulate an object passing is that the noise produced by it will first increase and then decrease. This is accomplished by the circuit around Q6. At high voltages Q6 will be fully on and the output is low. At low voltages Q6 will be off but the output will again be low. As the voltage applied to the circuit increases, (until the voltage on the base of Q6 is sufficient to make it conduct) the output voltage will be the same as the input. When, however, Q6 starts to conduct, the junction of R38, R41, and R43 will stay at a constant potential. The reason for this is that as the input voltage rises, more current will flow into the circuit via R38. A small amount of this will go through R41 to drive Q6 further into conduction, drawing the rest out via R43. This action will continue until the voltage across Q6 is virtually zero again. The output from Q6 drives the volume control pin of IC10 via IC9a.

The last effect is of a decaying explosion. While IC10 will produce the noise of the explosion, the decay envelope has to be generated by Q3 and Q4. Most of the time the base of Q3 is held at 5V6 by the output of IC7d (part of the "HIT" latch). In the event of a "HIT" being registered, the base of Q3 will now be driven low. C6, which previously was held at about 5 V by Q3, will start to discharge via R35. The voltage on C6 is buffered by Q4 and fed to IC10 by IC9a. Also for the explosion effect, R44 is connected into circuit by IC9b. This changes the noise slightly to give a more realistic sound. C11 and R48 are included in the amplifier circuit feedback to give more prominence to the mid-frequencies and cut down on the hiss effect of the digital generation of the various noises.



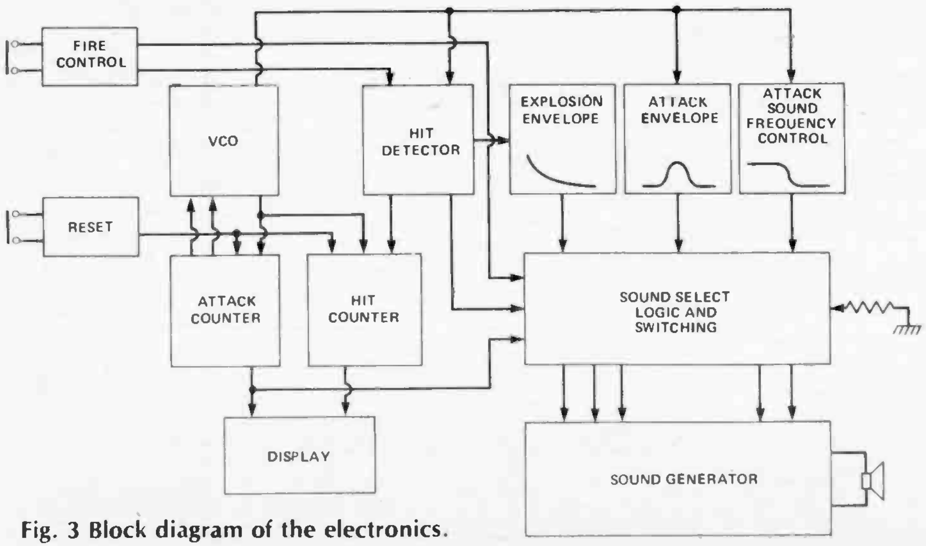
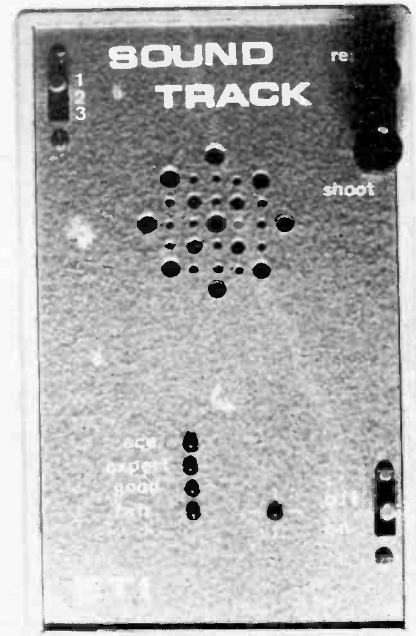


Fig. 3 Block diagram of the electronics.



The on-off switch, SW2, is mounted on the front panel at the bottom right-hand side such that it will be over R38, 41-43.

BUYLINES

Not too much here that's hard to find. The sound generator chip is one of the latest ones from the Texas Instruments range, so it should be available from TI stockists such as Technomatic and Watford Electronics. The Pac-tec case is available from Watford or direct from OK Machine and Tool Ltd, 22 Dutton Lane, Eastleigh, Hants. The speaker is type R3812 from Henry's Radio, while the PCB is available from us via our PCB Service (see page 91).

CAR ALARM

This cunning car alarm uses the battery earth strap as a sensor to detect when a courtesy light or other electrical load occurs if a thief enters the vehicle. Design and development by Phil Wait.

A significant proportion of cars are stolen at least once in their lifetime. The thieves are generally 'joyriders' who use them for a few hours and then abandon them after vandalising such items as wheels, seats, stereo/radios and so on. If you fit a good, reliable alarm you're bound to deter all but the most determined of criminals — who are usually professionals out to 'redo' the car or strip it completely for parts. There's almost nothing that will stop the latter type of thief — alarms, steering locks or any other deterrents notwithstanding.

Early car alarms were electromechanical by nature. They generally had a balanced cantilever or a pendulum with a switch contact attached. Any movement of the vehicle would close the contact and latch on a relay sounding the horn. Simple and effective — but prone to false triggering. They've all but disappeared. Others operated from a series of hood and door switches, but installation often proved a major undertaking.

Drop Detectors

Later alarms became more sophisticated — one type sensed the slight voltage-drop pulse that appears across the vehicle battery's terminals when a load is connected — such as a 'courtesy' light being operated when a door is opened. Reliability often proved a problem with these alarms as they depended on the internal resistance of the battery, which causes the voltage-drop pulse following the connection of a load (the battery terminal voltage drops momentarily and then rises again). Any variation in the terminal clamp resistance produces the same effect — giving rise to false triggering problems.

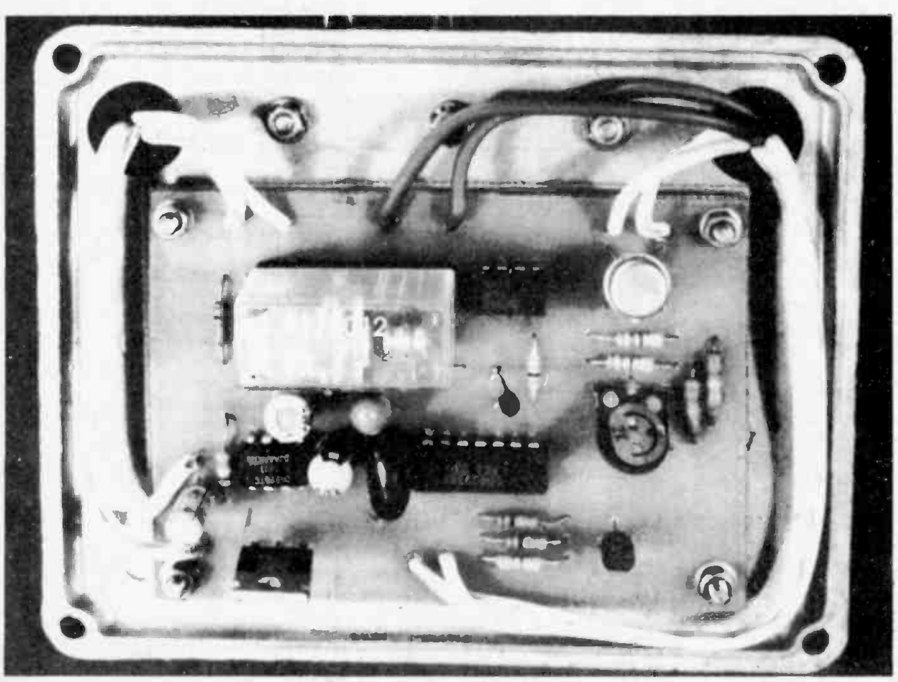
A cunning variation on this is to detect a voltage drop anywhere in the vehicle's electrical system. The battery 'earth' strap has a small, but finite, resistance. Any load on the battery will cause a current to flow through the earth strap (since the vehicle's chassis is used as the return circuit). The current causes a small voltage drop across the earth strap resistance. This is detected and used to trip the alarm circuit that sounds the car's horn. A thief entering the vehicle will inevitably operate a 'courtesy' light or something that draws current, thus tripping the alarm. As the 'sensing' input is essentially a very low impedance input, false triggering from magnetic induction or other sources is avoided. Other voltage drop sensing schemes essentially have a medium to high input impedance, hence their susceptibility to false triggering.

Pulsed Protection

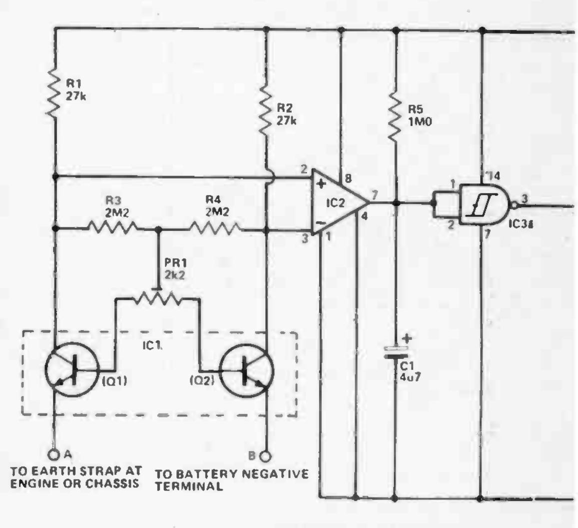
The sense and trigger circuit detects when the voltage drop across the battery earth strap rises above a predetermined amount. When triggered, this then arms the entry/exit delay. If the alarm does not remain triggered after the delay period nothing further will happen. If it does remain triggered, the delay circuit will trip the latch and start the alarm period timer. The alarm trip indicator will also light. When the alarm period timer is activated the relay driver is also activated. The relay pulser will then turn the relay on and off at one-second intervals, pulsing the horn on and off too.

The relay pulser circuit operates continuously and flashes a dashmounted LED to indicate that the alarm unit is 'armed'.

After the alarm period timer completes its period, the relay driver is turned off and the horn will cease pulsing on and off. If someone attempts to steal your car, trips the alarm and then abandons the attempt, the alarm trip indicator LED will remain on, telling you that the alarm was tripped in your absence.



(Left) The completed unit was mounted on the lid of a diecast box with a scrap of blank PCB substrate underneath as an insulating spacer. The external leads are passed through two grommetted holes to a terminal block on the outside.



Construction

Our prototype was constructed on a PCB; while this is not absolutely necessary — the project could be constructed on matrix board — a PCB does reduce the possibility of wiring errors which have to be sorted out when you first power up the project.

There is no particular order for assembling the components but it is usually easier to solder the resistors and capacitors in place first. Take care with the orientation of the tantalum and electrolytic capacitors. Follow with the semiconductors. Again, watch orientation of these components. The relay should be mounted last of all.

The completed board can be mounted in any convenient case — we housed ours in a diecast box measuring 120 x 40 x 95 mm. A diecast box was chosen because it can be effectively sealed against the ingress of dirt, moisture and other undesirable substances.

We mounted the PCB on the underside of the diecast box's lid and fitted a 10-way terminal block on the outside of the lid for all the external connections. Leads from the PCB to the 10-way terminal block are passed through grommetted holes.

Installation

First, mount the two LEDs on the dash in convenient positions where they can be seen from outside the vehicle. The alarm is switched on by a concealed switch which may be located under the dash or under the driver's seat. Alternatively, an externally mounted keyswitch may be used. If you install the latter, entry and exit delay may be reduced to about half a second by changing the value of C1 to 1uF.

We used a two-pole switch for SW1, one pole to switch the supply to the alarm, the other to short out the points when the alarm is switched on. Thus, if somebody does gain entry to the car and ignores the alarm or disconnects the horn, they will not be able to start the car even if they hot-wire the ignition!

Connection to the earth strap is quite straightforward. Take a wire from terminal A and solder it to the end of the earth strap. A wire from terminal B is soldered to the battery terminal connection. It's a good idea to keep these leads fairly short to reduce noise pick-up. Our were about 1 m long. Note that a poorly connecting or frayed earth strap may lead to damage of IC1.

The positive supply, via the alarm

HOW IT WORKS

The current in the earth strap is sensed by a pair of transistors connected in a common base configuration. These two transistors, Q1 and Q2, are encapsulated in an integrated circuit package (IC1) and are on a single chip of silicon, ensuring that they have very closely matched characteristics. The base-emitter voltages of each transistor will track within 50 uV of each other, a characteristic which is exploited here.

When no current is being drawn from the battery there will be no potential drop across the resistance of the battery earth strap (ignoring the miniscule current drawn by this alarm). Thus, the emitters of each transistor in IC1 will be at the same potential. As the base-emitter voltage of each is virtually identical the collector currents will be identical. Thus initially, the collector-emitter voltage of each transistor will be the same.

When current is drawn from the battery (when a courtesy lamp is operated, for example), a small voltage drop will appear across the battery earth strap. Thus, point A (emitter of Q1) will be raised to a *higher* potential than point B (emitter of Q2). That is, point A will be more positive than point B. The voltage on the collector of Q1 will thus rise (a common base amplifier is a *non-inverting* amplifier).

The voltage on the collector of each transistor in IC1 (Q1 and Q2) is initially set by a preset, which varies the current fed to each base. This compensates for any slight mismatch between Q1 and Q2 (the DC gain of this circuit is very high) and also acts as a 'sensitivity threshold' control by introducing an offset which must be overcome by a certain level of current through the battery earth strap before the alarm will trigger.

The voltage difference between the collectors of Q1 and Q2 is monitored by a differential input comparator (IC2). When the voltage on its non-inverting input exceeds the voltage on its inverting input the comparator's output switches high. IC2 has an open collector output requiring an external load resistor (R5). When the output of IC2 is low the timing capacitor, C1, is held discharged by IC2's output circuitry. When the alarm is tripped and the output of IC2 goes high C1 starts to charge through R5. After a time determined by the time constant of R5 and C1 the Schmitt gate IC3a toggles over and its output, pin 3, goes low.

The Schmitt gates IC3b and IC3c form a latch circuit. On power up, the latch is automatically reset by R6 and C2 placing a momentary low on pin 8. The output of IC3c is high and the output of IC3b is low, Q3 is turned off and LED1 is not lit.

When the output of IC3a goes low the latch toggles over. The output of IC3b goes high, turning on Q3 and lighting LED1. The output of IC3c goes low at the same time. The latch remains in this state until it is reset when the power is turned off and then on again.

Before the alarm is triggered the output of IC3c is high and the input of IC3d is held high by R9. As IC3d is wired as an inverter its output is low and Q4 is turned off. When the alarm is triggered the output of IC3c goes low, and since C3 is discharged, the input of IC3d is pulled low, its output goes high and Q4 is switched on, allowing the relay to operate.

The timing capacitor, C3, slowly charges through R9 and, after a period determined by the time constant of C3/R9, IC3d switches over, turning Q4 off again and stopping the horn.

The relay RLA, and therefore the horn, is pulsed on and off about once per second during the horn timing period. IC5 is a 555 timer wired as a free-running oscillator. The frequency of oscillation is determined by the time constant of R12 and C5. As the 555 is capable of driving quite high currents it is connected directly to the relay, which is then switched by Q4. In other words, the 555 pulses the supply to the relay.

The output from the 555 is also used to pulse LED2 (mounted on the dash) as a warning to would-be car thieves and as an indication that the alarm is on.

A three-terminal regulator, IC4, drops the battery voltage down to 5 V to supply the sense and timing circuits. This protects against false triggering from battery voltage variations and also helps to remove noise from the supply.

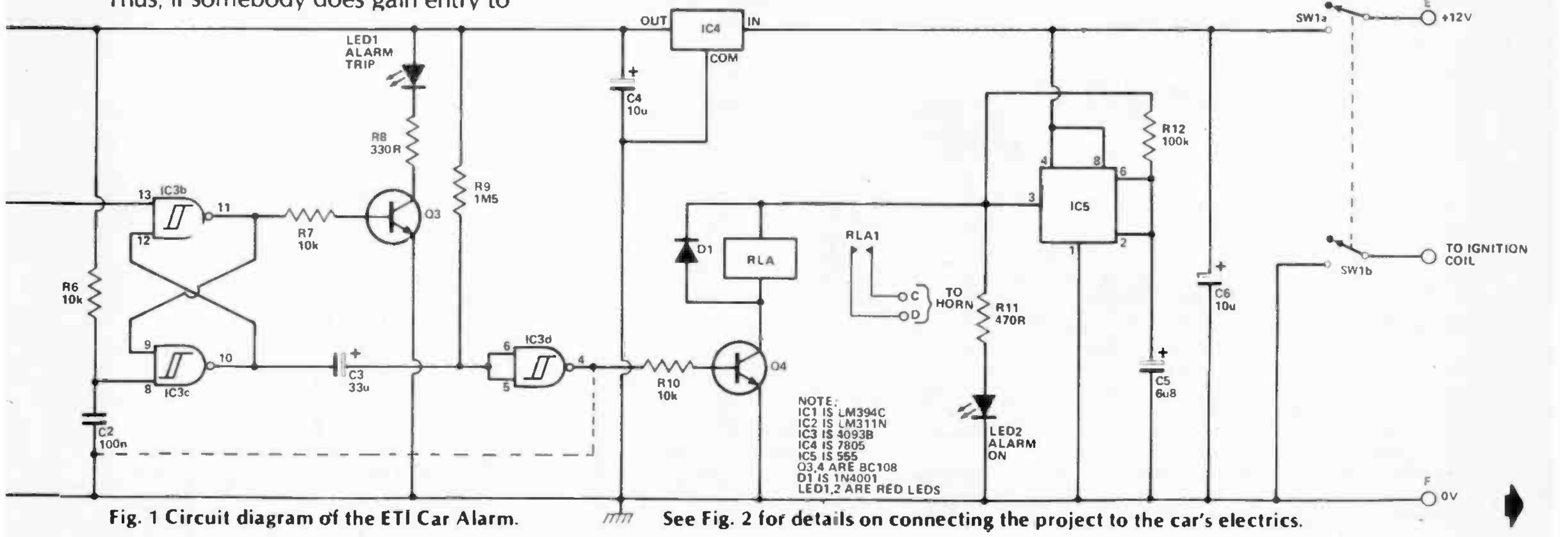


Fig. 1 Circuit diagram of the ETI Car Alarm. See Fig. 2 for details on connecting the project to the car's electrics.

switch, should be taken through an in-line fuse holder, directly from the battery positive terminal.

The output from the alarm is a pair of switched contacts, which operate the horn by bypassing the horn switch or, on some cars, the horn relay. We have shown two common horn circuits. In the first circuit the horn switch is bypassed by the relay contacts. The second circuit is a little more complex and requires an extra pole on the alarm switch. If you want to short out the points as well you will need a three-pole double-throw switch. Make sure you break the connection from the ignition switch to the horn as shown, or when you switch the alarm on you will also switch on the ignition!

Try to make all wiring as neat as possible and try to blend it in with the car's wiring so it is not obvious to a thief what wire he has to pull out to stop the alarm.

Setting Up

When all the wiring is complete, all that remains is to set the sensitivity preset. Disable the entry and exit delay by removing C1, or alternatively connect a high impedance voltmeter across C1. With no current being drawn from the battery, adjust PR1 until the alarm just fails to trip or C1 fails to start to charge. Note the position on the preset. Turn on the interior light and the alarm should trip. If it does not, check your first adjustment; if it is correct, you probably have the leads to the earth strap and the battery negative terminal swapped.

Turn the preset until the alarm just won't trip or C1 doesn't charge when the interior light is turned on. Note this position. The correct position for PR1 is midway between these limits, for reliable operation.

Next check that the alarm doesn't trip on the car radio or the electric clock. Some mechanical clocks are rewound by a motor every few hours, or even days, and these are often a cause of false triggering. If false triggering occurs from the radio or the clock, reduce the sensitivity. In some extreme cases it may be necessary to use a higher wattage interior light, though we found operation to be extremely reliable with a 5 W light, and there was sensitivity to spare!

BUYLINES

The only tricky components are IC1 and IC2 — these are both available from Technomatic. The relay is type RL111 from Watford Electronics. The case can be any diecast aluminium box about 120 mm x 95 mm x 40 mm; suitable ones are available from West Hyde Developments (order as EDD40) and Watford (DCB8). The PCB is available from our PCB Service (see page 91).

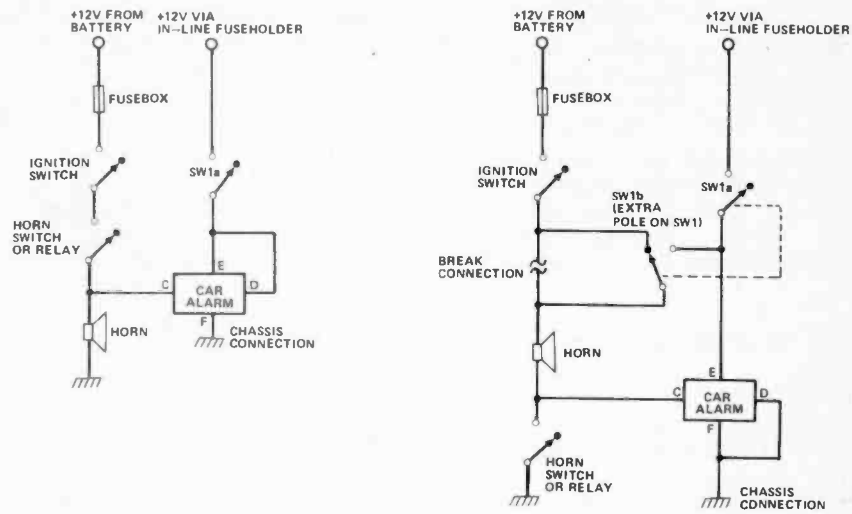


Fig. 2 Common horn circuits and how to connect the alarm to them. The circuit on the left will have only one wire from the horn. The one on the right is found in cars such as the Leyland Mini LS and is rather more complex.

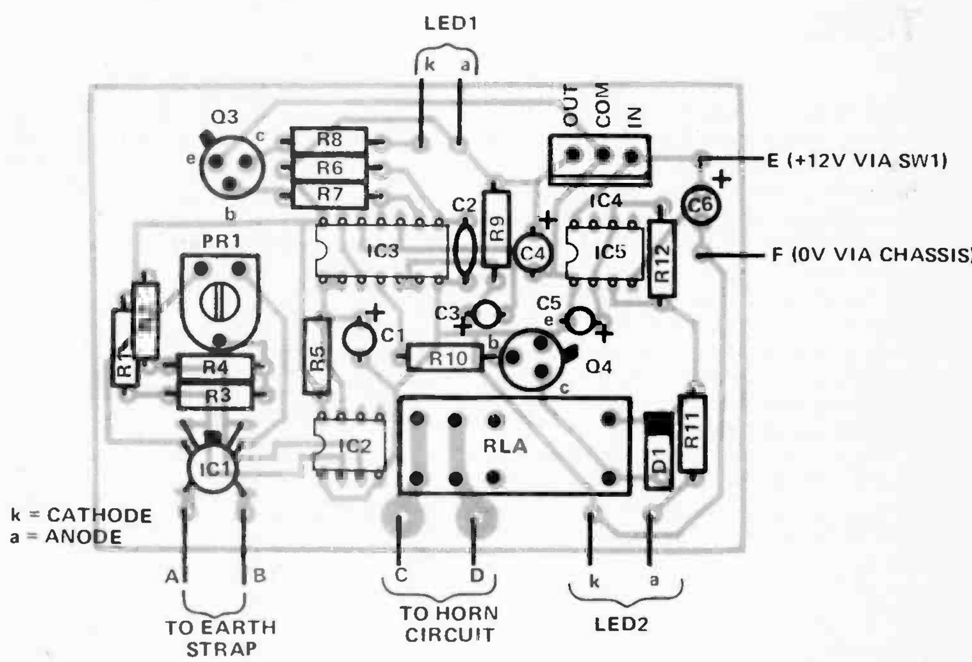


Fig. 3 Component overlay for the Car Alarm. There are a lot of polarised capacitors so make sure you get them all the right way round.

PARTS LIST

Resistors (all ¼ W, 5%)

R1,2	27k
R3,4	2M2
R5	1M0
R6,7,10	10k
R8	330R
R9	1M5
R11	470R
R12	100k

Potentiometer

PR1	2k2 miniature horizontal preset

Capacitors

C1	4u7 16 V tantalum
C2	100n polyester
C3	33u 16 V tantalum
C4,6	10u 16 V PCB electrolytic
C5	6u8 16 V tantalum

Semiconductors

IC1 (=Q1,2)	LM394C
IC2	LM311N
IC3	4093B
IC4	7805
IC5	555
Q3,4	BC108
D1	1N4001
LED1,2	0.2" red LED

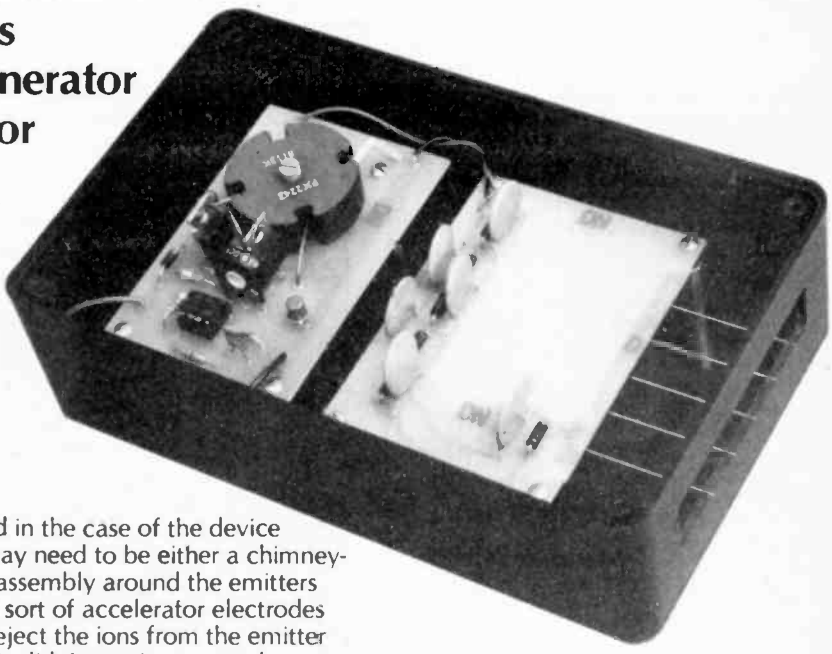
Miscellaneous

RLA	12 V DPCO PCB-mounting relay
SW1	DPST or 3PST toggle switch

PCB (see Buylines); case (see Buylines); 10-way terminal strip.

NEGATIVE ION GENERATOR

For readers who just have to find out for themselves what this subject is all about, this negative ion generator should provide a good basis for experiment. Design by Jonathan Scott. Development by Graeme Teesdale.



The rise in popularity of negative ion generators, the claims made for them, and the attention they have received in newspapers and magazines recently has undoubtedly intrigued many readers with a technical background or interest. As the electronics associated with a negative ion generator is relatively simple, generally employing readily available components, this article describes how to build a unit that can be used as the basis for experiment.

Design Of The Emitter Head

The object of the emitter head is to take in the HT, in our case about 3 kV, and produce a stream of negative ions flowing forwards into the room in which the generator is placed. The ions are produced by a very intense field gradient, which is induced by the high voltage and the geometry of the head assembly. This ion flow is a corona wind.

It is a basic principle of electrostatic physics that the field gradient is stronger in the immediate vicinity of a point projection, the gradient being greater when the point is sharper so most ion generators employ some combination of sharp projections and high voltage.

Pointing the Way

If the points are spaced well away from other parts of the unit the ions will naturally repel themselves away from the region of emission. However, if the point or points are partially enclosed in the case of the device there may need to be either a chimney-shaped assembly around the emitters or some sort of accelerator electrodes to help eject the ions from the emitter head. We didn't require an accelerator as the points protrude beyond the slot in the case.

Wherever there is ion production there will be ozone production. Ozone is corrosive as well as a strong antibacterial agent, and is poisonous in sufficient concentration. In order to keep it to a minimum, as low a voltage as possible should be used. Our project has been designed to give the lowest voltage compatible with adequate ion production. The design should be such as not to allow any arcing or serious breakdown; this is really only likely if you try using an "accelerator", as there will be no metal in close proximity to the emitter otherwise.

The best metal for the points which is easily obtainable is steel, preferably stainless. This is hard enough to hold an edge and will resist the effects of cathode stripping. The latter is undesirable both because the fine point will be eroded away, and also because the heavy metal ions which are ejected are undesirable agents in the air we breathe (stick to getting your minerals from cornflakes).

There is no shock hazard as the unit is not mains powered and there is a very large series resistance between the points and the multiplier output. At most, there results something between a nip and a tickle if you touch the emitter points.

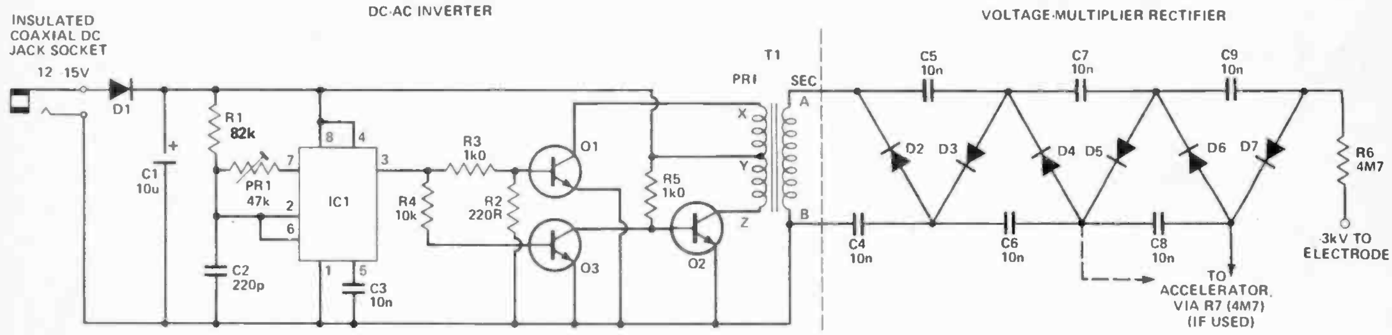
Construction

First stage of construction is to assemble the components on the PCBs; commence with the inverter board. Insert the resistors, capacitors, IC and transistors before asembling the transformer to it. As usual, take care with the orientation of the diode, IC1 and the transistors. Next, wind the transformer — details are given in the box. The transformer employs a potcore and this can be held on to the PCB with a nylon bolt — do not use a metal bolt. Cut the transformer coil wires to length, scrape off the insulation and solder them in place. The TIP31C transistors, Q1 and Q2, do not actually require any heatsink, though they do get warm in operation.

The high voltage board may be assembled next. Take care with the orientation of the diodes. Stand the capacitors erect on the board so that they do not touch each other or you may have arc-over problems between these components.

Mount the appropriate components on the 'blinker' board

Ion Generator



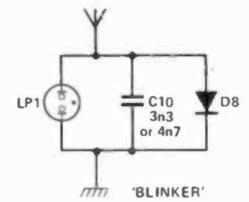


Fig. 1 Circuit diagram for all three sections of the negative ion generator. Current consumption of the unit is a little too high for use with ordinary batteries; we recommend you run it from a battery eliminator pack or a 12 V car battery.

PARTS LIST

Resistors (all ¼ W, 5%)

R1	82k
R2	220R
R3,5	1k0
R4	10k
R6,7	4M7

Potentiometer

PR1	47k miniature vertical preset

Capacitors

C1	10u 16 V tantalum
C2	220p ceramic
C3	10n polycarbonate
C4-9	10n 1 kV ceramic
C10	3n3 or 4n7 polystyrene or ceramic

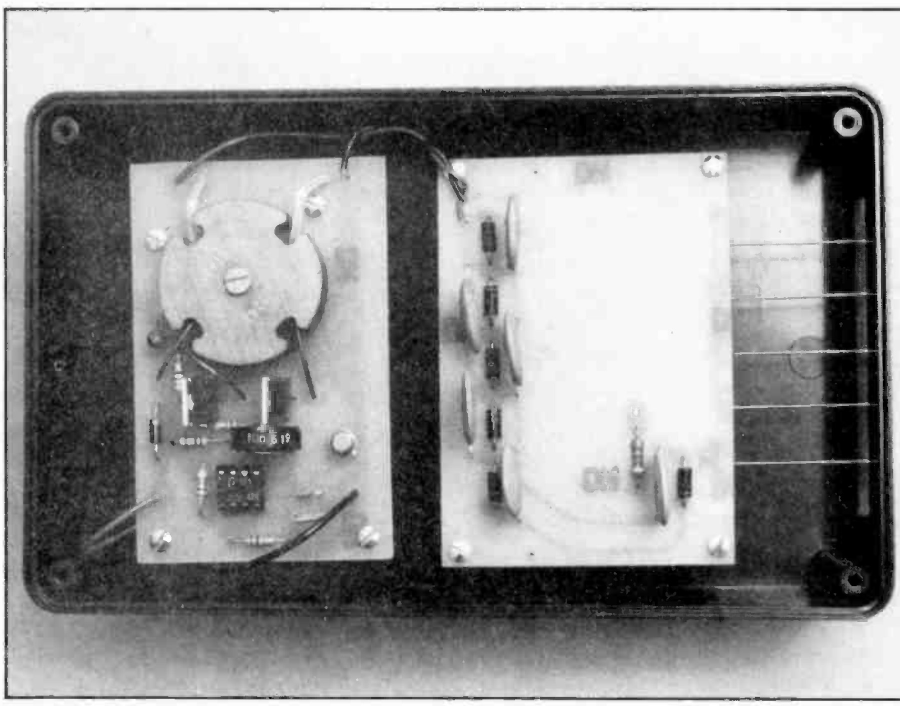
Semiconductors

IC1	555
Q1,2	TIP31C
Q3	BC107 or BC547
D1,8	1N4001
D2-7	BY208-800

Miscellaneous

LP1	wire-ended neon bulb (no series resistor)

PCBs (see Buylines); potcore and former (FX2242 — see Buylines); coaxial DC jack socket; case (180 x 110 x 55 mm, Vero ref. 75-2861D); five needles.



The inverter and high voltage boards mounted side by side in the case, with the emitter needles protruding through the slot in the side of the case.

next, as you'll need this for a testing aid. It is important to watch the diode polarity here. The cathode of the diode goes to the pad marked with the 'ground' symbol. Note that the components are mounted on the *copper* side.

The emitter points are steel needles soldered directly to the PCB. The easiest method is to tape the needles, parallel and the correct distance apart, so they overhang the end of a wooden block etc. Support the board underneath, and touching, the overhanging needles and solder them in place before removing the tape. Since ions will be ejected from any sharp point we recommend you cut all the component leads on the high voltage board and then resolder them, using enough extra solder to give rounded solder blobs. (Make sure the same is true of the needle connections). This will prevent unwanted ion leakage.

The DC input socket we mounted on one side of the box. Exactly how the DC coaxial jack socket is wired will depend on how your plug pack output plug is wired. Some have the outer connector connected to positive, while others have it connected to the negative.

Getting It Going

You will need a multimeter and a supply of between 9 V DC and 14 V DC. Switch the meter to the current range to read 300 mA full scale or more, and connect it in series with the DC supply input. Switch the supply on and, assuming all is well, adjust PR1 on the inverter board for *minimum* current. This could be between about 220-280 mA.

Run the unit for a few minutes, then switch off, discharge the rectifier capacitors and feel Q1 and Q2. One should not be markedly hotter than the other, otherwise you have adjusted PR1 incorrectly or you have a fault — most likely a transistor inserted incorrectly or a dry joint between the output of IC1 (pin 3) and the bases of Q1, Q2, or Q3.

Having confirmed everything

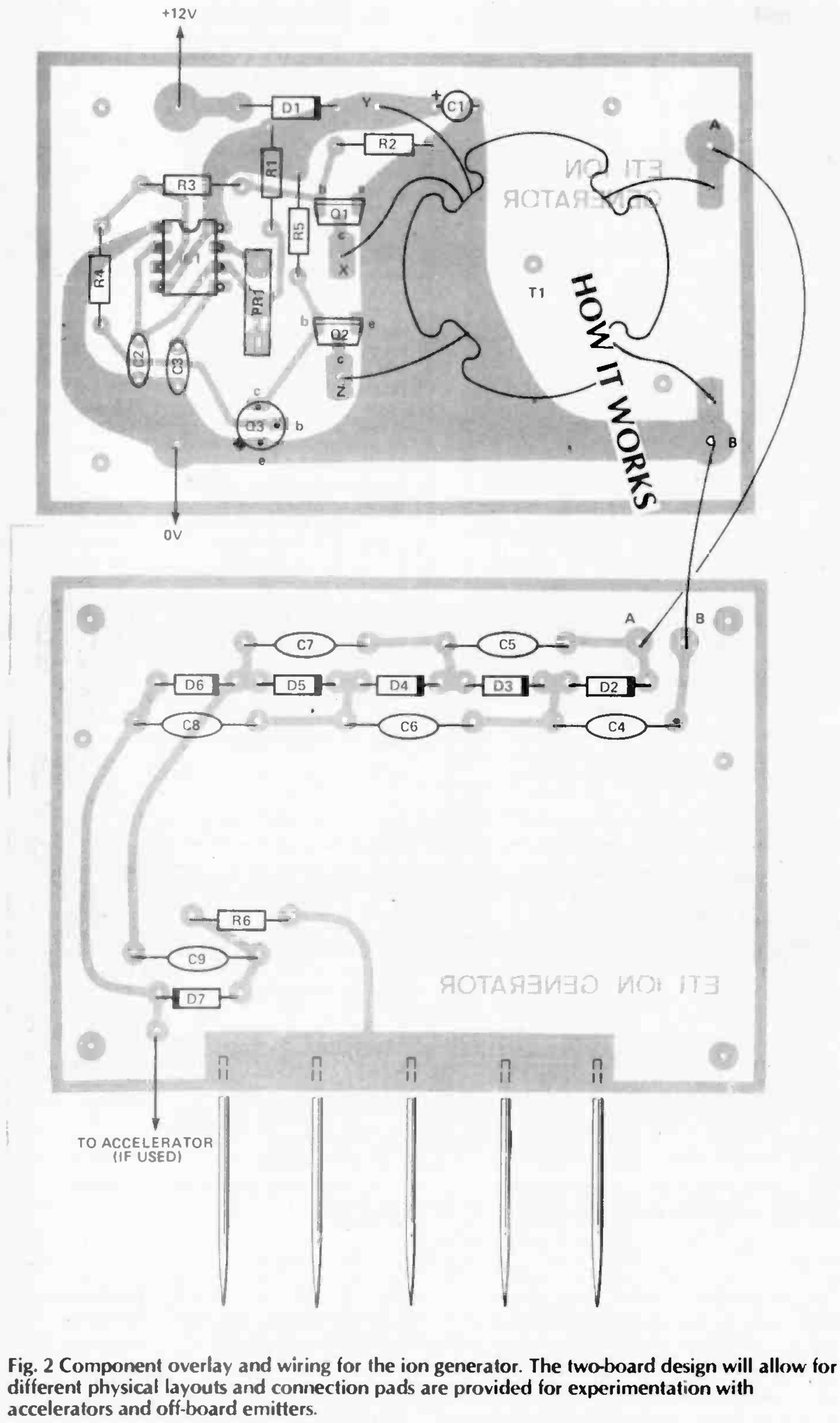


Fig. 2 Component overlay and wiring for the ion generator. The two-board design will allow for different physical layouts and connection pads are provided for experimentation with accelerators and off-board emitters.

HOW IT WORKS

The DC-to-AC inverter consists of a 555 astable multivibrator, the output of which is used to drive two transistors operated in push-pull. The collectors of Q1 and Q2 switch current through each side of the transformer (T1) primary in turn. Diode D1 prevents any damage from a supply connected with reverse polarity. Capacitor C1 is a bypass. IC1 oscillates at around 25 kHz, determined by R1 and C2. The exact frequency is unimportant. The mark-to-space ratio of the output of IC1 (at pin 3) may be adjusted by PR1, which is connected in series with pin 7 of IC1.

The output of IC1 drives the base of Q1 directly, via R3 and R2. Q1 turns on when the output of IC1 goes high. Resistor R3 is there principally to limit the base current supplied to Q1, while R2 serves to discharge the base emitter junction capacitance so that Q1 turns off quickly when the output of IC3 goes low.

When pin 3 of IC1 goes high, Q3 also turns on, preventing Q2 from turning on. When pin 3 of IC1 goes low, Q1 and Q3 turn off and Q2 will turn on as base bias will be supplied via R5. Thus current is alternatively switched through each side of the primary T1. The secondary provides a voltage step-up of 25:1. If the supply voltage is 12 V DC, then the peak-to-peak output from the secondary of T1 will be 600 V.

The voltage-multiplier rectifier employs the well-known Cockcroft-Walton circuit, where the output of successive half-wave rectifiers is connected in series with the previous one. This circuit provides a multiplication of six times. Thus, with a 12 V DC supply, the output will be about − 3.6 kV. With a 10 V DC supply (as can be obtained from a 9 V DC battery eliminator), about − 3 kV is obtained. An output for an 'accelerator' is provided.

The high voltage output to the emitter head is taken via a 4M7 resistor to ensure that only low short-circuit current occurs if the emitter head is accidentally contacted or excessively humid air causes 'flashover' from the emitter.

The blinker is simply a crude relaxation oscillator. When a charge builds up on the 'antenna' pad, it will charge C10. When the voltage on C10 reaches the breakdown voltage of LP1 (about 70 V), the neon will conduct. This will discharge the capacitor, the voltage across it falling until it reaches the extinguishing voltage of the neon (about 30-40 V), which will then cease conduction. While the neon conducts, it will emit light, but as it discharges C10 fairly rapidly, all you will see is a brief flash from the neon. Diode D8 ensures only negative charges operate the blinker.

When the neon ceases conduction, the charge on C10 will build up again and the whole process will be repeated.

works as it should, and having adjusted PR1, assemble it all into the case and you can check its operation with the blinker.

Turn the ioniser on and grasp the blinker so that your thumb is in good contact with the pad marked by the 'ground' symbol. Hold the blinker such that the 'antenna' pad is about 10 mm in front of the emitter. You should be able to count around one blink per second if all is well and this is a good 'benchmark' for successful operation when you experiment with different head designs and geometries.

Notes On Experimentation

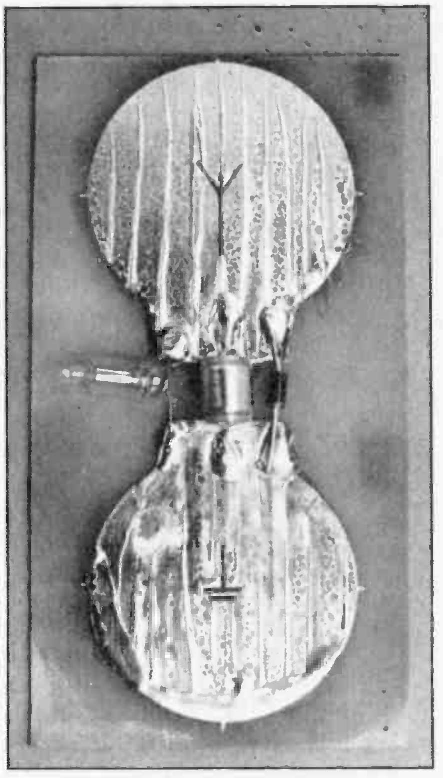
This project shows but one way to construct a negative ion generator and the electronics can readily serve as the basis for experimenting with different designs. Higher voltages are unnecessary — and are not usual in commercial designs — and can lead to problems with ozone generation, breakdown, etc. A connection is available on the high voltage board for supplying an 'accelerator' on an emitter head. It should be connected via a 4M7, ¼ W resistor. The accelerator voltage could be tapped off lower down the rectifier chain if desired — we suggest at the junction of C6 and C8.

The exact value of capacitors C4 to C9 on the high voltage board is not important and may be any value between about 1nF and 22nF or so, but should not be lower than 1nF. The voltage rating of these capacitors should not be less than 1000 V.

The DC supply should not be greater than 15 V or more turns be wound on the secondary of T1, else you may experience insulation breakdown within the transformer.

BUYLINES

The boards for this project are available from our PCB service — see p 91. Most other parts are readily obtainable except for the FX2242 for which you can try the B65611 K0000 R048 and B65612 A0000 T001 combination from Electrovalue Ltd, 28 St Judes Rd, Englefield Green, Egham, Surrey, TW20 0HB. BY208-800 are advertised by P.M. Components Ltd, Selectron House, Wrotham Road, Meopham, Kent, DA13 0QY.



The assembled blinker, which we tinned with solder to avoid sweaty finger marks and oxidation. The cathode of D8 is at the bottom.

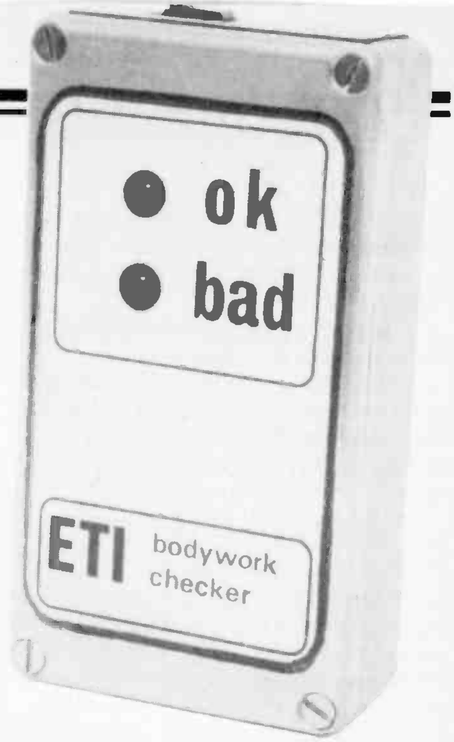
TRANSFORMER WINDING DETAILS

Potcore: FX2242
Secondary: 125 turns of 0.2 mm diameter enamelled copper wire.
Primary: 10 turns, centre-tapped, of 1.0 mm diameter enamelled copper wire.

The secondary is wound on the potcore bobbin first. Wind it in five or six neat layers. Slip thin plastic sleeving over the start and finish leads so that the sleeving is held well inside the bobbin. As you finish winding each layer, insulate it with 1 mm mylar sticky tape (if you can obtain it) or electrical insulation tape (a bit heavy, but it will do the job). Wind the next layer on the insulation of the previous layer, and so on until you finish winding. Wind several layers of insulation over the completed secondary. Leave the start and finish wires protruding from the different sides of the bobbin so that they exit via different slots of the assembled potcore.

Wind the primary over the secondary; it can be wound bifilar (two wires together, five turns, connect finish of one to start of other to provide centre tap) or in one winding — but don't forget the centre tap. Wind the primary so that its wires exit the potcore opposite the secondary wires.

In operation, if you have breakdown problems (arcing sounds inside the potcore) it means you have not wound or insulated your secondary carefully enough and you'll have to rewind the transformer.



BODYWORK CHECKER

Don't go out and buy a second-hand car without building this handy little gadget. It'll point out any problems under the paintwork. Design by Rory Holmes. Development by Tony Alston.

The purpose of this project is to help the selective second-hand car buyer detect the amount of body-filler used under well-disguised repair jobs. The unit gives a two-state indication of metal or plastic, ('OK' or 'BAD' respectively).

Our metal detector uses a capacitive sensing principle, which will detect the presence of *any* conductive object. Because of this the circuitry is much simpler and more reliable than metal detectors working on an inductive principle. It is also more suitable in this type of application where large areas of metal must be checked.

In use the device is switched on and lightly run over the car panels; if it runs over an area of body-filler the 'BAD' light will come on, otherwise it should read 'OK'.

Construction

The case is the most important part of this project as it is also part of the electronic sensing circuit. Take a careful look at the photographs of the finished project and you can clearly see the sensor area at the bottom rear of the case. First cut a rectangular hole (30 x 35 mm) about 8 mm from the bottom edge of the case and 14 mm from either side — make sure to clean off any burrs from the hole. A piece of single sided copper clad board (24 x 30 mm) is used for the sensor plate — this is centrally glued (copper side out) to a piece of plain paxolin or similar material (35 x 45 mm). This assembly is then glued to the case from the inside, so that the copper clad board will then be flush with case surface.

A small hole is drilled through to the copper side of the sensor plate and a short length of insulated wire, long enough to reach the main PCB, is soldered to the copper surface of the sensor plate.

The components can now be assembled and soldered to the main PCB as shown on the overlay diagram, making sure to correctly orientate D1,D2, IC1 and IC2 and the LEDs. Make sure to fit the link adjacent to IC1.

A short length of insulated wire is connected from the PCB to a solder tag fixed to the case — make sure this is a good connection as it forms part of the detecting circuit. The connecting lead from the sensor plate is soldered to the main PCB as indicated. A further insulated lead is taken from this same point on the PCB and held against the side of the case by a piece of insulating tape to form a capacitive trimming circuit (see photograph and refer to the setting up procedure). The LEDs are

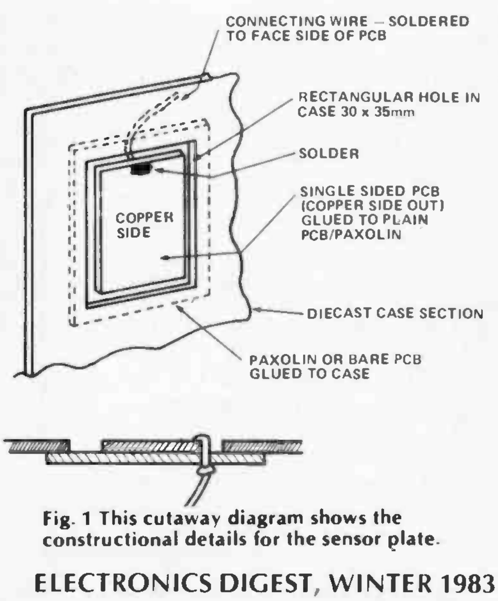


Fig. 1 This cutaway diagram shows the constructional details for the sensor plate.

With the protective felt peeled back to reveal the sensor, you can see how the fixing screws should be countersunk so they lie flush. In the internal shot (right), note how the trimming wire is taped to one side of the case.

directly mounted on the PCB and appropriate holes are drilled in the front case panel to allow these to pass through.

Finally, a piece of felt cut to size is then glued to the rear of the case, covering the sensor plate; this prevents the case from scratching the car bodywork and upsetting your friendly second-hand car dealer!

Setting Up

Setting up the circuit is straightforward; PR1 controls the detecting sensitivity and PR2 the metal/plastic switching threshold. When altering the presets bear in mind that replacing the case lid will slightly offset the adjustments, so replace the lid after each adjustment to check the effect.

Start with maximum sensitivity, ie set PR1 to its full resistance (anti-clockwise). Then place the case, sensor side down, onto a non-conductive object. With the lid off, PR2 can now be adjusted until the switching threshold is found. When the 'OK' LED is on, back off preset PR2 until it just extinguishes and the 'BAD' LED comes on (indicating no metal). The unit can now be placed against a metal surface and the 'OK' LED should re-light.

The trimming wire capacitively couples a small degree of HF voltage into the detector, effectively altering the switching threshold. Its effect can be varied by trimming the length. By experimenting with this if necessary, together with PR1 and PR2, a suitable switching action can easily be found.

Note that the human body is a fairly good conductor — you can prove this by holding your hand against the sensor, when the 'OK' LED should come on. This resulted in one member of staff wandering round the office, checking out the female employees and reassuring them that all was well!

HOW IT WORKS

CMOS inverter gates IC2a and IC2b form a high frequency oscillator of about 150 kHz. This signal is connected directly to the case, which in turn is capacitively coupled via the sensor to the high-impedance detector circuitry based around IC1. This unusual way of screening the circuit prevents the user's hand from affecting the capacitance between the detector input and the 0 V ground rail.

D1, D2, C1, and PR1 rectify the signal from the sensor and pass this voltage to the positive input of the op-amp, which is configured as a simple comparator. PR1 is used to set the input impedance and hence the sensitivity of the sensor. PR2 sets the switching threshold voltage on the non-inverting input to the comparator. When the coupling capacitance is increased, due to a conductive object lying across the case and sensor, the high frequency signal strength arriving at the detector will increase, raising the voltage on pin 3 of the comparator above the threshold, and switching the output from pin 6 fully positive.

IC2c,d are connected as a Schmitt trigger with R4 supplying positive feedback. This sharpens up the switching action coming from the comparator and further provides suitable drive signals for the two LEDs. These drive signals are buffered and current-limited by IC2e,f which power the LEDs. When metal is detected LED2 is lit and LED1 is off; the converse is true if metal is absent.

BUYLINES

All ICs and other components for this project are readily available. Most mail-order who advertise within these pages, eg Bi-Pak, will be able to supply all that is necessary. The PCB is available from our PCB Service as advertised on page 91.

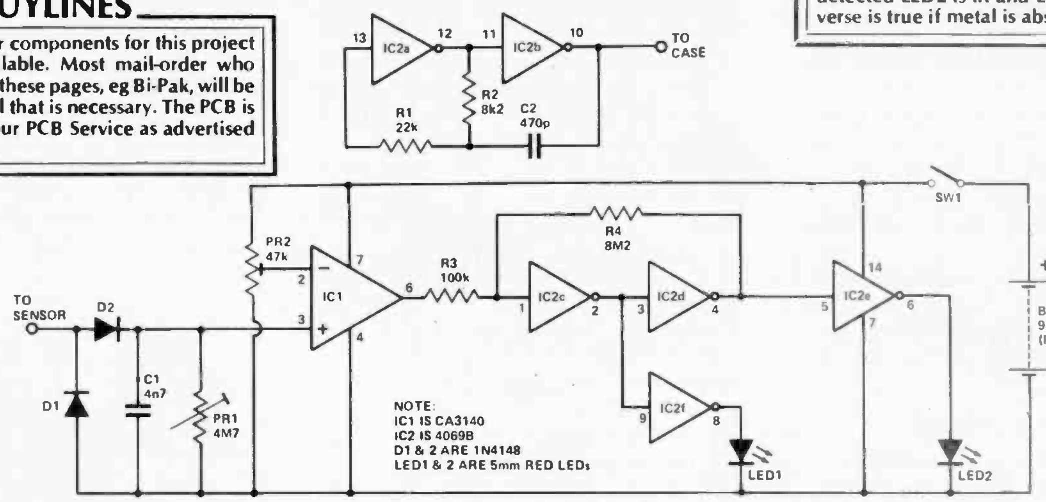


Fig. 2 Circuit diagram.

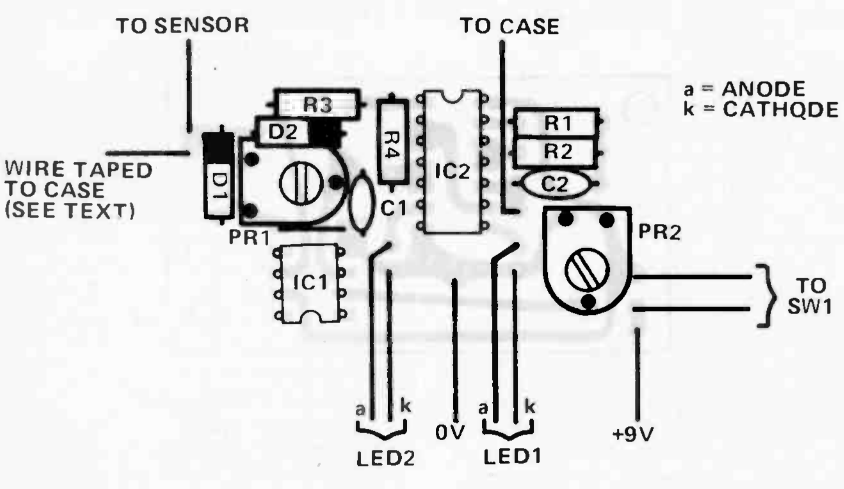


Fig. 3 Component overlay of the ETI Bodywork Checker.

PARTS LIST

Resistors (all ¼ W, 5%)

R1	22k
R2	8k2
R3	100k
R4	8M2

Potentiometers

PR1	4M7 miniature horizontal preset
PR2	47k miniature horizontal preset

Capacitors

C1	4n7 disc ceramic
C2	470p polystyrene

Semiconductors

IC1	CA3140
IC2	4069B
D1,2	1N4148
LED1,2	5 mm red LEDs

Miscellaneous

SW1	miniature rocker switch

Battery and clip (PP3); diecast case, approximate size 114 x 64 x 30 mm (RS 509-939 or similar — see Buylines).

PARKING METER TIMER

Living in fear of parking tickets is a thing of the past now that the ETI Parking Meter Timer is here. It'll remind you to get back to the car before the wardens do. Design and development by Rory Holmes.

After much design research ETI have come up with the best parking meter timer that has ever been offered to the home constructor. It is a truly functional design, and due to a careful selection of circuit components it is built to fit an extremely small and readily available case which can be clipped neatly on to a key ring. No awkward miniature switches are required; a single pin head touch switch allows the selection of all four time periods, including on-off control and an alarm test option.

The time periods are arranged in 20 minute increments, giving 20, 40, 60, or 80 minute options to suit modern parking meters. Three minutes are allowed for each 20 minute period for returning to the car, so giving more time the further away you go. The unit provides a high efficiency pulsed tone alarm using the Toko slim-line transducer.

By the use of only three CMOS chips and special power control circuitry, the current consumption when not in use is completely negligible.

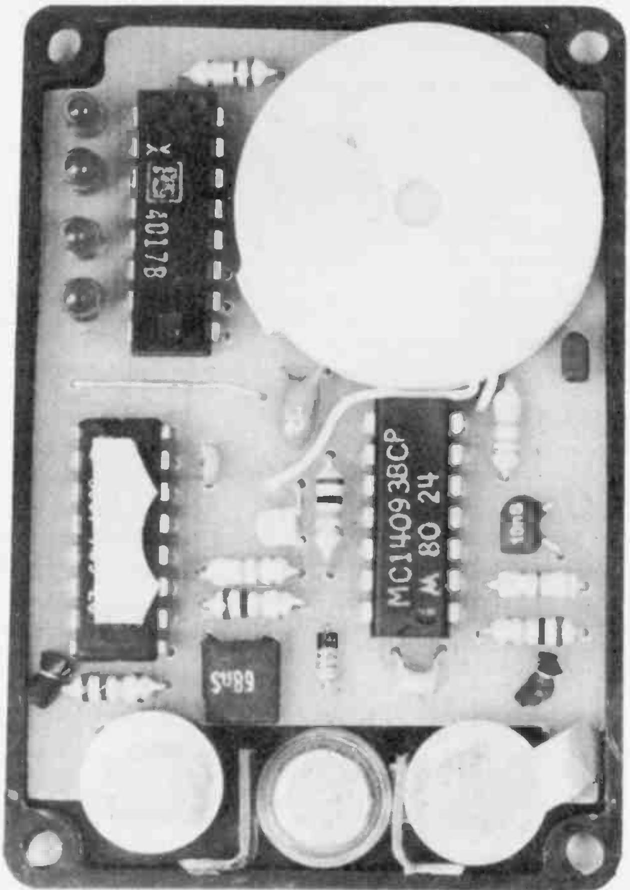
Power is provided by three ordinary hearing-aid batteries (available from most chemists). The LED display for indicating the time period selected also includes automatic blanking for further battery economy.

Construction

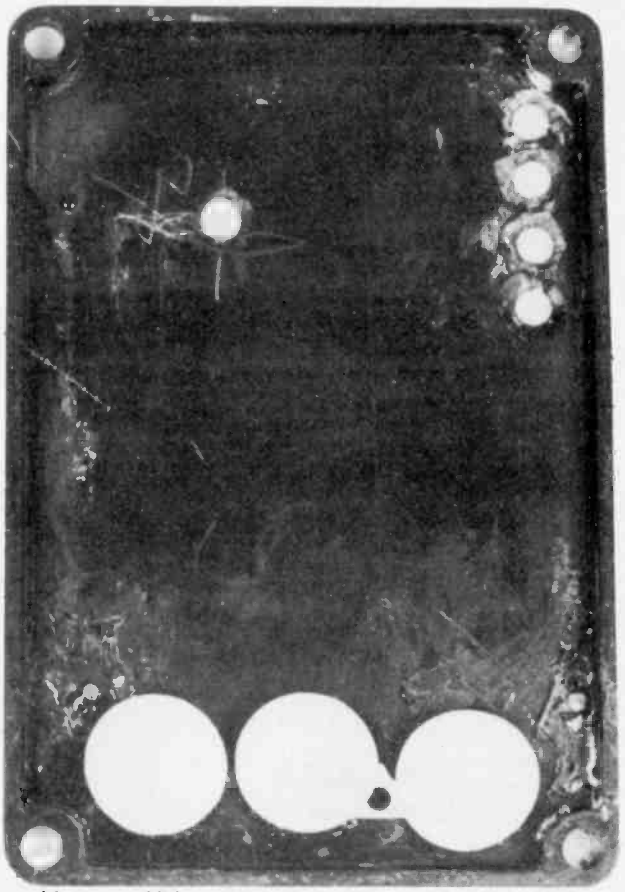
The Parking Meter Timer has been specifically designed to fit into a very small case. In fact it's not a standard case at all, but two lids from the smallest Vero potting box fitted back-to-back. We're trying to think of a project that uses two lidless boxes! The PCB is cut to fit inside the bottom lid with a cut-out to hold the three batteries in a line; see the overlay diagram and photographs.

The board should be assembled first, checking component orientation from the overlay and carefully observing the following points. When mounting the components on the

Parking Meter Timer



Overleaf: The parking Meter Timer being modelled by the lovely Sonia. Provided you're careful with the construction, everything can be fitted into the lid of a small Vero potting box . . .



. . . with a second lid completing 'case'. Countersink the holes for the LEDs as shown, so as to get a neat finish, and remember to insulate the battery sides from one another.

board it is essential to crop the leads very close to the tracks before soldering, to reduce the resulting solder lump. Apart from C3 and C5, all the passive components need to be mounted flat against the board. The LEDs (high efficiency 2 mm types) must have their leads bent apart slightly before they are pushed through the PCB holes. They are soldered in (observing polarity) with a 1 mm gap between the LED base and board. The piezo transducer is supplied with two mounting flanges; these need to be cut off before the transducer can be stuck to the PCB. This should only be done after the surrounding components are soldered in and with the PCB in the case, to ensure that the transducer clears the case rim. The two leads are then soldered to the board at the points marked on the overlay. Sockets must not be used on the ICs (there isn't enough room under the lid), but remember the static precautions for the CMOS chips.

Make Contact

The touch contacts are formed by two ordinary pins soldered vertically into the PCB at the points marked. The pin heads should be above the component side and at a height to just protrude through two holes drilled in the front lid. The timing resistor R2 should initially be left off the PCB (its exact value is set during testing). The batteries used are ordinary 1V4 hearing-aid types; they are held in place by the PCB cut-out and the two case halves. We used suitably bent pieces of copper contact strip soldered to the PCB for connecting our batteries so they could be removed. Two additional copper strips were glued to the inside of each lid, to connect the batteries in series when sandwiched by the lids. An alternative method is to solder the batteries permanently in circuit with thin wires. Remember that the sides of the batteries must not touch each other; spacers made of cardboard or plastic will keep them apart.

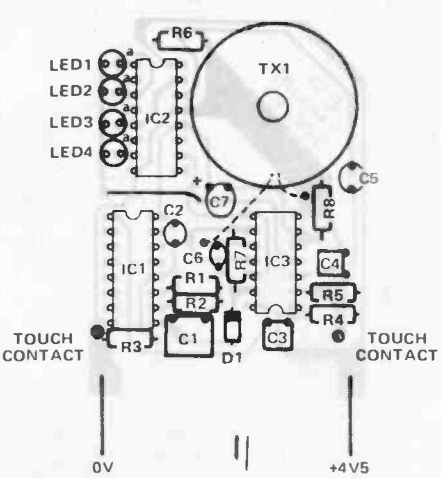


Fig. 1 Component overlay for the ETI Parking Meter Timer. You can't use sockets, so take care when soldering the ICs.

Testing

Initially a 2M0 pot should be wired with flying leads to the PCB connections for R2, and set at about 1M0 (half travel). 4V2 can now be applied to the circuit (at the supply pads by the battery cut-out), either

PARTS LIST

Resistors (all ¼ W, 5%)	
R1,8	4M7
R2*,3,4,7	1M0
R5	6M8
R6	100k
* See text	
Capacitors	
C1	68n ceramic
C2,3	2n2 ceramic
C4,5	10n ceramic
C6	470p ceramic
C7	3u3 16 V tantalum
Semiconductors	
IC1	4060B
IC2	4017B
IC3	4093B
D1	1N4148
LED1-4	2 mm red LEDs
Miscellaneous	
TX1	PB-2720

PCB (see Buylines); Veropins; three off 1V4 hearing-aid batteries; two off potting box lids (Vero order ref. 75-1413E).

from a variable supply, or by wiring three spare batteries together. The LEDs should now illuminate sequentially each time the touch pins are bridged by two fingers (remember they will turn off after about 20 seconds). On the fifth touch the alarm should sound, and on the sixth the unit is in the off state. The basic time period can now be set for exactly 17 minutes. Connect a voltmeter to the Q_9 output of IC1 (pin 13); after turning on to the first time period the meter will be at 0 V. It should switch to 4 V after 1 minute 4 seconds, and then back to zero after the same amount of time. By adjusting the pot and checking with a stopwatch, this time can be set exactly. The pot is then disconnected and its resistance measured; a standard resistor closest to this value can then be soldered into the PCB.

The case is assembled with adhesive pads stuck inside the lids to firmly sandwich the PCB and batteries. Holes must be drilled for the LEDs and we recommend that these are countersunk from the inside of the lid. A hole drilled directly above the transducer will increase the volume. 6 BA bolts about 5/16" long are ideal for securing the case halves; if one bolt at the transducer corner of the case is left out, the timer can then be fitted directly onto a key ring!

BUYLINES

Apart from the PCB, which is available from our PCB Service as usual (see page 91), should be no problems obtaining any of the parts for this project from the usual mail order sources.

HOW IT WORKS

The heart of this circuit is IC2, a 4017 decimal counter decoder; it is used for a number of different functions in a rather unusual way. Each of the decimal decoded outputs goes high in succession as the counter receives clock pulses on pin 14. The '6' output (pin 5) is wired to reset pin 15, thus the counter will have six different states and each clock pulse will shift it to the next state, counting from 0 to 5 and back to 0 again. The 5 output is not shown on the circuit diagram because it is not actually used; however, the divide-by-10 output on pin 12 (which is high for states 0 through 4 and low for states 5 to 9) is used, and in this configuration for IC2 it will only be low for state 5.

The counter is used both for counting the timing pulses generated by IC1, and selecting and setting the number of count periods before the alarm goes off. IC1 (CD4060B) provides the timing pulses for IC2; it consists of an internal RC-controlled oscillator followed by 14 stages of binary division. The oscillator is set to a frequency of 4 Hz by R2 and C1. When IC2 moves from state 5 to 0 (ie when switching on), IC1 is powered up from the divide-by-10 line of IC2 and it receives a positive reset pulse to pin 12 from C2 and R3. It then starts counting and provides a negative-going edge on the Q_{12} output every 17 minutes. This Q_{12} output is passed as a negative pulse through C3 and R4 to one input of the NAND gate IC3a; the output of IC3a goes briefly high and a slightly shorter positive pulse is applied through C5 to the clock input of IC2. Thus IC2 moves one count state every 17 minutes.

When IC2 reaches the fourth state the pin 10 output goes high and is used to enable the alarm. The alarm is a standard gated astable built around IC3a and b, driving an acoustic transducer. One input from each gate is wired to the '4' output of IC2. When this line goes high the oscillator is enabled and produces a tone pulsed at about 4 Hz. This modulation is derived from the master timing clock via D1, which switches the oscillator on and off. If left unattended it continues sounding for a further 17 minutes until IC2 receives another clock pulse from the timing generator IC1, at which point IC2 moves to state '5', turning off the alarm and powering down the rest of the circuit (the cycle can only be started again with a clock pulse from the touch switch).

The touch switch, built around IC3a, will also clock IC2 to select the number of 17 minute periods before the alarm sounds. IC3a is a NAND Schmitt trigger, with both its inputs normally held high by R4 and R5. Its output is thus low and C5 is discharged through R6. The pin 2 input is taken low when the touch pins are bridged by the skin resistance; this switching action is debounced by C4, and the output of IC3a will now go high, supplying a brief positive pulse through C5 to clock IC2. Thus the touch switch is used to turn the circuit on and initially set the number of time periods; thereafter clock pulses are supplied by the time generator.

IC2 also acts as the LED display driver for indicating the time period selected. The LED anodes are connected to the decoded ouput lines of IC2. When a state is selected via the touch switch, its output line goes high and applies a positive voltage to the corresponding LED. The cathode driving is arranged so they will only remain on for about half a minute after IC2 switches out of state '5'. As IC2 switches from 5 to 0 the divide-by-10 line goes from low to high and the positive end of C7 (which was previously discharged) is taken to the positive rail. This puts a high logic level on IC3d's inputs which drives its output low, thus providing a current-limited ground return for the LED which is selected. Meanwhile C7 is slowly charged up via R8 providing a long time constant of 20 seconds. When the voltage at the the negative end of C7 goes below IC3d's switching threshold, the output of the Schmitt gate will revert high again thus extinguishing the LED.

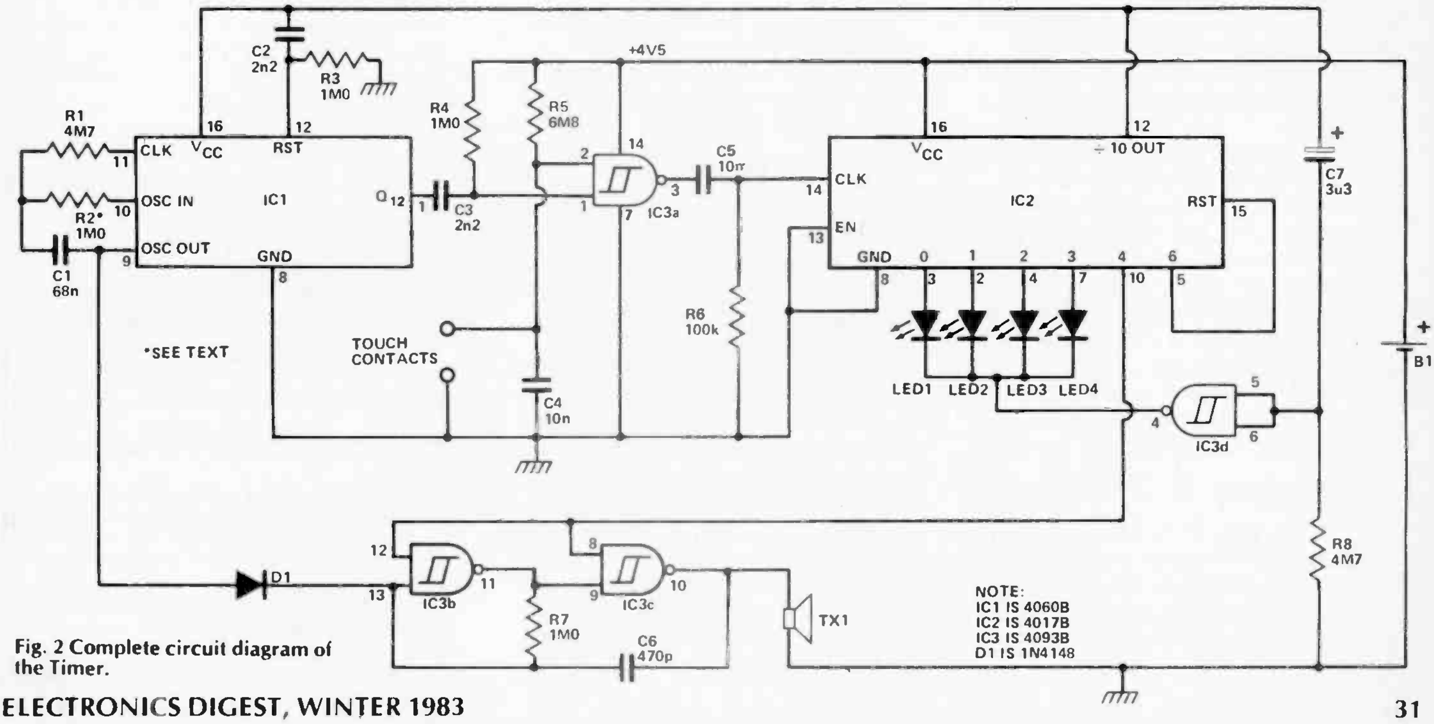
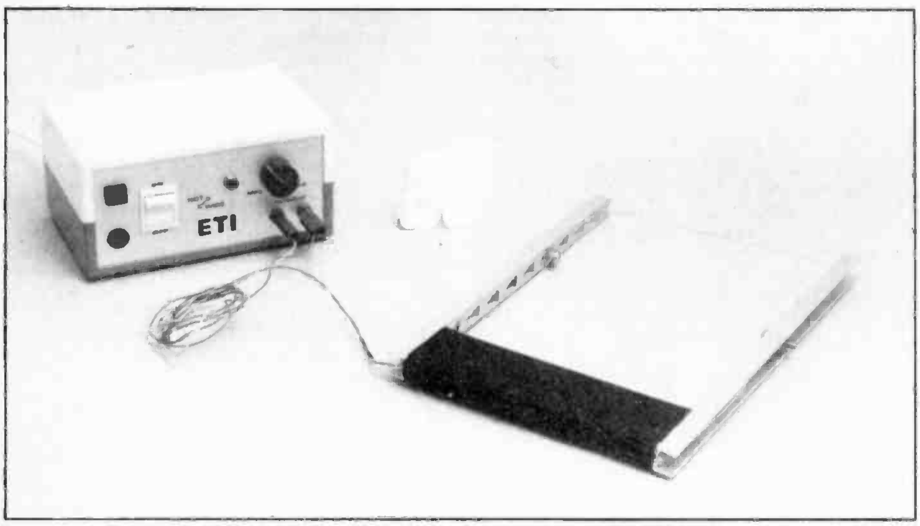


Fig. 2 Complete circuit diagram of the Timer.

POLYSTYRENE CUTTER

The ETI Hotwire is just the thing to get you going. No, it's not for stealing cars, it's for modelling. Turn that waste polystyrene packing into beautiful models with the Hotwire and some imagination. Design and development by Phil Walker.



This easy-to-construct project is a controller for a hot-wire polystyrene cutter. This method of cutting foam polystyrene is probably better than most others as it does not create any rough edges or crumbs; it actually works by melting the material as it comes into contact with the hot wire.

The object of the controller is to maintain the wire at a fairly constant temperature sufficient to melt the polystyrene foam quickly but without charring. This is accomplished by using a simple type of phase controller to regulate the power applied to the wire. The circuit employs a 747 dual op-amp, both parts of which are used as comparators. Speed of operation is not critical here as the circuit is operating at mains frequency (50 Hz).

Taking A Pulse

The first part of the circuit produces a 100 Hz pulse signal which synchronises the rest of the circuit to the output from the bridge rectifier. The second part generates a variable time delay which is used to regulate the amount of power developed in the cutting wire. The longer the time delay, the less power is developed and vice versa.

The control element used in this project is a thyristor as this will withstand the high peak currents in the circuit without the necessity for large drive currents.

Construction

This is fairly simple since most of the components are mounted on the PCB. Make sure that the diodes and IC are the right way round. Bolt the small heatsink to the rectifier bridge using some heatsink compound before mounting it on the board. Allow it to stand about 6 mm away from the board to avoid thermal stress effects. The thyristor is mounted on top of the larger heatsink, both being held by the same screw. Heat conductive paste should be used here as well. R9 will get quite hot in operation and should be stood away from the board if possible to allow air flow around it.

When mounting the PCB in the case, it is advisable to do so with the capacitor C1 at the bottom so that it is not heated by the other components.

Fairly thick wire should be used for connecting to the transformer and output sockets as they will be carrying several amps. RV1 is wired so that *minimum* resistance occurs at *clockwise* rotation.

Some Cutting Remarks

In our prototype the cutting head was made from two short pieces of slotted aluminium extrusion of the type sold for shelving systems. These were screwed to a piece of wood to form a handle while also insulating them from each other. The steel wire was clamped with some large nuts and bolts so that it was under some tension. The wires to the control unit were also clamped to the large bolts and held in place along the arms of the head with sticky tape.

It is recommended that the ceramic insulators sold by good electrical shops be used for the ends of the cutting wires in order to keep the metalwork isolated. Plastic connector block could be used but may melt under extreme circumstances.

Once everything is working correctly you can begin to exercise your creative talents on the nearest piece of polystyrene. Apart from a modelling tool, a gadget for 3-D doodling and something to keep the kids quiet during the summer holidays, you could use the Hotwire for cutting out large letters — ideal for advertising displays or exhibition stands.

BUYLINES

All of the parts for this project should be readily available from the usual outlets. The thyristor, SCR1, can be either a 2N4443 or a 2N4444 — the latter has a higher voltage rating and a higher price. The PCB can be obtained using the order form on page 91.

HOW IT WORKS

The 15 V AC from the transformer is rectified by BR1 to give a raw 100 Hz pulsating DC supply. C1 is charged to the peak voltage of this supply via D1 and provides the power for the circuitry. The raw DC supply is taken via R2 to IC1a where it is compared with the voltage across ZD1. The output from IC1a consists of a train of negative-going pulses which occur around the zero crossings of the AC input. These pulses are used to synchronise the variable time delay circuit by discharging C2 at the zero crossing of the AC input. The capacitor then charges at a rate set by R4 and RV1 until its voltage reaches the level set by R5 and R6. At this point the output of IC1b changes from its low to high state and switches SCR1 into conduction.

Once SCR1 has been switched on it causes the raw DC supply to be applied across R9 and the cutting wire until the voltage falls to zero at the end of the half cycle. At this time the thyristor turns off, the variable time delay circuit is reset and starts again. The proportion of the total time for which the output is on is determined by the time delay set by RV1; hence this controls the amount of power dissipated in R9 and the cutting wire. The main function of R9 is to reduce the peak surge current which would flow in the circuit, but it will also give some protection against inadvertent short circuit (the wire itself has a resistance of a couple of ohms). LED1 is incorporated to indicate when the output is operating and gives a visual indication of the power setting.

PARTS LIST

Resistors (all ¼ W, 5% unless stated otherwise)

R1	6k8
R2	3k3
R3,7	1k0
R4	5k6
R5	2k2
R6	15k
R8	1k8
R9	1R0 10 W wirewound
R10	180R 2 W wirewound

Potentiometer

RV1	100k linear

Capacitors

C1	1000u 25 V axial electrolytic
C2	68n ceramic

Semiconductors

IC1	747
D1	1N4002
D2	1N4148
BR1	6 A bridge rectifier, square package, 50 V or greater
ZD1	3V6 400 mW zener
SCR1	2N4443 (see Buylines)
LED1	5 mm red LED

Miscellaneous

FS1	20 mm 1A6 slow-blow fuse and holder
SW1	Double pole rectangular mains rocker switch
LP1	Mains panel-mounting neon indicator with integral resistor
T1	15 V 60 VA mains transformer

Heatsinks (finger-style TV21 for thyristor, TV4 for rectifier); PCB (see Buylines); case (Verobox 21039, 180 x 120 x 90 mm); panel mounting socket for LED1; two off 4 mm banana sockets, grommet, wire, nuts, bolts, brackets etc; 0.010" steel wire (guitar top 'E').

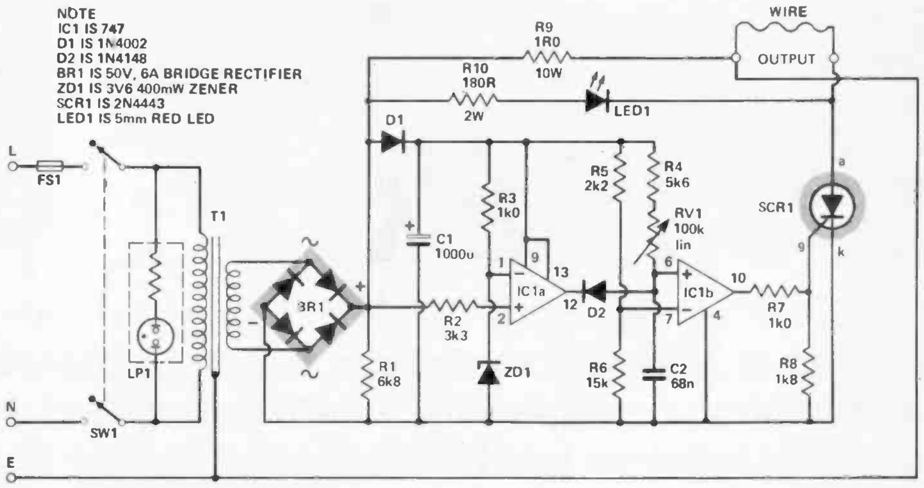
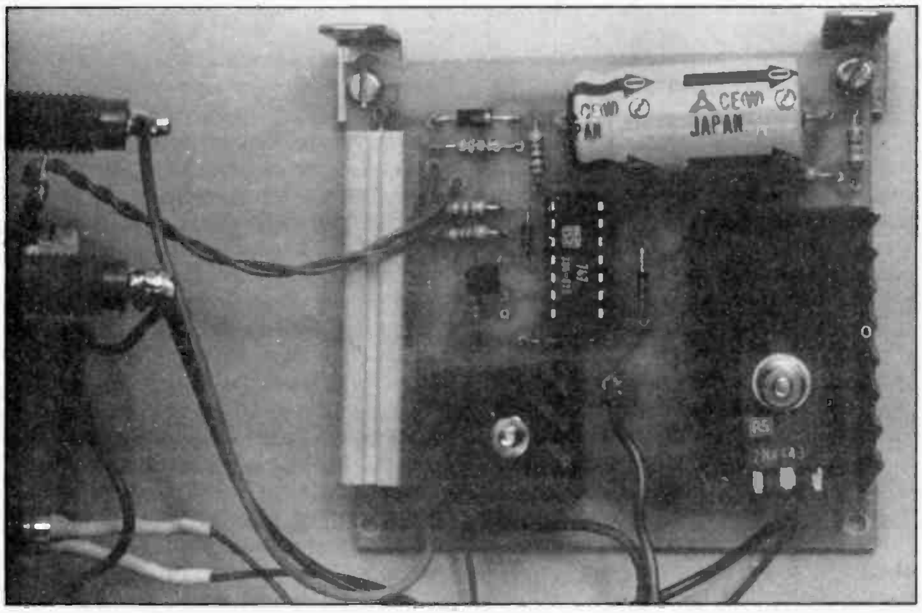


Fig. 1 Circuit diagram of the ETI Hotwire.



The Hotwire PCB. On the left you can see the earth connection to the pot case.

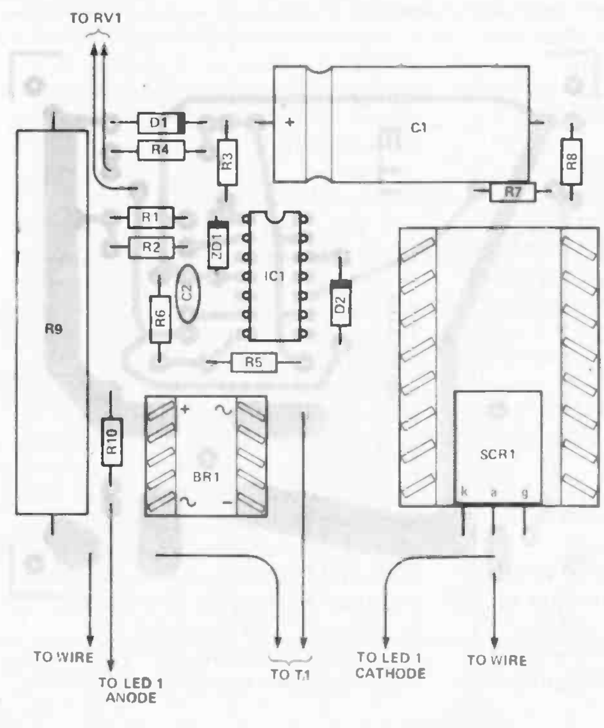


Fig. 2 Component overlay for the polystyrene cutter.

SOUND BENDER

This neat little ring modulator has a built-in wide-range sine/triangle modulation oscillator and a 'pan pot' output mixer, but can be built (less the case) for under £10. Design by Ray Marston. Development by Steve Ramsahadeo.

One of the most popular types of cheap sound-effects units is the so-called 'ring modulator' or four-quadrant multiplier. These units have two inputs, one being a voice or music audio signal and the other being a simple sine or triangle oscillator waveform: the output of the unit is equal to the product of the two instantaneous signal amplitudes. In other words, the oscillator effectively amplitude-modulates the voice/music signal, to give some very interesting changes in the apparent signal content of the original voice/music material.

The ETI Sound Bender is a fully self-contained version of the popular ring modulator circuit. Naturally, however, our project has few special features. First, it has a built-in modulation oscillator that can span the frequency range 3 Hz to 5 kHz using a single control pot and which can produce either sine or symmetrical-triangle output waveforms. Second, the actual ring modulator is based on a precision four-quadrant multiplier circuit that is integrated into the oscillator chip; the multiplier balance is externally adjustable, enabling the unit to be used either as a 'sound bender' or as a simple sine/triangle audio generator. Finally, the unit incorporates a two-channel audio mixer in its output stage, which enables the original and modulated audio signals to be mixed in any desired ratio (ranging from 'all original' to 'all modulated') by a single pan-pot type control.

Our unit is designed to operate from nominal audio input signal levels of about 100 mV RMS or greater and can simply be interposed between the output of the preamplifier and the input of the main amplifer of an existing audio system. The unit is battery powered by a stack of eight 1V5 cells and typically consumes about 12 mA.

Construction

The ETI Sound Bender is a fairly simple project and construction should present very few problems. Build up the PCB as shown by the overlay, noting the use of 16 Veropins to facilitate the circuit interwiring, then fit the PCB into a suitable case and complete the interwiring to the off-board components, noting that the two halves of RV4 are contra-connected. On our prototype unit the four control pots are fitted on the unit's front panel and the two switches and the input/output terminals are fitted on the

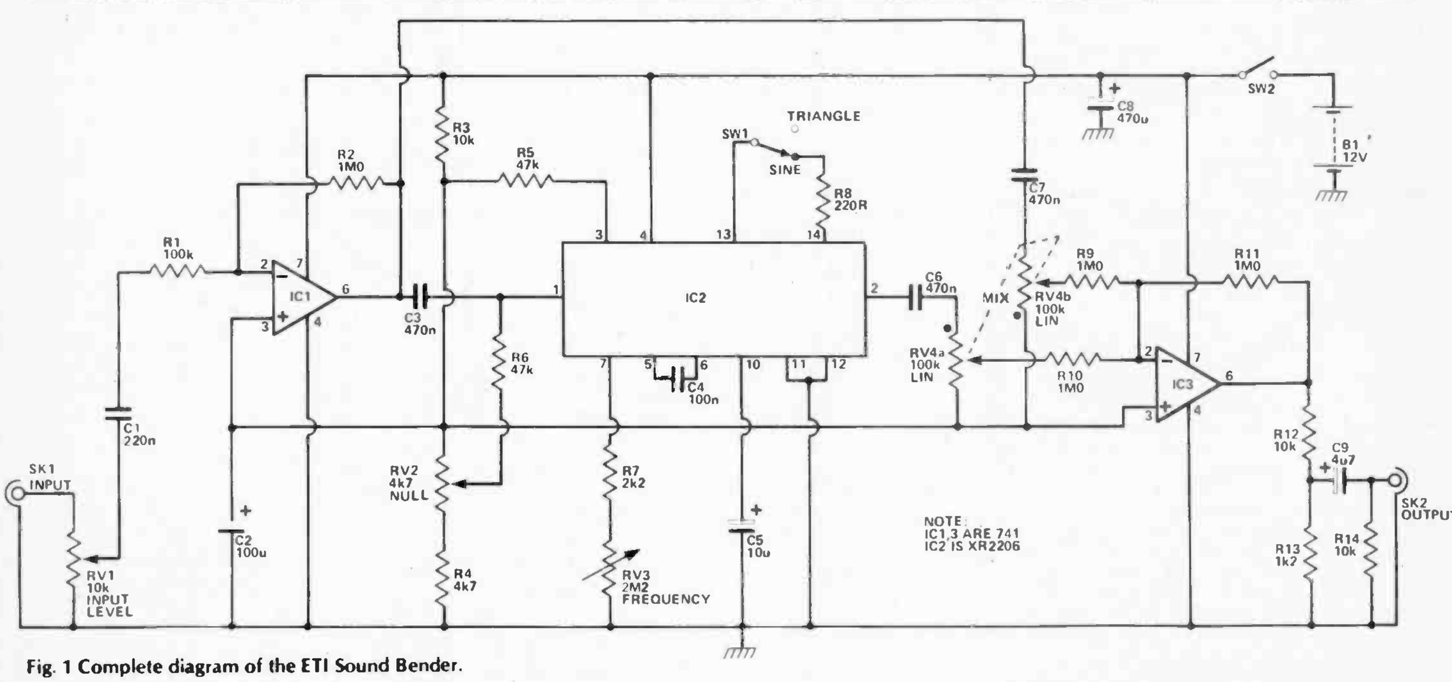


Fig. 1 Complete diagram of the ETI Sound Bender.

rear panel. As you can see from the photographs, the circuitry and battery pack make a fairly tight fit in the specified case.

The unit is very easy to use. Simply connect the output to an audio power amplifier/speaker combination, adjust RV2 (null) for zero output tone, then connect a voice or music input signal and see how the sound can be 'bent' using the frequency and mix controls. Level control RV1 is simply adjusted to give good sensitivity without amplitude limiting (clipping).

To use the unit as a simple audio generator, turn the input level control down and set the mix control (RV4) to give a 'modulation only' output, then adjust null control RV2 to give the desired output signal amplitude. RV3 then acts as the frequency control and SW1 gives selection of either sine or triangle output waveforms.

BUYLINES

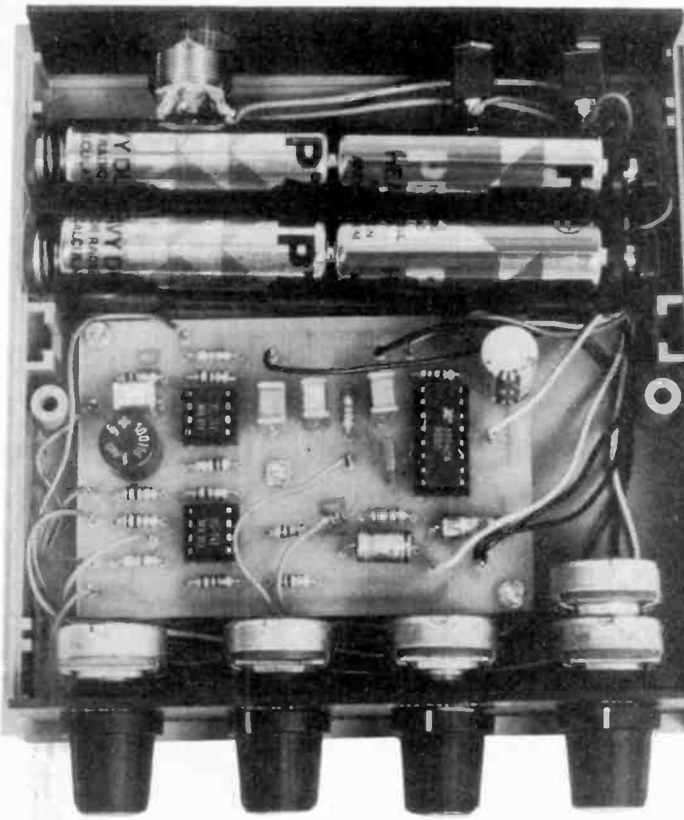
There are no unusual components used in this project — the XR2206 should be available from most major mail order companies advertising in this issue. The case used can be obtained from Watford Electronics or OK Machine and Tool Ltd — order as CM5-125. The PCB is available from our PCB service as advertised on page 91.

HOW IT WORKS

The heart of this unit is IC2, an XR2206 function generator chip that incorporates a wide-range sine/triangle waveform generator and a precision four-quadrant multiplier within a single package. The output of the waveform generator is internally connected to one input of the multiplier, and the other input of the multiplier is accessible at piin 1: the output is available at pin 2.

In our application, the generator can produce either sine or symmetrical-triangle waveforms, depending on the setting of SW1, and its frequency (determined by C4-R7-RV3) can be varied over the range 3 Hz to 5 kHz via RV3. The pin 1 input of the multiplier is biased by RV2, which is normally adjusted to balance the multiplier so that it produces zero output when zero signal input is applied to pin 1.

The audio input signal is applied across RV1 and a fraction of this signal is tapped off and applied to x 10 amplifier IC1. The output of IC1 splits into two paths, with one path passing to one input of two-channel audio mixer IC3 via RV4b, and with the other path passing to the input (pin 1) of IC2, which has its output (pin 2) taken to the other input of the IC3 mixer via RV4a. Note that mix controls RV4a and RV4b are contra-connected, so that they control the mixing action in 'pan pot' fashion, giving a final output from IC3 that ranges from 'all original signal' to 'all modulated signal' in the extreme settings of RV4. The output amplitude of IC3 is divided by 10 (by R12-R13), so that the final output signal has an amplitude roughly equal to that of the input signal feeding IC1, thereby giving the Sound Bender a good overall signal-to-noise ratio.



Everything does fit in the case specified, but only just!

PARTS LIST

Resistors (all ¼ W, 5%)	
R1	100k
R2,9,10,11	1M0
R3,12,14	10k
R4	4k7
R5,6	47k
R7	2k2
R8	220R
Potentiometers	
RV1	10k linear
RV2	4k7 linear
RV3	2M2 linear
RV4	100k dual linear
Capacitors	
C1	220n polycarbonate
C2	100u 16 V PCB electrolytic
C3,6,7	470n polycarbonate
C4	100n ceramic
C5	10u 25 V axial electrolytic
C8	470u 16 V PCB electrolytic
C9	4u7 16 V axial electrolytic
Semiconductors	
IC1,3	741
IC2	XR2206
Miscellaneous	
SW1,2	SPDT miniature toggle
SK1,2	phono sockets
PCB (see Buylines); four-section HP7 battery holders (two off).	

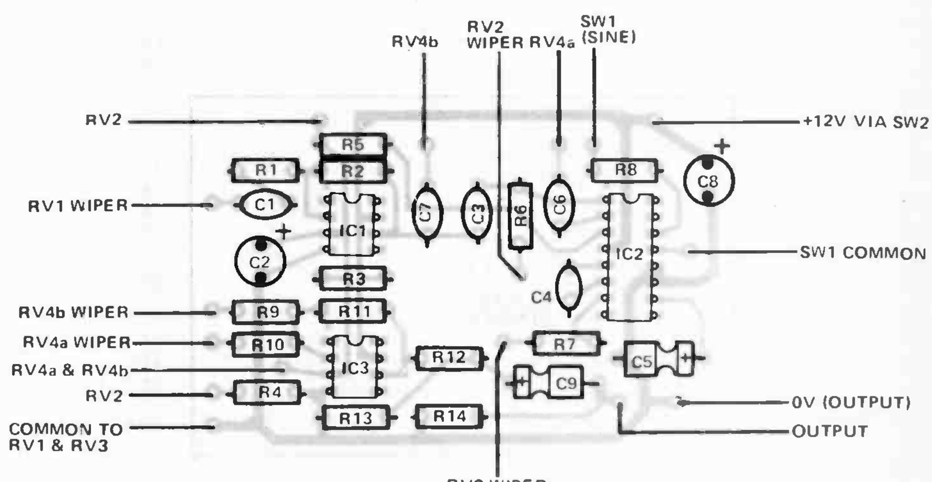
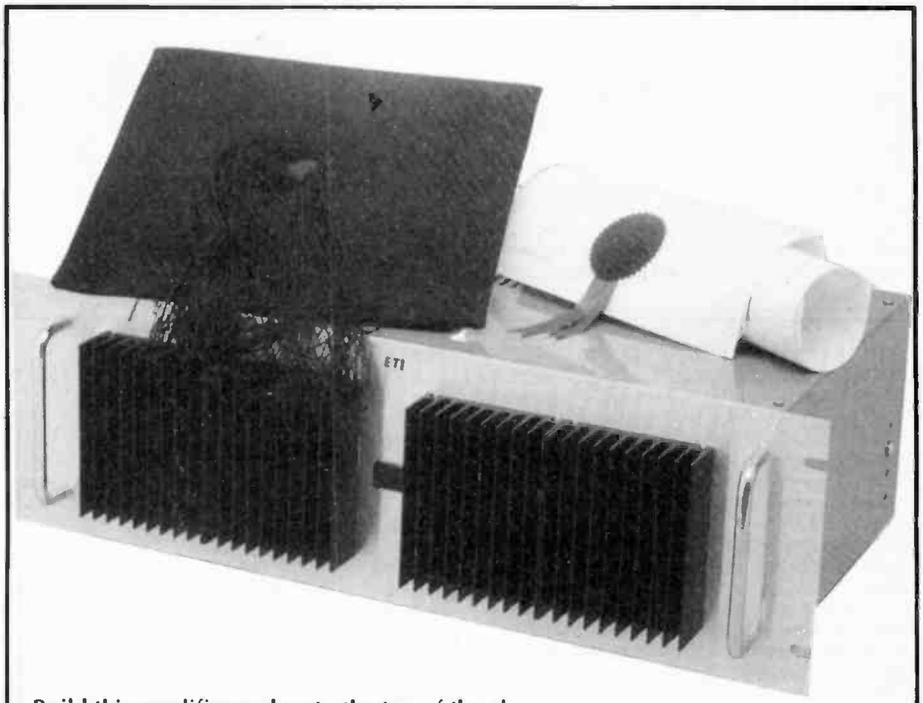


Fig. 2 Component overlay.

150 W MOSFET AMPLIFER

Employing MOSFETs, this power amplifier features a 'no compromise' design from Dave Tilbrook and is rated to deliver 150 W RMS maximum; it features extremely low harmonic, transient and intermodulation distortion. This is achieved by overcoming the many basic problems encountered in the use of MOSFETs for audio amplification.



Build this amplifier and go to the top of the class.

The objective of this project is to provide a power amplifier module of the highest possible performance. Ideally the power amp should produce an amplified version of its input signal and contribute no sound of its own. In order to design a practical amplifier that will come as close as possible to this ideal, it is necessary to 'define' limits on the input signal characteristic and then ensure that the power amp exceeds these limits.

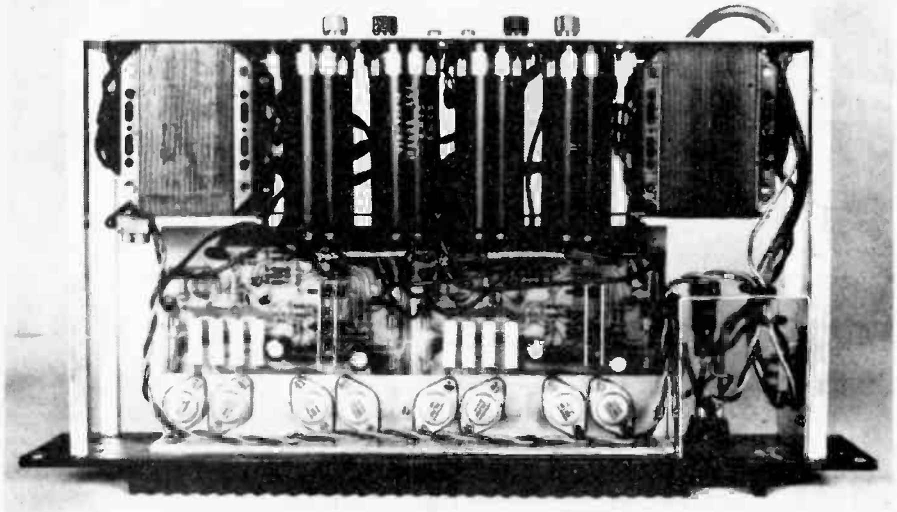
The problem of amplitude overload cannot be eliminated, since no practical power amplifier has access to infinite supply voltage. In order to overcome this problem, the ETI-5000 module has been designed to handle in excess of ± 50 V rails, giving it a conservative power rating of 100 W RMS into 8 ohms. The output stage has been designed so that the MOSFETs will not operate outside their safe operating area on any load in which the effective series resistance does not drop excessively below 8 ohms. Increasing the supply rails will increase the audio power output (up to 150 W RMS max.) but for normal use, we recommend sticking to ± 55 V.

Similarly, since no power amp has an infinite slew rate or infinite frequency response, the input signal has been limited by a passive input filter. It can be easily demonstrated by experiment that the introduction of a passive filter that does not excessively affect the frequency response within the audio passband will not affect the sound of the input signal. This filter will define a maximum possible input slope. It is therefore only necessary to design the amplifier with a slew rate that exceeds this by a sufficient margin to ensure freedom from slew-induced distortion. Since the amplifier is operated below its slew rate limit, the application of negative feedback will decrease distortion produced as a result of the signal slope approaching the slew rate (TIM).

Pair Difference

Differential pairs have been used throughout the design to form not only the input stage but also the voltage gain stage. This ensures that the distortion characteristics of the input and voltage gain stages are low enough so that the open loop characteristics of the amplifier will be determined by the output stage. The improved frequency and phase linearity of the differential pair make it considerably easier to ensure that the amplifier meets the Nyquist stability criterion. Another advantage of the differential pair is its relatively high supply rejection, a parameter which is often not given sufficient attention in power amp design.

Careful control of the feedback loop and the use of a passive filter/load on the output of the module, coupled with the design points mentioned above, have yielded an amplifier with particularly low dynamic distortion characteristics. An amplifier that has been designed with these objectives in mind will automatically have low THD and TID figures. The ETI-5000 is no exception, with a THD at 1 kHz and 10 W RMS of less than 0.001%, rising



A general view showing the internal layout of the amplifier. Toroidal transformers can be used instead.

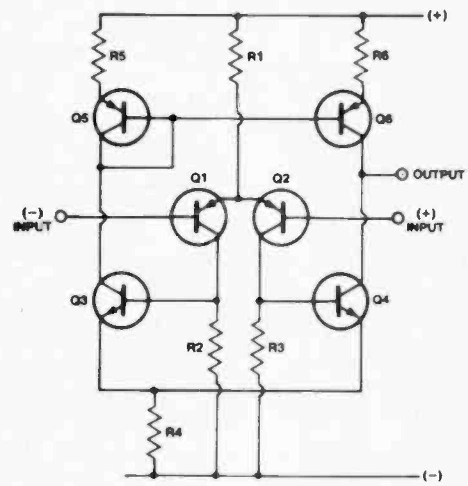


Fig. 1 The basic voltage gain stage of the ETI-5000 MOSFET amplifier.

SPECIFICATIONS

Power output
100 W RMS into 8 ohms
(± 55 V supply)
(up to 150 W with suitable PSU)

Frequency response
8 Hz to 20 kHz, +0 −0.4 dB
2.8 Hz to 65 kHz, +0 −3 dB
NOTE: These figures are determined solely by passive filters.

Input sensitivity
1 V RMS for 100 W output

Hum
−100 dB below full output (flat)

Noise
−116 dB below full output
(full, 20 kHz bandwidth)

2nd harmonic distortion
<0.001% at 1 kHz
(0.0007% on prototypes)
at 100 W output using a
± 56 V supply rated at 4 A continuous
<0.003% at 10 kHz and 100 W

3rd harmonic distortion
<0.0003% for all frequencies less than 10 kHz and all powers below clipping.

Total harmonic distortion
Determined by 2nd harmonic distortion (see above).

Intermodulation distortion
<0.003% at 100 W
(50 Hz and 7 kHz mixed 4:1)

Stability
Unconditional

slightly to around 0.003% at 10 kHz (top end distortion figures are a function of bias current). It should be remembered, however, that obtaining low THD figures should not be the prime objective of a good power amplifier design, but results from the reduction of dynamic distortion mechanisms.

Tested and Trying

The module has been tested exhaustively and all prototypes have performed with negligible differences.

When attempting to measure distortion figures as low as these, great care must be taken with the earthing arrangement to the test equipment. The amplifier module will give its lowest distortion figures only when measured with respect to the correct earth. It may be necessary to remove the connection between mains earth and signal earth inside some distortion analysers. This problem will not arise when the amplifier is connected to a loudspeaker. This condition is not unique to the 5000 module, but will occur whenever an alternative earth path is provided to the output signal earth.

The sound is clean with no sign of the aggressive high frequency performance common to many transistor amplifiers. There are some amplifiers that give the subjective impression of being 'over smooth', ie the amplifier on first listening sounds clean and unobtrusive. Further listening test reveals, however, that these amplifiers lack detail, and complex sounds like a symphony orchestra tend to beome a single mass of sound rather than being rendered as single instruments. The ETI-5000 does not suffer from this problem!

Carefully Does It

Particular care has been taken to minimise slew-rate limiting and harmonic distortions. An inspection of the 'How It Works' section will reveal the techniques employed.

Follow the suggested constructional method and no problems will be encountered.

Construction — Module

The construction of the power amp module is not difficult since all the components are mounted on a single board. Since the design employs a fairly large amount of negative feedback, the board pattern is a critical factor in attaining the maximum theoretical performance. It would be virtually impossible to achieve the same performance if the board pattern were altered, without recourse to a distortion analyser with a sensitivity of at least 0.005% and a very good spectrum analyser. The board pattern shown ensures freedom from earth path interaction and therefore does not degrade the distortion performance of the design.

Commence construction by soldering all the resistors onto the circuit board. The 0R22 (0.22 ohm), 5 W source resistors in the output stage get warm if the amplifier is operated for extended periods at high power.

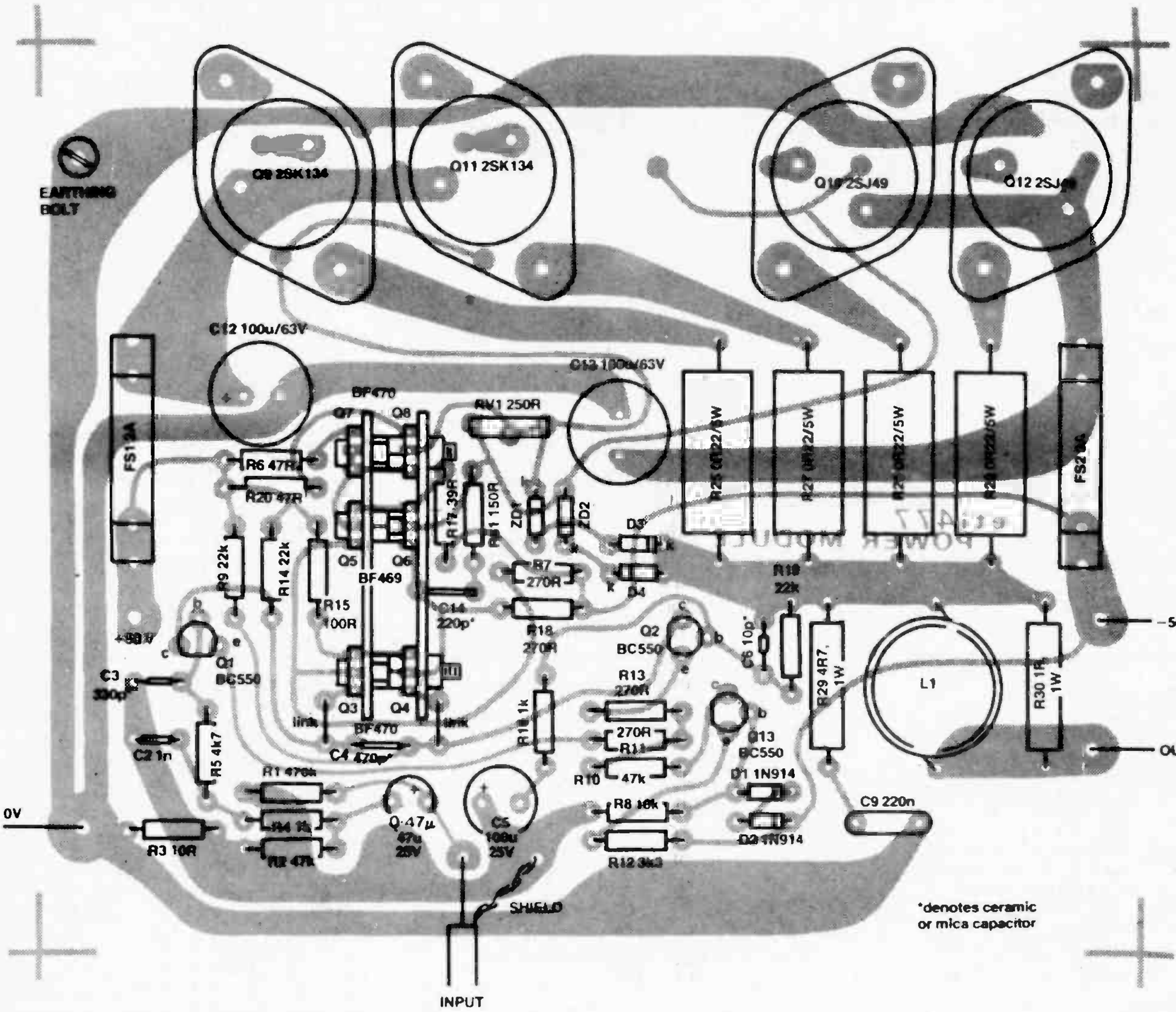


Fig. 2 Component overlay of the MOSFET amp module.

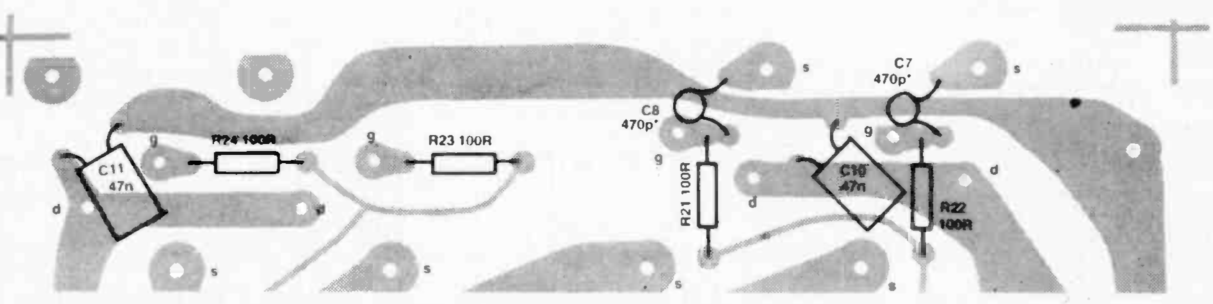


Fig. 3. This overlay shows the components which are mounted directly on the copper side of the PCB.

They should never get hot enough to burn the circuit board, since any fault capable of causing this much power dissipation should blow the supply fuses first. Nevertheless, it is good construction practice to space these resistors a few millimeters off the surface of the board. The 4.7 ohm, 1 W resistor R29 should *definitely* be spaced off the board since it will overheat if a fault condition should cause oscillation of the amplifier at high frequencies. Do not mount the four 100 ohm resistors R21, R22, R23, R24 at this stage. These are mounted on the rear of the circuit board and are best left until after the MOSFETs are mounted.

Solder the four fuse clips into the board next. Now mount all of the capacitors, with the exception of C7, 8, 10, and 11. Once again, these mount on the rear of the board. Make sure the electrolytic capacitors C1, C5, C12 and C13 are inserted with the correct orientation as these are polarised components. Mount the 1N914s and zener diodes, taking care to orient them correctly. Solder the trimpot RV1 into place and then the small-signal transistors, Q1, Q2 and Q13.

Next step is to mount the six voltage amp transistors, Q3 through Q8. These are situated on the board in two parallel rows, each row with three transistors. In the prototype modules, the heatsinks were constructed from two pieces of aluminium, as can be seen from the photographs. The transistors are mounted using 6BA bolts, each passing through a pair of transistors. This forms a very strong assembly which can then be soldered onto the board. Insulating mica or plastic washers should be used between the metal side of the trasnsistors and the heatsink strip, using a small quantity of heatsink compound between each mating surface. When this transistor-heatsink assembly is completed, but before soldering it into the circuit board, check that each transistor is effectively insulated from the heatsink. Using a multimeter on the resistance range, check for shorts between the centre lead (collector) of each transistor and the heatsink strip. Note that the bolts through the six transistors are automatically insulated from the metal rear of the transistor by the plastic body of the device so no additional insulation of the bolts should be necessary.

Before mounting the MOSFET output devices it is necessary to make the heatsink bracket. This is cut from a suitable aluminium extrusion. The board has been designed to suit extrusions with one of the sides at least 40 mm wide. The transistor mounting holes have been placed so that the heatsink brackets used in the ETI 300 W modules (April '80) are compatible, although there will be some unused holes.

The output assembly should now be checked for shorts. Remove the earthing bolt first (see overlay). The resistance between the case of each MOSFET and the bracket should be checked with a multimeter. If any device shows a short to the bracket it should be disassembled and the short found. Usually it is necessary to replace the TO-3 insulating washer as most faults of this type are the result of small metal burrs cutting through the washer when mounting the device.

Once the MOSFETs are mounted, the last passive components — resistors R21, R22, R23 and R24 plus capacitors C7, C8, C10 and C11 can be mounted on the rear of the circuit board. These are positioned on the rear of the board so that lead length is kept as short as possible. Cut the leads just short enough to mount the components in place.

Set-up Procedure

The recommended supply voltage for the modules is around ±55 V. With this voltage and reasonable supply regulation, the module will deliver around 100 W RMS into a nominal 8 ohm load.

First, re-check that the output devices are not shorted to the heatsink bracket. This is best done with the earthing bolt removed as mentioned earlier. If no shorts are found, replace the earthing bolt.

Do the same check for shorts between the six voltage amp transistor collectors and their heatsinks.

Check the polarity of all polarised components. It is often difficult to tell one end from the other on diodes since the markings are easily rubbed off. If in doubt, check these with a multimeter. Wind the wiper of the trimpot RV1 fully *counterclockwise* (least resistance). This ensures no bias is applied to the output stage. Now, remove the fuses from the board if they have been fitted and replace them with 10 ohm, ½ W resistors.

The module can now be connected to a power supply.

Make sure that the power supply connections are sound, with good solder joints. If you have access to a current limited bench supply it is best to connect the module to this for the set-up and initial test. If you can do this, set the current limit to around 200 mA. *Do not* connect a load to the output of the module at this stage.

If the power is now turned on, the current through the two 10 ohm resistors replacing the fuses should be low. If these resistors start to smoke, this indicates a fault condition — turn the power off immediately.

PARTS LIST

Resistors (all ½ W, 5%)

R1	470k
R2,11	47k
R3	10R
R4,16	1k0
R5	4k7
R6,20	47R
R7,10,13,18	270R
R8	10k
R9,14,19	22k
R12	3k3
R15,21-24	100R
R17	39R
R25-28	0R22, 5 W
R29	4R7, 1 W
R30	1R0, 1 W
R31	150R

Potentiometer

RV1	250R vertical trimpot

Capacitors

C1	470n, 25 V PCB electrolytic
C2	1n0 polyester
C3	330p ceramic or mica
C4,7,8	470p ceramic or mica
C5	100u 25 V PCB electrolytic
C6	10p ceramic or mica
C9	220n polyester
C10,11	47n polyester
C12,13	100u 63 V PCB electrolytic
C14	220p ceramic or mica

Semiconductors

Q1,2,13	BC550
Q3,4,7,8	BF470
Q5,6	BF469
Q9,10	2SK134
Q11,12	2SJ49
D1,2,3,4	1N914 or similar
ZD1,ZD2	12 V, 400 mW zener

Miscellaneous

PCB; four fuse clips; two 3 A fuses; one plastic bobbin or similar former, 15 mm diameter; one metre of 0.8 mm dia. enamelled copper wire; two strips of 20g aluminium, each 15 mm wide by 47 mm long (for voltage amp heatsink); heatsinks, case to suit.

Semiconductors

BR1	200 V, 35 A bridge rectifier

Capacitors

C1,2,3,4	10,000 uF, 80 V can electrolytics
C5,6,7,8	100n polyester
C9	470n polyester

Transformers

T1,T2	two x 35 V secondaries

Miscellaneous

SW1 . . . . illuminated rocker switch, 240 V AC rated; 1 off 2A fuse and fuseholder; 1 off 3-pin DIN socket; 2 off 2-way plastic terminal blocks; 2 off phono sockets; 2 off red and 2 off black heavy duty screw terminals; clamp grommet and sundry rubber grommets; hookup wire; nuts, bolts etc; heatsink/front panel, metalwork etc.

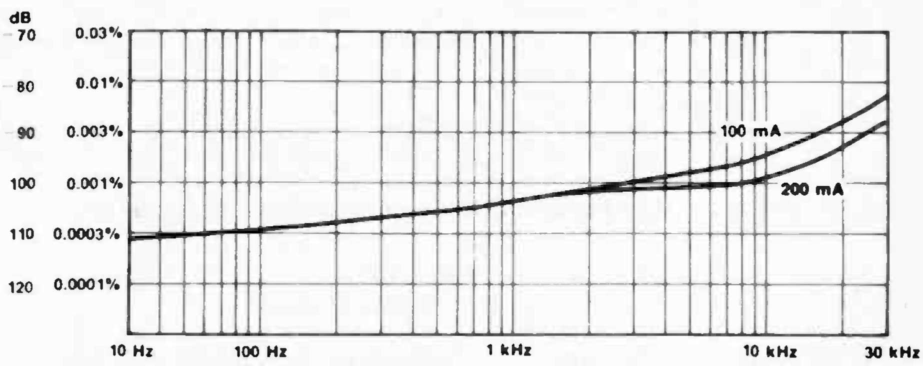


Fig. 4 This graph shows the measured distortion versus frequency for two values of quiescent current in the output stage.

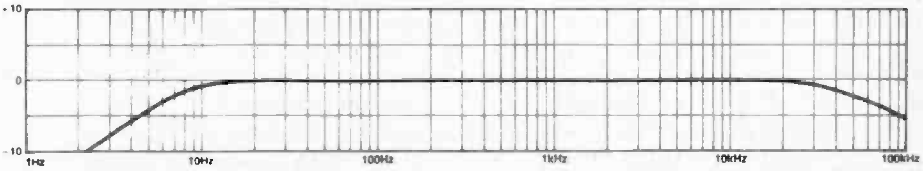


Fig. 5 The measured frequency response of the amplifier (single module). Roll-off points are defined by the input filter (low end) and output compensation network (high end).

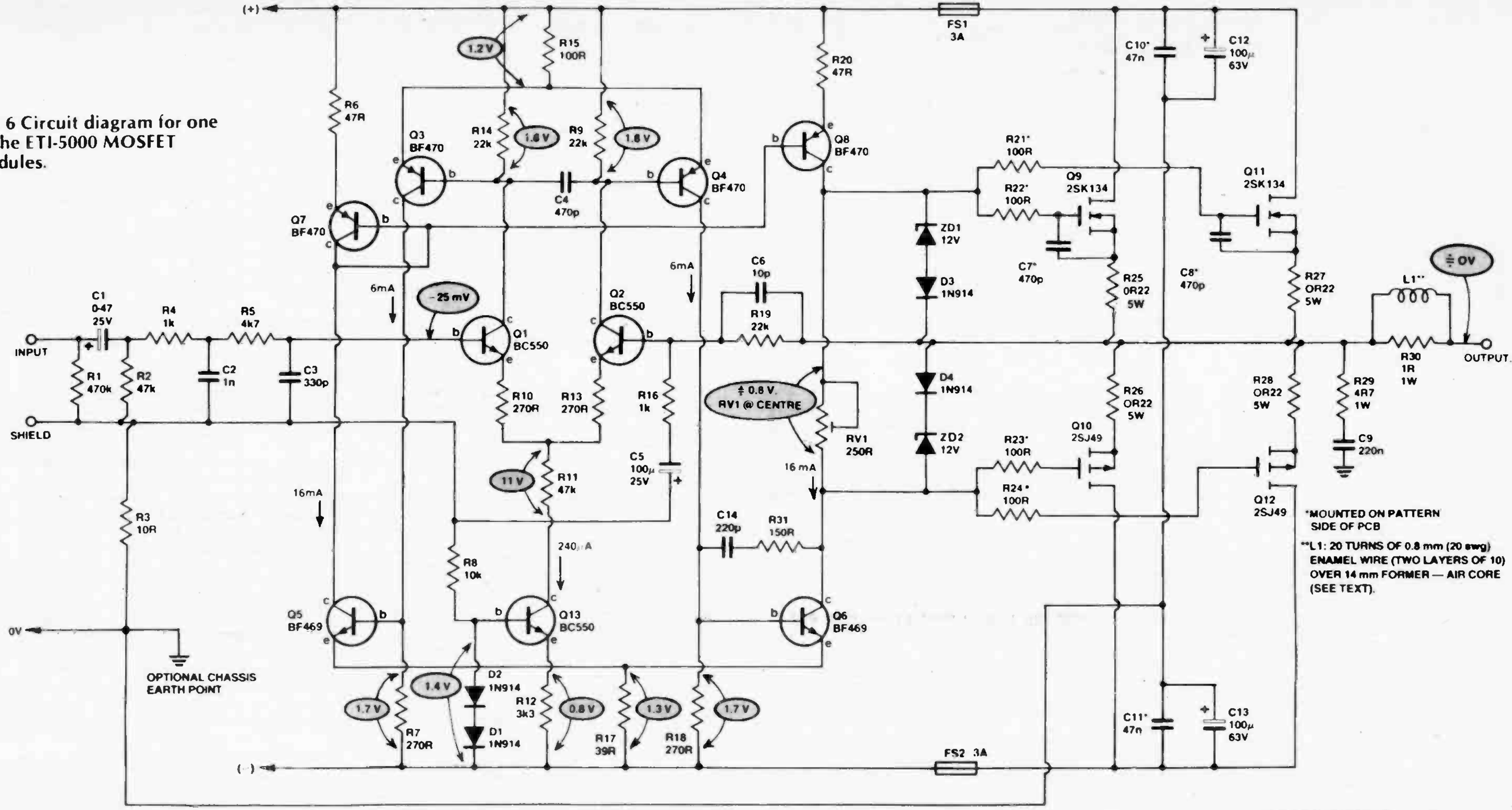


Fig. 6 Circuit diagram for one of the ETI-5000 MOSFET modules.

This very brief DC analysis of the circuit is intended to help the constructor rationalise the voltages quoted on the circuit diagram. The voltages are the result of averaging voltage measurements on a number of prototypes, and slight deviations from these should be expected.

The input signal is fed to three RC filter sections formed by C1/R2, R4/C2 and R5/C3. The C1/R2 filter defines the low end 3 dB point of the amplifier at around 7 Hz, with an attenuation rate below this frequency of 6 dB/octave. The two sections R4/C2 and R5/C3 both have 3 dB points of around 30 kHz defining the top end response of the amplifier. This filter limits the maximum signal slope to less than the slowest stage in the amplifier. It also provides protection against RF interference.

Transistor Q13 and the associated components R8, R12 and D1, D2 form a constant current source (or sink), maintaining a final DC current through the differential pair Q1, Q2 of around 240 mA. Under no-signal conditions this current is shared equally between Q1 and Q2. The resistances of R9 and R14 are in parallel with the equivalent input resistances of the transistors Q3, Q4, decreasing the effective load resistances of the differential pair to around 15k. The voltage drop across R9 and R14, should therefore be 1V8 approximately. This voltage, minus the 0V6 drop of Q3 and Q4, will cause a voltage drop of 1V2 across resistor R15, causing a current of 12 mA to flow in this resistor. This current is shared by Q3 and Q4 and causes a 1V7 drop across resistors R7 and R18. Once again, the effective input impedance of Q5 and Q6 is in parallel with the resistors, but in this case the value of R7 and R18 is very much lower than the base impedances of Q5 and Q6, so this effect can be ignored in a DC analysis like this. The voltage across R7 and R18 minus the 0V6 drop of Q5 and Q6 will cause 1V3 to be dropped across 32 mA through the resistor. Again Q5 and Q6 form a differential pair, and this current is shared equally by the two transistors. The load for these devices is formed by a current mirror, Q7-Q8, that ensures the current through Q5 and Q6 will remain the same. Transistors Q4 and Q5 therefore form the main voltage gain section of the amplifier and have a typical emitter-collector current of 16 mA. The preset RV1 will drop nominally 1 V across it when the output stage quiescent current has been set.

Diodes D3, D4 and zeners ZD1, ZD2 protect the MOSFET output devices from being overdriven, as described in the text. The RC-RL network on the output ensures that the amplifier has a correct load at all frequencies thereby eliminating the problem of oscillation that could otherwise result.

BUYLINES

When designing amplifiers intended, as this project is, for extremely low distortion, it is essential that all stages in the amplifier be designed for as low a distortion as possible. When the negative feedback is applied, truly excellent performance can be expected.

In any practical power amplifier design the bulk of the open loop gain is provided by the main voltage gain stage. The difference amplifier will generally provide some voltage gain but its main objective is to provide a linear difference signal with respect to its two inputs. In some power amps the output stage is a common drain or source follower MOSFET design and consequently has a voltage gain slightly less than unity.

An analysis of the distortion characteristics of the differential pair reveals that it is significantly better than a single transistor operated in common emitter configuration. If we assume that the distortion in a bipolar transistor is due exclusively to the exponential relationship between collector current and base-emitter voltage, the distortion generated by a differential pair and a single transistor can be calculated by techniques of Fourier analysis.

As well as having low distortion itself, a differential voltage amplifier will enable both outputs of the input differential pair to be used, thus giving a balanced output. This overcomes the problem of asymmetrical loading of the input pair by the input impedance of a single-ended voltage amplifier, a problem that would otherwise lead to increased distortion in the first stage.

Figure 1 is a circuit diagram of the voltage gain stage developed for this project. The stage is really a double differential pair with 'current mirror' load. Transistors Q1 and Q2 form the first differential pair with R1 as the common emitter resistor. The output of the Q1, Q2 pair provides a differential drive to Q3 and Q4. The latter pair, however, must be converted into a single-ended stage suitable for driving the output stage. This is the job of the current mirror formed by Q5, Q6, R5 and R6. Transistor Q5 has its base-collector junction shorted and therefore acts like a diode, but with the same characteristics as the base-emitter junction of Q6. The bases of Q5 and Q6 are connected together and therefore the voltage on the bases is identical. Since Q5 forms a 'mirror image' of the base-emitter junction of Q6, the voltage drop across the two transistors will be almost identical, depending on how well the two transistors are matched. Since the voltage on the bases and the voltage drop across the base-emitter junctions is almost identical, the voltage on the bottoms of R5 and R6 will be almost identical. If these two resistors are made the same value, the current through each resistor, and therefore the current in the collectors of Q3 and Q4 will be identical. This ensures the Q3, Q4 pair will operate symmetrically, even when a single-ended load is attached to the pair.

Furthermore, since the collector of Q3 is connected directly to the base of Q6, the transistors Q6 and Q4 combine to form a push-pull pair with very high gain.

In order to achieve good transient performance when drivng the slightly capacitive load of the output stage it is necessary to run a fairly large amount of current through this stage, especially the final differential pair and the associated current mirror. In the 5000 there is approximately 16 mA through these transistors, and the average power dissipation is therefore around 0.8 W.

These transistors will run fairly hot, approximately 60°C on the small heatsink shown, but the transistors are well inside their maximum ratings. The result is a voltage amplifier stage of exceptional linearity and very high gain. Coupled with a well-designed input differential amplifier and a good ouput stage, the phase linearity produced by this voltage stage is excellent, and makes it an easy matter to ensure total stability of the amplifier.

Since the output of the voltage gain stage has been designed to have sufficiently low output impedance to drive the gates of two MOSFETs in parallel, the output stage consists simply of the MOSFETs themselves. If a preset pot is inserted between the collectors of Q6 and Q4 in the voltage amplifier stage of Fig. 1 the voltage across this preset can be used as the bias voltage for the output stage. This is shown in the circuit diagram.

The gates of the output divices are connected to either side of the preset via resistors R21, R22, R23 and R24. As mentioned earlier, these resistors increase the time constant associated with the MOSFET gate capacitance, reducing the frequency response of the output stage slightly but ensuring stability.

Both N-channel and P-channel MOSFETs are used in this project. The important difference is the value of gate-source capactance for the two devices. If this is not equalised the presence of asymmetrical reactance in the output stage makes it almost impossible to ensure stable operation. The only cure is to decrease the open-loop gain of the whole amplifier, so that the negative feedback is reduced, and accept the consequent increase in distortion.

The most common method employed to achieve this is to increase the value of the emitter resistors in the input stage. This reduces the voltage gain of the input stage, at the same time increasing its small-signal bandwidth. Consequently, in a power amplifier employing MOSFETs in the output stage, the relatively small amount of negative feedback available will not be able to linearise the transfer characteristics of the output devices, and the result is an amplifier with only mediocre performance.

The problem of the asymmetry of the gate-source capacitance is cured by the addition of the capacitors C7 and C8 shown in the circuit diagram. With these capacitors in the circuit and with the 100R resistors R21, R22, R23 and R24, the only other component required to ensure total stability of the output stage is the power RLC combination formed by R29, C9, R30, L1 and the supply bypass capacitors C10, 11, 12 and 13.

When a square wave, for example, is fed into a purely capacitive load, the output stage will virtually see a short circuit since the high frequency (>100 kHz) fourier components of the sine wave will see very little impedance in the capacitive load.

The inductor L1 ensures that this does not happen by inserting a reactance that increases at high frequencies. The resistor R30 is placed in parallel with the inductance so the top end frequency response of the amplifier is not unduly affected.

The other two components of this network, R29 and C9, ensure that the amplifier always has a load at high frequencies. Without this 4R7 load the gain of the output stage is slightly greater than unity, due to the presence of positive feedback caused by the effective capacitance around the output devices, and this causes oscillation.

The 220nF capacitor is necessary since the resistor must be removed from the circuit for frequencies inside the audio passband or the high power dissipation in the resistor would destroy it instantly.

Resistors R25, R26, R27 and R28 provide slight emitter degeneration to the output devices. This helps to linearise the output transfer characteristic of the amplifier and thereby assists in ensuring stability of the output stage.

The final components needed to ensure stability of the output stage are the supply bypass capacitors C10, C11, C12 and C13. Capacitors C12 and C13 are 100uF electrolytics that provide general supply bypassing to frequencies inside the audio passband, but have little effect at 1 MHz, where the MOSFET would tend to oscillate.

For this reason capacitors C10 and C11 have been included also. These capacitors must be positioned very near the output devices, since the resistance of only several centimetres of track will greatly decrese their effectiveness.

In the 5000 these capacitors are mounted on the rear of the circuit board with one lead soldered directly to the drains of the output devices.

The gate resistors R21, R22, R23 and R24 and the gate power capacitors C7 and C8 are also mounted on the rear of the circuit board, again soldered directly to the MOSFETs.

The remaining components in the output stage are the zener diodes ZD1 and ZD2 and their associated diodes D3 and D4, and the bias preset RV1. The maximum gate to source voltage of these MOSFETs is 14 V; if this voltage is exceeded the MOSFET can be destroyed, so the zeners and diodes are placed between the gate and the output of the power amp to prevent the drive voltage ever becoming more than 12V6 above or below the output voltage.

This condition would mainly occur when the output is driving a short circuit or a capacitance load big enough to look like a short circuit. Under these conditions the output voltage cannot deviate much from 0 V since it is shorted to 0 V, and the negative feedback loop will drive the output of the voltage gain stage into clipping in an attempt to compensate for the error.

The gate to source voltage is now around ± 50 V, well above the absolute maximum voltage for the output devices. The diodes D3 and D4 are necessary to prevent the zeners from shorting out the gate drive under normal operating conditions.

The current flowing through the preset RV1 will give rise to a voltage drop across the preset that can be varied by adjusting RV1; this acts as the bias voltage for the output devices. If the diodes D3 and D4 are not present, the gates of Q10 and Q12 for example can never go more than 0V6 above the output voltage, due to the 0V6 forward voltage drop of ZD2. Since the voltage drop across RV1 is around 1 V, the drive voltage to the gates of Q9 and Q11 can never go higher than the sum of these two voltages, ie above the output voltage, and this limits drive to the MOSFET. The same occurs to negative-going signals due to the 0V6 forward turn-on voltage of ZD1.

The combination of all these techniques yields an output stage that is totally free of instability and with a bandwidth of a round 5 MHz! Furthermore, the transfer characteristics and phase response of the output stage are predictable, making it relatively easy to ensure stability of the overall feedback loop.

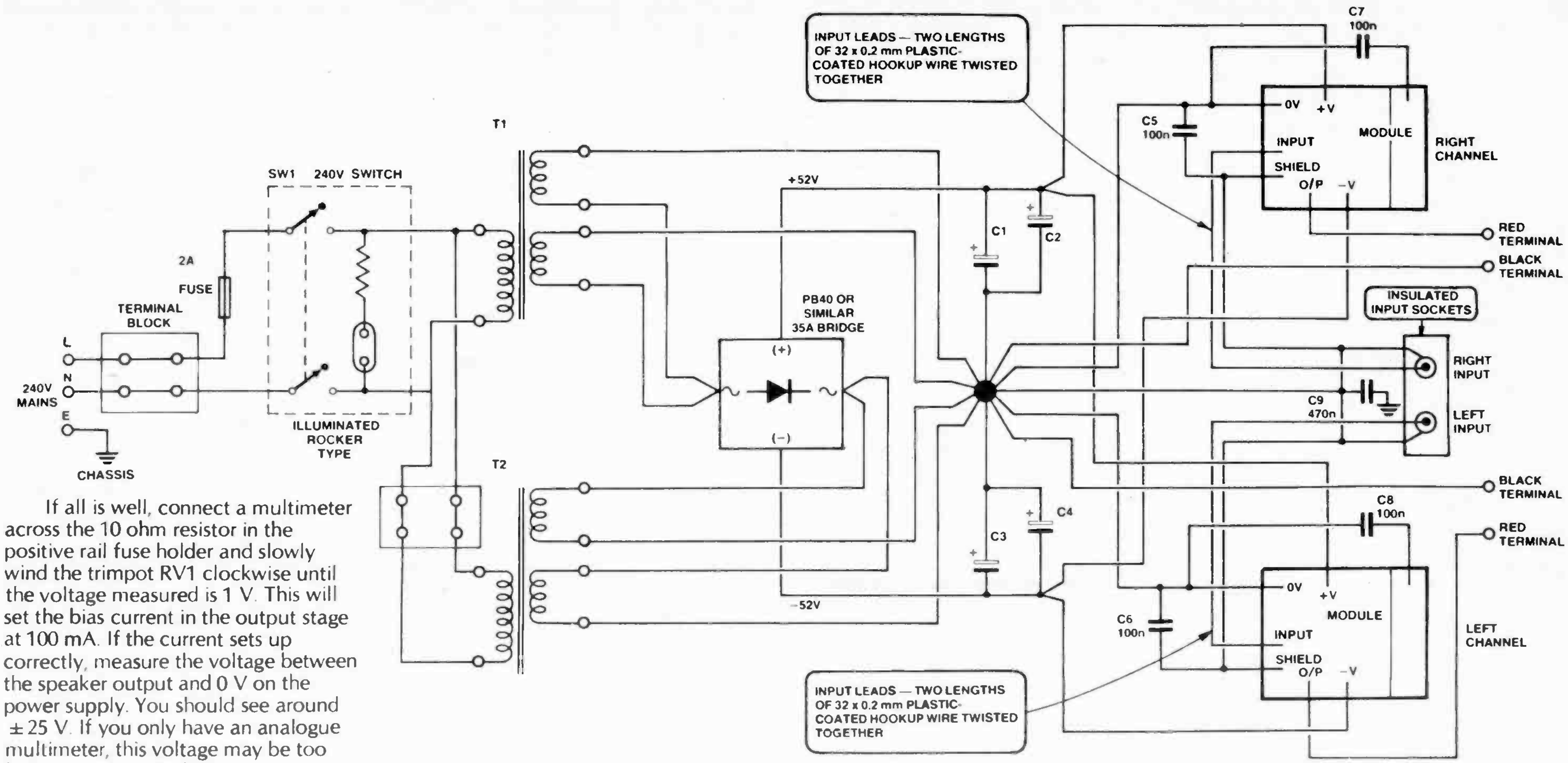


Fig. 7 Interwiring diagram for the complete (stereo) amplifier.

If all is well, connect a multimeter across the 10 ohm resistor in the positive rail fuse holder and slowly wind the trimpot RV1 clockwise until the voltage measured is 1 V. This will set the bias current in the output stage at 100 mA. If the current sets up correctly, measure the voltage between the speaker output and 0 V on the power supply. You should see around ±25 V. If you only have an analogue multimeter, this voltage may be too low to measure; in this case it is sufficient to show that the output is at 0 V.

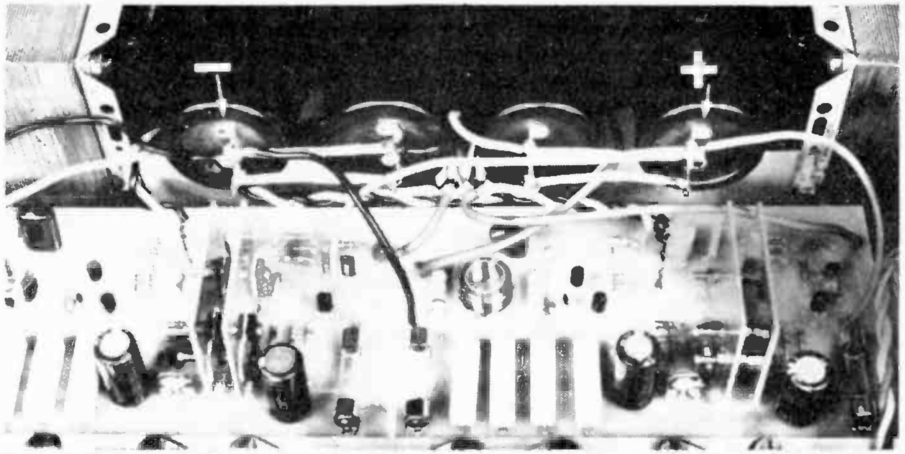
If the module passes all these tests, it is safe to replace the fuses and connect the load.

Construction — PSU

The advantage normally associated with independent power supplies is the reduction of crosstalk between channels. In the case of the ETI module however, the high supply rejection of the design reduces crosstalk to a level that is completely insignificant (ie around the noise level), so independent supplies offer no real advantage. On the other hand the use of two power transformers in a single power supply yields a supply capable of more than 100 V at over 7 A continuous. The transformer we used has two independent 35 V windings. These are connected in parallel to produce a single 35 V RMS winding capable of supplying 5 A RMS. It is unimportant whether these are toroidal or ordinary types — use whatever you can get.

If the same type of transformer cannot be obtained locally, then any type which can supply 35 V at 5 A is compatible.

One terminal of each transformer is connected to the bridge rectifier, a 35 A type. The filtering for the power supply is done with two 8000uF capacitors to form a total of 16 000uF across each half of the DC supply rails. The resulting supply voltage should be approximately ±52 V, unloaded. At full power this will drop to around ±50 V. With a 10 V drop across the output devices the peak signal voltage before clipping is around 40 V, which gives 100 W into a eight ohm load. In reality, the voltage drop across the MOSFETs is not as high as this since the module uses two devices in parallel. The maximum output power of the prototype unit using the power supply shown was 112 W single channel and 105 W both channels driven (RMS).



Wiring of the four filter capacitors. Note the common O V point between the two inner capacitors.

Interwiring

By far the biggest problem in the design and construction of any amplifier is that of earthing. If maximum performance is to be obtained from the ETI-5000 modules great care must be taken to ensure complete isolation of high current earths from low current ones such as the input signal earth. If this is not done the large currents flowing in the speaker return earths, for example, will interact with the input and distortion results. Similarly, if the earth current from the electrolytic capacitors is allowed to interact with any low current signal earth amplifier will have degraded hum figures and may even be unstable. The board layout has been designed to overcome these problems through the use of a *single-point* earthing arrangement. Earth lines from the output devices and power earth lines from the on-board electrolytic capacitors are kept separate until they reach the 0 V point on the circuit board.

The main input signal earth is the most critical. To overcome any problem the input earth is isolated from the 0 V track on the circuit board by the 10 ohm resistor R3, shown on the circuit diagram. The input wiring to the module is done with a twisted pair of 10 amp hookup cable and the connection for the input earth is done at the input sockets. This is shown in the circuit diagram.

The 10 amp hook-up cable is used instead of the more usual shielded cable, since in this application the lower resistance of the hookup cable results in better hum rejection.

The remaining earth problem is the possibility of hum loops caused by the fact that both the power amplifier and the preamplifier used to drive it must be connected to the same chassis ground point via their power cables. If the chassis of both the preamp and the power amp are connected to the 0 V point on their respective power supplies and the two 0 V points are connected together via the shielded cables between the preamp and power amp, a closed circuit is formed. Any hum currents induced into the earth lead of the three-core power cable, for example, can flow through the chassis of the power amp to the power amp 0 V point, down the shielded cable at the power amp input to the 0 V point in the preamp and via the preamp chassis around the loop again. The presence of this hum current in the power amp input earth will be seen as an input by the power amp and output hum results. The cure is to open-circuit this loop so that hum current cannot flow in the input signal earth line. The best way to do this is to break the connection between the chassis of the power amp and the 0 V point on the power supply. In this way the power amp still has a valid earth reference at its input but the possibility of a hum loop is eliminated.

The disadvantage of this technique is that the chassis can no longer act as an effective shield to external electrical noise sources, but this problem can be overcome by capacitively coupling the chassis to the 0 V track at selected places in the power amplifier. The relatively high impedance of these capacitors at 50 Hz still maintains an effective open circuit to prevent the hum loop problem.

The earthing procedure outlined above has consistently given good results both in the prototype and in numerous other power amps, and provides the power amplifier with good earthing that is not affected excessively by the earthing configuration in the preamp.

Lead Astray

The input wires to each module should be attached at this stage. We used a twisted pair of 32 x 0.2 mm plastic-coated hookup wire. This is superior to standard shielded cable for this application. The input wiring must be kept away from the 240 V wiring at the rear of the power switch.

The input 'earth' on each board has to be AC-coupled to the 0 V line on each board for the reasons discussed earlier. This is done by soldering a 100n capacitor on the rear of each board, immediately beneath R3. The 'earthing bolt', which makes connection to the heatsink bracket, is assembled with a transistor mounting insulator on the underside of the board so that the bolt is insulated from the 0 V line on the board. A solder lug is placed under the nut. A 100n capacitor is then soldered between this lug and the 0 V track adjacent.

Follow the interwiring diagram carefully and recheck at each stage.

Do the bridge rectifier and transformer wiring first. Then do the capacitors. The lower terminals of all four capacitors are connected together using heavy braid stripped from a piece of RF type coax cable. The centre of this bus becomes the central 0 V return point (refer to the photograph). The two right hand capacitors also have their terminals bridged by a length of braid, as do the two negative terminals of the left hand capacitors. The positive output terminal from the bridge rectifier then connects to the positive terminal of the innermost right hand capacitor. The negative terminal of the bridge rectifier connects to the negative terminal of the innermost left hand capacitor. Two wires from each transformer secondary are wired directly to the central 0 V point (see wiring diagram).

Next wire in the amp modules, the speaker terminals and the 240 V circuitry.

Check everything carefully once you have finished.

Getting It Going

Having satisfied yourself that all is well, remove the fuses on each board, arm yourself with a multimeter, hold your breath . . . and switch on. Assuming no disasters occur, measure the supply rail voltages. They should be around 52 V. If you have previously set up your modules then you can replace the four fuses and proceed with listening tests. Before replacing the fuses allow sufficient time for the electrolytic capacitors to discharge. This will take several minutes.

Once you have completed the set-up procedure, your amplifier is ready for listening tests.

The top and bottom covers can be screwed in place once you've confirmed all is well. We recommend you use aluminium for these items as steel plates will react with the field of the transformers and produce quite a loud hum.

MOSFET AMPLIFIER BRIDGING ADAPTOR

Some like it loud. Here's how to operate the two 150W power amp modules in the MOSFET power amp in bridge configuration with the addition of a simple, inexpensive module.

This project consists of a unity gain phase inverter that can be installed within the Series 5000 power amp. The input to one of the power amps is disconnected from the input socket and is wired to the output of the bridging adaptor. The input of the bridging adaptor is connected in parallel with the input of the other channel. This leaves one of the input sockets unused, although it could be connected to the other input socket if required.

The bridging adaptor must not degrade the distortion figures of the amplifier to which it is connected. Similarly good noise figures and freedom from slew-induced distortions must be ensured through careful design of the unity gain amplifier stages. Unfortunately, amplifiers with a gain of one tend to be the most difficult to stabilise because of the relatively high amounts of negative feedback. To overcome this problem and to maintain good noise figures, NE5534N op-amps were used in the design.

Noise Problems

The conventional way to achieve an inverting amplifier is to ground the non-inverting input and insert the input signal into the inverting input via a resistor. In this configuration the inverting input is also connected to the output of the op-amp through another resistor and forms a virtual earth point. The input resistor therefore forms the input resistance of the stage. Since this is connected to the output of the preamplifier the value of this resistor must be high, ie around 10k-100k. Unfortunately, this would seriously degrade the noise performance. To overcome this problem the bridging adaptor has been broken into two stages. The first is simply a unity gain buffer. This stage has low noise figures and an output impedance low enough to drive the following inverter stage. Since the input resistor has been kept to a small value in the second stage a good noise figure results.

PARTS LIST

Resistors (all ¼ W, 5% except where stated)	
R1,2	100k
R3-6	1k0
R7,10,11	100R
R8,9	1k5 2 W
Capacitors	
C1	220n polyester
C2	1n0 ceramic
C3,4	10p ceramic
C5	10n ceramic
C6	100u 25 V PCB electrolytic
C7,8	1000u 63 V PCB electrolytic
C9,10	10u 25 V tantalum
Semiconductors	
IC1,2	NE5534N
D1-4	1N4001 or equivalent
ZD1,2	12 V 400 mW zener
Miscellaneous	
PCB (see Buylines); mounting hardware; hookup wire.	

The board installed in the Series 5000 amp at the left hand end of the chassis.

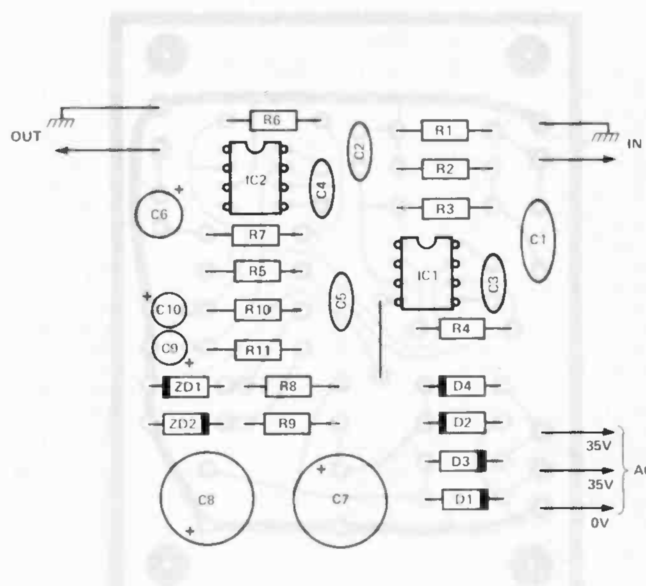


Fig. 1 Component overlay for the bridging adaptor.

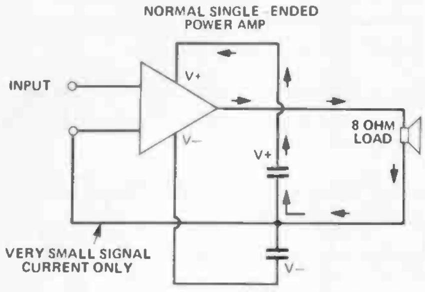


Fig. 2 Single-ended power amp showing how current flows in the power supply and the load.

Construction

Construction of the bridging adaptor is not difficult since all components are mounted on the PCB. The components can be mounted on the board in any order, although it is probably best to leave the two large electrolytic capacitors until last. As usual, be careful of the orientation of all polarised components such as the electrolytic capacitors, ICs and diodes.

Solder input and output leads to the board and bolt to the side bars on the left hand side of the power amp (viewed from the front), as shown in the accompanying photograph. Use twisted pairs of 32 x 0.2 mm plastic-covered hookup wire, as with the existing input wiring. Solder the output directly to the input of the power amp closest to the bridging adaptor. Solder the input leads of the bridging adaptor to the input socket of the other power amp. Included here is a block diagram of the Series 5000 power amplifier showing suitable modifications to incorporate the bridging adaptor.

Performance

The prototype bridged Series 5000

HOW BRIDGING WORKS

The amount of power an amplifier can deliver into a certain load is determined by the simple equation:

$$P = V^2/R$$

where V is the supply voltage and R is the resistance of the load. To achieve more power we must either decrease the resistance of the load or increase the supply voltage. Either of these will cause an increase in the amount of current to flow, and this must be catered for in the design. Unfortunately, power transistors are limited by the maximum voltage they can withstand so the supply voltage cannot be increased indefinitely. An amplifier with a supply voltage around 50 V is probably capable of supplying around 40 V peak to the load, the remaining 10 V being dropped by the output transistors, driver transistors and the power supply. This corresponds to a power level of around 100 W RMS into an 8 ohm load.

In order to increase this the load could be decreased to 4 ohms, for example. The simple equation above predicts a power level twice that of the 8 ohm case. In practice this ideal is never met since the increased current causes increased voltage drops. In the case of a MOSFET output stage such as the ETI-5000, the relatively high on resistance will cause quite a high voltage drop, decreasing the maximum output power to around 150 W for a 4 ohm load.

In order to increase the power of audio amplifiers it would seem we must increase the supply voltage and design the amplifier so that it is capable of withstanding higher signal currents. A closer inspection of Fig. 2, however, reveals another alternative. The conventional power amplifier consists of the amplifier itself and a power supply, as shown in the digram. The power supply is represented by the pair of capacitors. These correspond to the main storage capacitors in the power amp. The rest of the power supply has been omitted since its purpose is simply to maintain the necessary DC voltage differential between the ends of the capacitors.

In a class B output stage only one of the output capacitors is supplying energy to the load at any given time. The arrows in the diagram indicate the direction of the current flow when the power amp is delivering a positive-going output signal. As can be seen, the large signal current flows from the positive supply capacitor to the power amplifier, through the load and via an earth return path to the electrolytic capacitors. Every wire in this current path has resistance, so voltage drops occur at all points in the circuit. These voltage drops can be extremely significant in the performance of the power amplifier.

The distortion figure for the ETI-5000 module, usually around 0.001%, can be degraded to worse than 0.3% if the resistance in the power supply leads exceeds a small fraction of an ohm. If extremely low distortion figures are required the entire heavy current path and earth leads should be wired with one of the very *low resistance* speaker cables available.

We have seen above that at any given time in a class B power amp only one of the capacitors is supplying power to the load. So the load has access to only one of the supply rails. If both supply rails could be used at the same time the voltage available to the load would be doubled without having to redesign the amplifier, so long as the resulting current were within its capabilities. This is the purpose of the bridge configuration with power amps, sometimes referred to as 'bridging'. The principle is shown in Fig. 3. Two identical power amplifiers have been used here, the output of each going to opposite ends of the load. The input signal is fed to the input of the first amp in exactly the same way as in the more conventional approach. The arrows indicate the direction of current flow for a positive-going signal voltage. At the same time, the input signal is fed to the second power amp via a unity gain phase inverter. A positive-going input signal voltage becomes a negative-going signal at the input of the second amp. While the output of the first power amp is swinging positive the output of the second amp is swinging negative, so the load experiences double the supply voltage (neglecting for a moment the increased voltage drop due to increased signal current).

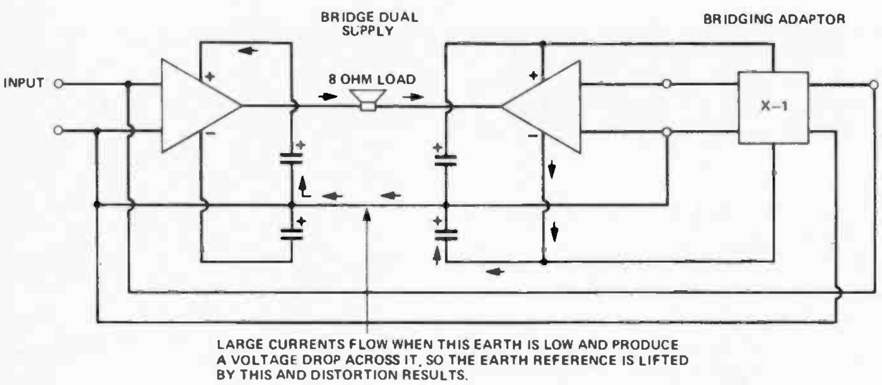
In the 4 ohm case discussed earlier the signal current is doubled, while the supply voltage remains much the same; the maximum power is therefore doubled. In the bridge case, however, the maximum signal voltage is doubled, thus also doubling the current. Since power is given by the product of voltage and current the power increases by a factor of *four*. In a real amplifier, of course, this power is never achieved. Once again the voltage drops across the output transistors, etc will decrease the power considerably, and this is especially true when using MOSFET output devices. To make a closer estimate of the power that can be expected of an amplifier when connected in bridge, determine the power delivered into a load of half that used in the bridge and double this value. If the bridge is to be used with an 8 ohm load, for example, determine the power delivered by one amplifier into a 4 ohm load and double this figure. In the case of the ETI-5000 module the power into 4 ohms is around 150 W RMS, so the power achieved by two 5000s in bridge should be around 300 W RMS. Measurements carried out with the bridging adaptor gave power figures between 280 and 300 W RMS, in good agreement with the estimate.

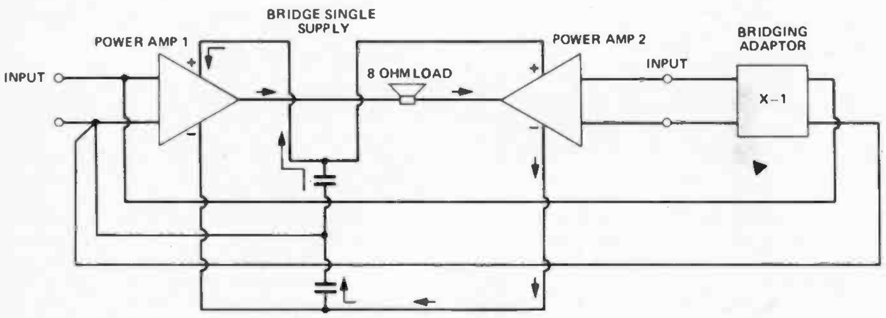
There are also limitations, however, which must be considered for successful operation of a bridge amplifier. First, since each amp is effectively driving a load half that of the real load, the load resistance connected to a bridge amplifier must be twice the minimum load specified for individual power amps. Since the minimum load recommended for the ETI-5000 module is 4 ohms the minimum load used in bridge should be 8 ohms.

Another problem associated with bridging is that both power amps used should share the same power supply to ensure the integrity of the earthing system. If this condition is not met, the distortion figure and stability margin of the amp will almost certainly be degraded. In Fig. 3, two independent power amplifiers are connected in bridge. This is done by joining their earth reference points together and driving the loudspeaker with out-of-phase signal voltages. Current resulting from a positive-going signal voltage flows from the positive supply through the first power amp and through the loudspeaker to the second power amp, and then to the negative supply rail of the second power amp. The circuit is completed by the connection between the two earth points. The problem is that, since this connection has a finite resistance, a voltage drop will occur across it, varying with the signal voltage and modulating the earth current for the second power amp.

The solution is to operate both power amps from a single power supply. Figure 4 shows a pair of amps connected in bridge and using a common supply. Once again, the arrows show the direction of current resulting from a positive-going signal voltage. Notice that in this case the connection between earth reference points has been eliminated and both power amps have access to the same single reference point. This is one of the reasons the Series 5000 power amplifier was configured with a single supply even though two power transformers and a total of four electrolytics were used; the two channels in a stereo power amp could be bridged, forming a mono power amp. For stereo operation two such amplifiers are required.

Bridge Module





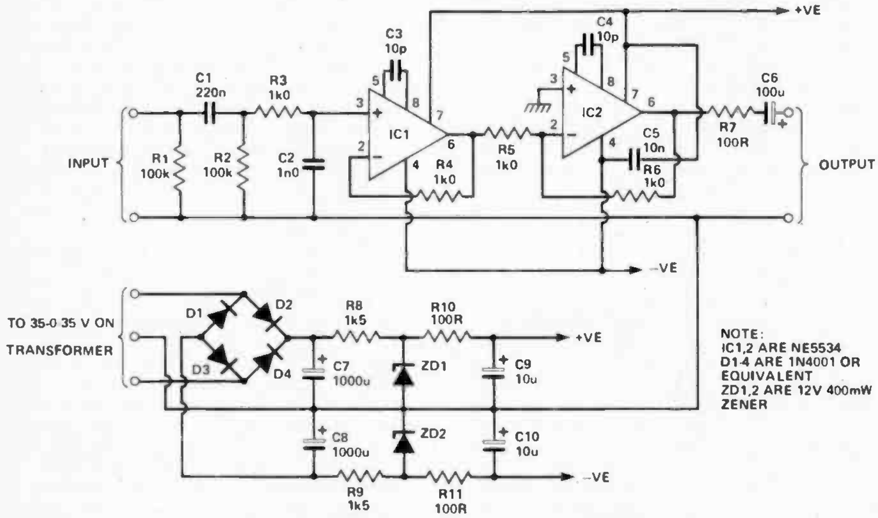


Fig. 3 (Top left) Two separate bridged power amps showing individual power supply and load currents.

Fig. 4 (Centre left) Bridged power amp and single supply showing load and supply current flow.

Fig. 5 (Bottom left) Circuit diagram of the bridging adaptor.

amp performed favourably and gave distortion figures around the resolution of our THD analyser (approx. 0.003%). Similarly, noise figures were not degraded and the adaptor tested was free of slew-induced distortion. The power output achieved was around 300 W RMS when connected to an 8 ohm load. Connection to a 4 ohm load is *not* recommended for the reasons given in the accompanying box.

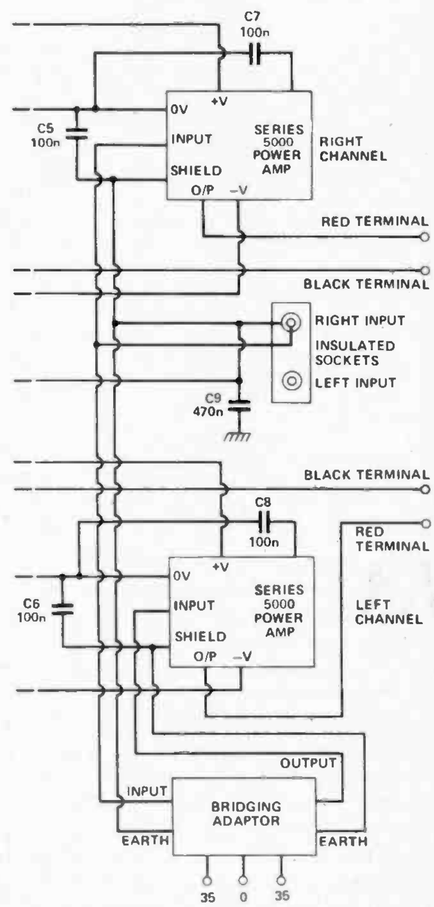


Fig. 6 How to wire the bridging adaptor into the Series 5000 amplifier for bridged operation.

HOW IT WORKS

The Bridging Adaptor is a unity gain (gain of x1) inverting stage that has its input in parallel with one power amplifier module and its output driving the other power amplifier module. Thus the power amp module it drives operates out of phase with the other power amp module.

The bridging adaptor has two stages — a non-inverting input buffer stage and an inverting output stage. The active device in each stage is an NE5534 high performance op-amp. A on-board rectifier provides dual supply rails regulated by two zeners.

Input is coupled to the non-inverting input of IC1 via an RC network consisting of C1, R2, R3, and C2. Resistor R1 provides a DC return for the input line. Resistor R3 is a low value to ensure good noise performance for IC1, and together with C2, a lowpass filter is established to limit the slew rate of incoming signals to prevent slew-induced distortions. Feedback for IC1 is provided by R4, connected between the output and the inverting input. The output of IC1 drives the inverting input of IC2 via R5. Feedback around IC2 is provided by R6. The feedback constants for both IC1 and IC2 are arranged so that each stage has a gain of one.

The output from IC2 is coupled via R7 and C6, which provide a low frequency rolloff, C6 also providing DC blocking.

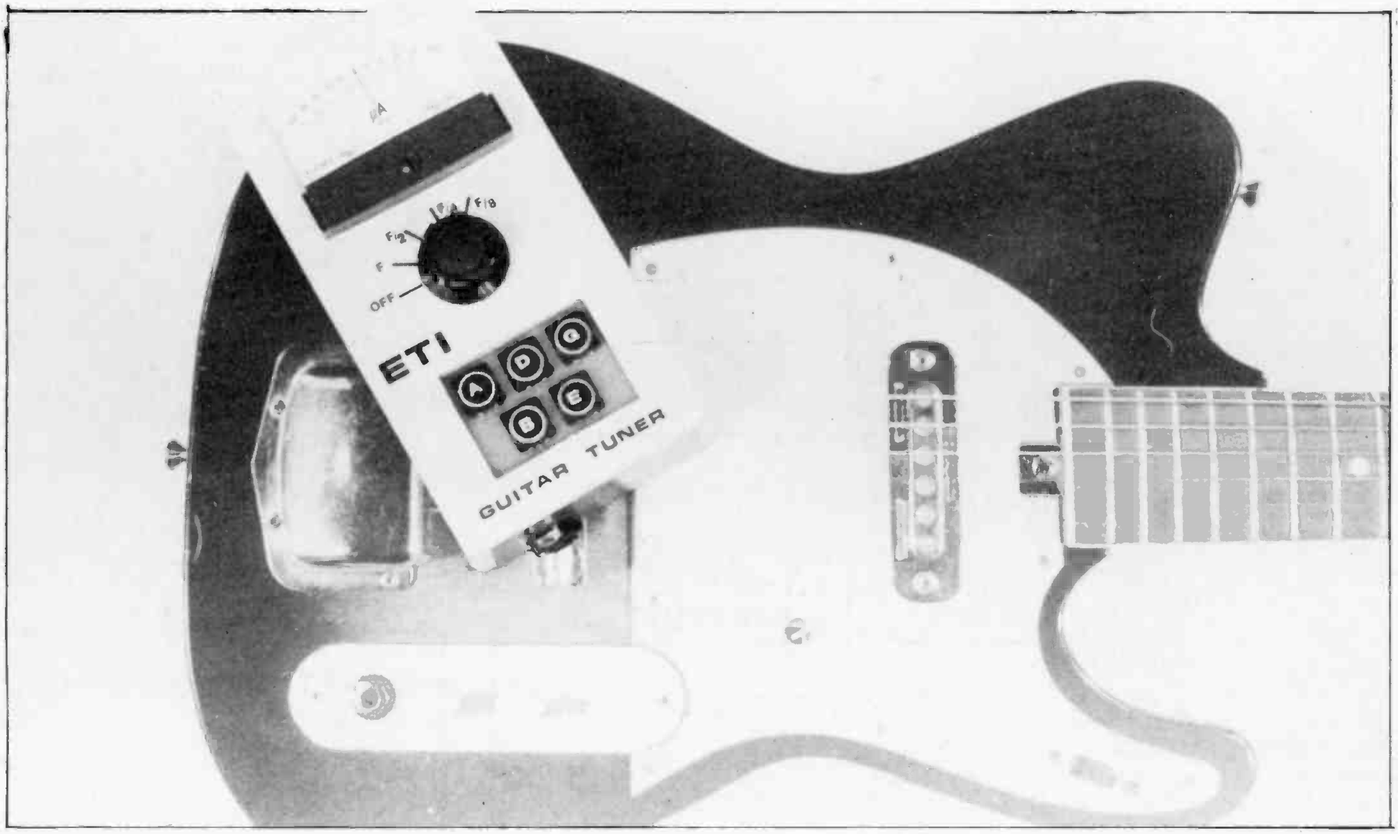
The bridging adaptor is powered from the Series 5000 amplifier power supply transformers. Diodes D1 to D4 form a bridge rectifier providing about ±52 V DC with respect to the winding centre tap. Capacitors C7 and C8 provide smoothing. Two zener diodes, ZD1 and ZD2, are used to provide regulated positive and negative 12 V DC supply rails for the two ICs. Resistors R8 and R9 provide current dropping for the two zeners and R10/C9, R11/C10 provide further filtering. Capacitor C5 provides a high frequency bypass for the supply rails. Capacitors C3 and C4 provide frequency compensation for IC1 and IC2 respectively.

BUYLINES

As usual, we can supply you with the PCBs; the order form is on page 91. Nothing else should cause any problems; the NE5534 is available from Watford Electronics, or as an alternative you could use the TDA1034 from Technomatic.

GUITAR TUNER

Roadies, perfectionists and the tone deaf can all benefit from this project. Our Guitar Tuner is both versatile and accurate. Design and development by Brian Brooks.



Musicians who wish to play together have to ensure that their instruments are tuned to the same fundamental pitch. If they aren't, the resulting sound is pretty awful. Listening to the radio these days can give the inpression that some groups actually *like* to play this way — nevertheless, we feel that the majority of musicians out there would be grateful for something that does away with the need for a tuning fork (or pitch pipes) and a good ear for pitch. Look no further — this is the project you've been waiting for.

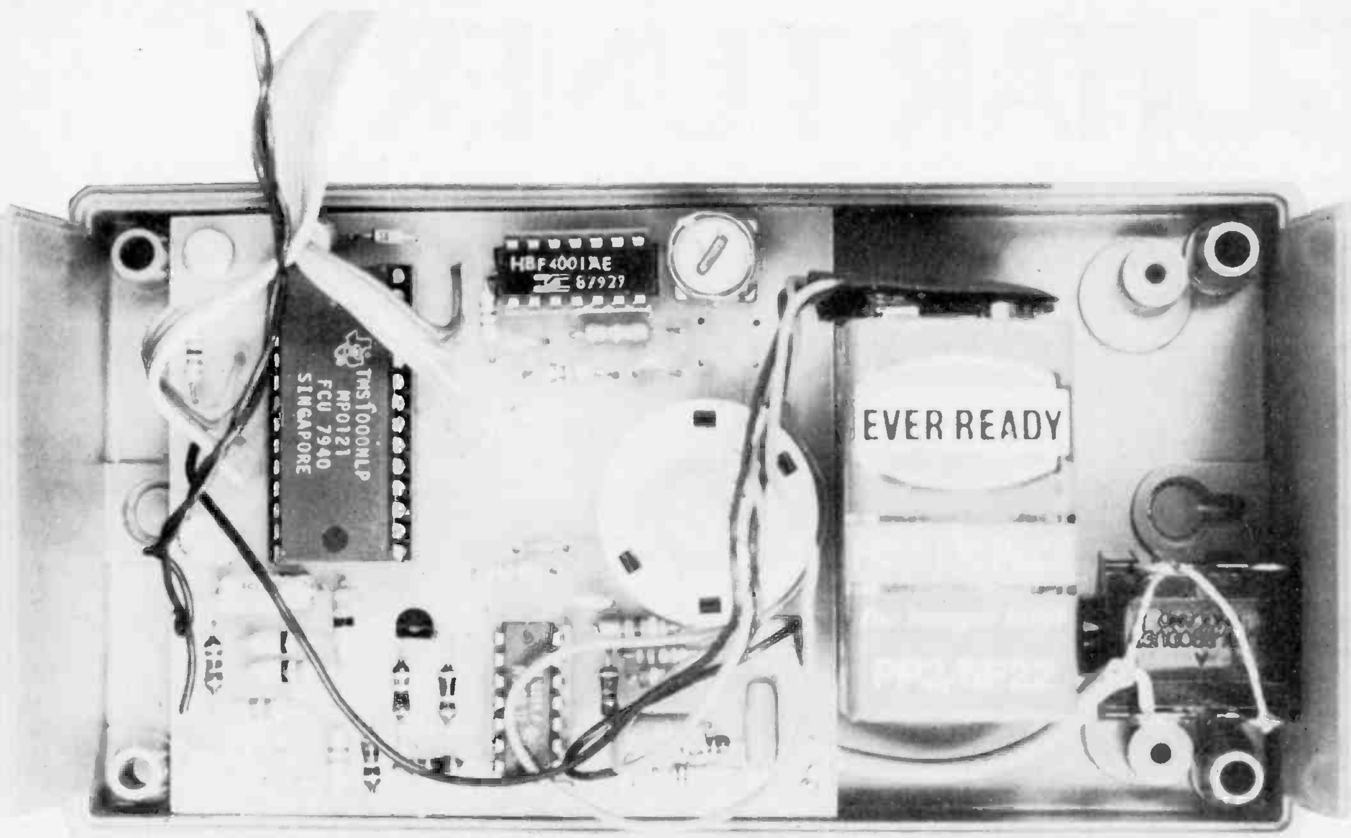
The ETI Guitar Tuner is quick, simple, versatile and highly accurate. Unlike other designs, reference frequencies are provided for all six strings of a guitar, and are selected by momentarily pressing one of the push-buttons. The note chosen is then synthesised by a frequency generator chip and remains latched until another pitch is selected. This overcomes the major disadvantage of the tuning fork, in that the note does not die away quickly. An octave selector switch allows the Tuner to be used with bass guitars or other instruments (anyone for synthesisers?), and even acoustic instruments can be tuned, by using a high impedance mike to provide an input.

Although it is the ideal project for anyone with a musical bent, or bent music, the people who will benefit most from this project are roadies, who often have to quickly tune a number of instruments before a concert while loud music is being played over the PA. Since the Guitar Tuner uses a meter to indicate beats visually, this daunting task becomes a piece of cake.

Construction

This is quite straightforward if the overlays are followed carefully. Start with the main board, soldering in the low-profile components first — wire links, resistors, diodes and IC sockets. Insert Veropins at the points where interwiring will take place later. Follow with the capacitors and the FET, and finally fit SW1 and L1. The tags on SW1 must be cut off so that it will fit the holes in the PCB. Check that all the polarised components are inserted the right way round; the coil has an asymmetrical pinout and will only fit one way. Now insert the ICs into their sockets, being *unbelievably* careful with IC1 — remember that this is an A-series *unprotected* CMOS chip and it will commit suicide at the slightest hint of static, so DON'T TOUCH THE PINS.

The completed board can now be fastened to the case bottom using short self-tapping screws. Although the PCB is only supported at one end (see photos), it is quite rigid enough. Mount the jack socket at the opposite end of the case to the PCB and link it to the input Veropins with a short piece of screened cable. Note in the photographs how two pins of the jack socket are connected to the screen of the cable — the socket is a shorting type so that if an instrument is not plugged in, the input is grounded to prevent noise pick-up.



Cut holes in the lid for the meter and the shaft of the rotary switch — the easiest way to accomplish the latter task is to smear ink on the tip and replace the lid to leave a drilling mark on the inside. Don't worry about missing slightly; the knob is wide enough to hide a multitude of sins!

Now solder the five push-buttons into place on the smaller PCB, making sure that the small dot on top is orientated as shown on the overlay, and solder the ribbon cable to the pads at the edge of the board. Make a rectangular cut-out in the lid of the box to allow access to the push-buttons (placing it far enough to one end to avoid fouling the rotary switch shaft) and fasten the keyboard inside the lid with double-sided tape or sticky pads. The corners of the board may be cut at an angle to fit between the moulded pillars.

After bolting the meter into place, the final interwiring can be completed. Solder the other end of the ribbon cable wires to the pins around IC2 — checking carefully that they're in the right order — and make the connections to the meter. All that remains now is to solder the battery leads, connect the two PP3s and fasten them (using sticky pads again) in the space between the main PCB and the jack socket.

PARTS LIST

Resistors (all ¼ W, 5%)	
R1	1k0
R2,3,8	100k
R4,12	10M
R5,10	1M0
R6	220k
R7	8M2
R9,11	47k
R13,14,18,20	4k7
R15,16,17,19	10k
R21	3k3
Capacitors	
C1,2	10u 16 V PCB electrolytic
C3	220p ceramic plate (low temp. coeff.)
C4,5	1n0 ceramic plate
C6	10u 25 V axial electrolytic
C7,9	100n polyester
C8	470p ceramic plate
C10	220p ceramic plate
Semiconductors	
IC1	4001A or 4001UB
IC2	TMS1000
IC3	TL064CP
Q1	2N3819
D1-3	1N4148
ZD1	12 V 400 mW zener
Miscellaneous	
L1	3.5 mH
SW1	2-pole 6-way rotary switch
PB1-5	push-button switches
SK1	¼" mono jack socket
M1	100 uA centre zero meter

Two off PP3 batteries and clips; PCBs (see Buylines); IC sockets; cable; case (Vero order ref. 202-21030); knob to suit.

TABLE 1

NOTE	OCTAVE	FREQUENCY (Hz)
E	F/4	82.4
A	F	110.0
D	F	146.8
G	F	196.0
B	F	246.9
E	F	329.6

(E's are 2 octaves apart)

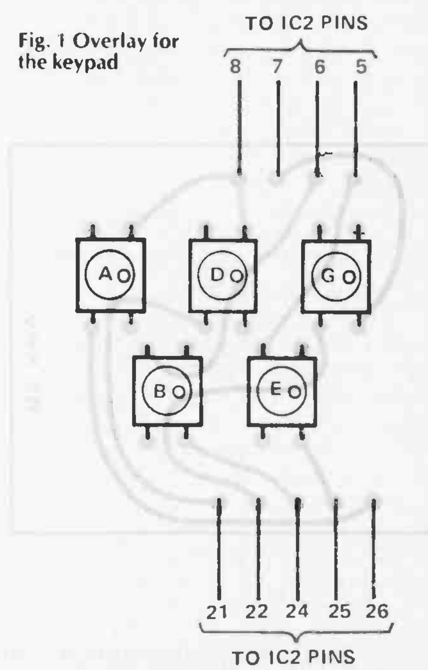


Fig. 1 Overlay for the keypad

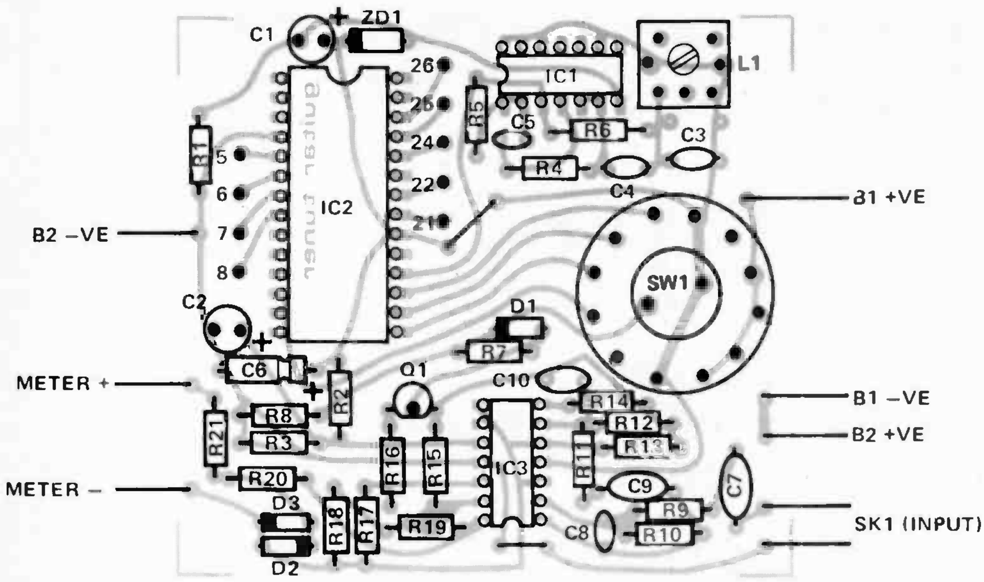


Fig. 2 Component overlay for the main PCB.

Setting Up

To calibrate the Guitar Tuner, select any of the frequencies in Table 1 and tune the coil while measuring the output of IC2 (pins 12-14) with an accurate frequency meter. Alternatively you may feed in a known original frequency to the jack socket (using a mike and a tuning fork, an accurate signal generator, an atomic clock or whatever else you happen to have handy) and adjust the coil for zero beats on the meter. Once set, the oscillator will stay very accurate regardless of the battery state or the ambient temperature.

To use the Guitar Tuner, plug your electric guitar (or microphone) into the jack socket, press the button corresponding to the string you want to tune, select the correct octave with the rotary switch (for example, use F/4 for the bottom E string), and strum the note. If you're way off tune the meter won't respond at all, but as you tune the string closer to the right pitch, the needle will start to register the beats by oscillating rapidly about the centre position. This oscillation will slow down as you approach the reference until the needle stops at zero, at which point the instrument is perfectly tuned.

BUYLINES

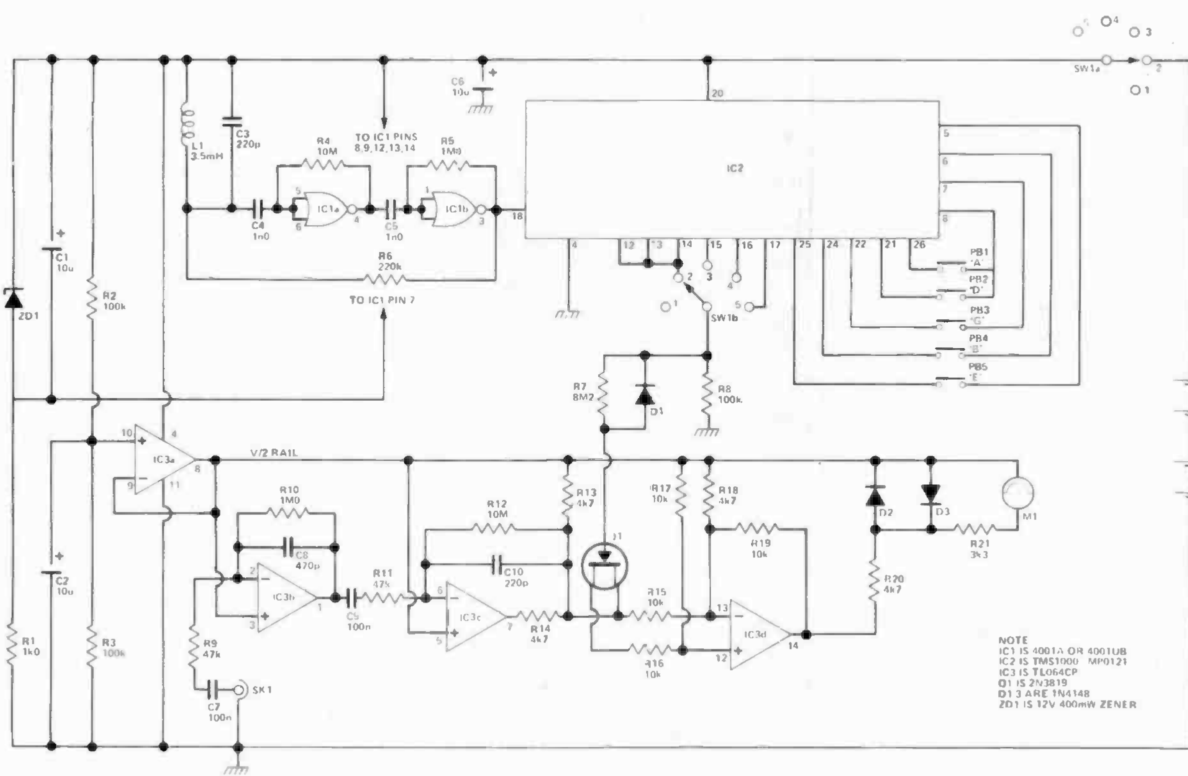
A complete kit of parts is available from Magenta Electronics, 135 Hunter Street, Burton-on-Trent, Staffs DE14 2ST for £32.37 all inclusive. IC1 is a special and is available from Magenta only for £6.48 all inclusive if bought separately from the kit.

HOW IT WORKS

The reference oscillator built around IC1a and b must be very stable, so a high-Q coil and capacitor design is used rather than the more familiar R/C astable. The IC is used here in the linear mode, so the A-series or unbuffered CMOS chip is better. Furthermore, this IC has its own stabilised 12 V supply, derived from R1-ZD1 and decoupled by C1.

IC2 is a frequency synthesiser chip, producing top octave notes A, D, G, B and E. These are selected by closing the appropriate push-button in the keyboard matrix driven by IC2; each note latches when the key is pressed so momentary action only is required. SW1b selects either the top octave (frequency F), or the three octaves below it, F/2, F/4 and F/8.

The remainder of the circuit requires a rail at half-supply voltage and this is achieved using potential divider R2-R3 and voltage follower IC3a. C2 provides decoupling for this rail. IC3b and c amplify the input signal; the input impedance is 47k, suitable for use with a guitar or high impedance mike if acoustic instruments are being tuned. IC3d is a synchronous rectifier and is used to compare the input frequency with the internal reference. When two signals are in phase the output of IC3d is positive, and when they are out of phase it is negative — the output swinging between the two at 90° and 270°. Q1 is driven from the reference to alternately invert the gain of IC3d at the appropriate pitch. Meter overload protection is provided by D2 and D3.



PLAYMATE GUITAR EFFECTS/AMP

The sounds of the superstar in your own room — or in the middle of a field! The PLAYMATE will help you on your way. Design and development by Phil Walker.

The Playmate is a small private amplifier for use with a guitar giving a few watts output for easy listening while also providing some of the basic effects used by many musicians. It is ideally suited to those who do not carry all the various effects units around in their guitar case but would like to be able to practice at odd moments or in out-of-the-way places.

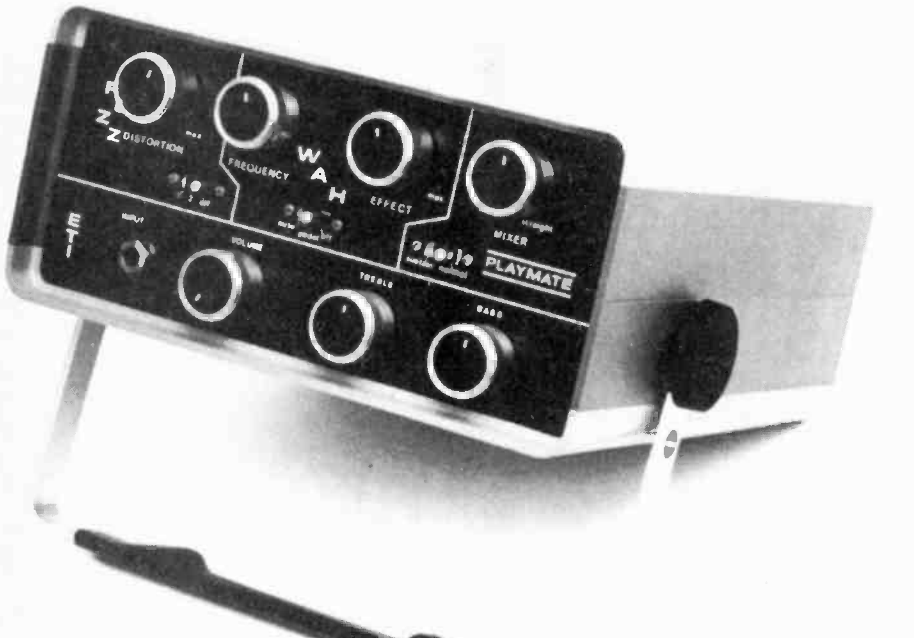
In addition to the amplifier and standard tone controls etc, various distortion and wah-wah effects are possible. As a by-product of the circuitry a sustain effect is also possible.

The sound output is provided by a small internal loudspeaker and the whole module is powered from a small mains unit or batteries. An external foot pedal could be used with the wah effect if required. This consists of a variable resistor and a couple of other resistors to provide the necessary control current. The internal control is still active at this time and can be used to set an operating range.

The Circuit

The circuit is in general straightforward. It consists of an input buffer with a gain of about 50 followed by a signal compression stage which reduces the dynamic range greatly in order to feed the effects circuitry at a constant level. The effects consist of a distortion-inducing stage for fuzz and a variable band-pass filter for the wah wah. After the effects stages, the dynamic range of the signal is restored to normal before being fed to the mixer, tone controls and power output stages.

The input buffer consists if a single 3140 CMOS op-amp whose gain is set at 48 by R2, R3. The following dynamic range compressor consists of one part of a LM13600 dual transconductance amplifier. The gain of this device is a function of the amplifier bias current, the input diode current and the load resistor. The output buffer of the device is used here as a peak detecting rectifier which charges a capacitor (C3) to the peak value of the output signal less two base-emitter drops (about 1V4). If this voltage is greater than about 0V7 the resulting current flowing through the input linearising diodes causes the effective stage gain to decrease and keep the output level constant.



Distorting The Facts

Distortion effects in this project are of two types. The first is mainly even harmonic generated by half-wave-rectifying the input, inverting it and then mixing it with the original signal to get from no distortion to complete frequency doubling. In addition to this, overload type distortion is provided by a high gain clipping amplifier using non-linear feedback (IC3a,3b).

Wah wah sound effects are produced by a current-controlled state variable filter. The control current determines the centre frequency of the pass band while a two-gang variable resistor sets the bandwidth and compensates for inevitable gain changes.

Tone controls are of a standard

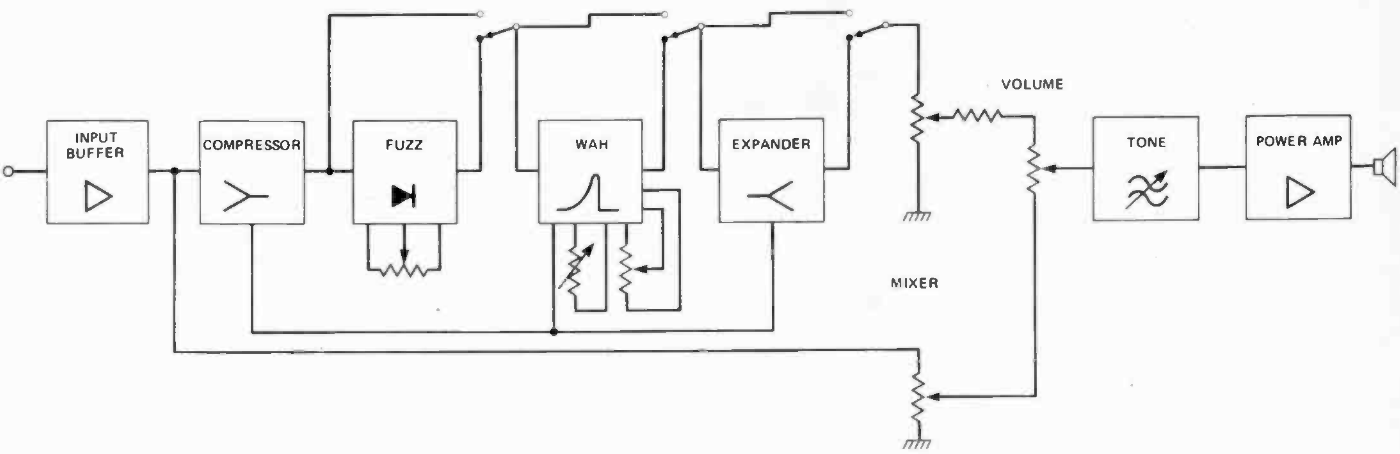


Fig. 1 Block diagram of the Playmate.

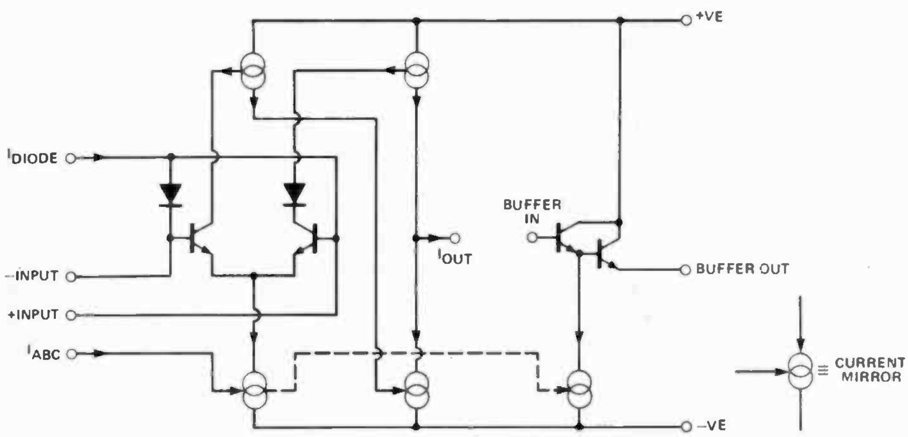
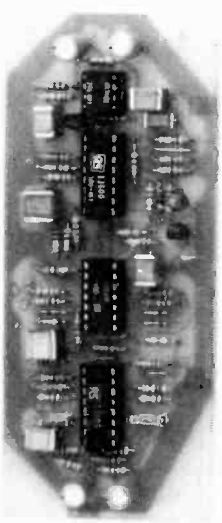
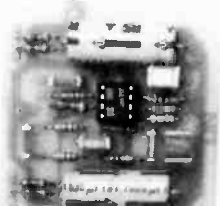
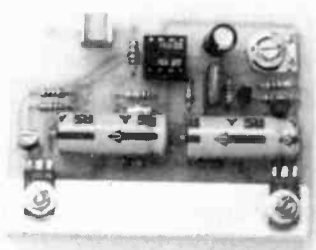


Fig. 2 Internal circuitry of the LM13600 — an operational transconductance amplifier!







type and use frequency-selective feedback networks around an op-amp. The following power amplifier has been designed to have a low quiescent current. This is important if batteries are to be used as many amps of the IC variety take 30 mA or more, or are designed for single tail working

THE LM13600

This device is used for two functions in this project. One of these is the compressor/expander while the other is the wah wah. In both of these, use is made of the fact that the gain of the device is dependent on the amplifier bias current and the linearising diode current (provided that the input current is less than half the diode current). In fact the output resistor also determines the gain but is not so easily varied.

If the diode current is zero then the manufacturers' data sheet shows that the transfer function of the device is:

$$I_{out} = \frac{I_{abc} \times q \times V_{in}}{2KT} = \frac{I_{abc}}{26 \times 10^{-3}} \times V_{in}$$

If the diode current is not zero and the signal current is less than $I_{D/2}$ then the transfer function is:—

$$I_{out} = \frac{2 \times I_{abc} I_s}{I_D}$$

where
I_s = signal current

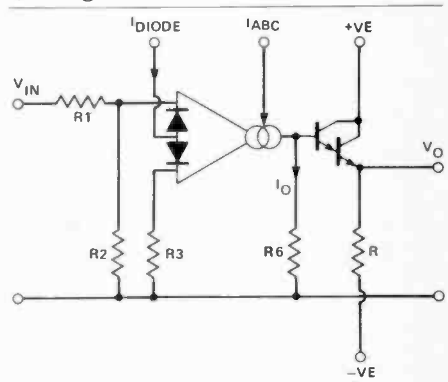
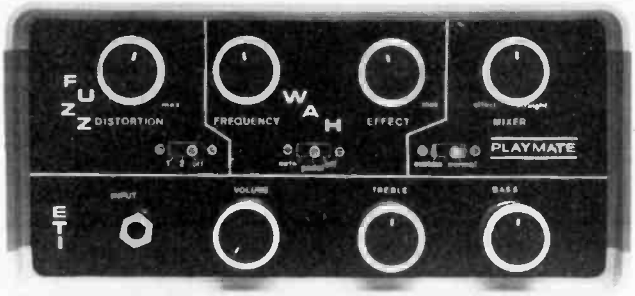


Fig. 3 Basic voltage amplifier circuit.

I_{abc} = amp bias current
I_D = lin. diode current
I_{out} = output current

If we use resistors for input and output, it can be seen that the voltage gain of a stage using this device can be controlled easily by use of the bias and linearising diode currents.

Figure 3 shows the basic circuit



HOW IT WORKS

The gain of the input buffer IC1 is set by R2 and R3 at 48. R1 determines the input impedance which C1 provides DC blocking. The output from this device goes to the dynamic range compressor IC2a and its buffer IC3a. This part of the circuit also provides control signals for the expander circuit and, if required, for the wah wah effect. The buffered output from the compressor then goes via C4 to the first part of the fuzz effect circuitry constructed around around IC3b. Here an inverted half-wave-rectified version of the input signal is produced by the action of D1 and D2 in the feedback netowrk of IC3b. This is applied to RV1 from which a portion is selected and mixed with a little of the original signal. As the half-wave-rectified signal at this point of the circuit is twice as great as the straight-through signal, by varying the setting of RV1, amounts of distortion varying from none to virtual frequency doubling can be selected.

The mixture of signals obtained above is now applied to IC3c where they are amplified. The amount of amplification is determined by the setting of SW1. In position 3 minimum gain is provided and in fact the whole fuzz section is bypassed. Position 2 gives the same gain, allowing the first distortion stage to be effective. The final position connects D3 and D4 via C5 and R19 into the feedback circuit of IC3c instead of R18. This has the effect of greatly increasing the small signal gain but causing the output to limit sharply, thus clipping and squaring the output. This facility is available on whatever output is coming from IC3b.

The output from the fuzz stages now passes to the wah wah. This effect is produced by the current controlled state variable filter used in a band-pass mode. The filter is realised by using a LM13600 device with a controlled bias current providing the variable centre frequency. The 'Q' factor is controlled by a dual gang potentiometer, half of which is used to control the 'Q' factor which the other half compensates for the effective gain change as this is altered. In this type of circuit the frequency range is determined by the values of C7, C8, R24 and R26, while the actual centre frequency is controlled by the amplifier bias current. If the bias current is allowed to become too small it is sometimes found that a thump is heard at the output; in order to prevent this R34, R35, D5 and R33 are used in the control circuitry to keep the current above this threshold.

SW2 selects between the control options for the wah wah circuit. The 'off' position bypasses the circuit altogether, the 'pedal' position makes access to an external foot pedal if fitted, while the 'auto' position connects to an output from the compressor stage. This control signal is a current which is proportional to the amount of signal compression being applied to the input signal. The magnitude of this current increases as the input signal increases. The result of this is that when the input signal is loud, the wah wah centre frequency is high and as the input decays, the wah wah frequency decreases with it. The effect of this is to make a wah sound automatically whenever a string or chord is played.

The output from this section is buffered and adjusted in level by IC3d. After this the signal passes to the signal expansion stage built around IC2B. C23 provides DC isolation and R36 converts the input voltage to a suitable drive current for the IC. For this application the linearising diode current is held constant while the amplifier bias current is varied. Q1 in the compressor circuit provides the control current for this sage allowing a good match in the attentuation/gain characteristics of the two stages. SW3 selects either the output from the expander or bypasses it as required to give normal or sustain on the effects channel.

A dual gang potentiometer RV4 allows mixing between the original signal and the effect-modified signal. This is followed by a volume control RV5 to set the output sound level.

After the volume control, IC5a buffers the signal before applying it to the tone control circuit around IC5b. The configuration used here is a very common type of feedback arrangement. As an approximation, the gain of an op-amp with feedback is taken as — (feedback resistor value)/(input resistor value). If we replace the feedback and input resistors with variable impedances, we find that wher the feedback impednace is greater thai the input impedance then the overall gaii is great than unity, and vice versa. As impedances vary with frequency, the gain , each frequency will tend to be different The only time the gain does not vary i: when the input and feedback impedances are equal whatever their magnitude. This is the general principle on which the tone control networks operate.

The final section to be considered is the power amplifier stage. Voltage amplification is provided by IC6 and the output from it drives two complementary compound Darlington pairs, Q4/Q6 and Q5/Q7. Quiescent current through the output devices is set by RV8 in conjunction with Q3, R54 and C19. R59 and R60 aid in maintaining bias stability and provide some protection to the output transistors in the event of a fault. R61 and C20 compensate for load impedance variations at high frequency and C18 reduces the high frequency gain of the power amplifier to reduce the possibility of RF oscillation. The large capacitors C21-25 are to reduce the effects of aging batteries and prevent low frequency oscillation or intermodulation distortion.

for a volage amplifier and from it we can show that the output voltage V_O is dependent on the bias and diode currents.

$$V_O = \frac{V_{in} \times 2 \times I_{abc} \times R_L}{R1 \times I_D}$$

$I_{in} = V_{in}/R_{in}$ Therefore

and the gain $\frac{V_O}{V_{in}} = 2 \times \frac{I_{abc}}{I_D} \times \frac{R_L}{R1}$

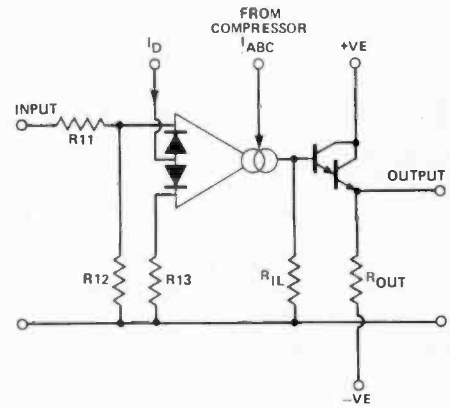


Fig. 4 The LM13600 as an expander.

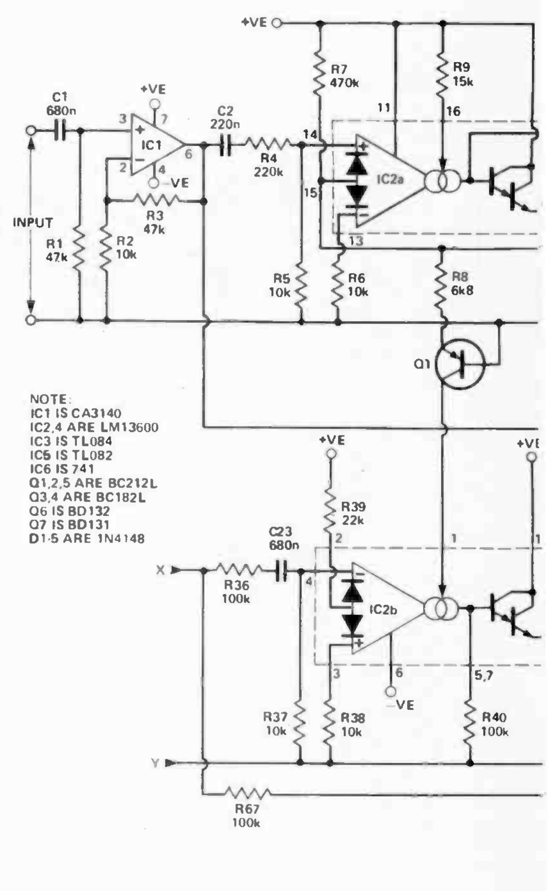


Fig. 6 Circuit diagram for the Playmate.

Compressing with The LM13600

Figure 5 shows the circuit used in this project to compress the dynamic range of the signal input. For very small signals I_D is virtually zero and the amplifier operates with a very high gain. As the signal increases, the output peak voltage will reach a level sufficient to charge the capacitor C to about one diode drop. If the input signal tries to increase further the resulting current into the input diodes will cause their impedence to fall, thus increasing the attenuation of the input and maintaining a constant output level.

At any time the current flowing into the diodes is:—

$$I_D = 2 \times \frac{(V_O - 3 \times 0.7)}{R2}$$

The 3 × 0.7 represents the voltage drops associated with the base-emitter junctions of the output buffer transistors and the voltage drop of the linearising diodes. This voltage does vary with temperature and current so since another control current is required for the expander function, this is derived by using a resistor and common base transistor. The configuration gives a current output which tracks the compressor control current very closely as it has the same number of junctions in series.

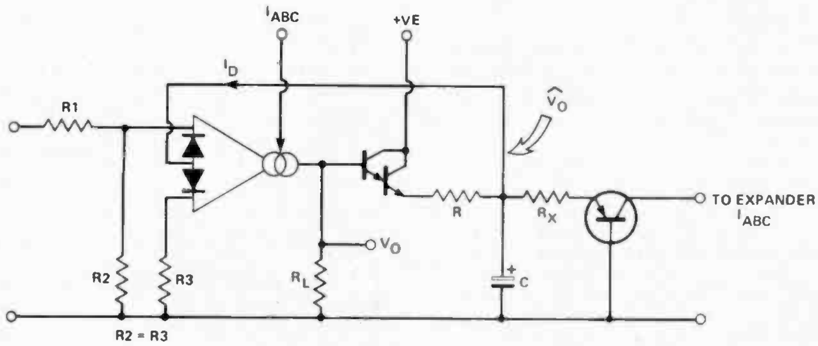
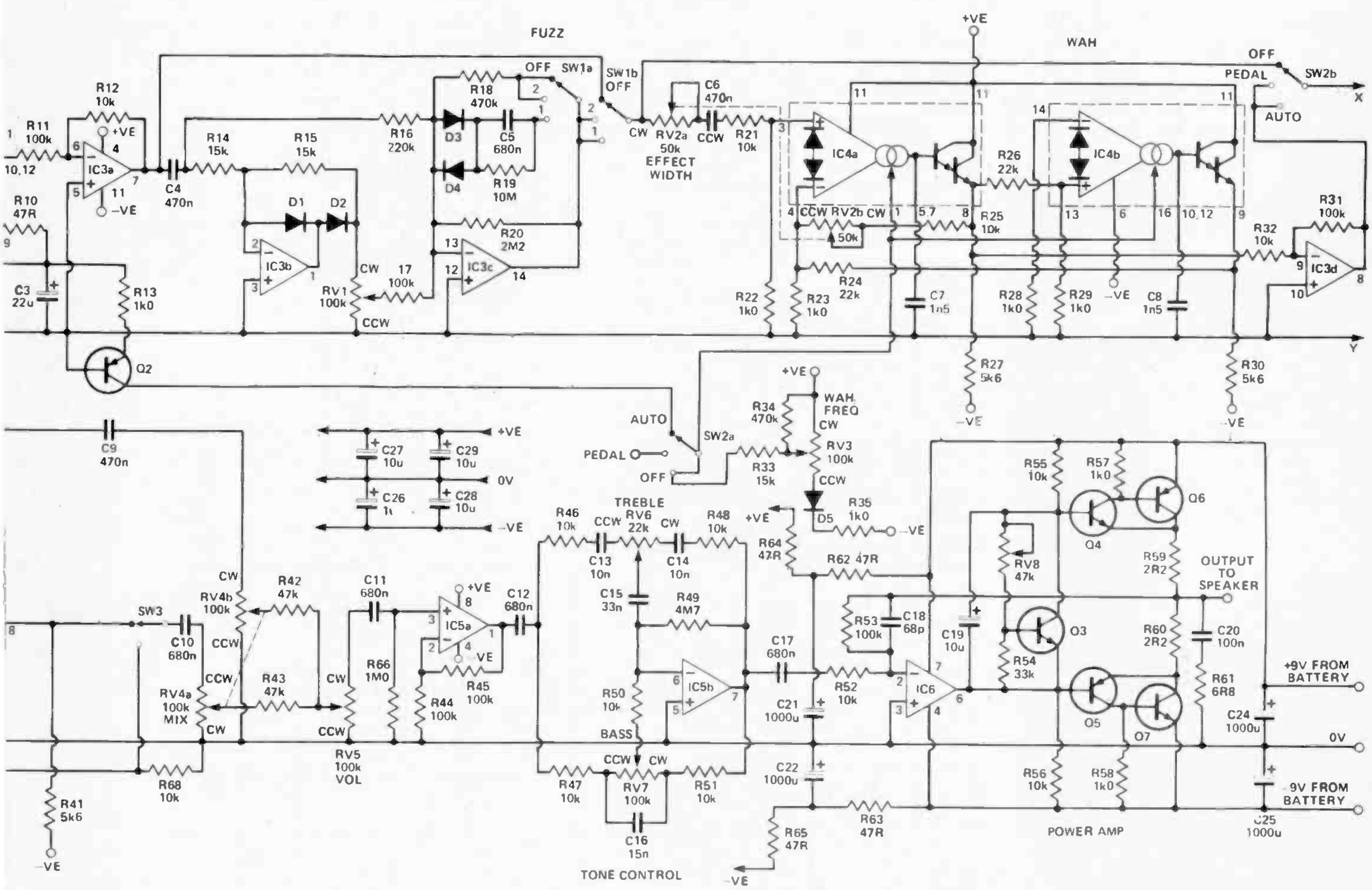


Fig. 5 Here the LM13600 is configured as a compressor.

The LM13600 As An Expander

If the current produced by the above circuit is fed into the bias current input of a virtually indentical stage while the diode current is held a constant then the voltage gain equation above shows tha the gain of the circuit will be increased as the current increases. Moreover the product of the two gains will be constant giving an invariant overall signal transfer function.



The State-variable Filter

Things start getting a bit heavy now! The following equations are the transfer equations for a state-variable filter such as that in Fig. 2:—

$V_x = [(R1/R3) \times V_{in} + (R2/R3) \times V_y] + R5/(R4 + R5) \times [1 + R3(R1 + R3)/(R1R2)] \times V_o$

If R1 = R2 = R3 = R5, then:

$V_x = -(V_{in} + V_y) + R1(1 + 2)/(R4 + R1) \times V_o$

$$V_x = \frac{3R1V_o}{R4 + R1} - (V_{in} + V_y) \quad (1)$$

$$V_o = -1/sCR \times V_x,$$

$$V_x = -sCR\,V_o \quad (2)$$

$$V_y = -1/sCR \times V_o \quad (3)$$

Substitute (2) and (3) in (1):—

$$-sCR\,V_o = \frac{3R1\,V_o}{(R1 + R4)} - (V_{in} + (-1)/sCR\,V_o)$$

$$\frac{V_{in}}{V_o} = (sCR + 3R1/(R1 + R4) + 1/sCR)$$

$$= j\omega CR + \frac{3R1}{(R1 + R4)} + 1/j\omega CR$$

Compare this with the equation for (s = jω, ω = 2π f, f = frequency) an LCR tuned circuit:—

$$V_{in} = (j\omega L + 1/(j\omega C) + R)\,I_{out} \quad \text{(Fig. 3)}$$

$$\frac{V_{in}}{I_{out}} = j\omega L + R + 1/(j\omega C) \quad \text{(LCR circuit)}$$

$$\frac{V_{in}}{V_{out}} = j\omega CR + \frac{3R1}{(R1 + R4)} + 1/(j\omega CR) \quad \text{(State variable filter)}$$

From this it is apparent that these responses are similar except that the LCR circuit gives a current output rather than a voltage.

For this type of LCR circuit, the frequency of minimum attenuation or maximum gain (the resonant frequency) is given by:—

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

For our circuit this is:—

$$\frac{1}{2\pi\sqrt{CR.CR}} = \frac{1}{2\pi CR}$$

The 'Q' factor influencing the bandwidth is given by:—

$$Q = \frac{2\pi f L}{R} = \frac{2\pi \times (1/2\pi CR) \times CR}{3R1/(R1 + R4)} = \frac{R1 + R4}{3R1}$$

Figure 4 shows the configuration necessary to use the LM13600 as a filter of this type. Last month we found that the transfer function of the LM13600 with no diode current is given by:—

$$I_{out} = \frac{I_{abc}}{26 \times 10^{-3}} \times V_{in}$$

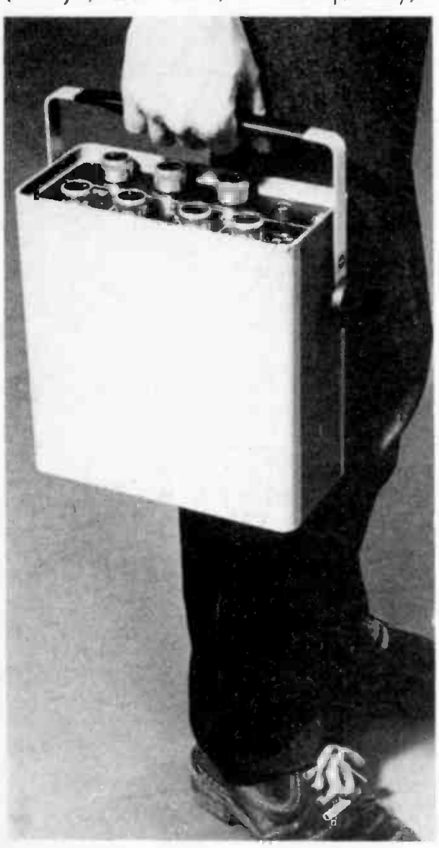
As we are using a capacitative load this output current will generate a voltage of:—

$$V_{out} = I_{out} \times \frac{1}{j\omega C}$$

$$= V \times \frac{I_{abc}}{26 \times 10^{-3}} \times \frac{1}{j\omega C}$$

$$= V_{in}\frac{R1}{R1 + R} \times \frac{I_{abc}}{26 \times 10^{-3}} \times \frac{1}{j\omega C}$$

Since we are not dealing with a normal type of op-amp, analysis of the circuit is not as easy as the normal filter but the result is an equation of much the same form. We do not show all the derivation here as it would occupy most of a page. However, the end result is that the centre frequncy of the filter pass band is proportional to the amplifier bias current. Therefore we have an easy way to control the wah wah effect.



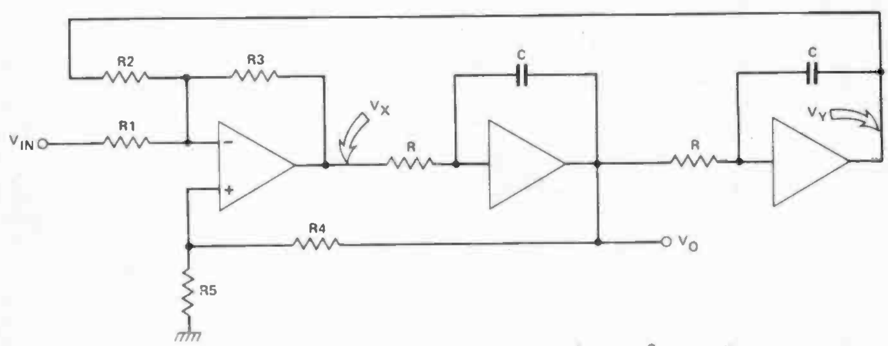


Fig. 2 (Above) Generalised state-variable filter.

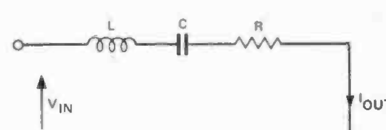


Fig. 3 (Right) Generalised LCR tuned circuit.

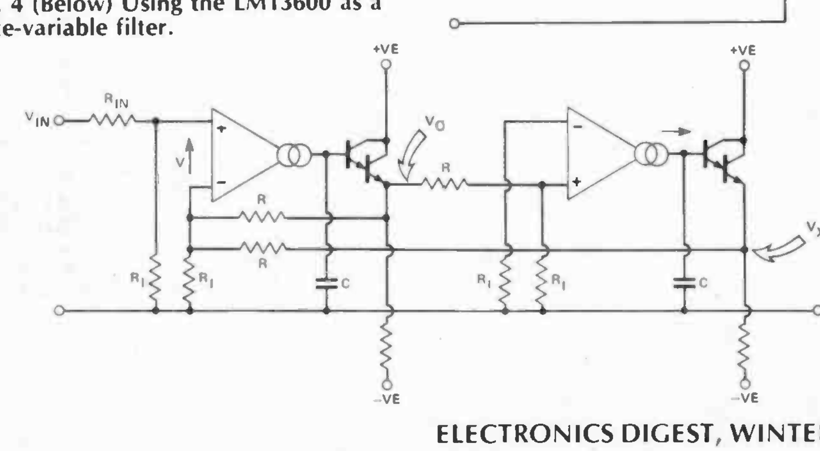


Fig. 4 (Below) Using the LM13600 as a state-variable filter.

PARTS LIST

Resistors (all ¼W 5%)

R1,3,42,43	47k
R2,5,6,12, 21,25,32,37, 38,46-48, 50-52,55,56, 68	10k
R4,16	220k
R7,18,34	470k
R8	6k8
R9,14,15,33	15k
R10,62-65	47R
R11,17,31, 36,40,44,45, 53,67	100k
R13,22,23, 28,29,35,57, 58	1k0
R19	10M
R20	2M2
R24,26,39	22k
R27,30,41	5k6
R49	4M7
R54	33k
R59,60	2R2
R61	6R8
R66	1M0

Potentiometers

RV1,3,7	100k linear
RV2	50k linear dual gang
RV4	100k linear dual gang
RV5	100k logarithmic with two-pole switch
RV6	22k linear
RV8	47k miniature horizontal preset

Capacitors (all polycarbonate except where stated)

C1,5,10-12, 17,23	680n
C2	220n
C3	22u 16 V tantalum
C4,6,9	470n
C7,8	1n5
C13,14	10n
C15	33n
C16	15n
C18	68p
C19	10u 35 V tantalum
C20	100n
C21,22,24, 25	1000u 10V axial electrolytic
26-29	47u 10 V PCB electrolytic

Semiconductors

IC1	CA3140
IC2,4	LM13600
IC3	TL084
IC5	TL082
IC6	741
Q1,2,5	BC212L
Q3,4	BC182L
Q6	BD132
Q7	BD131
D1-5	1N4148

Miscellaneous

SW1,2	2-pole, 3-way miniature slide switch
SW3	1-pole, 2-way miniature slide switch

PCBs (see Buylines); case (220 × 105 × 230 mm), Vero ref. 75-2443A; wire (single, single screened and double screened); 4" or 5" loudspeaker (8 ohms), 5 W); standard ¼" jack socket for input; stereo jack socket for foot switch (if required); 75 mm of 12 × 12 mm aluminium angle; two PP9 batteries and clips.

Construction

Except for the controls, almost all the components for this project are mounted on the three PCBs. The preamplifier board is the largest and most densely packed. It is advisable to use sockets for the ICs and don't forget the links. The capacitors used in the project should be as small as possible otherwise you will have difficulty fitting them. A fine tipped soldering iron will be useful when assembling the boards and care should be taken to avoid creating short circuits between tracks with accidental solder splashes.

Ensure that all the diodes, transistors and other polarised components (especially IC3) are fitted the correct way round. On the power amplifier board the output transistors are mounted on top of a

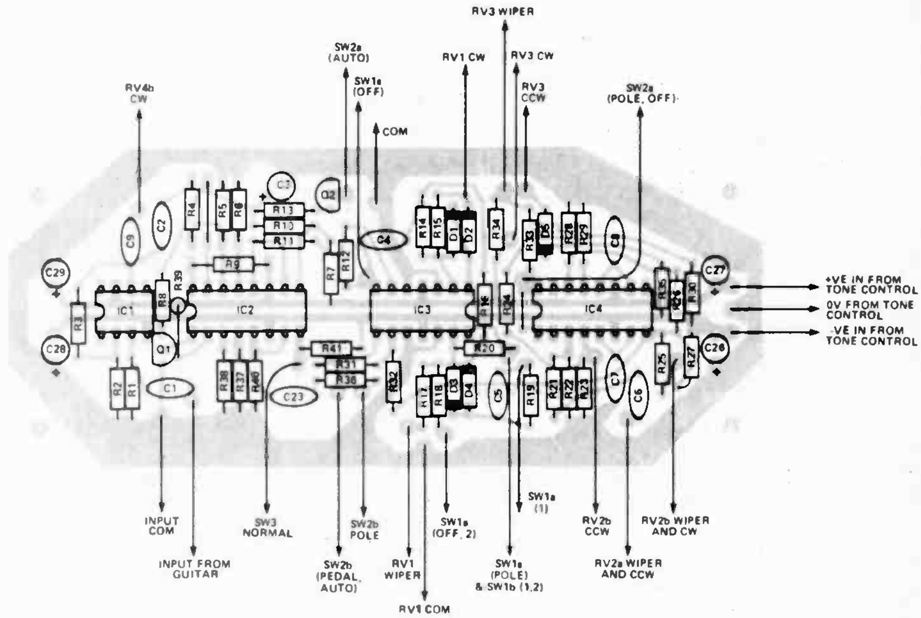


Fig. 5 (Above) Component overlay for the preamplifier board.

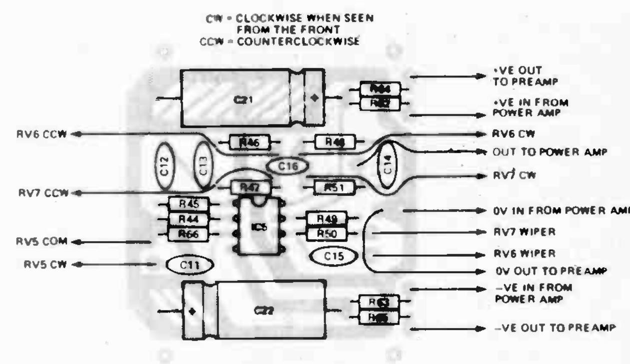


Fig. 6 (Right) Overlay for the tone control board.

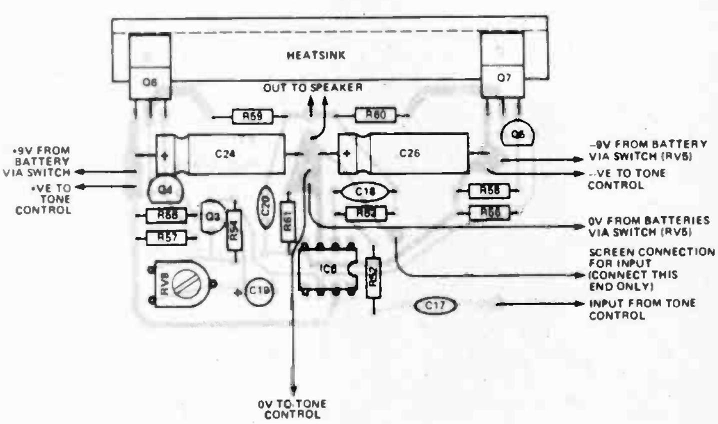
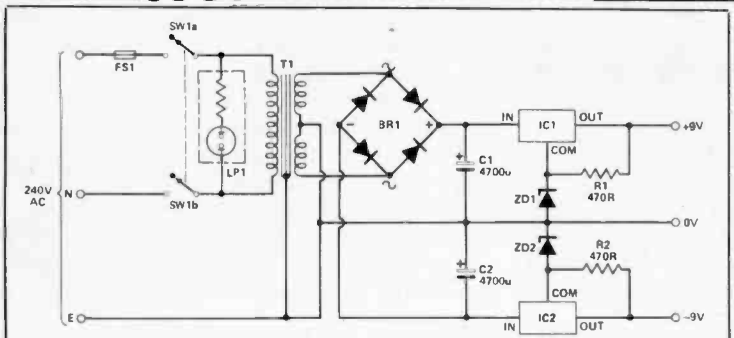


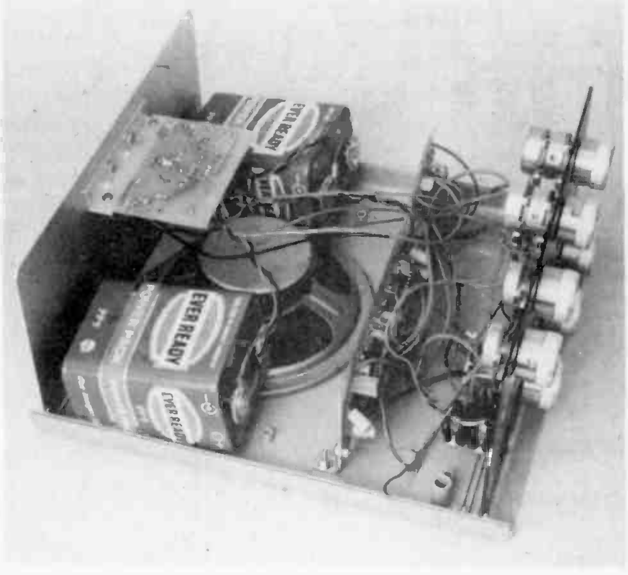
Fig. 7 (Right) Overlay for the power amp of the Playmate.

SUGGESTED MAINS UNIT



Parts List

R1,2	**470R ¼W 5%**	**T1**	**9-0-9 12 VA transformer**
C1,2	**4700u 25 V electrolytic**	**SW1**	**DPDT miniature rocker or slide switch**
IC1	**7805**	**FS1**	**1 A 20 mm fuse and holder**
IC2	**7905**	**LP1**	**240 V panel neon with integral resistor**
ZD1,2	**3V9 400 mW zener**		
BR1	**W01 (1 amp bridge rectifier)**		



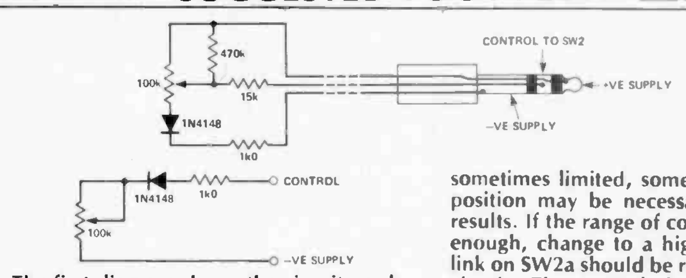
short length of ½" × ½" aluminium angle which acts as a heatsink. The transistors and the angle are held in position by the transistor mounting screws.

Mount the control switches and potentiometers on the front panel (see photo for our layout) and make the necessary interconnections and fit the three components needed around the balance control. The wiring from the front panel to the preamplifier/effects and the tone control boards should be carried out using thin flexible wire for control signals and miniature screened cable for any sound signals. These should be short but allow enough slack to be able to fit them in position easily. (It is probably easier to connect all the wires to the circuit boards first.) The power amplifier board was fitted to the metal base plate of our case using the angle on which the transistors were mounted. It was positioned so that it fitted neatly between the two batteries in the bottom of the case. Together with the loudspeaker the amplifier board holds the batteries without any further help.

The small loudspeaker for this project was mounted on the plastic case of the box through which a number of holes were drilled to let the sound out. If required a small mains power unit capable of 9-0-9 V at 100 mA or more may be used to power the unit.

A foot pedal control for controlling the wah wah effect could be plugged into a jack socket. This would need to be one of the three-connection or stereo type so that positive and negative supplies as well as the control signal could be connected.

SUGGESTED FOOT PEDAL



The first diagram shows the circuit used with a three-connection (stereo) jack. This will duplicate the effect of the panel-mounted control. As the angular rotation of foot pedal controls is sometimes limited, some adjustment of position may be necessary to get best results. If the range of control isn't wide enough, change to a higher value. The link on SW2a should be removed for this circuit. The second diagram shows a simple circuit for use with ordinary jack connectors which can be used if the link between SW2a pole and off position is retained.

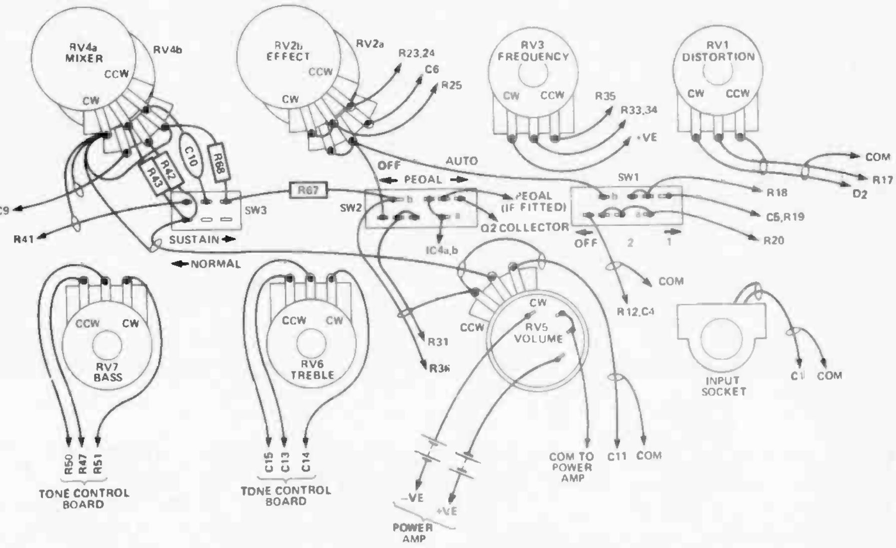


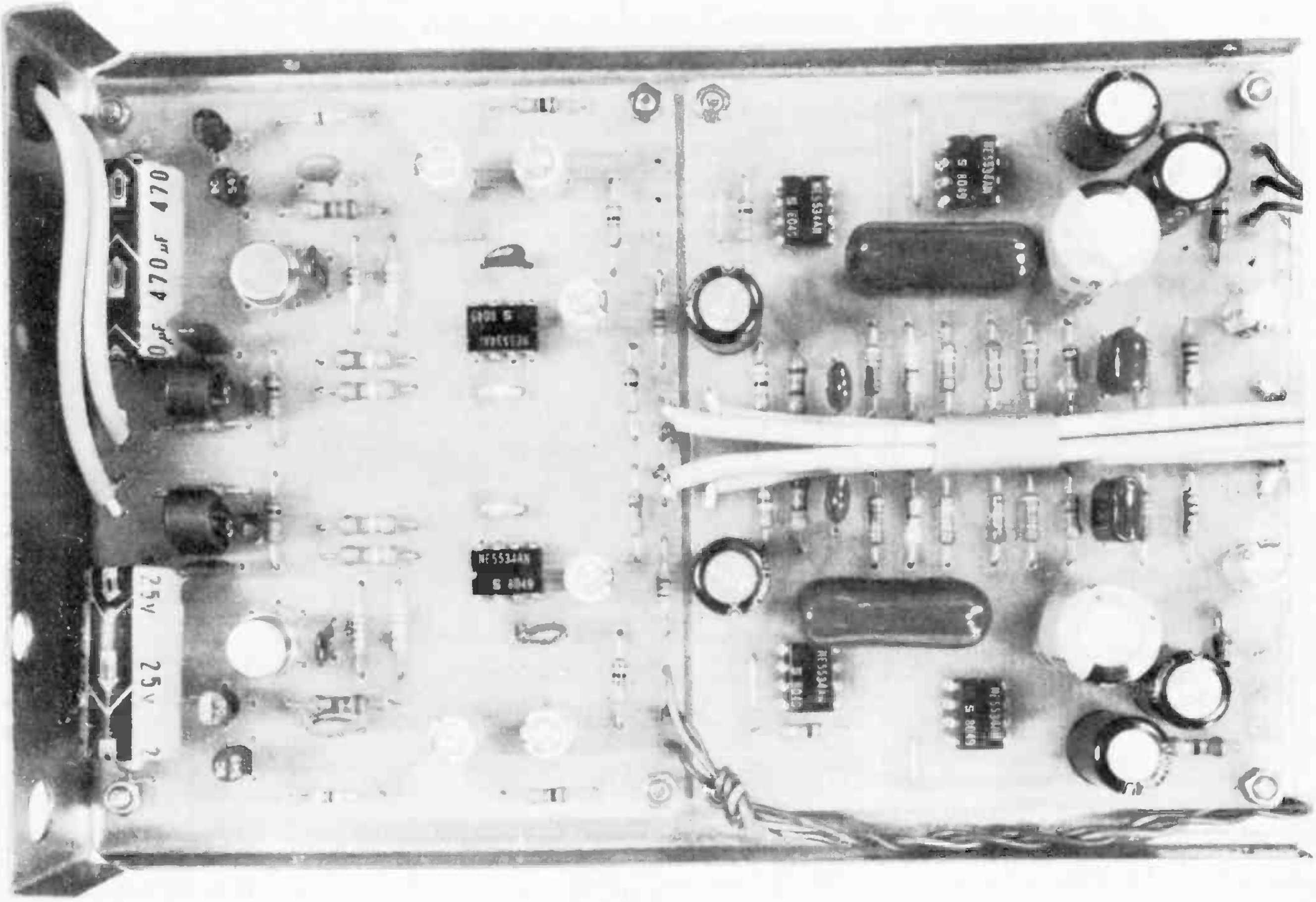
Fig. 8 How to wire the front panel. Compare with the front panel photo last month.

BUYLINES

There shouldn't be anything here to cause problems; most of the components are standard items and several mail order companies now stock the LM13600. Shop around in our advertisements for the best price. The set of three PCBs can be ordered from our PCB Service as advertised on page 91, should you not wish to etch your own.

HIGH QUALITY PHONO AMPLIFIERS

In ETI in January 1982 we gave an insight into the design procedures for low level phono signals. Here the theory is put into practice with these outstanding input amplifiers. Design and development by David Tilbrook.



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

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

Equal Change

In an attempt to overcome bass problems the RIAA has proposed a change to its playback equalisation curve. The extreme bass frequencies are attenuated on playback by the addition of another time constant. This takes the form of a single-pole RC filter with a time constant of 7950 uS, ie: a −3 dB point of 20 Hz. Since the

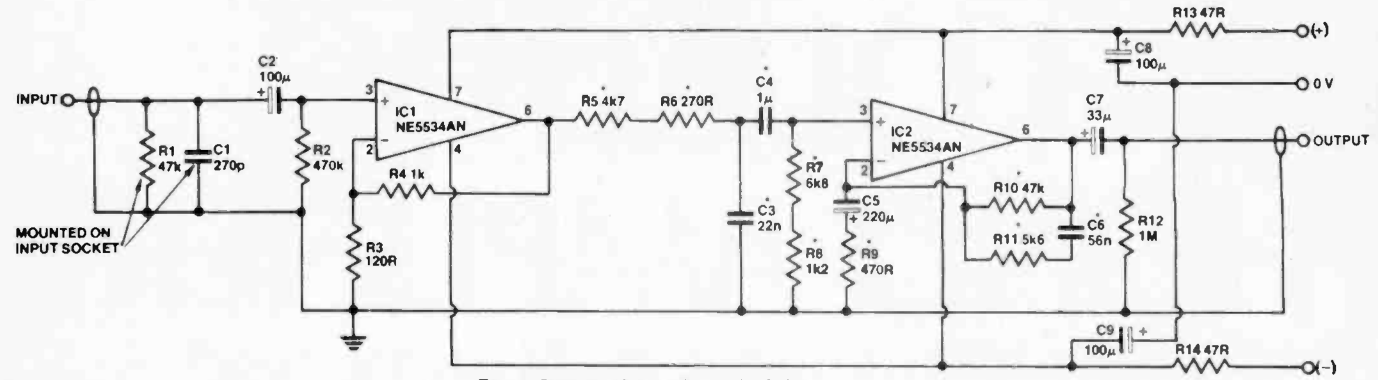


Fig. 1 Circuit of one channel of the moving magnet input stage. Note that the RIAA equalisation is incorporated in this stage.

* DENOTES COMPONENTS ASSOCIATED WITH THE RIAA EQUALISATION

frequency response is already flattened by the 3150 uS time constant, this new time constant gives a 6 dB attenuation rate below about 20 Hz. The resulting RIAA playback equalisation is shown in Fig. 2. Note that there are four time constants associated with the proposed RIAA equalisation: 7950 uS, 3150 uS, 318 uS and 75 uS. These are shown on the Bode plot, which is the dotted line in Fig. 2. It should be emphasised, however, that the introduction of this low frequency time constant is not sufficient to remove severe cases of turntable or tonearm resonance. Some preamps incorporate multiple-order subsonic filters that offer a very fast roll-off below 20 Hz. The problem with this is that severe cases of tonearm resonance or rumble generate distortion harmonics well above 20 Hz and into the audio spectrum. The only real cure is to remove the problem at the turntable or tonearm.

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

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

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

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

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

HOW IT WORKS

MOVING-MAGNET STAGE

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

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

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

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

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

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

The 1M0 resistor R12 ensures that the DC voltage on the output remains at 0 V. This is important so that operation of the selector switch following the stage will not cause thumps in the output.

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

Moving Coil Input

The complete circuit diagram for the moving coil input stage is shown in Fig. 4. The collectors of the LM394 are connected to the input of an NE5534, which functions as a high-gain differential amplifier, providing adequate open loop gain to ensure low distortion and a flat frequency response when negative feedback is applied. The input choke is used to minimise the stage's susceptibility to RF noise.

The input impedance of the stage is determined by the parallel combination of R1 and R2, around 65 ohms for the values shown. This should be suitable for most moving coil cartridges, but is easily changed if required. The DC operating point of the LM394 is determined by the constant current source formed by Q1, Q2, R3 and R6. So the current in resistor R2 is determined by this constant current source and the DC current gain of the LM394. Hence the value of R2 can be increased, in order to increase the input

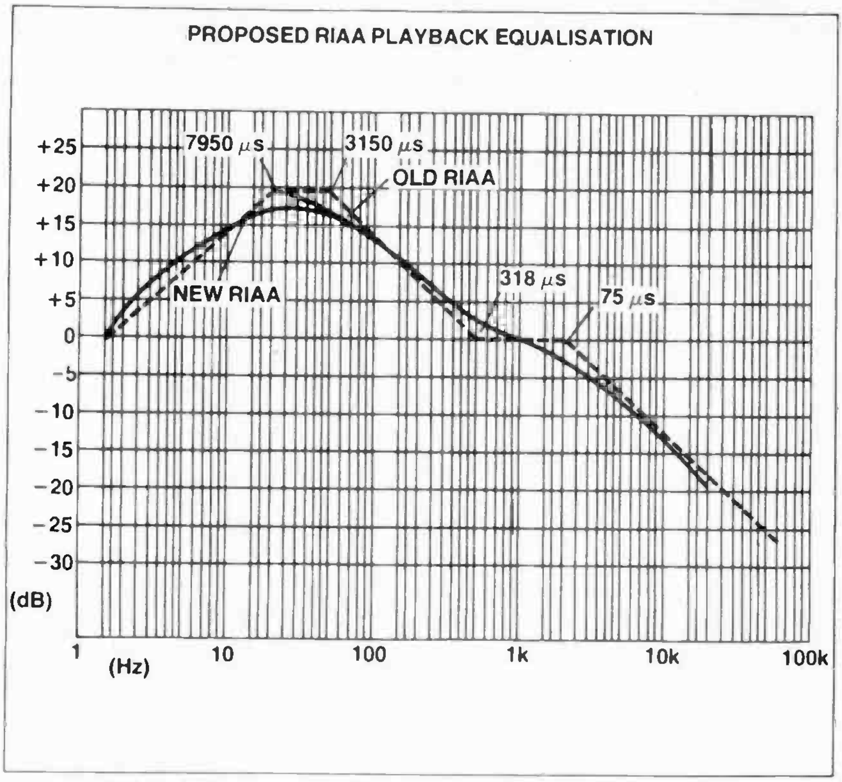


Fig. 2 Old and 'new' RIAA equalisation curves (solid line). The individual time constants (Bode plot—dotted lines) to produce the response are also shown.

PARTS LIST

MOVING-MAGNET STAGE

Resistors (all ¼W metal film, 5% unless stated otherwise)

R1,101	47k
R2,102	470k
R3,103	120R
R4,104	1k0
R5,105	4k7 1%
R6,106	270R 1%
R7,107	6k8 1%
R8,108	1k2 1%
R9,109	470R 1%
R10,110	47k 1%
R11,111	5k6 1%
R12,112	1M0
R13,14, 113,114	47R

Capacitors

C1,101	270p ceramic
C2,102	100u 16 V PCB electrolytic
C3,103	22n polyester
C4,104	1u0 polyester
C5,105	220u 16 V PCB electrolytic
C6,106	56n polyester
C7,107	33u 25 V PCB electrolytic
C8,9,108,109	100u 25 V PCB electrolytic

Semiconductors

IC1,2,101,102	NE5534AN

Miscellaneous

PCB (see Buylines); assorted mounting hardware; shielded cable.

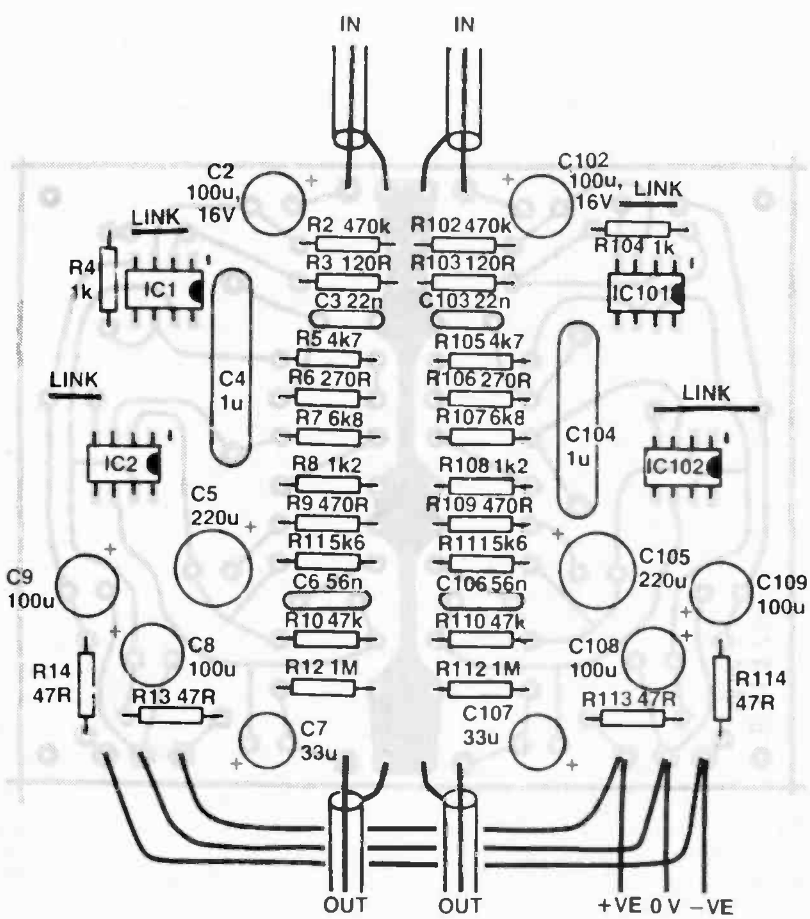
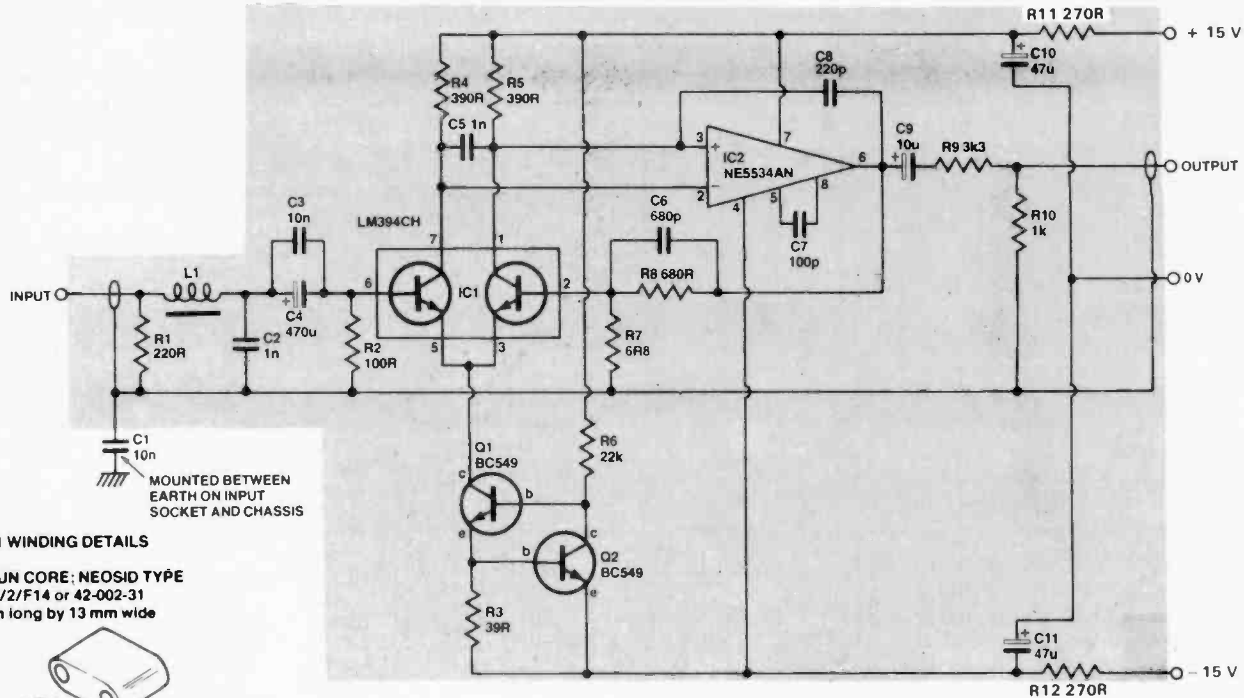


Fig. 3 Component overlay for the moving magnet stage.

impedance, over a fairly wide range of values without affecting the operation of the circuit.

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

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



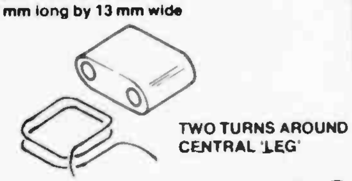


Fig. 4 Circuit of one channel of the moving coil input stage. Components for the other channel are designated R101, C101, IC101, etc.

HOW IT WORKS

MOVING-COIL PREAMP

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

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

The collectors of the LM394 are shunted by the 1n0 capacitor C5. This decreases the gain of the first stage at high frequencies and helps to ensure stability (ie: freedom from high frequency oscillations).

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

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

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

BUYLINES

These low level input stages have been designed to deliver state-of-the-art performance — as we've not compromised on the design we suggest you don't compromise on the components. Both the MM and MC stages use high performance NE5534AN op-amps; a possible alternative to this is the TDA1034 op-amp. Accept no substitutes. The LM394 and NE5534 are available from Watford Electronics, as is the Neosid balun core; the latter item can also be obtained from Neosid. Small Orders, PO Box 86, Welwyn Garden City, Herts AL7 1AS. Technomatic stock the LM394 and the TDA1034. The PCBs are essential to preserve the layout and earthing; a very necessary requirement if the full performance is to be achieved. Boards will be available from our PCB Service at the prices listed on page 91.

Construction

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

The inductor on the input of the MC stage consists of two turns wound on a ferrite balun core, 6 mm long by 13 mm wide. We used the type given in the Parts List.

Each of the PCBs is a stereo input amplifier, with each channel sharing a common input earth track running down the centre of the board. The power supply wiring from each channel on the board can be connected in parallel, so only three wires (+,0,−) need to be brought out for power.

The input earth is *not* connected to the 0 V line from the power supply at any place on the PCBs. This means that without a separate 0 V connection added to the input stage they will *not* work. This has been done deliberately

SPECIFICATION

MOVING-MAGNET INPUT STAGE

Gain:	74, 1 kHz
Frequency response:	Conforms to RIAA Equalisation ±0.2 dB (This is the performance of the prototype. The actual figure obtained will be determined by the accuracy and longterm stability of the components used.)
Total harmonic distortion:	<0.001%, 1 kHz, 10 mV RMS input
Headroom:	>28 dB with respect to 5 mV RMS input signal ie:135 mV RMS max.
Noise:	Total equivalent input noise: 112 nV 'A', input shorted, 216 nV flat, input shorted.

S/N ratio:

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

MOVING-COIL INPUT STAGE

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

S/N ratio of MC stage after RIAA Equalisation:

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

RESPONSE

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

PARTS LIST

MOVING-COIL STAGE

Resistors (all ¼ W metal film, 5%)

R1,101	220R
R2,102	100R
R3,103	39R
R4,5,104,105	390R
R6,106	22k
R7,107	6R8
R8,108	680R
R9,109	3k3
R10,110	1k0
R11,12, 111,112	270R

Capacitors

C1,3,101,103	10n polyester
C2,5,102,105	1n0 polyester
C4,104	470u 16 V axial electrolytic
C6,106	680p ceramic
C7,107	100p ceramic
C8,108	220p ceramic
C9,109	10u 16 V PCB electrolytic
C10,11, 110,111	47u 25 V PCB electrolytic

Semiconductors

IC1,101	LM394CH
IC2,102	NE5534AN
Q1,2,101,102	BC549

Miscellaneous

L1	Two turns on ferrite balun core, Neosid type 1050/2/F14 or 42-002-31

PCB (see Buylines); assorted mounting hardware; shielded cable.

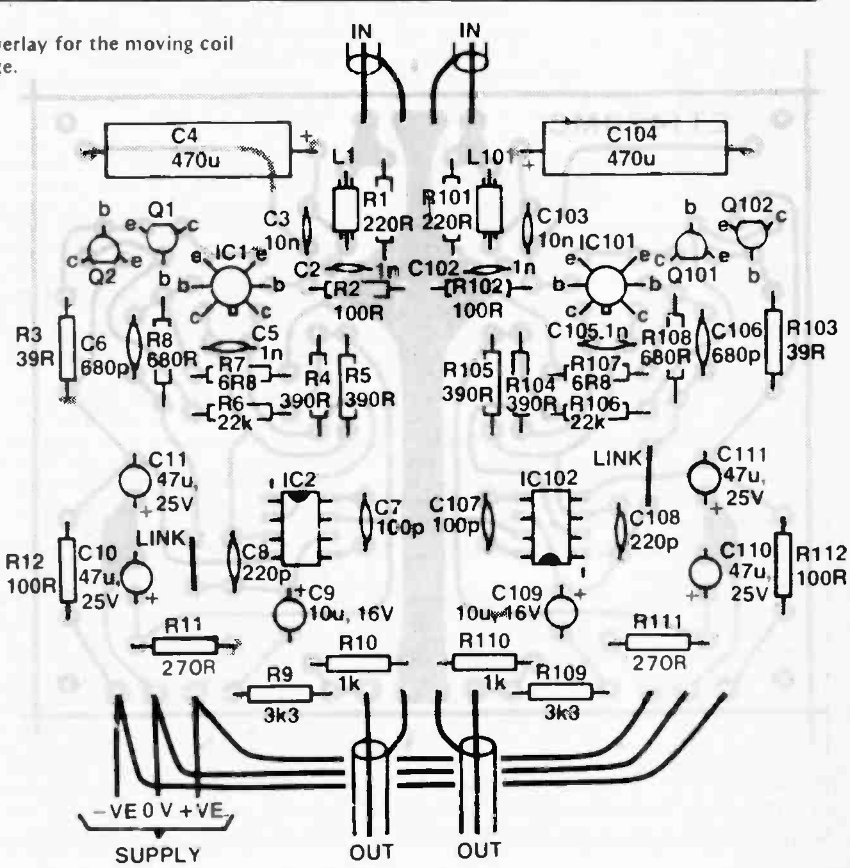
to ensure that hum present on the earth line, due to supply bypass capacitors for example, cannot modulate the signal earth, producing hum in the output. The 0 V line on the boards is, in fact, a separate supply bypass earth line and is not equivalent to the signal earth. A separate wire should be run from the centre point (0 V point) of the power supply used to signal earth at the input sockets.

Both boards should be mounted in a *steel* box which can be mounted as a unit inside the main preamp chassis. This greatly improves the rejection to 50 Hz magnetic fields generated by nearby power transformers or 240 V cables.

Powering Up

No setting-up procedure is required for either stage, but make a final check of all components before applying power to the unit.

Fig. 5 Overlay for the moving coil input stage.



UPGRADING AMPLIFIER PSUs

Even hardened DIY types may quail at the thought of tinkering with their expensive commercial hi-fi, but an improvement can often be made simply by beefing up the power supply. Phil Walker discusses PSU requirements and offers a ripple monitor to check whether your supply is up to the job.

Perhaps it is time to blow the dust and pine needles out of the hi-fi power supply corner. Is it really up to the job or is the mud at the bottom of the audio pond?

What's in a power supply? Or what SHOULD be in a power supply?

Here we examine what steps we must take to determine exactly what is necessary to get the best results out of your audio system.

In order to get anywhere we must first decide or find out what power output is required and into what load impedance we are going to put it. For an existing amplifier these are already defined, so don't be greedy!

From these facts we can determine the voltage and current flowing in the load, thus making the estimate of the power supply requirements. From basic theory we have:-

$V = I \times R$ (Ohm's Law)
$P = V \times I$ (Power Law)
(V = volts, I = amps, R = ohms, P = watts)
From this we get:-

$P = V^2/R$ or $V = \sqrt{(P \times R)}$
also
$P = I^2 \times R$ or $I = \sqrt{(P/R)}$

To get an estimate of the peak voltage required we approximate the signal to a sine wave. For this particular case the peak voltage is $\sqrt{2}$ (or 1.414) times the RMS value given by the above equations.

To go further now we must know what sort of output stage your amplifier has. In general there are three main types.

Split Rail

Each power rail supplies the load voltage and current for approximately half the cycle (Fig. 1.). The voltage required from each rail is:-

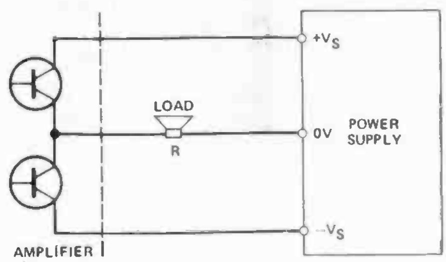


Fig. 1 Split rail power supply configuration.

$$V = \sqrt{(P \times R)} \times \sqrt{2} + X$$

(X is a factor to account for practical conditions — see later)
The average current required from each rail is:-

$$I = \sqrt{(P/R)} \times \sqrt{2} \times 0.6 \times 0.5$$

(The 0.5 factor is because the load current flows for half the output cycle only; the 0.6 factor is because the average of a sine wave is about 0.6 of the peak value) A reasonably safe approximation for the current is:-

$$I = \tfrac{1}{2} \times \sqrt{(P/R)}$$

Single Rail

The power rail supplies current on the positive half cycle of the signal only (Fig. 2). This drives the load while at the same time recharging the output capacitor C which has previously supplied the negative half cycle of the output signal. The positive end of C in the diagram normally rests at about half the supply voltage when no signal is applied. The supply voltage for this configuration is the same as the total for the split rail version and the voltage and current values found there apply.

Bridge

In this configuration the power supply provides the output current over the whole of the cycle. Therefore the average current is:-

$$I = \sqrt{(P/R)} \text{ (approximated as before)}$$

The supply voltage is effectively switched so that the positive and negative half cycles of the signal are drawn from the same supply rails. Therefore we have:-

$$V = \sqrt{(P \times R)} \times \sqrt{2} + Y$$

(Y is a factor to account for practical conditions — not the same as X above.)

What About X and Y?

Well, these are just factors which are included to make up for unavoidable imperfections of practical semiconductor devices. Bipolar and field-effect transistors always drop some volts when conducting and their driving circuits often require even more.

As a general approximation:-

for bipolar transistors	$X = 4$	$Y = 8$
and for MOSFETs	$X = 8$	$Y = 16$

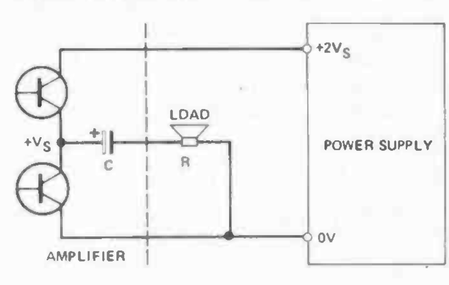
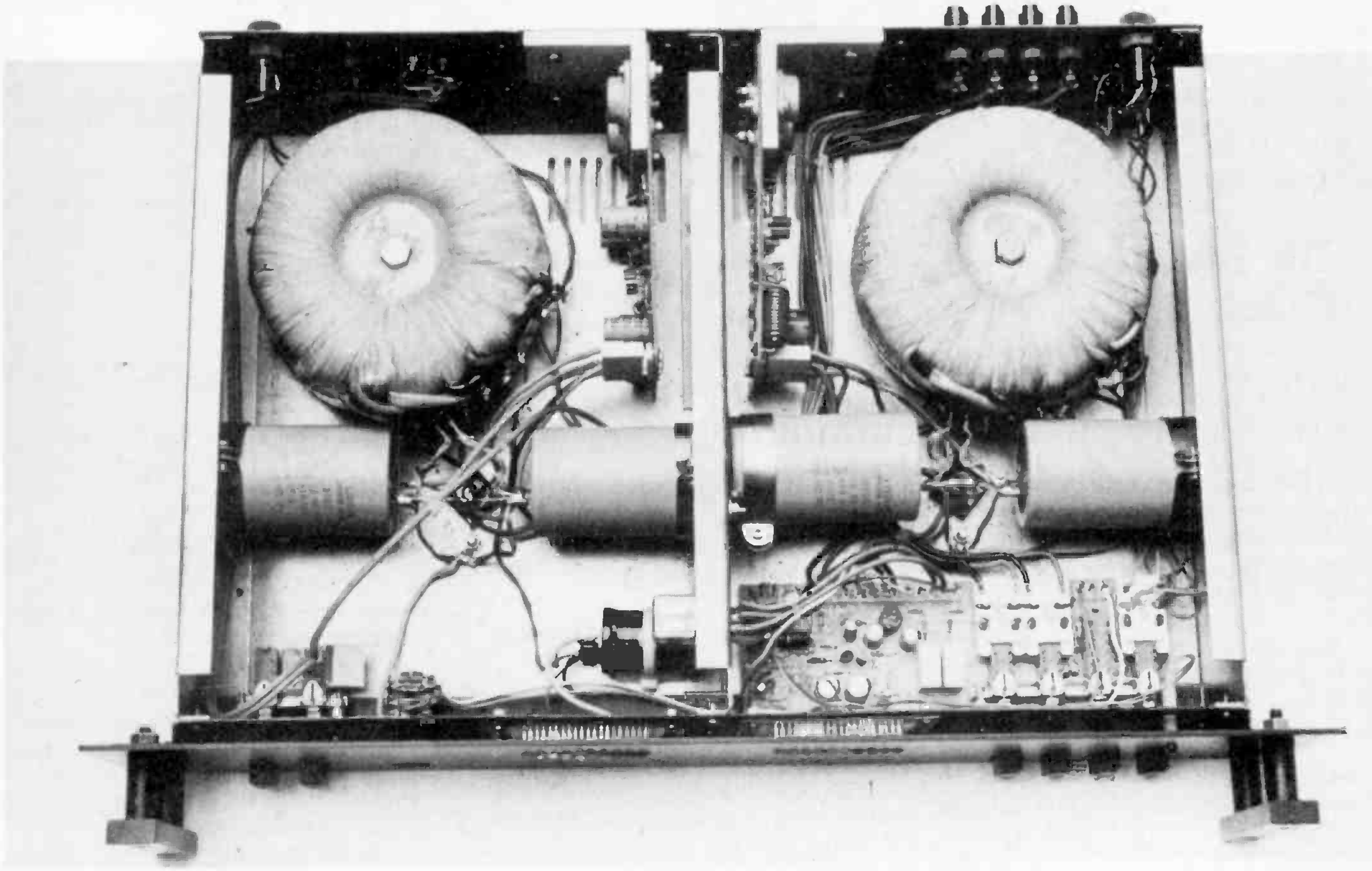


Fig. 2 Single rail configuration.



Most of the space in this 100 W power amplifier is taken up by the power supply components — and quite right too!

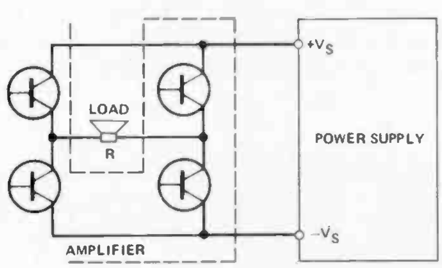


Fig. 3 (Left) Power supply configuration for a bridge amplifier.

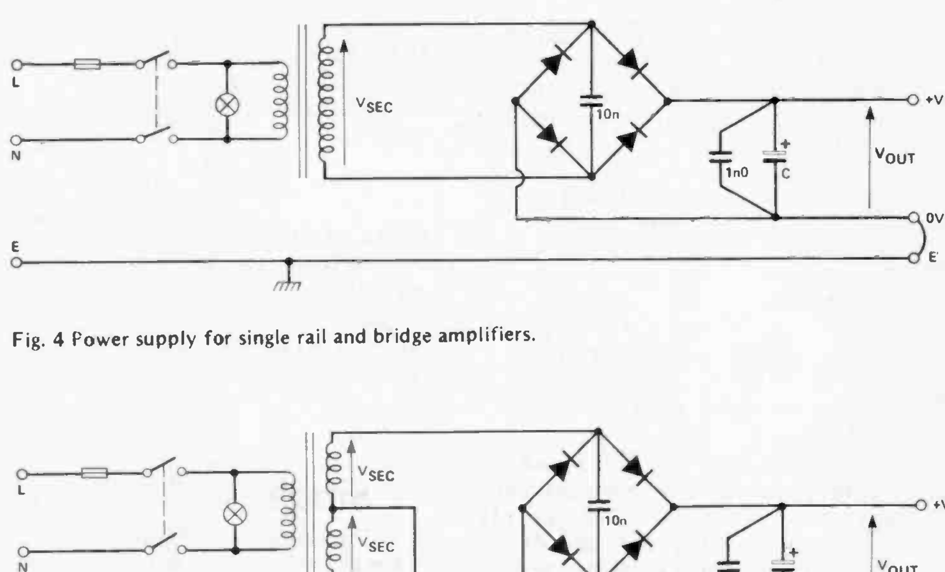


Fig. 4 Power supply for single rail and bridge amplifiers.

What About Class A?

Up to now we have only considered class B or similar output stages but some people may well have class A amplifiers. If this is the case, the power supply will have to be rated to provide a continuous current √2 or 1.4 times as great as found above.

The Power Supply Transformer

Now that we know the voltage and current required, we can determine what transformer will be required. As a general rule the current rating of the transformer secondary will be the same or slightly greater than the required DC output.

The voltage rating will be 1/√2 (0.707) times the required rail voltage plus a volt or two to make up for rectifier losses. The configurations shown in Figs. 4 and 5 are the most usual and will cater for most needs.

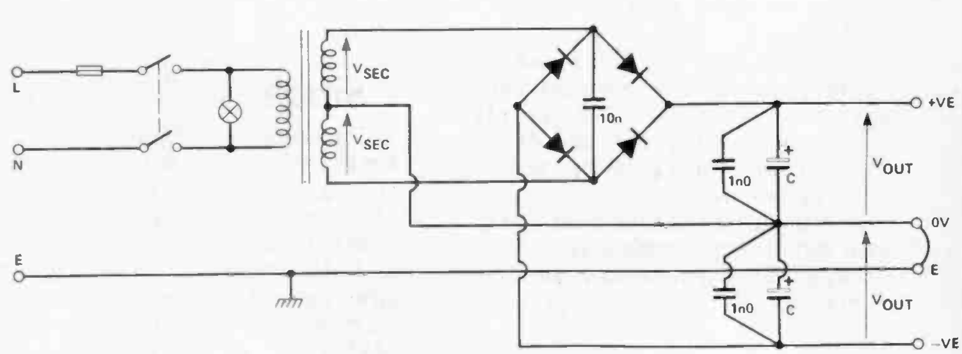


Fig. 5 Power supply for split rail amplifiers.

For single rail & bridge amplifiers:

$$V_{sec} = (V_{out} + 2)/\sqrt{2}$$

For split rail amplifiers:

$$V_{sec} = (V_{out} + 1)/\sqrt{2}$$

The Smoothing Capacitor

The main thing to bear in mind about this component is that the larger its capacitance and the better its quality, the better it will work.

The effect of this component is to reduce the ripple voltages due to the rectified mains and signal frequency currents flowing through it. If either of these currents produce an excessive voltage, the performance of the ampilifer is likely to be impaired.

An approximation to the mains derived ripple voltage developed across the capacitor can be found from the basic expression for capacitance:-

$C = Q/V$

(C in Farads, Q in Columbos, V in Volts)

Also $Q = I \times t$

(I in Amps, t in seconds)

Therefore:-

$$C = \frac{I \times t}{V}$$

For 50 Hz mains input the maximum time between capacitor charges is 10 ms. (Assuming full wave rectification.) The amount by which the capacitor discharges between charges is the ripple voltage and is dependent on the load current. Rearranging the last equation we have:-

$$V = \frac{I \times t}{C}$$ ($t = 10^{-2}$ for 50 Hz, full wave rectified)

From this it is obvious that C should be as large as possible to keep the ripple small. For example, if the load current is 2 A and we wish to know what the ripple voltage is with a 10,000uF capacitor, we have:-

$$V = \frac{2 \times 10^{-2}}{10{,}000 \times 10^{-6}} = 2 \text{ Volts (peak to peak)}$$

This about the figure required for a reasonable compromise between cost and effectiveness.

The capacitor should have a ripple current rating of at least three times the DC output current.

The Rectifier

Whether this is to be a single unit or separate diodes, the working voltage for each section must be a minimum of 1.5 times the *total* secondary voltage of the transformer and for preference threetimes it. It must have a current rating at least equal to the load current and a surge capability of:-

$$I_{surge} = \frac{V_{DC} \times C}{t}$$

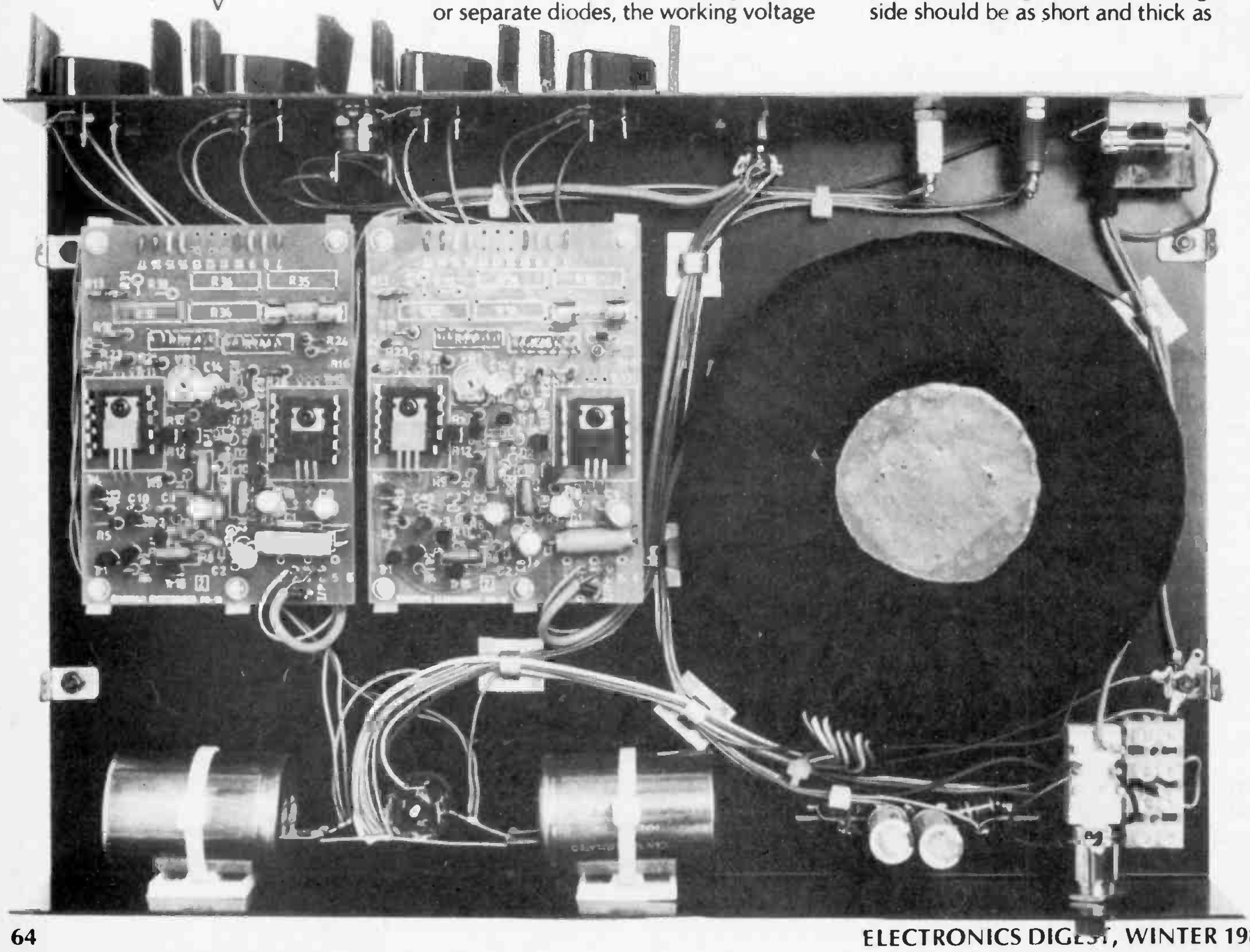
(I_{surge} in Amps, V_{DC} in Volts, $t = 10^{-2}$)

Bear in mind that this device may require some form of heatsink when supplying large currents.

Other Points

The rest of the power supply is mainly up to the individual but some things are worth mentioning.

The mains wiring should, of course, be completely safe with no exposed connections accessible even with the covers removed. A suitable slow-blow fuse, switch and indicator lamp should be included. The mains earth wire should be securely connected to all the metalwork and provision for connection to the amplifier brought to a convenient point. The wiring on the low voltage side should be as short and thick as



practicable. The possibility of earth loops must be avoided as far as possible.

The mains input to the transformer may be shunted with a capacitor of 10nF or so, to clean up any high frequency noise on the mains. *These must be made for this purpose* — 250 V AC working or better.

The rectifier bridge may be shunted with a 10nF capacitor of suitable voltage rating to reduce switching noise. It is recommended that the main reservoir capacitors also be shunted with non-electrolytic capacitors of 1uF or so to reduce their high frequency impedance.

For the best results each amplifier should be fed from a separate supply so that loading on one is not coupled to another.

For the purist, each power rail would be fully regulated and a proper mains filter would be used to eliminate any remaining mains-borne interference.

Power Supply Ripple Monitor

The purpose of this circuit is to continuously monitor the ripple voltage on a capacitor and light an LED if that voltage exceeds about 2 V peak to peak. This will show up those peaks in your programme material which cause distress in the power supply and which may be detracting from the overall clarity of reproduction.

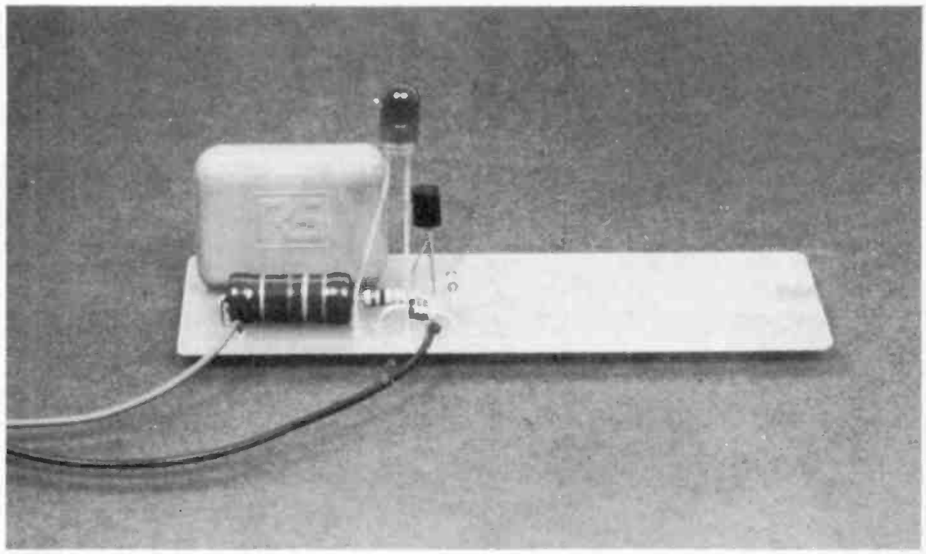
Construction

Construction is very simple and straightforward, provided care is taken to ensure that the polarity of each semiconductor is correct.

The PCB is laid out with two circuits in mirror image to allow stereo pairs of amplifiers or single split rail amplifiers to be accommodated easily.

If desired, the polarity of all the semiconductors may be reversed using a BC212L transistor instead of the BC182L.

Resistors (all ¼W, 5%)	
R1	**1k0**
R2	**10k**
R3	**See Table 1**
Capacitors	
C1	**1u0 250V polycarbonate**
Semiconductors	
Q1	**BC182L**
ZD1	**5V6 400 mW zener**
ZD2	**27v 400 mW zener (required if supply voltage over 35v)**
LED1	**red LED**



The dual ripple monitor PCB with one circuit built onto it. For use with two mono single rail amps, you can saw the board in half - or use two boards for a stereo split rail amplifier.

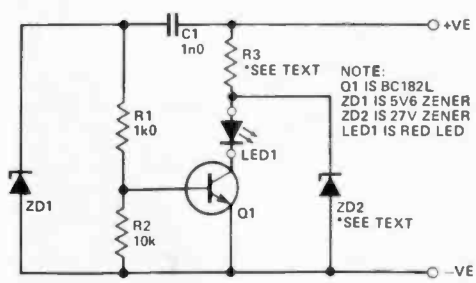


Fig. 6 (Left) Circuit diagram of the ETI Ripple Monitor.

HOW IT WORKS

The circuit is connected across the reservoir capacitor of the power supply. C1 is normally charged to about the same voltage as the reservoir capacitor and is also discharged by the same amount between charges. If the subsequent recharge is great enough, the current which flows through R1 and R2 will be enough to turn on Q1 hard and illuminate the LED. ZD1 is a zener diode which provides protection for Q1 base at switch-on and also a convenient discharge path for C1 in normal operation.

R3 limits the current in the LED and in conjunction with ZD2, limits the voltage reaching Q1 collector to a safe value.

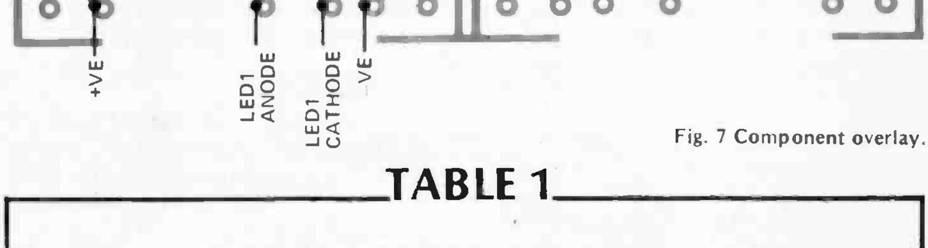


Fig. 7 Component overlay.

TABLE 1

Supply Volts	20	25	30	35	40	45	50
R3	1k2	1k5	1k8	2k2	3k3	3k9	3k9

R3 must be a 1 W resistor.

For more brightness reduce R3 to approx ⅔ of the value shown.

AUTORANGING CAPACITANCE METER

This project is designed to measure capacitance in the range of 100pF to 1000uF with no help (or hindrance) from the operator once the component is connected. Apart from the power switch, there are no external switches or knobs to adjust and only one internal adjustment. Once the instrument is working, setting up and using it is very simple.

The measurement capability of the circuit covers eight decades of capacitance and all but the lowest range should give true readings. The limitation on the 100pF range is due to the stray capacitance at the measurement terminals and the very high impedances necessary to give any sensible result.

The accuracy of the whole instrument depends on the quality of the five range resistors and the adjustment of PR1. The range resistors should preferably be 1% tolerance and of good quality. The threshold setting resistors in the comparator stage (R24, R27, PR1) should also be close tolerance devices if possible, although 5% components would be adequate in most cases.

There is a mixture of technologies used in the project; bipolar and TTL devices have been used where speed of operation, high current or low leakage were of greatest importance, while CMOS logic devices have been used to keep power requirements low wherever possible. The range switching logic and measurement sequencer are fairly complex but this was found to be necessary to avoid using very high frequencies, excessively high or low resistances, or waiting too long for a reading. The maximum response time for this device is about one second to get a reading on the 1000uF range. The lower ranges respond in milliseconds — faster than the display can follow.

Theory

If a voltage is applied via a resistor to a discharged capacitor, the voltage across the capacitor rises exponentially towards that of the supply (Fig. 1). From basic principles:

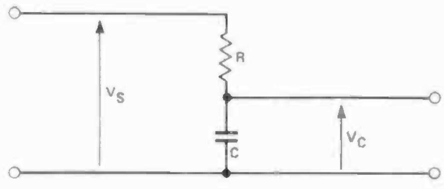


Fig. 1 Basic capacitor charging circuit.

$$V_C = V_S(1 - e^{-t/RC})$$

$$\frac{V_S - V_C}{V_S} = e^{-t/RC}$$

$$\log_e \frac{V_S - V_C}{V_S} = -t/RC$$

$$t = -RC\log_e \frac{V_S - V_C}{V_S}$$

$$t = RC\log_e \frac{V_S}{V_S - V_C} \quad \ldots\ldots (1)$$

This equation shows that the time to charge to a fixed proportion of the supply voltage is independent of the actual supply voltage and only proportional to the RC product (or "time constant"). This also means that the time to charge between any two fixed proportions of the supply voltage will similarly be independent of the supply voltage.

There are two main operations to be performed by the circuitry; the first is to determine what range of measurement to use for any particular component; the second is to use that range and give a measure of the component's capacitance. To do this the relationship found in equation (1) above is used twice, but in slightly different ways.

For any measurement 1600 clock cycles are allocated out of each 2000. These may be at 100 kHz or 100 Hz depending on the current range. At the start of the 1600 clock cycles, the unknown capacitor is in a discharged state and will charge via one of the range resistors at a rate depending on its capacitance. If the range is correct, its voltage will reach an upper threshold between 100 and 1600 clock cycles later. If the range is too low, the rate of charge will be too slow and the upper threshold will not be reached in time. The range will be changed for a higher one and the process repeated until it is correct or the maximum range is reached.

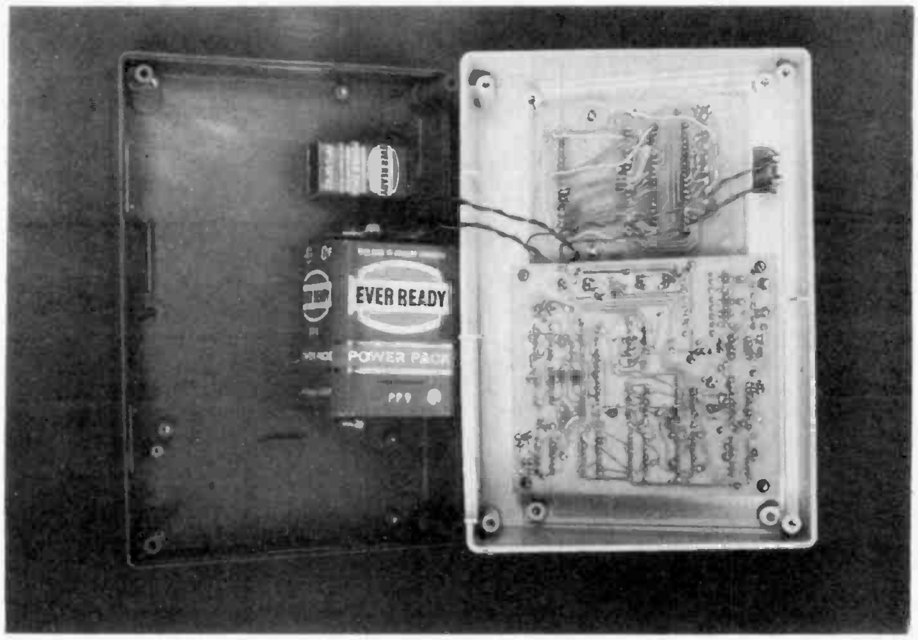
If the range is too large, the rate of charge will be very fast and the upper threshold will be reached before 100 clock cycles have elapsed. This time the range will be changed for a lower one and the process repeated until it is correct or the minimum range is reached.

In order to determine what the upper threshold will be, we choose — not entirely arbitrarily — that the range change point will be at 120% of the nominal for that range. This means that, with a capacitor 1.2 times the nominal, it will just reach the upper threshold in 1600 clock cycles. The resistors we use are in powers of 10 and the clock is also in powers of 10, so we can assume that the CR product will also be 1.2 times some power of 10. For convenience we can assume 1200 clock cycles. This also avoids thresholds ridiculously close to the supply rails. From (1) and the above we have:

$$1600 = 1200\log_e \frac{V_S}{V_S - V_C}$$

$$\frac{V_S}{V_S - V_C} = e^{1600/1200} = e^{4/3}$$

$$\frac{V_S - V_C}{V_S} = e^{-4/3}$$



The inside of the meter.

$$1 - \frac{V_C}{V_S} = e^{-4/3}$$

$$\frac{V_C}{V_S} = 1 - e^{-4/3} = 0.736 \text{ (upper threshold)}$$

This shows that the upper threshold should be at 73.6% of the supply for autoranging up at 120% of nominal.

To find the lower limit of the range we know that if the capacitor voltage reaches the threshold voltage just calculated within 100 clock cycles, the range will change down. From (1) we have:

$$100 = 1000C\log_e \frac{V_S}{V_S - V_C}$$

$$C = \frac{100}{1000\log_e \frac{1}{1 - V_C/V_S}}$$

$$C = \frac{1}{10\log_e (1/0.264)}$$

C = 0.075 or 7.5% of nominal for the range.

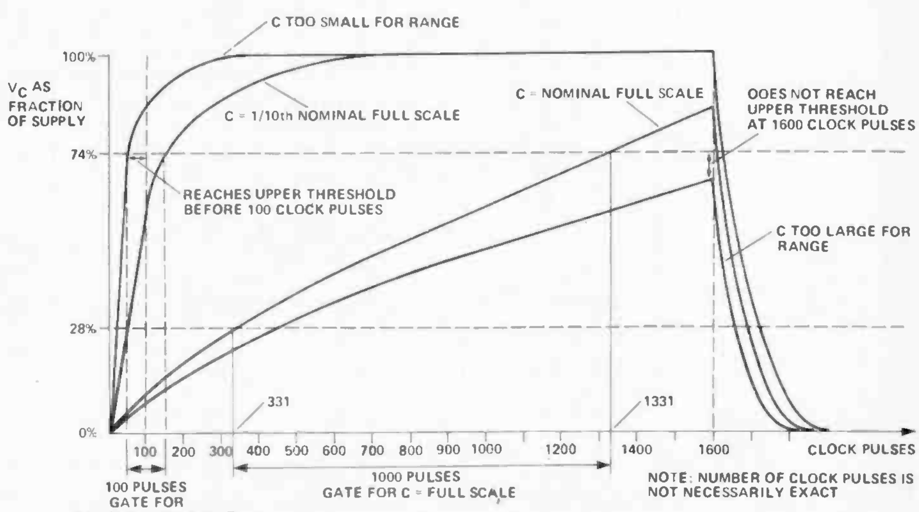


Fig. 2 The various modes of operation of the meter.

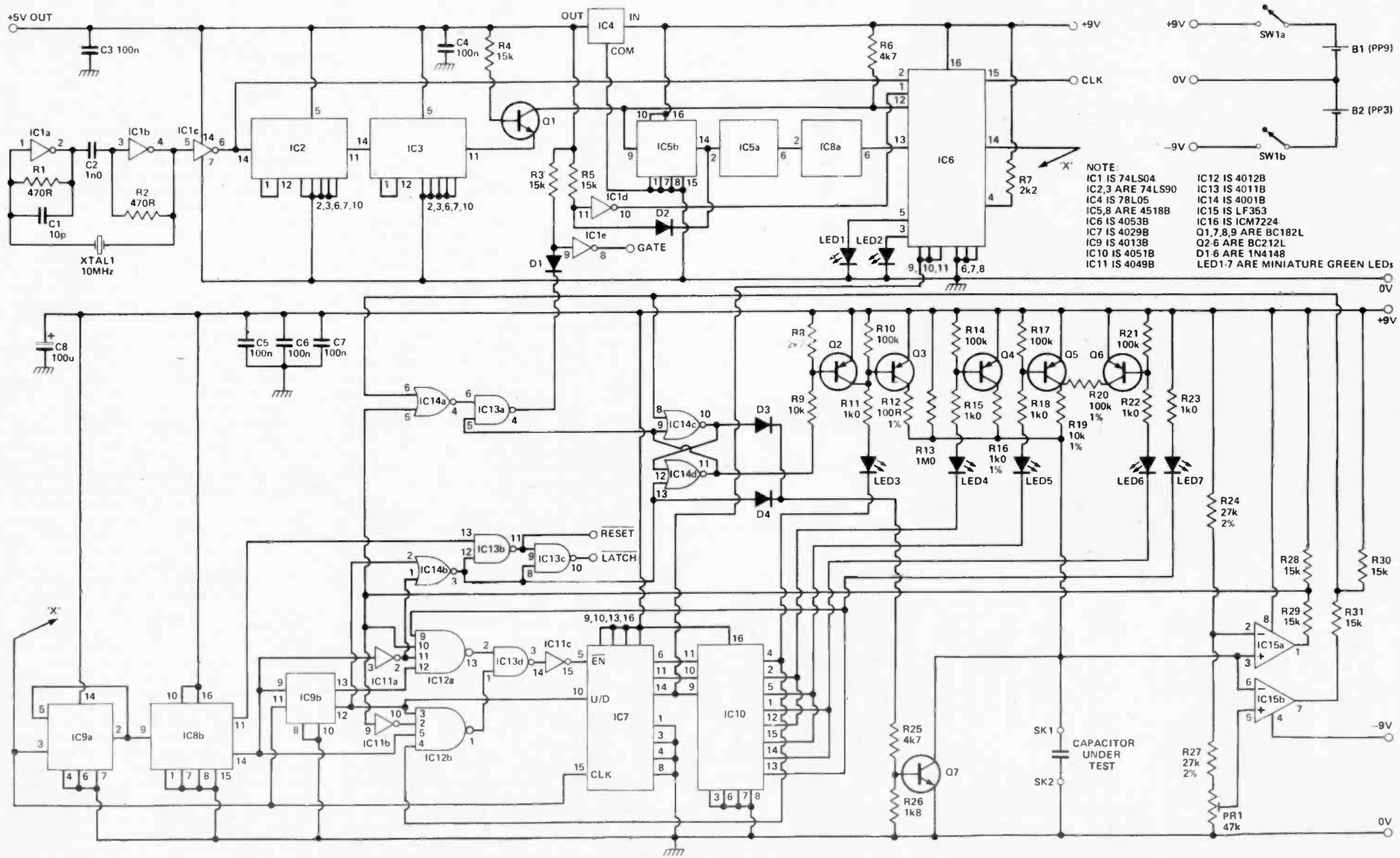


Fig. 3 Circuit diagram for the main board of the Autoranging Capacitance Meter

HOW IT WORKS

The master clock for the whole instrument is provided by a 10 MHz crystal-controlled oscillator made from three sections of IC1. This is divided by 10 in IC2 and IC3 (74LS90). TTL devices have been used here because operation of CMOS devices is not certain at these frequencies with a 9 V supply. Further division is provided by IC5 and IC8a which are CMOS devices. From these divider stages, frequencies of 10 MHz, 100 kHz, 10 kHz and 100 Hz are passed to IC6, which selects the output and sequencer clock frequencies of 10 MHz/100 kHz or 10 kHz/100 Hz as required. IC6 also drives two LEDs to show which set of ranges is in use.

The measurement sequencer is probably the most complex part of the circuit and consists of IC9, IC8b and parts of IC11, IC13 and IC14. These provide the signals which discharge the capacitor between counts 8 and 9 of IC8b while also controlling the counter-display chip via outputs from IC13. IC14c, d are connected as a set/reset latch and control Q7 via D3 and D4 to ensure that the test capacitor is discharged to below the lower threshold before each measurement begins. IC9, IC12 and parts of IC11 and IC13 form the gating logic which determines whether the range. The other part of IC12 (IC12b) is enabled for one clock period immediately after period immediately after IC8b counts from 9 to 0 and the range counter IC7 will be enabled if the upper voltage level comparator output goes high before the next clock pulse arrives. Under these conditions IC7 will count down a range. The other part of IC12 (IC12b) is enabled for one clock period immediately after IC8b counts from 7 to 8 and the range counter will be enabled if the output from the upper voltage level comparator is still low when the next clock pulse arrives. In this case the range counter will be incremented to the next higher range.

If the range counter IC7 is already on its maximum or minimum range, further counting in that direction would result in completely the wrong range being selected, so the extreme range outputs from IC10, which is the range resistor switch selector, are taken back to the relevant sections of IC12 to prevent further counting in the wrong direction.

Before going on it may be worth saying that IC9a and IC8b form a divide-by-20 circuit which, together with the fact that the output clock runs 100 times faster than the input to this circuit, means that there are 2000 clock pulses to each measurement. IC9b is a one bit shift register which delays the Q_4 output bit from IC8b for one clock cycle and with the logic of IC11, IC12, IC13 and IC14 does all the timing described above.

The next section is the part which actually provides the measuring function. IC10 is a one-of-eight analogue switch which is used here to drive the four resistor-switching transistors and also the LED range indicators. Six of the outputs of IC10 are connected in pairs to cope with the range overlap mentioned elsewhere. The lowest range does not have a transistor switch and its resistor (R13) is permanently in circuit. The effect of this on the next range is offset by connecting the next range resistor to the switch end of the one above it (R20 to R19). This enables decades of resistors to be used without needing any odd values.

Q2 in the resistor switch section serves to switch off Q3 which controls the smallest value resistor while the test capacitor is being discharged. This is necessary because, on the highest capacitance range, there is only a relatively short time in which to complete the discharge and Q7 does not have enough base drive or enough current capability to do its job if it has to sink the current which would flow through R12 as well. With this in mind it would be wise to ensure that Q7 has a high gain.

The last part of the circuit is the window comparator formed by IC15 and IC14a. These are the components which detect when the voltage on the test capacitor is between the upper and lower voltage limits. If this is the case and the capacitor is not being discharged, the display counter is enabled and the number of clock cycles counted is the measure of the capacitance. Outputs from the upper voltage level comparator feed back into the autoranging circuit and from the lower level comparator into the discharge latch. It was unfortunately found necessary to supply the comparator (a dual JFET op-amp) with a negative supply in order to get it to work at the voltage levels required, but this disadvantage was offset by its high input impedance and fast slew rate.

THE DISPLAY

IC16 consists of a 4½ digit counter, latches, display decoder and LCD driver. The logic signals from the main circuit control its sequence of operation but in some cases need to be conditioned to match the logic levels of the device. There are also two transistors on the display board; one drives the decimal point segment while the other drives the half digit segment.

The display itself is a 3½ digit LCD device. The three full digits are driven directly from the display driver least significant outputs while the half digit is driven via an inverter from segment 'a' of the fourth digit. This is necessary to get the half digit to respond at the correct time.

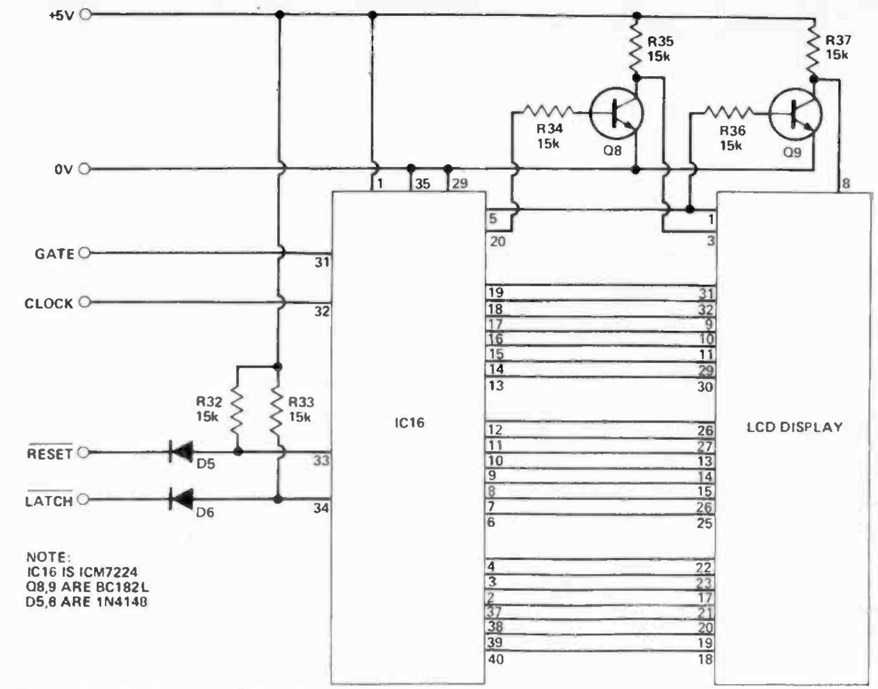


Fig. 4 Circuit diagram of the display board.

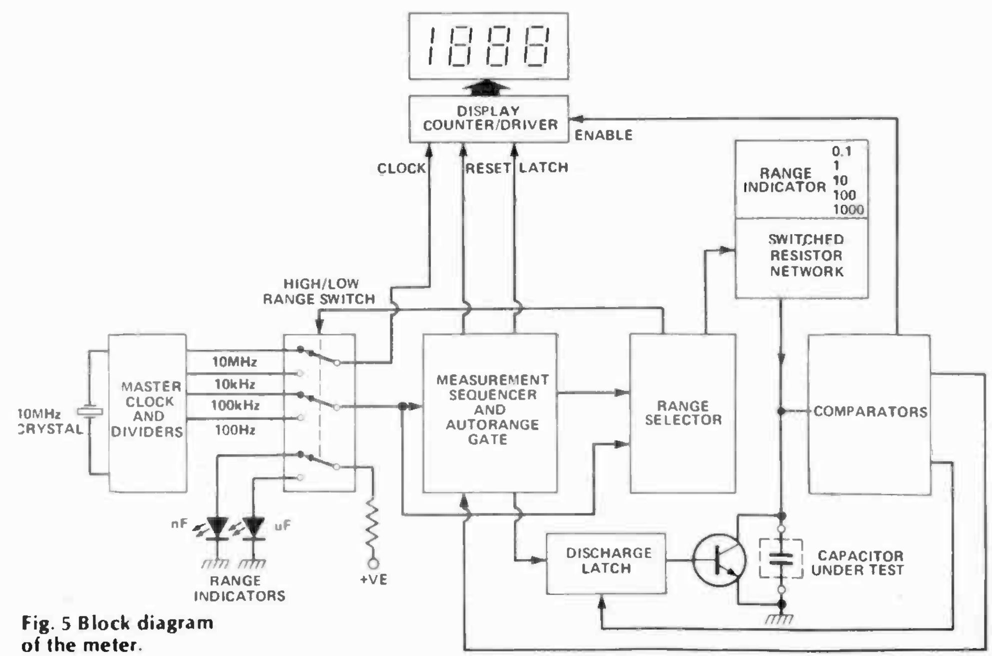


Fig. 5 Block diagram of the meter.

Having decided on the upper threshold using the requirements for autoranging, we must now consider what the lower threshold will be. With a nominal maximum capacitance for the range connected there must be 1000 clock cycles between the capacitor voltage crossing the lower threshold and it crossing the upper threshold. The time to reach the upper threshold is:

$$t = RC\log_e \frac{V_s}{V_s - V_c}$$

$$= 1000\log_e \frac{1}{1 - V_c/V_s}$$

$$= 1000\log_e \frac{1}{1 - 0.736}$$

$$= 1333.3 \text{ clock cycles.}$$

Therefore the number of clock cycles to reach the lower threshold is:

$$1333.3 - 1000 = 333.3$$

and the threshold voltage will be found from:

$$333.3 = 1000\log e \frac{1}{1 - V_c/V_s}$$

$$\frac{V_c}{V_s} = 1 - 1/e^{0.33} = 0.283$$

This shows that the lower threshold will be at 28.3% of the supply voltage.

Once we have set the upper and lower thresholds, the relationship between the measurement range and the autoranging function is defined. In this case the nominal range is 0.1 to 1 with an underrange down to 0.075 and an overrange up to 1.2 times the nominal. The relationship between ranges is defined by the range resistors and the digital division ratio. The various modes of operation are shown in Fig. 2.

In this instrument we use only five range resistors to give a coverage of eight ranges. This is desirable to avoid the use of very large resistors with small capacitors, or the use of very small resistors with very large capacitors and consequent battery drain. It is made possible by sharing three of the range resistors between six ranges and reducing the clock speed by a factor of 1000 on the high capacitance ranges to compensate.

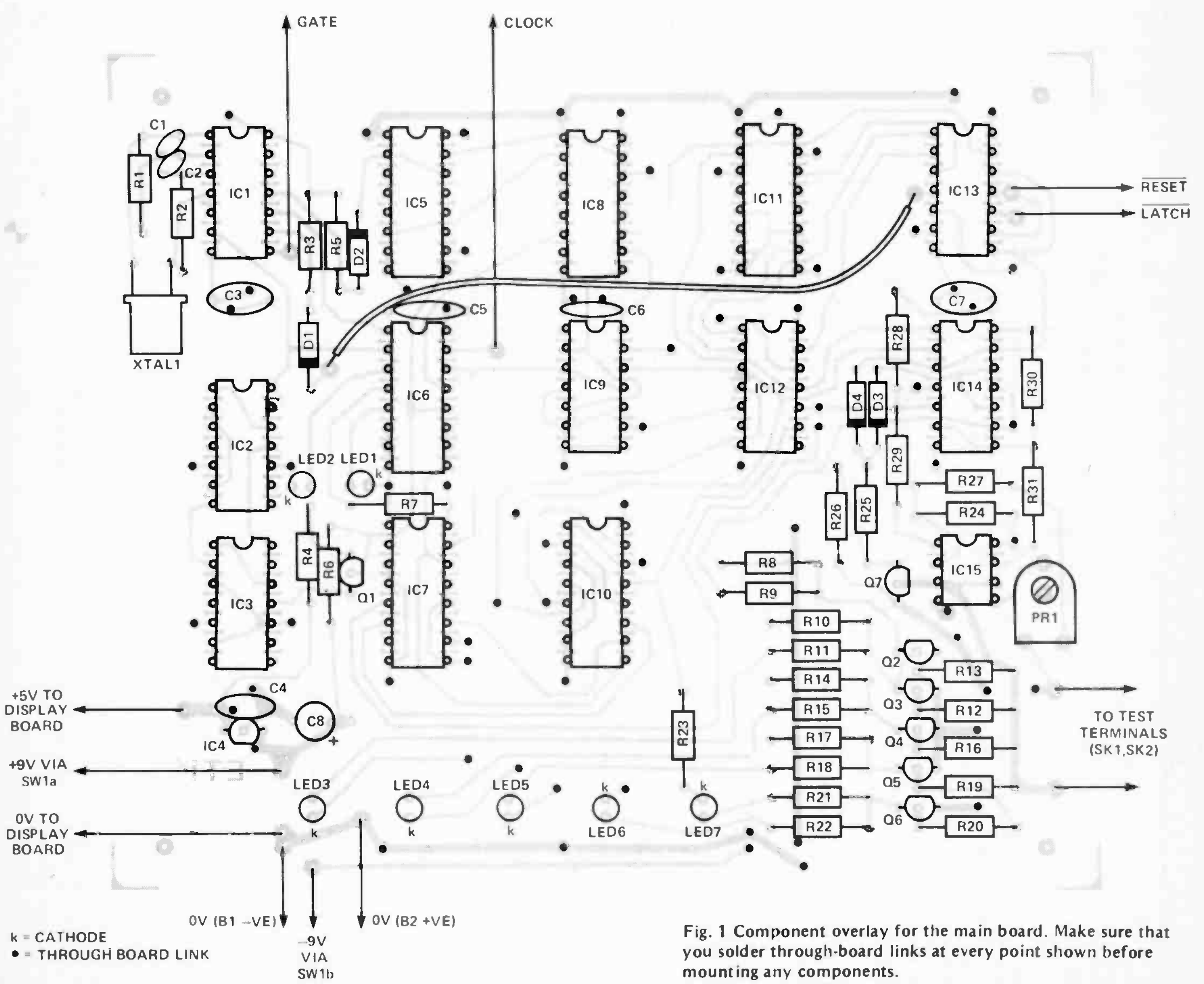
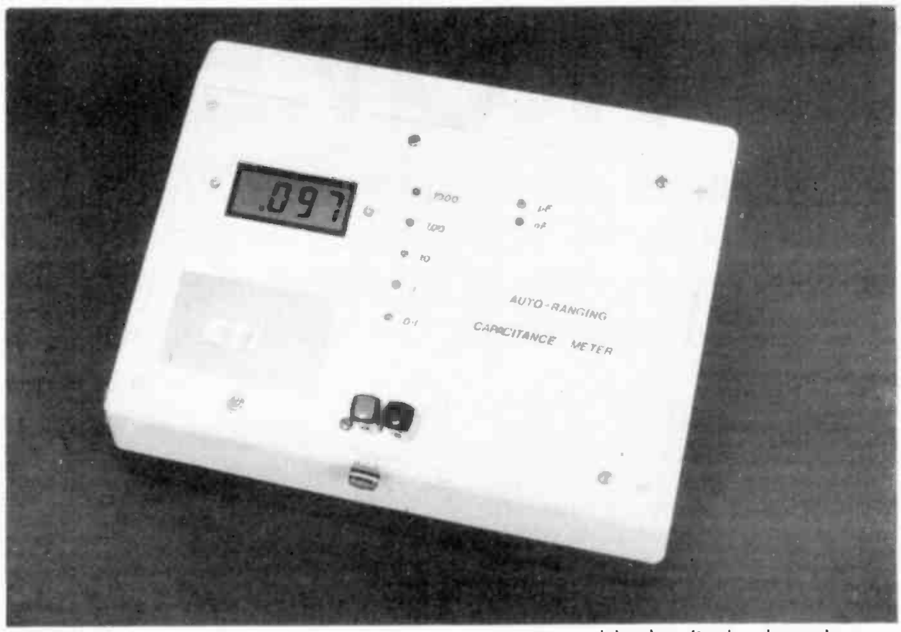


Fig. 1 Component overlay for the main board. Make sure that you solder through-board links at every point shown before mounting any components.

It is well worthwhile checking the PCB for shorts between tracks before doing anything else. Ensure that there is a hole through the board under the PR1 position to facilitate adjustment later.

Put links through the board at all positions marked with a dot on the overlay and solder on BOTH sides of the board. The other components may now be inserted into the board — preferably using sockets for all the ICs except IC4 and IC15. IC4 is a T092-type package 100 mA regulator and does not need a socket, while IC15 may foul PR1 if a socket is used.

The LEDs should not be fitted until the board is test-fitted in position as they are intended to protrude through the panel as indicators. Attach power supply wires and fit up to the panel, position the LEDs and solder in position.



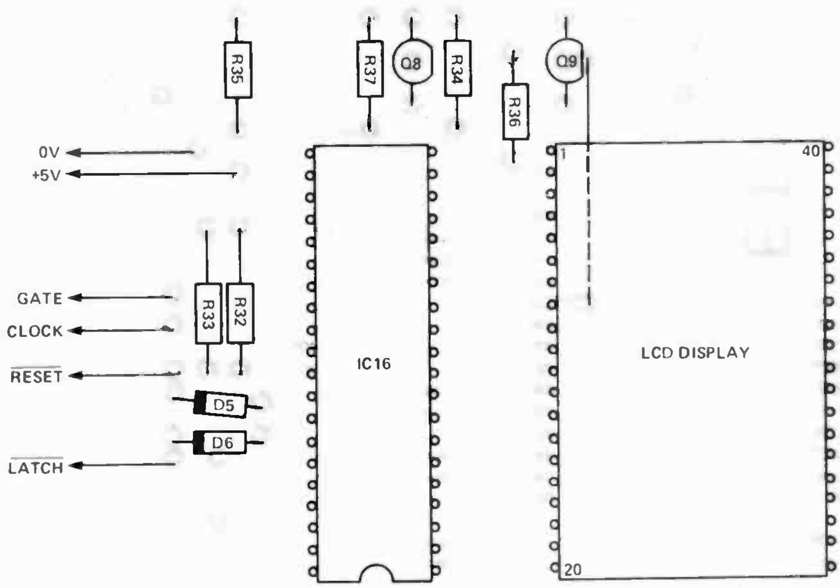


Fig. 2 Component overlay for the display board. Insert the link under the display first.

Assemble the display board components and attach the logic and power supply wires from the main board. Wire the remaining power leads via the on/off switch to the battery connectors and attach the two boards to the front panel using pillars or long bolts and lock nuts. Our prototype just fitted into a slope fronted instrument case made by Vero Industries (see Parts List).

Finally, the one calibration needed is the setting of PR1. Select a capacitor with an accurately known value (around 100 u would be ideal) and adjust PR1 to give the correct meter reading. And that's it!

BUYLINES

Very few unusual components in this project; all the logic is standard CMOS. The ICM7224 and the LCD display is stocked by Watford Electronics, while the LF353 is available from Rapid Electronics. The two PCBs can be obtained from our PCB Service, advertised on page 91.

PARTS LIST

Resistors (all ¼ W, 5% except where stated)

R1,2	470R
R3,4,5, 28-37	15k
R6,25	4k7
R7,8	2k2
R9	10k
R10,14,17, 21	100k
R11,15,18, 22,23	1k0
R12	100R 1%
R13	1M0 5% or better
R16	1k0 1%
R19	10k 1%
R20	100k 1%
R24,27	27k 2%
R26	1k8

Potentiometer

PR1	47k miniature horizontal preset

Capacitors

C1	10p ceramic
C2	1n0 ceramic
C3-7	100n ceramic
C8	100u 10 V tantalum

Semiconductors

IC1	74LS04
IC2,3	74LS90
IC4	78L05
IC5,8	4518B
IC6	4053B
IC7	4029B
IC9	4013B
IC10	4051B
IC11	4049B
IC12	4012B
IC13	4011B
IC14	4001B
IC15	LF353
IC16	ICM7224
Q1,7,8,9,	BC182L
Q2-6	BC212L
D1-6	1N4148
LED1-7	miniature green LEDs

Miscellaneous

XTAL1	miniature 10 MHz crystal
SW1	DPDT toggle switch
SK1,2	press terminals (one red, one black)

PCBs (see Buylines); 3½ digit LCD display; IC sockets; display socket (if required); PP9 battery and connectors; PP3 battery and connectors; mounting hardware; case (Vero 220 x 156 mm sloping front box, order no. 65-2523E).

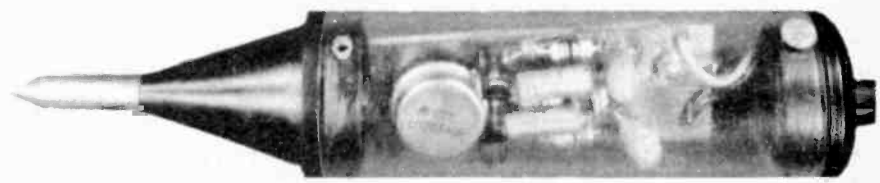
INSTRUMENT PROBE

This probe will allow you to make CRO or frequency meter/timer measurements on high impedance circuits with waveforms having rise times as fast as three or four nanoseconds. Cost is well below commercial equivalents. Design by Jonathan Scott.

Most readers would be aware that, when taking a measurement on electronic circuitry, the input impedance of the measuring instrument must be much greater than the impedance of the circuit to which it is attached, otherwise the accuracy of the measurement suffers. The input impedance of the majority of oscilloscopes is generally 1M0 with a parallel capacitance of between 20pF and 40pF. For a wide variety of applications this is perfectly adequate and will suffice for measurements of frequencies up to 5 MHz or so. The input impedance of the CRO falls with increasing frequency owing to the falling reactance of the input capacitance. For example, a capacitance of 30pF — which may be made up of direct input capacitance plus cable capacitance — has a reactance of only 500 ohms at 10 MHz. The input capacitance also affects the rise time of the input — that is, the speed at which a 'step' input will rise from the 10% amplitude value to the 90% amplitude value.

The input impedance of an oscilloscope can be effectively raised, and the capacitance decreased, by using a 'stepdown' probe. For example, a 'x10' probe will generally have an input impedance of 10M and a parallel capacitance of between 5pF and 15pF. While this improves the input impedance there are two trade-offs. Firstly, unless elaborate (and expensive) compensation is employed, the rise time is degraded, and secondly, maximum sensitivity is decreased by a factor of 10. As Murphy's law would have it, your CRO will run out of grunt just when you need it most.

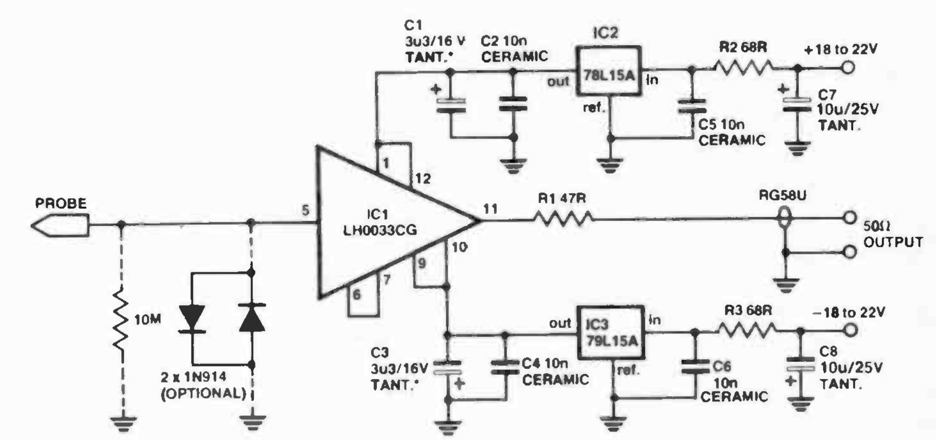
Taking the situation with digital counter/timers, we find similar problems. Those that operate beyond 30 MHz or 50 MHz generally employ a prescaler with an input impedance of 50 ohms — which is perfectly all right if you're working on low impedance circuits and/or with high signal levels. But there are those occasions when you need a high impedance input and a fast (high frequency) rise time. As with the CRO, this is where your counter/timer runs out of grunt.



It's times like these you need this project; a x 1 active instrument probe using a special buffer IC with an input impedance of typically 100,000 megohms! — that's 10^{11} ohms — a very low input capacitance of around four to five picofarads, a fast rise time (around three nanoseconds) and a bandwidth of 100 MHz. Output impedance is around 50 ohms and the device is capable of driving capacitive loads up to several thousand picofarads. Thus it is eminently suited for use with high speed, wide bandwidth oscilloscopes and digital frequency meter/timers at frequencies up to 100 MHz. Output impedance is close to 50 ohms and it is thus suited tc drive both high impedance instrument inputs and low impedance inputs (which are generally 50 ohms).

Design

It's all done inside a special IC — an LH0033CG from National Semiconductors. This is described as a 'fast buffer amplifier'. (It has a companion designated LH0063, described as a 'damn fast buffer amplifier'!). The LH0033 is a direct-coupled FET-input voltage follower/buffer (gain ≃1) designed to provide high current drive at frequencies from DC to over 100 MHz. It will provide ±10 mA into 1k0 loads (±100 mA peak) at slew rates up to 1500 V/uS, and the chip exhibits excellent phase linearity up to 20 MHz. No offset voltage adjustment is required as the unit is constructed using specially selected FETs and is laser-trimmed during construction. Input is directly to the gate of a



*AXIAL LEAD SOLID TANTALUM CAPACITOR

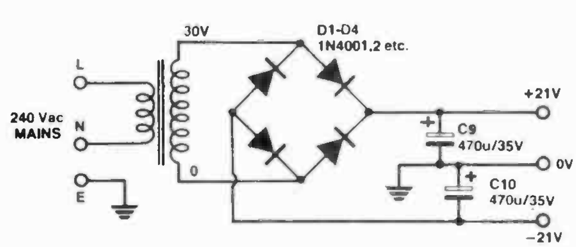


Fig. 1 Circuit diagram for the probe. C2 and C4 need to be ceramic

Fig. 1 Circuit diagram for the probe. C2 and C4 need to be 10n ceramic chip or 1n0 ceramic disc or plate types. C5 and C6 need only be disc or plate ceramic. See 'Bypassing' over the page.

junction FET, operated as a source follower, driving a complementary output pair of bipolar transistors.

Regulated plus and minus supplies of 15 V each provide power to the IC. Low-power three-terminal regulators are used to keep the unit compact. An external unregulated supply of between 18 and 22 V at around 50 mA is required to power the probe.

The supply pins on the IC need to be well bypassed over a wide frequency range so that the IC can maintain its characteristics, and the construction has been specially arranged to achieve this. Axial lead solid tantalum capacitors are used to bypass the IC's supply pins at the lower frequencies, while low inductance ceramic capacitors are employed as bypasses for the higher frequencies. A double-sided fibreglass PCB is used to preserve the high frequency response and the high input impedance, and the layout is arranged to permit direct connection to the probe tip and provide low input capacitance.

However, the presence of the PCB substrate will degrade the input impedance, surprisingly enough, and you can drill out the area of board immediately beneath pin 5 of the IC and solder the pin directly to the probe tip. For those who wish to go 'all the way' (as Frank Sinatra sings), the plastic insulation of the probe tip can be replaced with a similar piece of Teflon — if you can afford it and have access to a lathe.

The maximum input voltage permissible, when driving a high impedance load, is plus or minus 15 V. When driving a 50 ohm load, maximum input voltage permissible is only plus or minus 10 V (limited by maximum output current). No input protection has been included. However, if you are only working with circuits where voltages are no greater than about 1 V peak-to-peak, protection can be added by putting two diodes back-to-back in parallel with the input, along with a 10M resistor. The maximum input voltage figures include any DC voltages present, *plus* the superimposed signal voltage.

At this stage it is only fair to tell you that the LH0033CG is an expensive device (by comparison). But — compare the total cost of this probe to a similar commercially-made type and you won't catch your breath a second time!

Construction

The project is constructed on a small double-sided fibreglass PCB *with*

BYPASSING

Supply lead bypassing is important in order that the LH0033 can operate correctly over the full bandwidth from DC to 100 MHz. To ensure this, the bypassing has been specially arranged and the techniques employed are probably unfamiliar to many readers.

The output circuit signal return path for the IC is via the ground and the two supply rails. Any significant impedance in series with this path (or paths) will subtract signal from the output load. Thus, the supply rail bypassing has to present an impedance which is a *fraction* (like one-tenth or better) that of the minimum output load impedance. Here, the minimum output load is about 100 ohms (R1 + 50 ohms instrument input impedance) and the supply bypassing impedance should ideally be less than 10 ohms across the frequency range.

The bypassing on each supply rail to the IC leads here takes advantage of the characteristics of three separate components to cover three sections of the frequency range.

From DC to around 100 kHz, each three-terminal regulator (IC2, IC3) has an output impedance well below one ohm, rising to four or five ohms at 1 MHz, as shown in Fig. 1. The two tantalum capacitors, C1 and C3, then take over.

Solid tantalum capacitors have a characteristic impedance that falls with frequency according to its value, which then 'flattens out' in the region around 500 kHz — 1 MHz, rising to a few ohms around 10 MHz, as can be seen in Fig. 2. Thus, C1 and C3 serve as effective bypasses across the range from around 100 kHz to around 10 MHz. Axial lead tantalum capacitors were chosen as their construction exhibits the slowest impedance rise following the minimum impedance value.

To provide bypassing over the decade from 10 MHz to 100 MHz, capacitors C2 and C4 have been specially chosen and positioned on the PCB. For the prototype, 'chip' ceramic capacitors were used. These tiny, 'naked' chips of ceramic with a capacitor embedded in them are probably the most effective bypass capacitors made. The leads and physical construction of all capacitors form an inductance which is effectively in series with the capacitance of the component. The combined effect forms a series resonant circuit, the frequency of which (that is, the self-resonant frequency of the component) is mainly dependent on the length of the connecting leads, the particular construction of the capacitor and the way in which it is mounted. Ceramic chip capacitors, being a tiny block with connecting pads or surfaces on each end, have extremely low values of series inductance and thus very high self-resonant frequencies — see Fig. 4. Now, any value of chip capacitor between 1n0 and 10n can be used for C2 and C4. The self-resonant frequency of a 1n0 chip capacitor is somewhat above 100 MHz (as per Fig. 4), but that of a 10n chip is between 40 MHz and 50 MHz. Now, this isn't a problem, for the chip's impedance falls with frequency as usual until near the self-resonant frequency where it falls rapidly, reaching a minimum at the self-resonant frequency. Above that frequency its impedance rises again, but is still low enough for effective bypassing.

Ordinary ceramic disc and plate capacitors behave in much the same way. The self-resonant frequency of a typical 5 mm diameter disc or 5 mm square plate capacitor depends on the lead length, as shown in Fig. 5. Thus, you could use 470pF or 1000pF (1n0) capacitors of this type for C2 and C4, provided you installed them on the underside of the board with *absolute minimum lead length*.

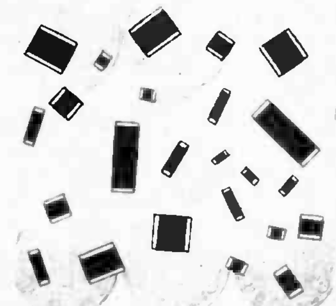


Fig. 3 Ceramic chip capacitors shown about actual size.

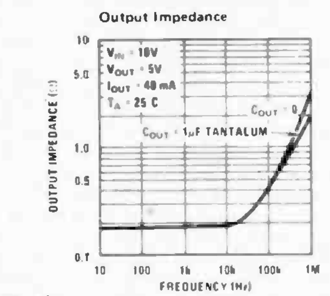


Fig. 1.

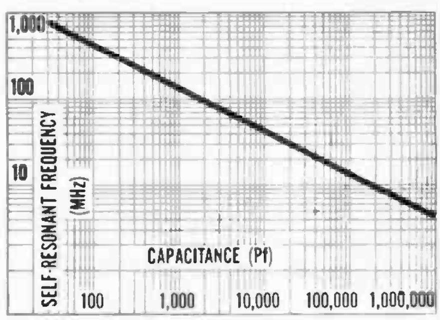


Fig. 4.

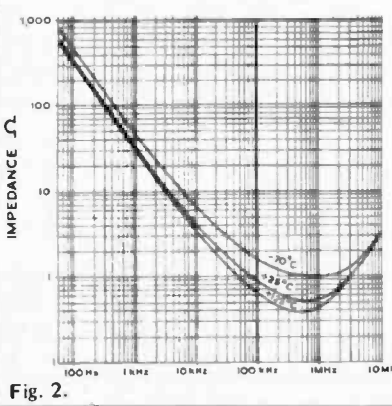


Fig. 2.

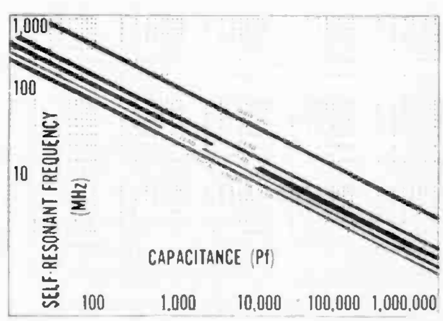


Fig. 5.

components mounted on both sides of the board. Commence by soldering in place the components that go on the top side of the board, leaving IC1 until last. Note that the positive leads of both C3 and C8 are soldered to the groundplane areas on both the top and the bottom sides of the board. Take care with the orientation of the tantalum capacitor, as well as IC2 and IC3. Having done that, solder C2, C4, C5 and C6 to the bottom side of the board. Now you can install IC1. You will have to juggle the legs a little. Push the can as far down on the board as you're able; its base should sit no more than 3 mm from the board.

Now that you have everything in place, *check it all.* It seems pretty simple, but Murphy's law will ensure that the simplest things have the highest stuff-up rates!

All's well? — now you attach the output coax cable to the underside of the board, plus the DC input and ground (0 V) wires. *But* — before you do, slip the output end piece of the probe case over the cable and supply wires, push it down about 150 mm or so and then slip the case of the probe case down the wires. This saves slipping them over the other end of the whole business and sliding them all the way to the probe.

The probe tip can be attached and soldered in place last of all. Now you can screw it all together and attach the appropriate plugs to the other end of the cable and supply wires.

With the construction completed, you can power up and try it out. Note that the transformer suggested in our power supply is but one of many suitable types. Any transformer that will deliver at least 26 V AC at a load of about 50 mA will suffice. Alternatively, any dual polarity DC supply having an output between 18 and 22 V at 250 mA will power the probe.

Note

Always take care that you don't exceed the input voltage limitation; LH0033s are expensive.

BUYLINES

Ceramic chip capacitors and solid tantalum axial capacitors are a trifle unusual; however, they are stocked by C.T. Electronics (Action) Ltd, 267 & 270 Acton Lane, London W4 5DG. (They also stock the BNC plug should you have any problems there). We will be selling the double-sided board through out PCB Service — the order form is on page 91.

PARTS LIST

Resistors (all ¼ W, 5%)		**Semiconductors**	
R1	47R	IC1	LH0033CG
R2, R3	68R	IC2	78L15A
		IC3	79L15A
		D1-D4	1N4001,2,etc. (if required)
Capacitors		**Miscellaneous**	
C1, C3	3u3 16 V solid tantalum axial leads	PCB (double-sided fibreglass); RG58U coax cable and BNC plug; T1 — (if required) 240 V to 30 V transformer or similar; optional 10M/¼ W 5% resistor and 2 x 1N914 diodes; wire; probe housing.	
C2, 4, 5, 6	10n ceramic block		
C7, C8	10u 25 V tantalum		
C9, C10	470u 35 V electrolytic (if required)		

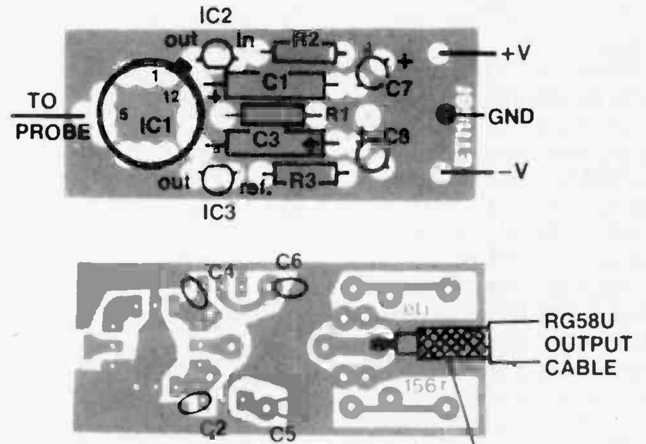


Fig. 2 Component overlays for the top of the board (top) and the bottom of the board (bottom!).

HOW IT WORKS

This instrument probe employs a wideband hybrid voltage follower/buffer IC, the LH0033, with very close to unity gain, that features a very high input impedance and a low output impedance. It requires regulated, well-bypassed supply rails. Two three-terminal low power regulators provide plus-and-minus 15 V supplies from an unregulated input.

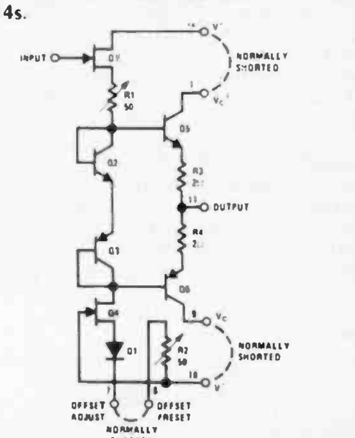
The internal circuit of the LH0033 is shown below. Basically, it consists of a FET input stage (Q1), operated as a source follower. The other FET, Q4, provides a constant current source for the source bias of Q1, while Q2 and Q3 are connected as diodes and provide bias for the bases of Q5 and Q6. Resistors R1 and R2 are laser trimmed in manufacture so that the IC meets the offset voltage specification. As Q1 has a constant current source load, the input impedance at the gate of Q1 is very low. The output of the source follower drives a complementary pair output stage, Q5-Q6. Thus the IC will have a very high input impedance, a very low output impedance and a gain very close to unity. With appropriate construction employed for the internal devices, the bandwidth over which the device will operate can be made very wide indeed. The −3dB point for the LH0033 is 100 MHz.

As the device is direct-coupled, DC levels will be maintained between input and output.

Bypassing requirements for the IC's supply leads are explained elsewhere in the article.

To provide regulated plus-and-minus 15 V rails for the IC, two three-terminal regulators are employed, a 78L15A for the positive rail and a 79L15A for the negative rail. These can supply up to 100 mA and have a very low output impedance up to several hundred kilohertz, which is exploited for low frequency bypassing. Each supply rail requires an unregulated input of between 18 V and 22 V. Decoupling of the supply leads provided by R2/C7 on the positive rail and R3/C8 on the negative rail. The input terminal of each regulator is bypassed to prevent instability.

As the input voltage is limited to a maximum equal to the supply rails (high impedance load), input protection may be added in applications where only low level signals are being examined. As shown in the main circuit, this protection consists of two 1N914 diodes connected back-to-back in parallel with a 10 M resistor across the input. Signals above 1 V peak-to-peak will be clipped, preventing any damage to the IC. If very fast rise time signals are to be examined then better protection for the IC can be obtained by using hot-carrier diodes such as the HP 5082-2800 instead of the 1N914s.



INSULATION TESTER

Some time ago ETI published a design for a very low resistance meter. Now we've built the DVMeg, for resistance up the other end of the scale. Design and development by Phil Walker.

The ohms range on most multimeters is fine for most electronics work but if you start dealing with hundreds rather than tens of volts then there is the possibility that damp, contaminated, or merely inadequate insulation could cause equipment failure or personal danger. In these circumstances the 1V5 or sometimes 15 V used by most multimeters is not sufficient to show up the dangers. To avoid the false sense of security which may be given by low voltage tests, a voltage somewhat higher than the normal working voltage of the circuit under test must be used. In the case of mains wiring this is often 500 V and this is about the level generated by our project. D is the Roman numeral for 500 and the instrument measures in the megohm range — hence the name, DVMeg.

HOW IT WORKS

When the device is turned on, Q2 will start to conduct due to current flowing via L1a and R1 into its base. This causes the supply voltage to appear across L1b, which by transformer action increases the voltage (and current) available in Q2's base circuit. This ensures that Q2 will be held on during this part of the cycle.

As L1 is an inductance and the voltage across it is by now fairly constant, the current flowing through it will rise linearly. However, the current also flows through R3, causing a voltage to appear across it which is proportional to the instantaneous value of this current. When the voltage across R3 becomes large enough, Q1 is turned on and robs Q2 of its base current. Q2 promptly turns off and its collector voltage rises sharply. This rise is coupled back to Q1 base via C2 to keep Q1 on during this period. At the same time, the voltage across L1c will be in a direction such that D3 conducts and charges C3. A short time later Q1 will turn off as it has no base drive once C2 is charged, and Q2 will turn on to start another cycle. D2 provides a discharge path for C2 at the start of a cycle as Q2 turns on.

If when C3 is being charged its voltage exceeds about 500 V, the voltage across L1a will be greater than the supply (due to transformer action with L1c) and D1 will conduct, diverting the excess energy back into C1 and the battery. The 500 V on C3 is applied via R5 to the circuit under test. The resulting current then passes through D5 in parallel with M1 and PR1. D4 provides protection if the probes are accidently connected to a live circuit, while D5 protects the meter from excess current and also modifies its response.

R4 in the high voltage section ensures that the output voltage dissipates quickly when the instrument is turned off.

The Instrument

The project is designed to use a standard PP6 9 V battery and contains a low power DC-DC converter to produce an output of about 500 V. This output is limited so that it does not rise too much even when off-load. The output current is also limited; about 500 uA maximum, even when short-circuited. Even so IT BITES! — so be careful!

In use the test leads are connected to the circuit under test and when the button is pressed, the circuit generates a high DC voltage which is applied to the test leads via a 1M0 resistor. The resulting current through the insulation is monitored by the meter and displayed as a resistance. When the button is released, the internal capacitor and the circuit under test are discharged fairly rapidly to avoid the risk of shock.

Construction

The coil is constructed using a

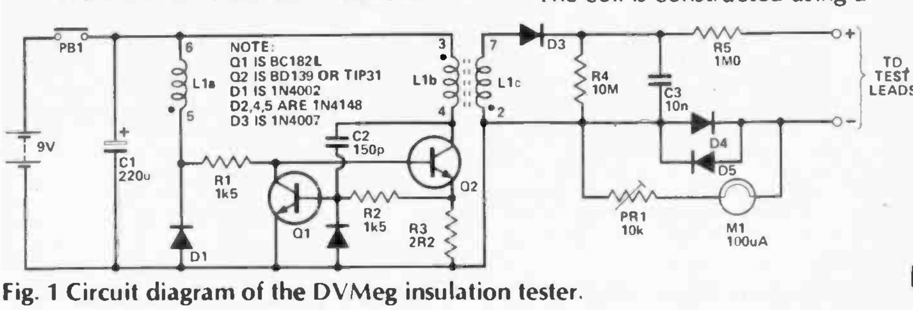


Fig. 1 Circuit diagram of the DVMeg insulation tester.

Neosid potcore. Wind 220 turns of 40 SWG enamelled copper wire on to the former in four layers; start at pin 2 and finish at pin 7. Each layer should be about 55 turns and as this is less than the width of the former, the space each side should be filled with a single layer of insulating tape 2-3 mm wide. A layer of tape the width of the former should be laid on top of each winding. Next wind on 22 turns of 32 SWG enamelled copper wire starting at pin 3 and finishing on pin 4. Insulate this as before and then wind another four turns of the same wire starting at pin 5 and finishing at pin 6.

Construction of the circuit board should pose no problems so long as component polarities are observed. The wires to the test probes should be flexible and well insulated.

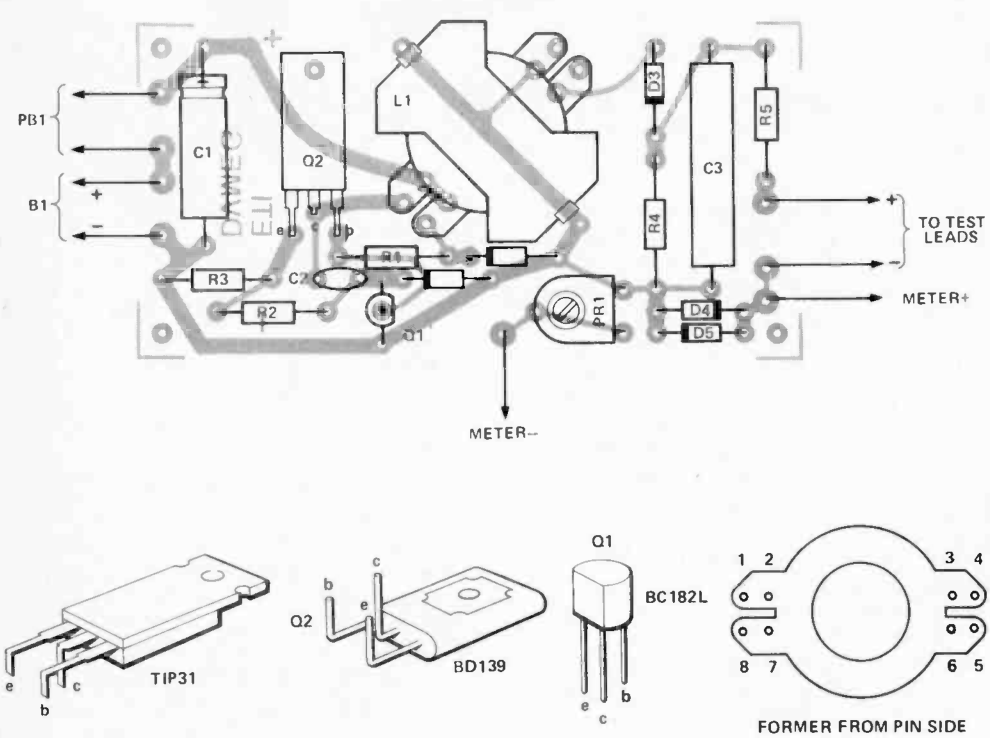
The two types of transistor specified for Q2 have different connections, so the TIP31 must be mounted 'upside-down' if used (see diagram).

Installation into the box is also fairly simple. The meter is fitted at one end of the box lid and the push-button just below it, but a little to one side to allow the battery to fit inside the case. The circuit board is stuck or bolted to the inside of the case lid and the battery leads connected via the switch. The meter leads are also connected, after which the battery may be connected and fixed into place with sticky pads or wedged with foam.

PS

It also tests neons!

Fig. 2 Component overlay and component pinouts for the DVMeg.



PARTS LIST

Resistors (¼ W 5% except where stated)

R1,2	1k5
R3	2R2
R4	10M ½ W
R5	1M0 ½ W
Potentiometer	
PR1	10k miniature horizontal preset
Capacitors	
C1	220u 16 V axial electrolytic
C2	150p 160 V ceramic
C3	10n 1000 V mixed dielectric
Semiconductors	
Q1	BC182L
Q2	BD139 or TIP31
D1	1N4002
D2,4,5	1N4148
D3	1N4007
Miscellaneous	
L1	RM10 pot-core, former, clips (A_l about 400)
PB1	push-to-make non-latching
M1	100 uA meter

PCB (see Buylines); case (West Hyde ref. BOC440); grommet; flexible wire; test prods; PP6 battery and connectors.

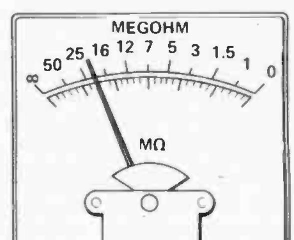
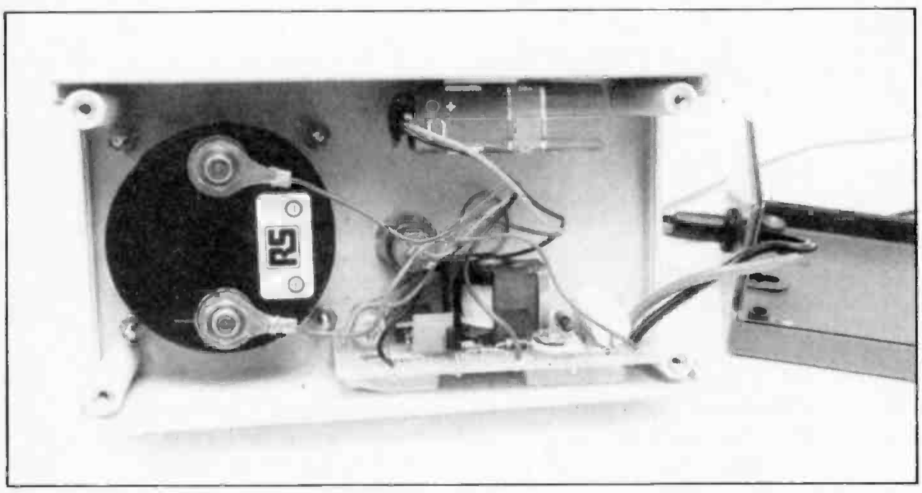


Fig. 3 A suitable meter scale.



BUYLINES

The pot core used to wind the inductor for this project is available from Neosid Small Orders, PO Box 86, Welwyn Garden City, Herts AL7 1AS; quote ref. 29-835-41 when ordering. West Hyde supply the case we used; for people who want to use substitutes the size is 150 x 80 x 55 mm. The PCB can be obtained using the order form on page 91.

LABORATORY PSU

This unique design can be used as a 0-30 V, 1A2 precision PSU project or as a 20 mA-1A2 constant-current generator. The design features electronic short-circuit protection, high transient current-drive capability, and a built-in audio-visual over-current alarm. Design by Ray Marston. Development by Steve Ramsahadeo.

A variable power supply unit (PSU) is one of the most basic and essential pieces of test gear in any amateur or professional laboratory or workshop. Trouble is, most cheap PSUs of the amateur type tend to suffer from quite severe performance restrictions, while professional PSUs tend to be very expensive and to suffer from various usage restrictions. Cheap PSUs, for example, often span the simple voltage range 1.2-25 V and have fixed 1 A current limiting, while professional units usually span the full 0-30 V range and have variable current limiting, but have no transient over-current driving capability.

A New PSU

We at ETI have recently taken a careful look at some of the basic concepts of conventional PSU design and, as a result, have come up with this brand new and ultra-practical PSU project. To arrive at our final design, we started off with the following basic precepts.

- Most of the cost of a PSU is attributable to the cost of the mechanical and electrical hardware (the case, transformer, moving-coil meter and so on), rather than to the electronics. Thus, PSU performance sophistication can be greatly increased, with little growth in total costs, by sensibly increasing the amount of electronics in the project.
- Our ideal PSU should fully span the 0-30 V range and be provided with fully-variable semi-precision 20 mA-1A2 current limiting. This variable current-limiting facility will also enable the PSU to serve the dual function of a variable constant-current generator or Ni-Cd charger.
- The ideal PSU should have two independent types of current-limiting facility. In our system, the variable (20 mA-1A2) limiter responds to *mean* currents and has a time constant of a few milliseconds; the second limiter is a fixed 1A8 instantaneously-acting type. The provision of the two types of limiter gives the completed PSU excellent mean-current-limiting characteristics combined with a high transient current-drive capability.
- The PSU should provide an audio-visual alarm indication of the variable over-current state, so that the user is given an instant warning of circuit defects when using the PSU to develop experimental circuits. The audible part of the alarm should be manually switchable, so that the alarm can be disabled when the PSU is used as a constant-current generator or Ni-Cd charger.

The results of our design efforts are shown in this article. You'll note from the circuit diagram that, although the design is moderately complex and unconventional, the design makes wide use of inexpensive ICs and transistors. As a result, you'll find that our PSU unit costs little more than the cheapest of PSUs to build, but gives a performance that is better than the best of professional units. Well, of course, you'd expect that from an ETI project, wouldn't you?

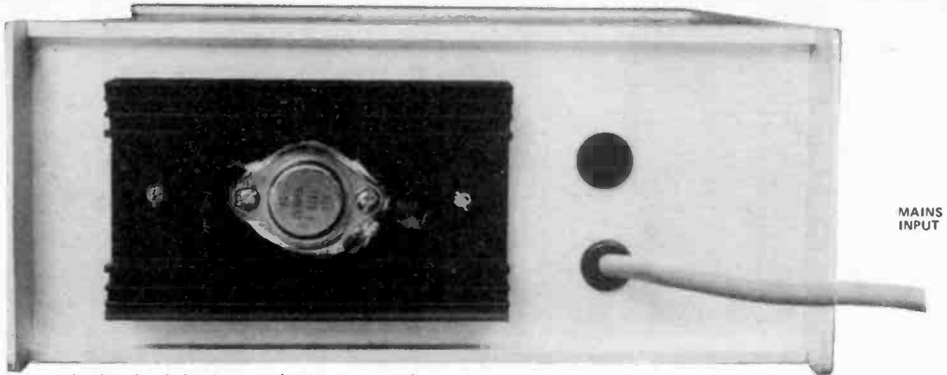
Construction

Using our PCB greatly simplifies construction and reduces the possibility of any errors. As usual, check the orientation of all semiconductors and capacitors against the overlay.

As you can see from the photographs, Q4 and Q8 are fitted with T05 style heatsinks; Q9 will dissipate approximately 45 W at the maximum setting and has to be mounted on a 100 mm x 60 mm finned heatsink. This heatsink must be insulated from the rear panel by plastic bushes. We also recommend the use of heatsink compound, to be applied on both mating surfaces. Heavy gauge wire (13/0.2 mm) should be used to connect Q9 and the output sockets to the PCB. As an added safety precaution all mains wiring, including the transformer terminals, should be insulated with rubber or PVC sleeving.

You may have difficulty obtaining a meter whose scale is printed with the required scale of 0-1.5 and 0-30. We overcame this problem by using a common-or-garden 100 uA meter and recalibrating the scale with dry transfer lettering (eg Letraset).

To set the FSD of the meter, connect a voltmeter across the output terminals and turn RV2 up to maximum. Now alter PR1 until the reading on M1 agrees with that of the voltmeter.



The back of the box with Q9 mounted on a substantial heatsink.

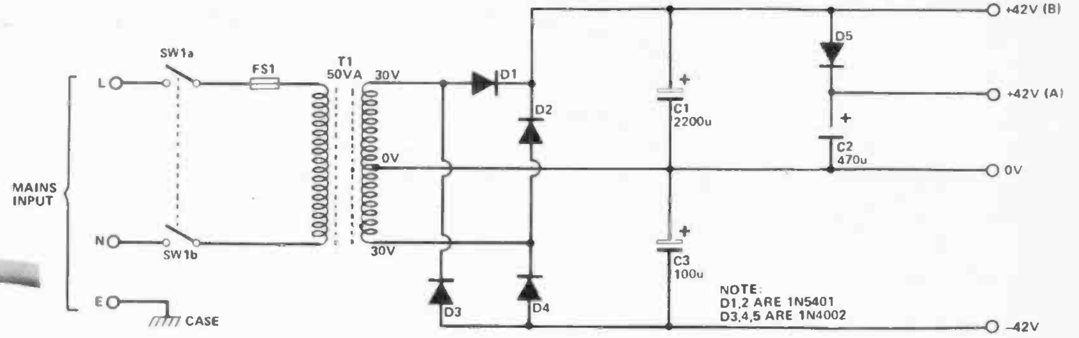


Fig.1 The basic power supply circuit.

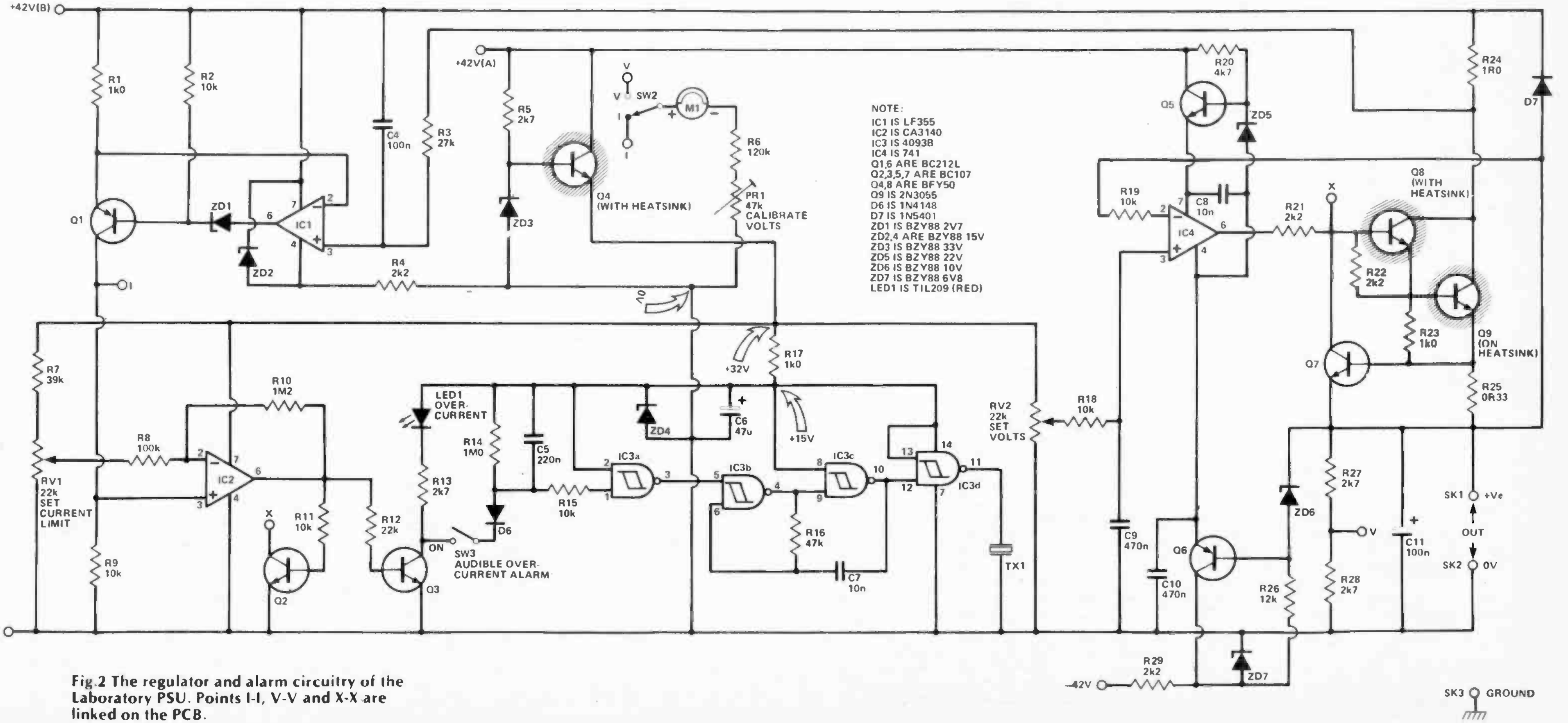


Fig.2 The regulator and alarm circuitry of the Laboratory PSU. Points I-I, V-V and X-X are linked on the PCB.

PARTS LIST

Resistors (all ¼ W, 5% except where stated)

R1,17,23	1k0
R2,9,11,15,18,19	10k
R3	27k
R4,21,22,29	2k2
R5,13,27,28	2k7
R6	120k
R7	39k
R8	100k
R10	1M2
R12	22k
R14	1M0
R16	47k
R20	4k7
R24	1R0 2W5
R25	0R33 ½ W
R26	12k

Potentiometers

RV1,2	22k linear
PR1	47k miniature horizontal preset

Capacitors

C1	2200u 63 V axial electrolytic
C2	470u 63 V electrolytic (PCB type)
C3	100u 63 V axial electrolytic
C4	100n ceramic
C5	220n polycarbonate
C6	47u 25 V axial electrolytic
C7	10n polycarbonate
C8	10n ceramic
C9,10	470n polycarbonate
C11	100u 63 V electrolytic (PCB type)

Semiconductors

IC1	LF355
IC2	CA3140
IC3	4093B
IC4	741
Q1,6	BC212L
Q2,3,5,7	BC107
Q4,8	BFY50
Q9	2N3055
D1,2,7	1N5401
D3,4,5	1N4002
D6	1N4148
ZD1	BZY88 2V7
ZD2,4	BZY88 15 V
ZD3	BZY88 33 V
ZD5	BZY88 22 V
ZD6	BZY88 10 V
ZD7	BZY88 6V8
LED1	TIL209 (red)

Miscellaneous

T1	30-0-30, 50 VA
SW1	DPDT miniature toggle
SW2,3	SPDT miniature toggle
M1	moving-coil meter, 100 uA FSD
SK1,2,3	4 mm terminal post
FS1	500 mA fuse and holder
TX1	PB-2720

Heatsinks (two off T05 and one off 60 x 100 mm finned), mounting kit for T03 transistor, case ref SWF 222X (see Buylines).

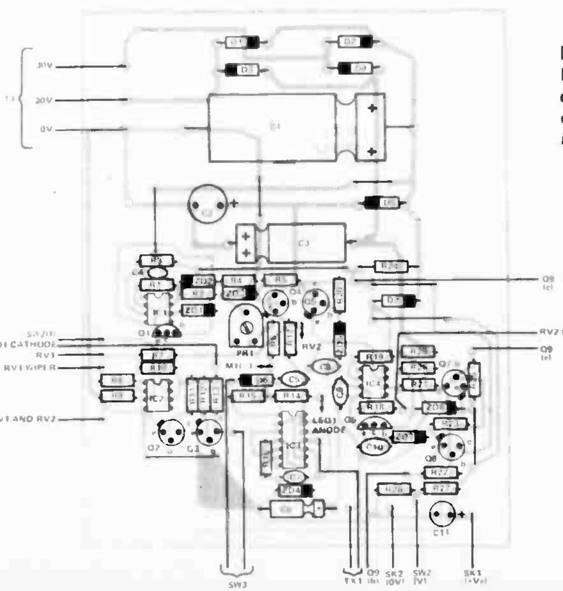
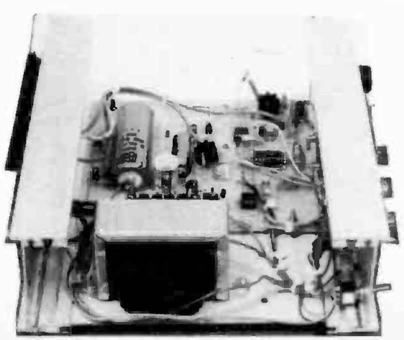


Fig.3 Component overlay for the Laboratory PSU. Don't miss the off-board connections in the centre of the PCB — the ones to M1(−), LED1 anode and RV2.



An internal shot of the case showing how we fitted everything in. The piezo-electric buzzer mounts on the right-hand end cheek. The case comes apart into several sections to make drilling and construction much easier.

HOW IT WORKS

At the outset of the design of this project it was decided that the PSU would be required to span the full 0-30 V output range and have semi-precision current limiting that is fully variable from 20 mA to 1A2. These requirements precluded the possibility of using reasonably-priced regulator ICs as the basis of the design, hence the use of a relatively large number of discrete transistors and IC op-amps in our project.

The main circuit can be broken down into a number of distinct blocks, as follows. ZD3-R5-Q4 and RV2 act as a variable reference-voltage generator, in which the slider voltage of RV2 is fully variable from zero to 32 V. IC4-Q8-Q9 act as a high-power voltage follower, with the output voltage across C11 closely following the voltage set on RV2 slider: Q9 has short-circuit protection provided by R25 and Q7, which automatically limit the peak output current of the circuit to about 1A8. To protect IC4 against excessive supply rail voltages without impairing its operating capabilities, the supply to this IC is provided via the Q5-Q6-ZD5-ZD6-ZD7 'sliding voltage generator' network, which holds the negative terminal of IC4 (pin 4) at roughly 9 V below the C11 output-terminal value and the positive terminal (pin 7) at 22 V above that of pin 4. All of the above circuitry comprises the variable voltage regulator section of the PSU.

The variable current-limiting part of the project is designed around R24-IC1-Q1 and RV1-IC2-Q2, with audio-visual over-current alarm indication given by the Q3-LED1-IC3 network. These sections of the PSU operate as follows.

The output current of the voltage regulator is monitored by R24; a voltage of 1 V/A is generated across this resistor. This voltage is integrated (to eliminate the effects of transients) by C4-R3 and the resulting voltage is fed to the input of the IC1-Q1 DC level translator, which has a gain of 10 and causes a voltage of 10 V/A to be generated across R9. This voltage is fed to one input of uni-directional differential amplifier IC2; the other input of this amplifier is derived from the slider of RV1, and the output of IC2 is coupled to point 'X' of the voltage regulator circuit via Q2.

The overall action of IC2 is such that, if the current-related R9 voltage is below that set via RV1 slider, Q2 is cut off and the voltage regulator acts in the normal way, but if the R9 voltage is greater than that of RV1 slider Q2 is driven on linearly and pulls point 'X' towards ground (0 V), thereby reducing the output voltage (and hence the current) of Q9. IC2-Q2 and the associated components are effectively coupled into a current-sensitive negative feedback loop with the voltage regulator, causing the output current of Q9 to self-limit at a mean current level (variable from 20 mA to 1A2) that is set using RV1.

Note that the output of IC2 is also coupled to Q3 via R12. Consequently, when the current-limiting circuitry becomes active, Q3 is driven on and causes LED1 to illuminate, giving a visual indication of the over-current state. Simultaneously, if SW3 is closed, the IC3 audible over-current alarm circuit is activated: in this circuit, IC3b-IC3c form a gated astable, with its output fed to the PB-2720 acoustic transducer via buffer stage IC3d and with gating provided from Q3 collector via the IC3a peak-detecting network.

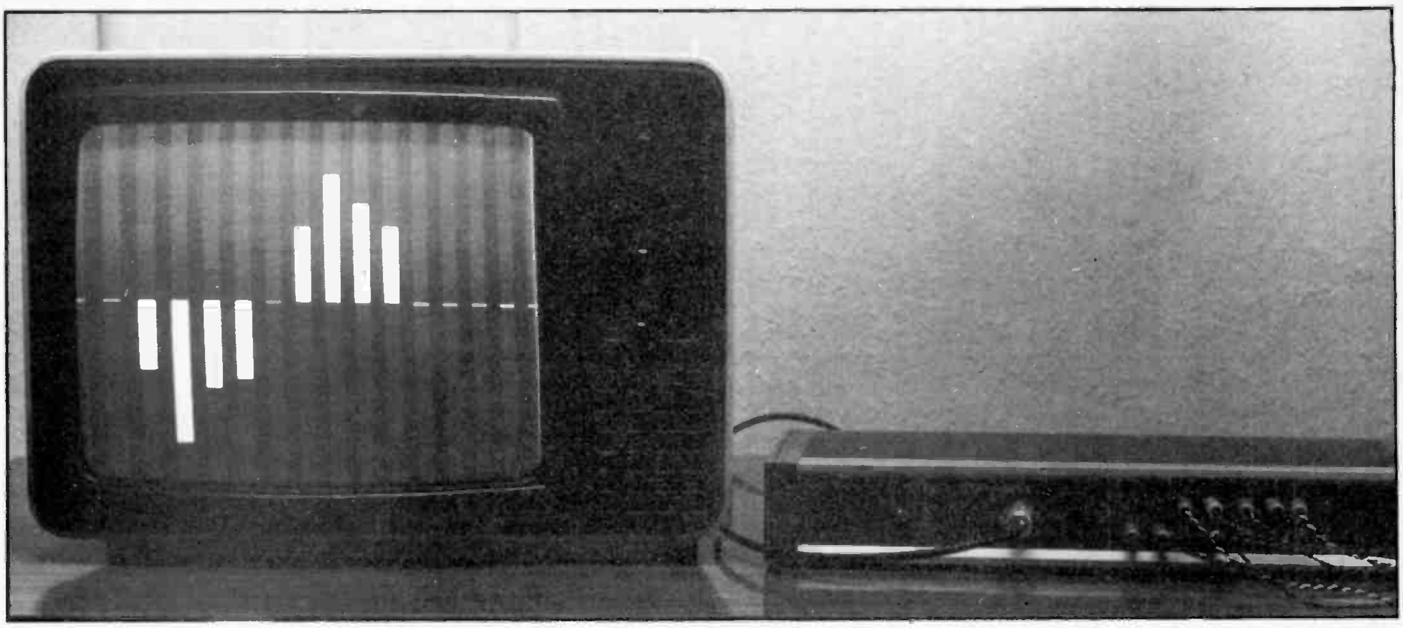
Note that the complete PSU circuit uses a total of three supply rails. The 42 V negative rail is used to enable pin 4 of IC4 to swing negative when required; this rail is required to provide only 20 mA or so of current. Two positive supply rails are used in the circuit: the + 42 V (A) rail drives most of the low-power sections of the circuitry and is required to provide only a few tens of milliamps. The + 42 V (B) rail powers IC1-Q1 and the high-power Q8-Q9 sections of the circuit and is required to provide mean currents up to 1A2. The use of the two positive supply rails enables power losses to be minimised and enables the main smoothing capacitor (C1) to be given a relatively low value.

BUYLINES

The case used in this project is one from the Swift range offered by West Hyde Developments and costs £11.96 including postage and VAT. The high power resistors are available from Electrovalue, the PB-2720 is stocked by Ambit, and the LF355 can be obtained from Marshall's. None of the other components should cause any problems.

The PCB for this project will be available from our new PCB service — see page 91 for details.

TV BARGRAPH



Lots of information to display? Make better use of the box in the corner; statistics, vital or otherwise, look good on the ETI TV Bargraph. Design and development by Phil Walker.

It's been some time since we did a really *unusual* piece of test gear. Here we are setting out to provide a solution to the problem of displaying many variables with good resolution. This sort of requirement can arise with spectrum analysis, statistical measurements, multi-point measurements of temperature, pressure, humidity, speed, current, voltage or indeed anything which can be converted to a proportional voltage.

It was realised some time ago by the electronic games and home computer manufacturers that most of the households who would be interested in their products would own at least one television set. This, we feel, is probably also true for you, our readers. So we have devised this instrument with that in mind.

The main problem in using a television set is that to obtain a high quality display on it the various synchronising signals normally produced by the gentlemen and ladies of the BBC and IBA must now be produced by our humble selves.

At this stage, we can go one of two ways. There are on the market several specially made integrated circuits for controlling VDU systems for home and commercial computers. However, this time we decided to steer clear of these and stay with gates and counters in the standard CMOS range.

The object of this project is to generate a display on a television screen consisting of a number of vertical columns. The height of each column is proportional to a specific input voltage. The columns may be upwards (positive) or downwards (negative) from a reference level. This reference may be changed if only positive-going signals are to be processed.

In order to generate the sync pulses we must first have some idea of their structure. Figure 1 shows the pattern of synchronising pulses aimed for in this design.

A normal TV picture consists of a total of 625 lines. Of these, 312½ are scanned each 20 mS such that each block covers the whole display screen, but the lines of one block fall neatly between those of the next. This reduces the impression of lines across the screen while also preventing objectionable flicker effects.

The apparent complexity of the sync pulse pattern is designed to ensure that a normal TV set can pick out accurately the right moment for line and frame flyback. The last thing necessary for a complete picture is tht the video signal shall appear at the correct times between the line sync pulses, and be blanked during the line flyback and frame flyback periods.

The Circuit

The basic timing for the whole circuit is derived from a 2.5 MHz crystal oscillator. This was found to be essential as minute changes on

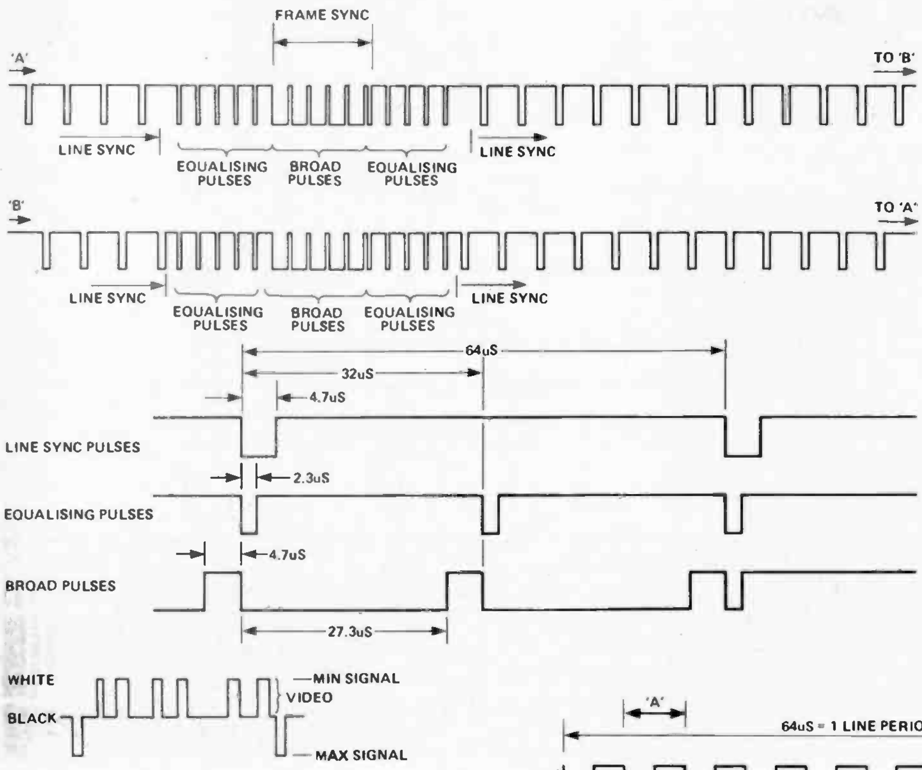


Fig. 1 Sync pulse structure.

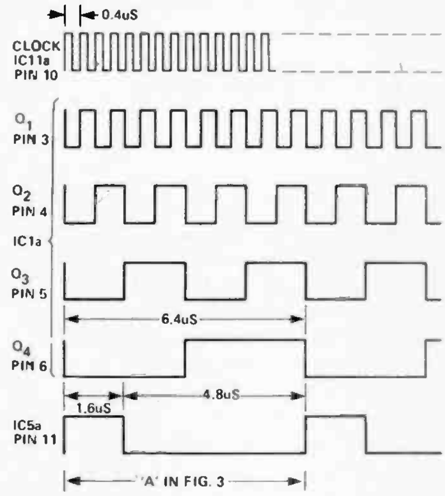


Fig. 2 Expanded section of part of Fig. 3.

the power supplies caused very visible display distortions when a free running oscillator was used. This 2.5 MHz signal is divided by 160 to give the basic line time of 64 uS. The logic associated with the line dividers gives two pulses per line period at IC5b which trigger the 'equalising' and 'broad' pulse monostables. The 4098/14528 dual monostable device (IC10) was used because derivation of these signals with normal logic would be very complicated (see Figs. 2 and 3).

The line sync pulse is generated by the logic (IC4a) and a choice between this and the other two pulses is effected in ICs 5c and 9.

The double line rate signal at IC5b drives a divide-by-5 prescalar (IC2b) and then a divide-by-125 device, IC3 (see Fig 4). This produces the 20 mS period for the frame scan. By virtue of the fact that the input frequency to these dividers is twice the line frequency and the total division ratio is 625, the number of lines scanned is 312½ in each vertical scan. This means that successive scans will be offset by half a line pitch vertically. The generation of the frame sync signal is accomplished by IC1b, IC7 and IC8.

When IC3's count reaches zero its output goes low for one input clock period. This causes IC1b to be reset to zeros and then to start counting and select the components of the frame sync at

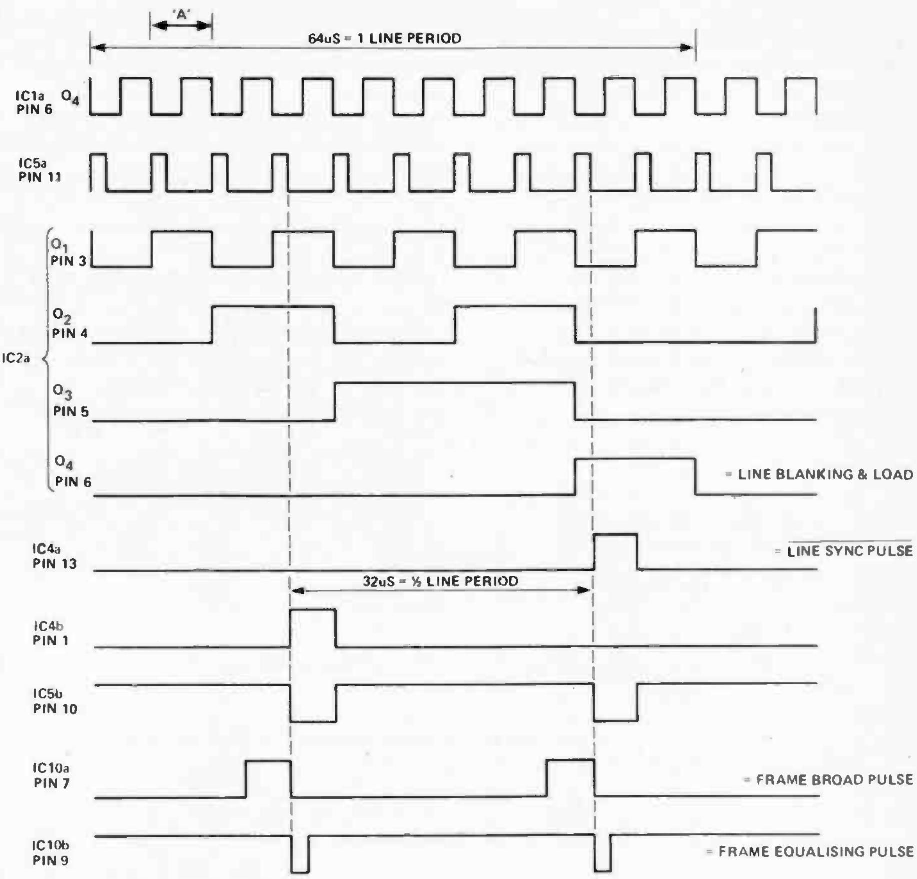


Fig. 3 Sync pulse generation.

the appropriate times. When IC1b reaches its all 1s state, further counting is inhibited (Fig. 5).

The video generation is performed mainly on the channel cards. These consist of eight comparators and an eight-bit shift register on each card. The cards are controlled from the main board by a clock signal which determines the number of bars to be displayed. The load signal is the same as the line blank signal. This effectively causes all the channels to be sampled (in the line flyback period) and then displayed.

The other necessary signals provided are the reference and a ramp signal to which the inputs are compared (see Fig. 5).

Construction

In order to keep costs down the PCBs for this project are single-sided. However, this has meant a number of links on the boards. Most of these are used to complete the power supply lines to the devices. The actual construction of the boards is not difficult but we suggest that IC sockets would be a good idea. Most of the links on the board can be put in place before any other components although the one near IC11 and C1 should be left a little longer to allow C1 to be inserted. Insulated wire is recommended for all the links. The link from CLK to 8, 16, 32 or 64 should be made when the number of channels is known. It is easier in fact to put thick wire posts or PCB pins through these holes and link up afterwards.

The eight channel position is useful for setting up the complete system initially and making sure that it works.

It is most desirable that a fine-tipped soldering iron is used during construction and that a check is made for solder bridges, especially where tracks lie between IC pins. It is also most important that all polarised components are fitted the right way round. Especially note that the TL084's and the 4014 on the channel cards are not the same orientation.

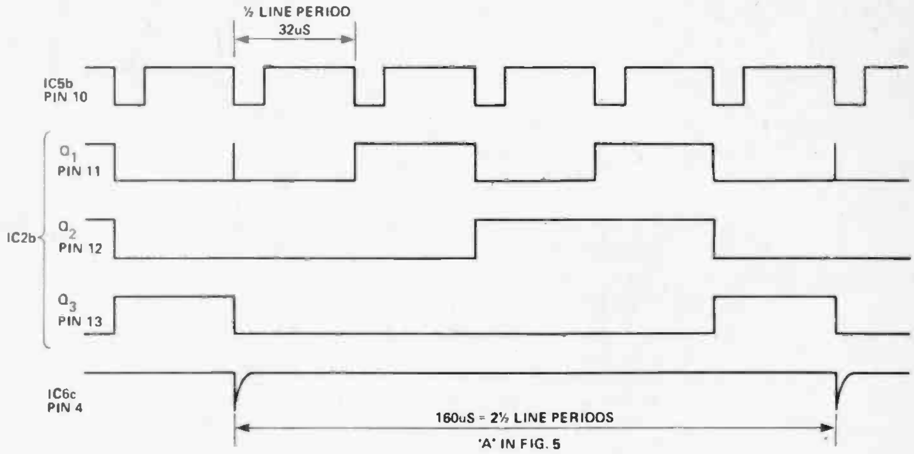


Fig. 4 The divide-by-5 effect of IC2b.

The completed boards in our prototype were fitted into a metal case made by Newrad; a small aluminium bracket was needed to support the rear of the main board as this overhangs the integral chassis member by about 1 cm. Support for the channel cards is by two pieces of PCB material with suitable holes drilled in them. One end of these members is bolted to the main board while the other is supported on pillars. When fitting the boards it is essential that no part of the 0 V supply line gets linked to the metalwork as this will short-circuit the reference supply.

On the channel cards, it is helpful if the channel inputs are fitted with PCB pins or similar to facilitate connection from the top of the board. The 2.5 mm jacks on the front panel used in our model could well be replaced with nine-way 'D' range connectors or anything else of that type, if desired.

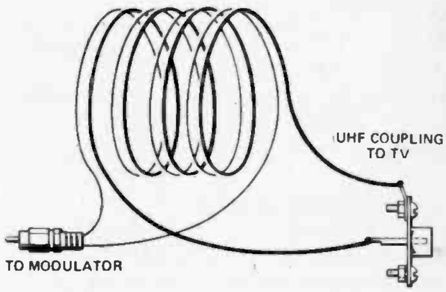


Fig. 6 The UHF coupling coil.

The output from the UHF modulator should be DC-isolated from the panel and our method, shown in Fig. 6, was to take about 20 cm of thin twin flexible wire and coil it about itself to a diameter of 1.5 mm (this makes a simple 1:1 transformer). Connect the ends of one wire to the coax socket pin and skirt, and the ends of the other wire to a phono plug pin and skirt. A possible improvement here would be to use a special two-hole ferrite core with two windings of four or five turns through the holes. Capacitance coupling of the UHF signal can cause stability problems in the voltage reference amplifier. Oscillation here may appear as broken or strange-shaped bars or possibly break-up of the picture.

Connection between the main board and channel cards and card to card is by 10-way jumper wire (or just link them direct). This is probably easiest done with the cards mounted on the support bars. Each channel card adds eight channels to the units' capability.

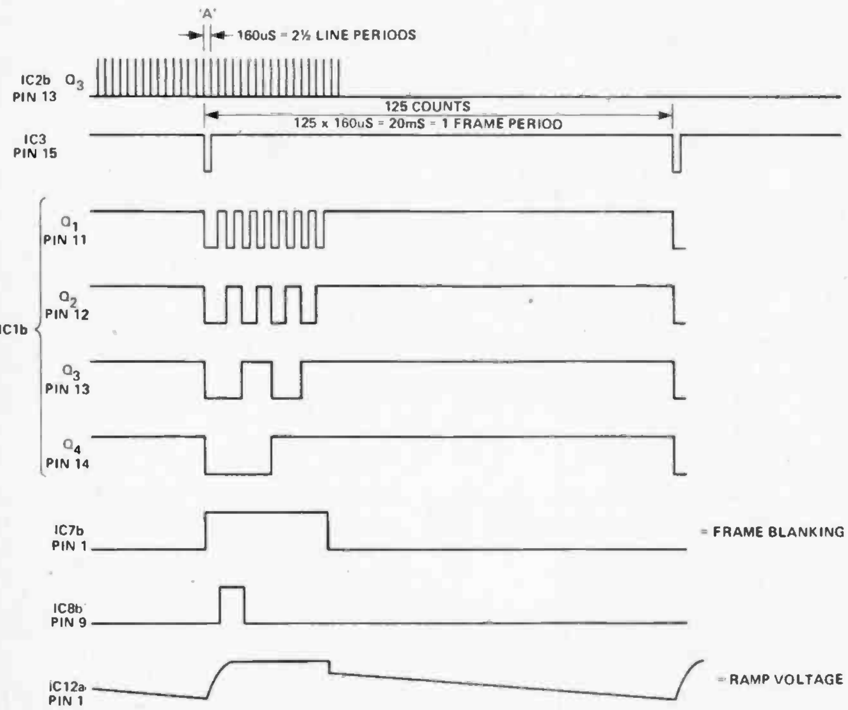
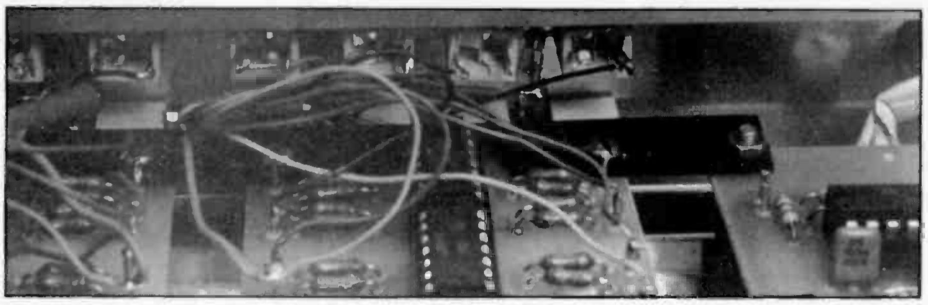


Fig. 5 Frame sync generation.



PARTS LIST

MAIN BOARD

Resistors (all ¼ W 5%)	
R1	1M0
R2	22k
R3,4,5	10k
R6,7,8	2k2
R9,10,12	10k
R11	100k
R13	Omit or select for reference voltage required

Potentiometer	
PR1	10k miniature horizontal preset

Capacitors	
C1	10p ceramic
C2	4n7 ceramic
C3	220p ceramic
C4	100n ceramic
C5,6	10u 35 V tantalum bead
C7	10u 25 V PCB aluminium electrolytic
C8	220u 25 V axial aluminium electrolytic
CV1	2-22p miniature trimmer

Semiconductors	
IC1	4520B
IC2	4518B
IC3	40103B
IC4	4002B
IC5	4001B
IC6	4011B
IC7	4012B
IC8	4025B
IC9	4023B
IC10	MC14528 or CD4098
IC11	4070B
IC12	TL084
D1-D4	1N4148
ZD1	4V7 400 mW zener

Miscellaneous
UM1233 UHF modulator (Aztec), 8 MHz bandwidth; 2.5000 MHz crystal.

CHANNEL CARD

Resistors (all ¼W, 5%)	
R101	56k
R102-117	27k

IC101	4014B
IC102,3	TL084

Miscellaneous
PCBs (see Buylines); 2.5mm jack sockets (eight per channel card); coaxial panel socket; miniature on/off toggle switch; PP9 battery and battery clips; phono plug; two off coaxial plugs and suitable length of coaxial cable; nuts, bolts, pillars, 12 x 12 mm aluminium angle or similar bracket; two off 12 x 100 mm (approx.) pieces of PCB laminate or similar (channel card support); wire, IC sockets etc; case (Newrad NP1426).

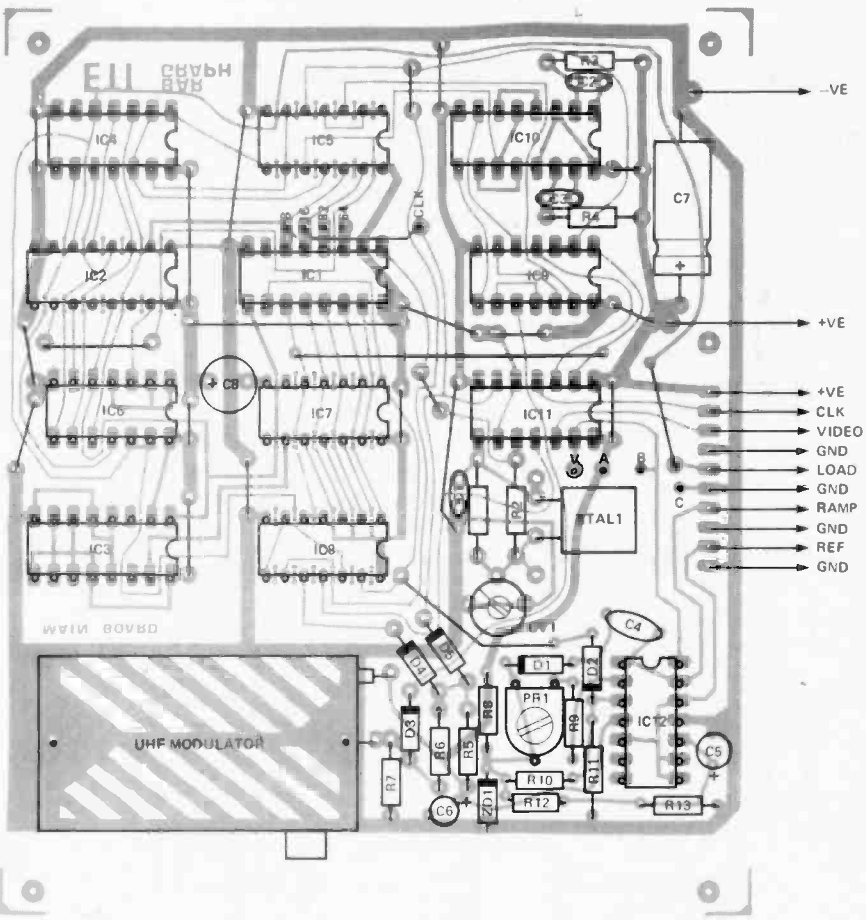


Fig. 7 Component overlay of the main board.

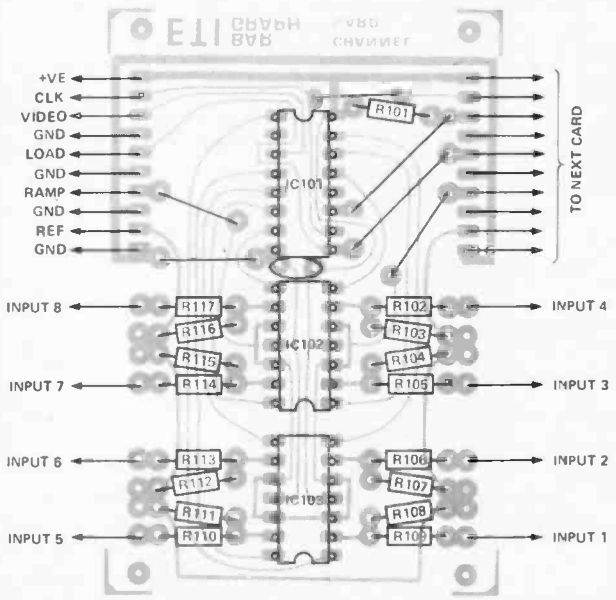
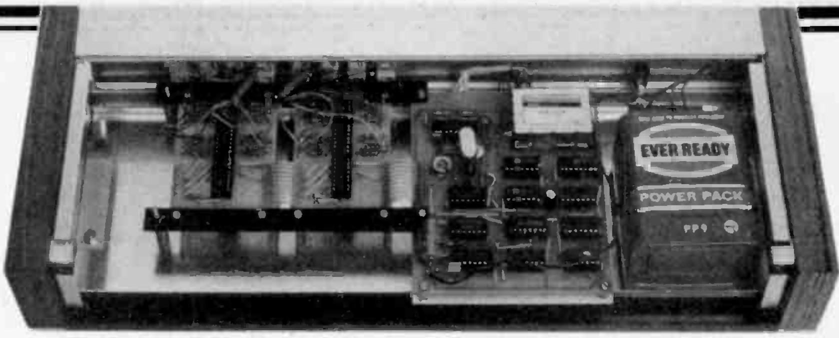


Fig. 8 Overlay for one of the channel cards.

BUYLINES

Since we elected to use standard components rather than fancy chips for this project, there should be absolutely no supply problems. The modulator is from Technomatic or Watford, but make sure you get the wide-band (8 MHz) version. The case we used came from Watford. The PCB Service order form is on page 91.



The maximum number of channels sensibly usable is, we feel, 64, although 128 is possible on a good television set with the modulator specified.

When constructing the case for this project, it was found that the application of a drop of cycle oil to the self-tapping screws holding the extruded front pieces made it very much easier to screw them in. As we assembled and disassembled this part several times during construction, this was most helpful. Also, when assembling the chassis member to the side panels of the case, put the piece with the side flanges upwards and bolt it on underneath the lugs on the side panels. This is necessary to allow sufficient room for the battery inside the case.

Setting Up

There is very little in the way of setting up to be done. The basic unit should give some sort of display when powered up although no bars (or very few) will be present. To see anything more, it is necessary to feed some voltage into the channel inputs. As specified, full-scale should be about 6 V. This sensitivity can be altered by changing the input resistors as required. The reference voltage can be adjusted slightly using R13 or overridden by driving IC12 pin 10. The ramp rate is controlled by PR1 and should provide a fair control range for many purposes enabling the reference voltage crossing to be positioned near mid-screen.

When first trying the board, set CV1 to about half-capacitance (plates half-meshed) and adjust it only if the picture is unstable or cannot be pulled in by the television line and frame hold controls. (Don't forget to tune it in as accurately as possible).

Use

When fully operational the device should give the required number of bars along the screen, with a vertical height proportional to the input voltage for each one. The vertical resolution is about 270 steps, corresponding to a pair of interlaced lines. In the display mode built into the board, bars are not visible until the input is greater or less than 0 V. They then appear above or below the centre line; if the input sensitivity is too low then extra conditioning amplifiers will be required.

To change the display mode,

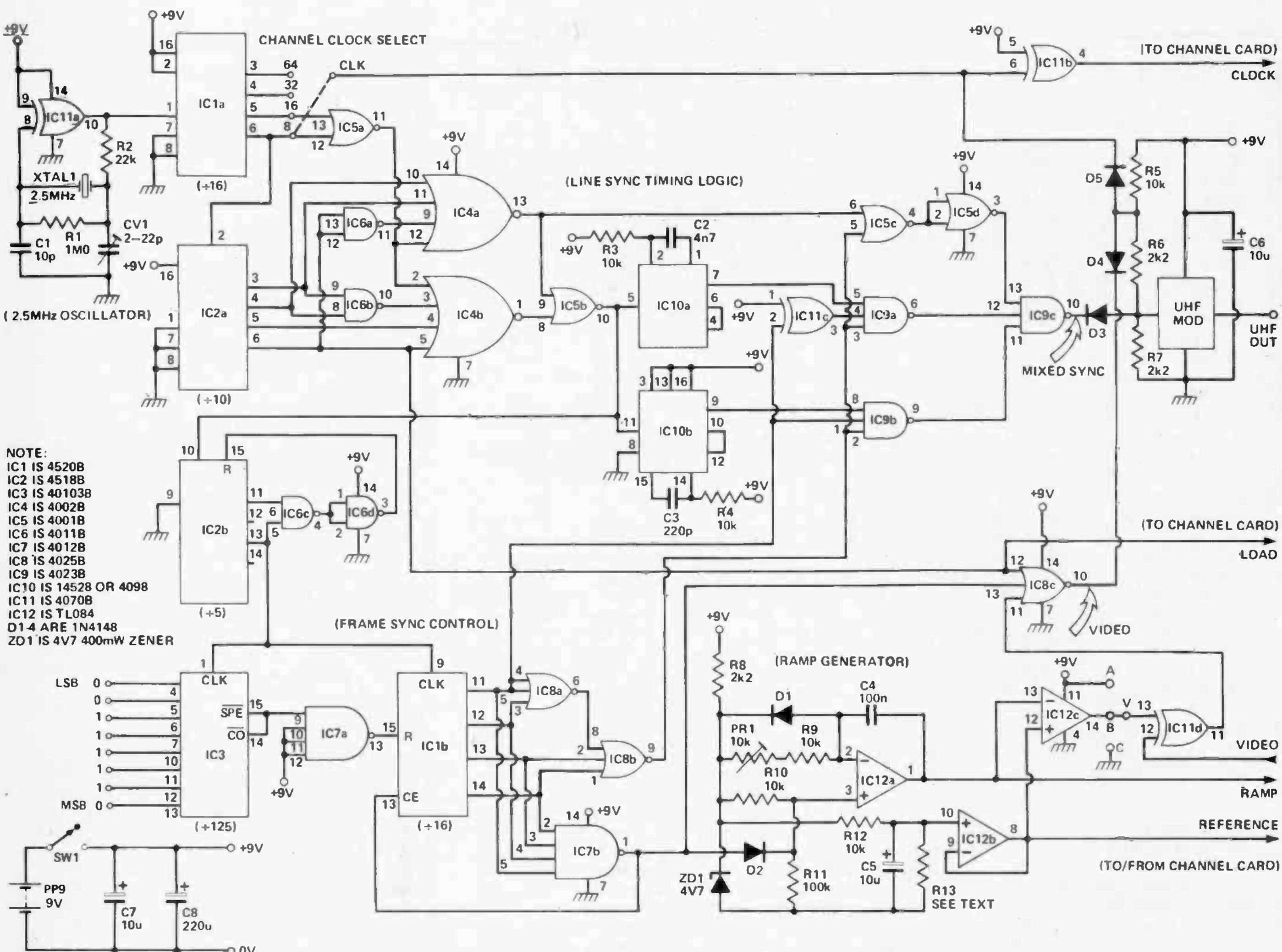


Fig. 9 Circuit diagram for the main board of the TV Bargraph.

break the track linking points V and B on the overlay. Linking V to A gives bars which start at the top of the screen and move down for decreasing input, while linking V to C gives bars which start at the bottom and move up with increasing input. Leaving out D5 will give double-width bars which merge with those on either side.

It is possible when using large numbers of channels that some will be off the edges of the display screen; this can only be cured by using a higher quality TV or monitor as the line flyback time in cheaper sets is often longer than that specified (12.8 uS). A way round this is to miss out the channels on the first card and start with the next. The device is ideal where a semi-graphical display of multiple inputs is required and has the advantages of easy expansion and no requirement for computing facilities.

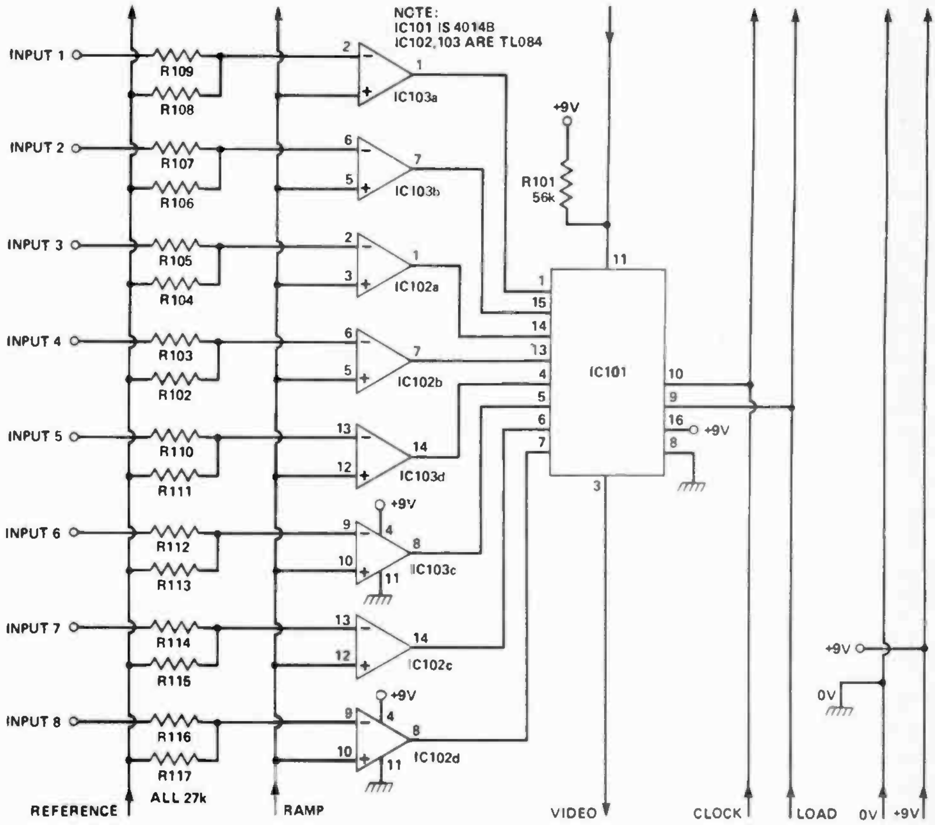


Fig. 10 Circuit diagram for one channel card; this provides eight input channels.

HOW IT WORKS

SYNC GENERATOR AND TIMING

IC11a is connected as an inverter and with XTAL1 and associated components forms a crystal-controlled oscillator working at 2.5 MHz. This is divided by 16 in IC1a and 10 in IC2a to give the 15,625 Hz for the line sync generator. The outputs from IC2a are decoded by IC6a, 4a and 4b to give two outputs. One is during count 1000 for the line sync, the other is at count 0011. The outputs from IC4a and 4b are disabled for about 1.6 uS after the active clock edge by IC5. This allow the counters and logic to settle and prevents glitches due to propogation delays. It also starts the line sync pulses at the right time relative to the line flyback blanking pulse taken from IC2a pin 6.

As IC2 is a dual decade counter, counts 1000 and 0011 occur at equal time intervals after each other. These outputs from IC4a and 4b are combined in IC5b to form a pulse chain at exactly douhle the frequency of the line sync pulses (Fig. 3). The regularity of this signal is important to get correct interlacing of the final picture. This signal triggers the two sections of IC10 and also IC2b. IC2b, IC6c and IC6d divide the input by 5 before driving the clock inputs of IC3 and IC1b. IC3 is a presettable eight-bit down counter which is configured to divide by 125, giving an output equal to the input clock period each 125 input cycles. Thus we get a low pulse 2½ line periods long (five double frequency pulses) every 312½ line periods (5 x 125 = 625 double frequency pulses). The low pulse from IC3 is inverted by IC7a and resets IC1b to all zeroes; it then holds it there for 2½ line periods. At this point IC1b will be incremented by the output from IC2b (ie every 2½ line periods) until it reaches the all 1s state. IC7b detects this condition and its output inhibits further counting. The output from IC7b is also used to blank the video signal during the frame flyback period. The actual frame sync signal is generated during counts 0001, 0010 and 0011 of IC1b. The logic for this is provided by IC8a and 8b. The output from IC8b determines whether normal line sync or frame sync is required and IC1b pin 11 decides which type of frame sync signal is to be sent. The signals are actually switched and combined in IC9, IC11c, IC5c, d.

IC10a and 10b are monostables triggered by the double line frequency pulse chain and provide the 2.3 uS 'equalising' pulses and 27.6 uS 'broad' pulses reqired for proper synchronisation in the 625 line system.

VIDEO AND CHANNEL SAMPLING

Having generated all the sync pulses to stabilize the display format, the video information must be generated. IC12 generates a negative-going ramp signal, synchronised with the frame blanking signal via D2. The frame blanking signal forces the non-inverting input of IC12 high, so the output of IC12 goes high to try and reduce the differential voltage between its inputs. However, the inverting input cannot go more than about 0V7 more positive than the cathode of ZD1. This means that C4 will charge very rapidly. When the frame blanking period is over, D2 is effectively out of circuit and the voltage on the non-inverting input to IC12 will be about 10/11 of V_{REF}. By normal op-amp operation the inverting input to IC12 will also be at this voltage, causing a current of 1/11 V_{REF} / (R10 + PR1) to flow through C4. The result of this is that the output voltage from IC12 will now fall linearly until it approaches the 0 V rail or another frame blanking pulse occurs.

Another section of IC12 provide a buffered reference voltage while a third acts as the comparator — switching as the ramp voltage passes the reference.

The ramp and reference voltages pass to the channel cards where the video signal is generated. Each input signal is compared with the ramp voltage in a section of IC102 or IC103. The more positive the input signal, the sooner its comparator will switch and the higher up the screen its bar will start.

The outputs of all the eight comparators on each card are fed to the parallel inputs of a 4014 eight-bit shift register (IC101). During the line flyback blanking pulse the data is loaded into the 4014; during the rest of the line time it is clocked out under the control of the channel clock. If more than one channel card is in use the output from each additional card is fed to the serial input of the next card along. This effectively extends the length of the shift register. For best results 8, 16, 32 or 64 channels should be used. Four or 128 channels are possible with modifications while 24, 40, 48, 56 etc will give poor display formats.

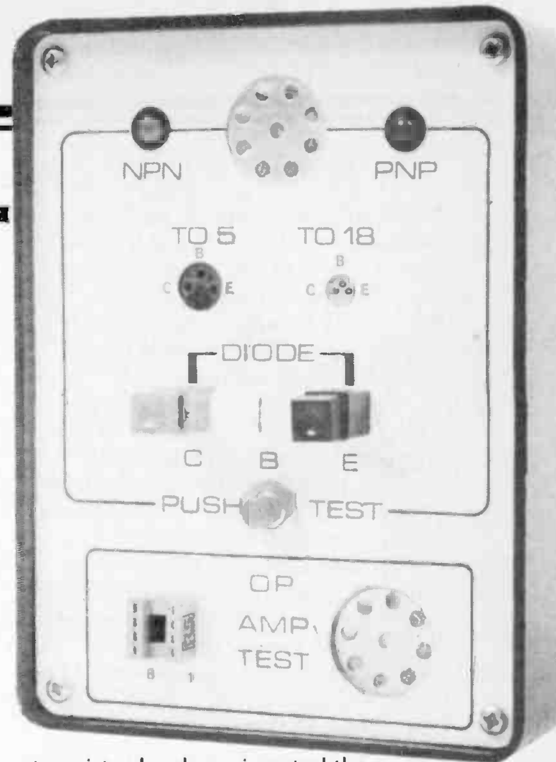
The video signal from the channel card(s) returns to the main board and, via IC11b, is mixed with the sync and blanking signals in D3,4,5 and R5,6,7 before going to the UHF modulator.

The line blank signal mentioned above is derived from IC2a pin 6 (Q_4) while the frame blank signal comes from IC7b. The output of this last device is high for a total of 40 lines in each half frame (16 x 2½) leaving 272½ lines for the actual display. This is still probably more than a normal portable TV will display vertically.

The feeding of the channel clock signal via D5 into the video mixer causes the bars to be separated and half width. Omitting D5 allows the bars to broaden and merge with each other. The jumper points V, A, B, C are pre-linked to give a video inversion at $V_{RAMP} = V_{REF}$. Alternative effects can be obtained by linking V to one of the other points, when bars starting from the top or bottom of the screen will be obtained instead of starting from the centre line.

COMPONENT TESTER

Check out your semiconductors with this cunning but simple project. It's brilliant, even if we do say so ourselves (and we do). Design by Rory Holmes. Development by Tony Alston.



When you've completed your latest design, a brilliant project which not only solves the world energy crisis but proves that Einstein made a small mathematical error as well, it can be very frustrating if you rush to your junk box and discover that you can't breadboard the circuit because the markings have rubbed off your transistors. To help with this problem we've come up with *our* latest design, a brilliant project which tells you which lead is which, whether the transistor is OK, what polarity it is and its approximate gain. Diodes and LEDs may also be tested, and for good measure we've thrown in an op-amp checker. The world energy crisis you'll have to figure out for yourself.

Construction

Assembly is straightforward if the recommended PCB is used. Make sure to orientate IC1, IC2, D1 and D2 correctly, and use sockets for the ICs to avoid damage by soldering them. Remember to put the three wire links on the PCB!

Although there are quite a few off-board connecting wires, these should not be a problem if the circuit diagram, overlay and internal photos are studied carefully. Only one transistor test socket is shown on the circuit diagram but several types can be wired in parallel (as we did) to accommodate various types of transistor. The TO-5 and TO-18 types were epoxied to the front panel, as was the eight-pin DIL socket for the op-amp tester. Three insulated test terminals were also included for testing other types of transistors, diodes and LEDs.

TX1 and TX2 are crystal mike inserts, Eagle type MC25 or similar. Warning! — most inserts have one terminal connected to their case and as we've used a metal front panel for this project, TX2 should be insulated from this panel. Otherwise, TX1 and TX2 will be common linked and as the circuit diagram shows that TX1 is connected to 0 V, TX2's connection to IC1, IC2 and C2 will be incorrectly taken to 0 V. We got round the problem when we glued a circular fibre washer to one insert before fixing it to the front panel.

Testing Times

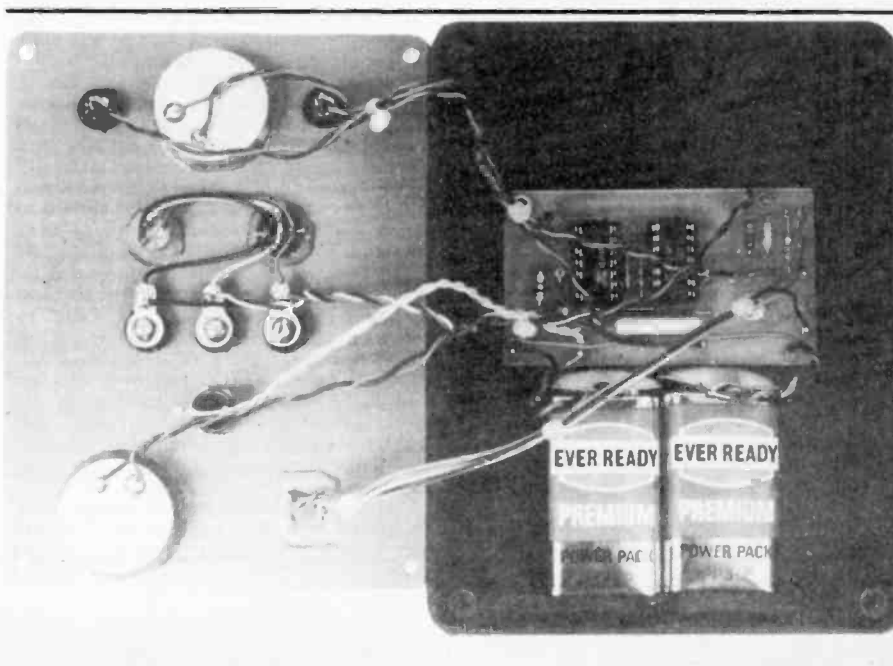
Transistors are plugged into the appropriate socket, and any type may be tested; NPN, PNP, small signal or power. No selection of NPN or PNP is necessary as this is done automatically by the tester. When the push-to-test button is pressed, an intermittent tone is produced. The frequency of the tone is proportional to the gain of the transistor, giving a rough guide. The LEDs also flash alternately in time with the pulsing tone; the LED that is on at the same time as the tone indicates the polarity of the transistor. If the transistor has no gain or is open circuit there will be no tone, although the LEDs will still flash. If the transistor has a large leakage current or is shorted, there will be a 'two-tone' sound. If the transistor has been inserted the wrong way there will be either no tone or a very high-pitched tone.

Diodes and LEDs may be tested across the 'C' and 'E' terminals. If it is OK, the LED under test will flash, accompanied by an intermittent high-pitched tone and flashing indicators. Ordinary diodes require a series resistor (any old value) and should then produce an intermittent tone and flashing LEDs as before; the coincidence of flashing LED and tone indicates the anode.

Op-amps are plugged into the IC socket and no push-switch is required; power is only applied when the IC is inserted, and a good IC produces a continuous tone from the second insert.

BUYLINES

No problems with anything used in this project; all components are standard items and are obtainable from the major mail order suppliers advertising in this issue. If you don't want to make your own PCB, you can obtain one from our PCB Service (see page 91).



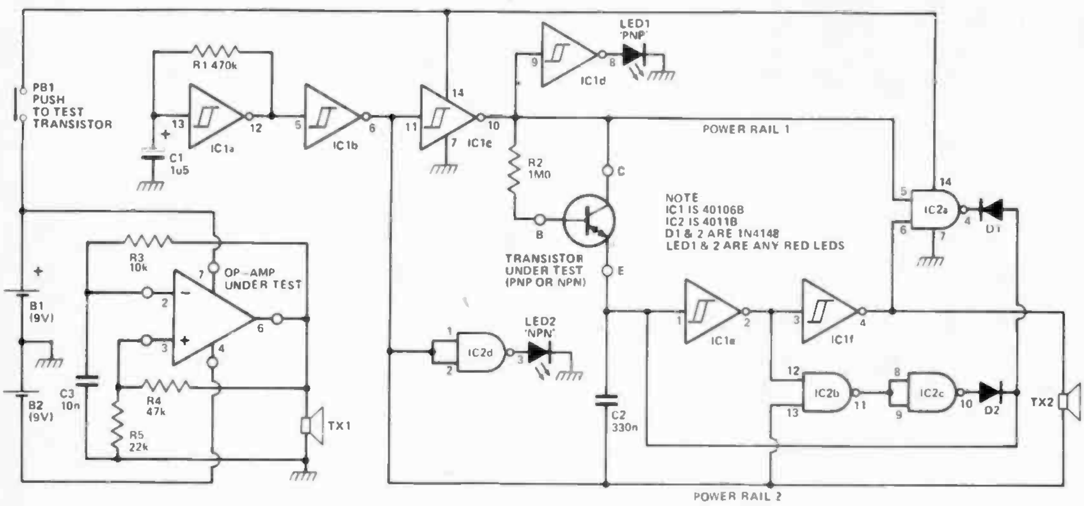


Fig. 1 Circuit diagram of the Component Tester.

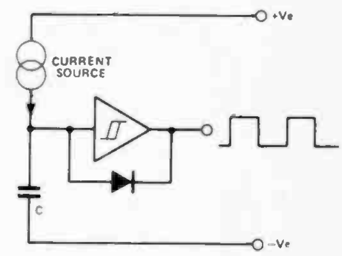


Fig. 2 Principle of the CCO.

HOW IT WORKS

The op-amp tester and transistor tester are completely separate circuits; we shall deal with the transistor tester first. IC1a, a Schmitt trigger inverter, forms a low frequency square wave oscillator with a period (determined by R1 and C1) of about 1 second. This square wave is used to switch the polarity of the 'power rails' (labelled 1 and 2 in the diagram) of the test transistor and its associated oscillator circuitry.

IC1b is used to buffer the square wave, and its output (on pin 6) is used to drive 'power rail 2'. This switching signal from IC1b is also fed to the input of IC1c, which inverts it and drives 'power rail 1'. Thus for half a second in each cycle rail 1 will be positive (high) and rail 2 (low); for the other half second rail 1 goes negative and rail 2 positive. Each power rail drives an LED (LEDs 1 and 2) via inverter gates IC1d and IC2d, such that an LED will be illuminated if its associated power rail is at 0 V. These LEDs will therefore flash alternately when the circuit is operating, providing an indication of the state of the power rails.

The oscillator circuit that is connected across these power rails is essentially the simple current-controlled oscillator shown in Fig. 2, but with some adaptations to enable it to work with either supply polarity. The oscillator of Fig. 2 works as follows. Assume C is initially discharged, so that the input to the Schmitt inverter is low; the output is thus high and the diode, being reverse biased, is effectively out of circuit. Capacitor C will now begin charging up from the current source and the input voltage to the Schmitt will be increasing. When the input passes the Schmitt threshold the inverter output will switch low; the diode is now forward biased and will rapidly discharge the capacitor. The process then repeats, producing a square wave output from the inverter with a frequency that is proportional to C and the current from the source. The bigger the current from the source, the faster C will charge and the higher the frequency will be.

The current source in our actual circuit is provided by the transistor under test. R2 supplies a small base current to the transistor, and the current flowing from the emitter will be proportional to the gain of the transistor. If the transistor is PNP it will only supply current to the CCO (current-controlled oscillator) when power rail 1 is negative with respect to power rail 2. Similarly, power rail 1 must be positive for the oscillator to function if the transistor is NPN. Thus the CCO will produce an intermittent frequency for either transistor polarity (assuming the transistor is a good one) with a frequency roughly proportional to the gain. If the frequency is audible when LED1 is on, the transistor is PNP, and if LED2 and the tone coincide then it is NPN.

Going back to the oscillator of Fig. 2, we see that if the oscillator is to work when the supply connections are reversed, then the diode polarity must also be reversed. In our circuit this is achieved by using two diodes, D1 and D2. When power rail 1 is at 0 V, the NAND gate IC2b will be disabled and its output (pin 11) will be high. This output is inverted by IC2c, thus reverse biasing D2 which is now effectively out of circuit. At the same time power rail 2 will be high, enabling NAND gate IC2a whose output (pin 4) will follow the logic level on the output of the Schmitt trigger IC1e via IC1f. Thus when IC1e goes low during an oscillation cycle, the cathode of D1 will also go low, forward biasing the diode and discharging C2 for the next cycle.

When the voltage on the two power rails is reversed a similar action occurs, but with D1 switched out of circuit and D2 providing the discharge path. The intermittent square wave produced at the output of IC1f is fed to crystal transducer TX2 which gives an audible note.

If an LED or diode is connected between the C and E terminals of the test socket, it appears to be a large-value current source in one direction only. Hence the circuit reacts as if a high-gain transistor were in circuit, and polarity is indicated as before.

The op-amp under test is configured as a simple RC relaxation oscillator. When the op-amp is plugged in, assume that its output (pin 6) is high (positive saturation). Then C3 will begin charging up to + 9 V through R3 with a time constant C3.R3. When the voltage on C3 reaches one-third of the positive supply (this fraction is set by R4 and R5), the op-amp output will switch low, with R4 and R5 providing positive feedback for Schmitt trigger action. C3 will then discharge towards − 9 V, until the op-amp switches back to positive saturation. This cycle repeats indefinitely, producing a square wave at the op-amp output which is fed to transducer TX1. This produces an audible note if the op-amp is good.

PARTS LIST

Resistors (all ¼ W, 5%)	
R1	470k
R2	1M0
R3	10k
R4	47k
R5	22k
Capacitors	
C1	1u5 25 V tantalum
C2	10n disc ceramic
C3	330n polyester
Semiconductors	
IC1	40106B
IC2	4011B
D1,2	1N4148
LED1	0.2" red LED
LED2	0.2" green LED
Miscellaneous	
PB1	momentary push-button
TX1,2	crystal mike inserts

2 off PP3 batteries and clips; transistor sockets; IC sockets; case to suit

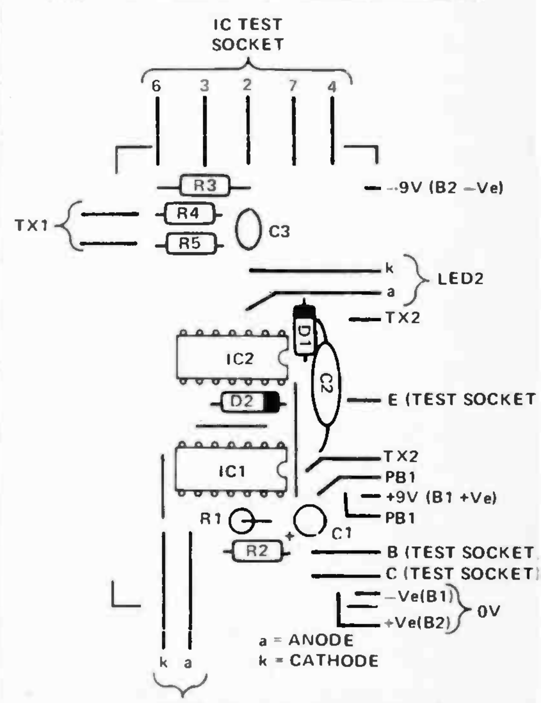
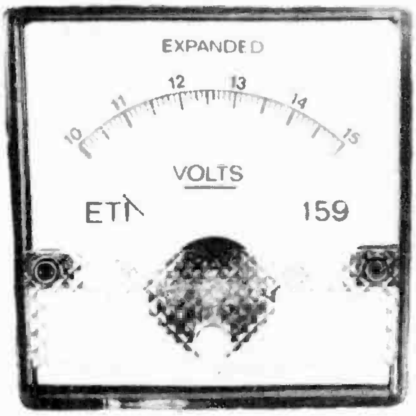


Fig. 3 Component overlay.

ACCURATE VOLTAGE MONITOR



This simple, low-cost instrument can be built into power supplies or used as a portable or fixed 'battery condition' monitoring meter. Design by Simon Campbell and Roger Harrison.

Common storage batteries to power nominal 12 V DC electrical systems have a terminal voltage that ranges from a little over 10 V when discharged to around 15 V when fully charged, the operating voltage being somewhere in the range 11V5 to 13V8. Lead-acid batteries, for example, may have a terminal voltage under rated discharge that commences at around 14V2 and drops to about 11V8. A 12 V (nominal) nickel-cadmium battery may typically have a terminal voltage under rated discharge that starts at 13 V, dropping to 11 V when discharged.

Equipment designed to operate from a nominal 12 V DC supply may only deliver its specified performance at a supply voltage of 13V8 — mobile CB and amateur transceivers being a case in point. Other DC operated equipment may perform properly at 12V5 but 'complain' when the supply reaches 14V5.

To monitor the state of charge/discharge of a battery, a battery-operated system or the output of power supplies, chargers, etc, a voltmeter which can be easily read to 100 mV over the range of interest (10 to 15 V) is an invaluable asset. This project does just that.

The Circuit

An LM723 variable voltage regulator is employed to set an accurate 'offset' voltage of 5 V, and the meter (M1) plus the trimpot RV2 and R3 make up a 5 V meter, with the trimpot allowing calibration. The negative terminal of the meter is connected to the output of the 723 so that it is always held at 5 V 'above' the circuit negative line. The positive end of the meter goes to a zener which will not conduct until more than 5 V appears between the circuit +ve and −ve lines. Thus the meter will not have forward current flowing through it until the voltage between the +ve and −ve rails is greater than 10 V, and will read full scale when it reaches 15 V (after RV2 is set correctly).

The meter scale limits may be adjusted by setting the output of the 723 higher or lower (adjusted by RV1) and setting RV2 so that the meter has an increased or decreased full-scale deflection range.

A variety of meter makes and sizes may be used.

Construction

Mechanical construction of this project has been arranged so that the PCB can be accommodated on the rear of any of the commonly available moving coil meter movements. We chose a meter with a 55 mm wide scale (overall panel width, 82 mm). A meter movement with a large scale is an advantage as it is considerably easier (and more accurate) to read than

HOW IT WORKS

The meter, M1, is a 1 mA meter with series resistance — made up of R3 and RV2 — so that it becomes a 0-5V voltmeter. The negative end of the meter is maintained at 5 V above the circuit negative line by the output of IC1, a 723 adjustable regulator. The positive end of the meter is connected to the circuit positive line via ZD1, a 4V7 zener diode. Thus, no 'forward' current will flow in the meter until the voltage between the circuit negative line and the circuit positive line is greater than 5 + 4.7 = 9V7.

Bias current for the zener is provided by a FET, Q1, connected as a constant current source so that the zener current is accurately maintained over the range of circuit input voltage. This ensures the zener voltage remains essentially constant so that meter reading accuracy is maintained.

The trimpot RV1 sets the output voltage of the 723. This determines the lower scale voltage. Trimpot RV2 sets the meter scale range, less resistance decreases it.

Diode D1 protects the circuit against damage from reverse connection.

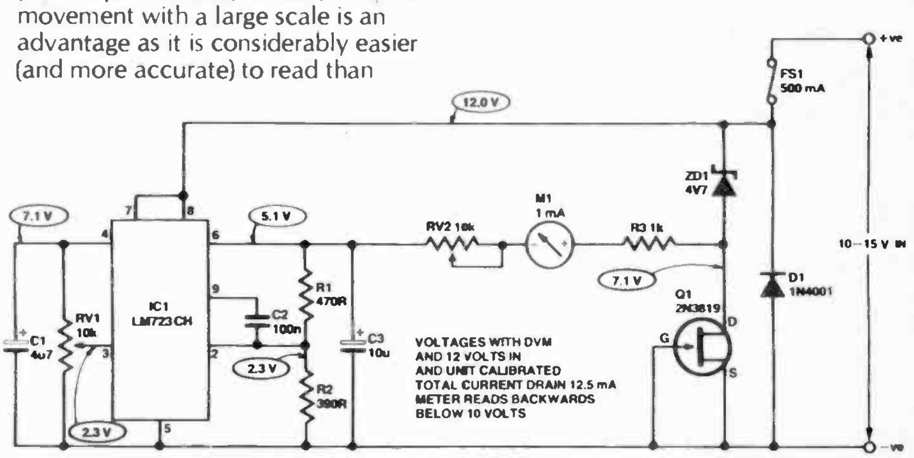


Fig. 1 Circuit diagram for the Voltage Monitor.

Having chosen your meter, drill out the PCB to suit the meter terminal spacing first. The components may then be assembled to the board in any particular order that suits you. Watch the orientation of the 723, ZD1, the FET and particularly D1. The latter is an 'idiot diode'. That is, if you have a lapse of concentration or forethought and connect your project backwards across a battery, the fuse will blow and not the project. Fuses are generally found to be cheaper than this project!

Seat all the components right down on the PCB as the board may be positioned on the rear of the meter with the components facing the meter. The size of C2 may give you a little trouble. Polyesters are generally too large and therefore unsuitable. We used a ceramic type capacitor — as commonly used on computer PCBs as bypasses. Alternatively, a 100n tantalum capacitor (+ve to pin 2 of IC1) may be used. The actual value or type of capacitor is not all that critical.

We have used multiturn trimpots for RV1 and RV2 as they make the setting up a whole lot easier.

Calibration

For this you will need a variable power supply covering 10 to 15 V and a digital multimeter (borrow one for the occasion).

First set the 10 V point. Connect the digital multimeter across the power supply output and adjust the power supply to obtain 10.00 V. Set the mechanical zero on the meter movement to zero the meter's pointer. Connect the unit to the power supply output and adjust RV1 to zero the meter needle.

Next, set the power supply to obtain 15.00 V. Now adjust RV2 so that the meter needle sits on 15 V (full scale). Check the meter reading with the power supply output set at various voltages across the range. We were able to obtain readings across the full scale within ± half a scale reading (±50 mV). With a 2% FSD accuracy meter the worst error may be about ± one scale division.

BUYLINES

Only one thing to comment on here; when you purchase your LM723 (or uA723 — same thing) make sure you get the version that comes in a T099 case, not the DIL version. The PCB is designed for the 10 pin version as shown in the overlay and the DIL type won't fit. Speaking of PCBs, as usual you can get it from us using the order form on page 91.

PARTS LIST

Resistors (all ¼ W, 5% metal film)		**Semiconductors**	
R1	470R	IC1	LM723 (see Buylines)
R2	390R	Q1	2N3819
R3	1k0	D1	1N4002 or similar
		ZD1	4V7 400 mW or 1 W zener
Potentiometers			
RV1,2	10k cermet multiturn horizontal trimpot	**Miscellaneous**	
		M1	1 mA meter (see text)
		FS1	500 mA fuse and in-line fuse holder
Capacitors		PCB (see Buylines); meter scale to suit meter; red and black cable, etc.	
C1	4u7 10 V tantalum		
C2	100n ceramic		
C3	10u 10 V tantalum		

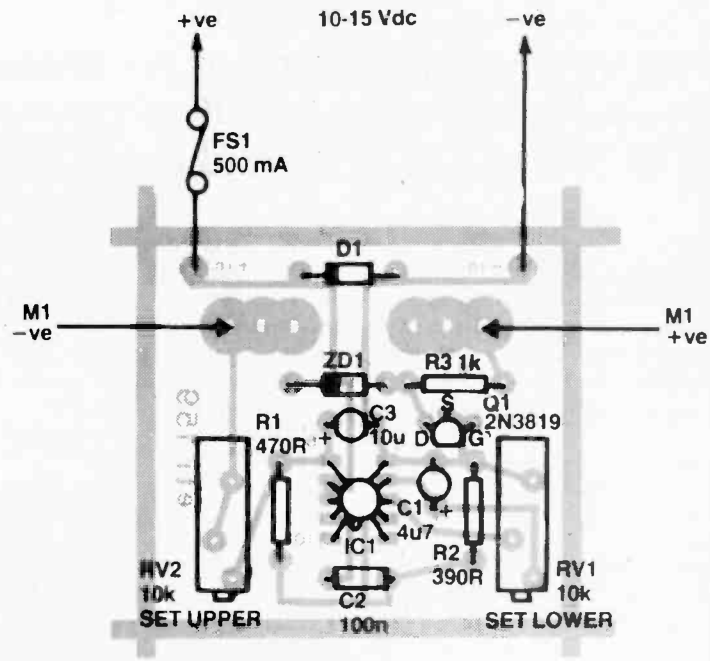


Fig. 2 Component overlay for the Voltage monitor. Note that IC1 is in a 10-pin T099 case.

BATTERY CONDITION AND TERMINAL VOLTAGE

The 12 V battery, in its many forms, is a pretty well universal source of mobile or portable electric power. There are lead-acid wet cell types, lead-acid gel electrolyte (sealed) types, sealed and vented nickel cadmium types, and so on. They are to be found in cars, trucks, tractors, portable lighting plants, receivers, transceivers, aircraft, electric fences and microwave relay stations — to name but a few areas.

No matter what the application, the occasion arises when you need to reliably determine the battery's condition — its state of charge, or discharge. With wet cell lead-acid types, the specific gravity of the electrolyte is one reliable indicator. However, it gets a bit confusing as the recommended electrolyte can have a different S.G. depending on the intended use. For example, a low duty lead-acid battery intended for lighting applications may have a recommended electrolyte S.G. of 1.210, while a heavy-duty truck or tractor battery may have a recommended electrolyte S.G. of 1.275. Car batteries generally have a recommended S.G. of 1.260. That's all very well for common wet cell batteries, but measuring the electrolyte S.G. of sealed lead-acid or nickel-cadmium batteries is out of the question.

With NiCads, the electrolyte doesn't change during charge or discharge.

Fortunately, the terminal voltage is a good indicator of the state of charge or discharge. In general, the terminal voltage of a battery will be at a defined minimum when discharged (generally between 10 and 11 V), and rise to a defined maximum when fully charged (generally around 15 V). Under load, the terminal voltage will vary between these limits, depending on the battery's condition.

Hence a voltmeter having a scale 'spread' to read between these two extremes is a very good and useful indicator of battery condition. It's a lot less messy and more convenient than wielding a hydrometer to measure specific gravity of the electrolyte!

NEXT MONTH

AN ARGUS SPECIALIST PUBLICATION

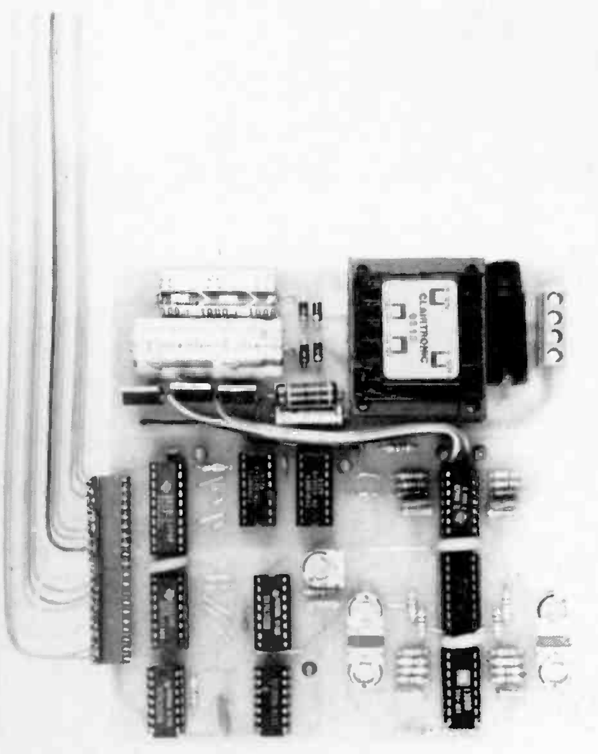
electronics today INTERNATIONAL

AUDIO DESIGN

This month, audio design reaches the seat of the mystique surrounding the whole topic of audio amplifiers — the design of the power section. In true ETI fashion, we intend to boldly separate fact from fiction, truth from fantasy, and infinitives from their verbs!

VECTORGRAPHICS

There is an alternative to using a raster-scanned display — that of using vectors drawn across the screen directly. Phil Walker has produced a system for experimenters to play with, using a ZX81.



EPROM PROGRAMMING

We were so overwhelmed with requests for photocopies of the assembler listing for the EPROM programmer published in our August issue that we asked Mike Bedford to produce some notes on how to interconvert the program for other home computers. We'll be published this, along with the listing for the benefit of the very few of our readers who have not already requested it from us!

CHORUS/FLANGER

This project is designed by Tim Orr — who has absented himself from our pages for a while, but has recently learned better. It's for a chorus/flanger unit that can be used with guitar, bass guitar, high-output microphone, keyboards, or, in fact, almost anything that you can attach to a jack-plug.

LOOK OUT FOR THE JANUARY ISSUE
ON SALE DECEMBER 2nd

Articles described here are in an advanced state of preparation. However, circumstances may dictate changes to the final contents.

Electronics Digest PCB Service

The best way to get a really professional finish to your project is to use a professionally made PCB — look no further, here they are!
In order to ensure that you get the correct board, you must quote the reference number when ordering — if you don't, you may get sent the wrong board.

Please note that only boards listed below are available.

- ☐ Watchdog Home Security System (2 boards) (E818-5) £5.31
- ☐ Allez Cat Pest Repeller (E822-2) £1.68
- ☐ Enlarger Timer (E8110-1) £3.40
- ☐ Phone Bell Shifter (E8111-3) £2.96
- ☐ Sound Track (E828-5) £4.25
- ☐ Car Alarm (E8111-1) £2.81
- ☐ Ion Generator (3 boards) (E826-1) £8.00
- ☐ Bodywork Checker (E8112-3) £1.75
- ☐ Parking Meter Timer (E821-1) £2.20
- ☐ Hotwire (E827-4) £2.63
- ☐ Sound Bender (E8110-2) £2.65
- ☐ MOSFET Amplifier Module (E826-4) £6.78
- ☐ Bridging Adaptor (E827-5) £2.38
- ☐ Guitar Tuner (2 boards) (E821-3) £5.55
- ☐ Playmate (3 boards) (E828-1) £7.20
- ☐ Moving Magnet Stage (E822-5) £3.49
- ☐ Moving Coil Stage (E822-6) £3.49
- ☐ Ripple Monitor (E822-1) £1.92
- ☐ Capacitance Meter (2 boards) (E823-4) £10.14
- ☐ DV Meg (E825-1) £2.72
- ☐ Laboratory PSU (E819-4) £4.53
- ☐ TV Bargraph Main (E827-7) £4.56
- ☐ TV Bargraph Channel (E827-3) £2.28
- ☐ Component Tester (E8112-4) £1.49
- ☐ Voltage Monitor (E824-5) £1.86

How to order: indicate the boards required by ticking the appropriate boxes and send this page, together with your payment, to: Electronics Digest PCB Service, Argus Specialist Publications Ltd, 145 Charing Cross Road, London WC2H 0EE.

Cheques should be made payable to ASP Ltd and should be in sterling only.

Please allow 28 days for delivery.

Total cost of PCBs £..........
p&p 45p
Total enclosed £..........

Name ....

Address ....

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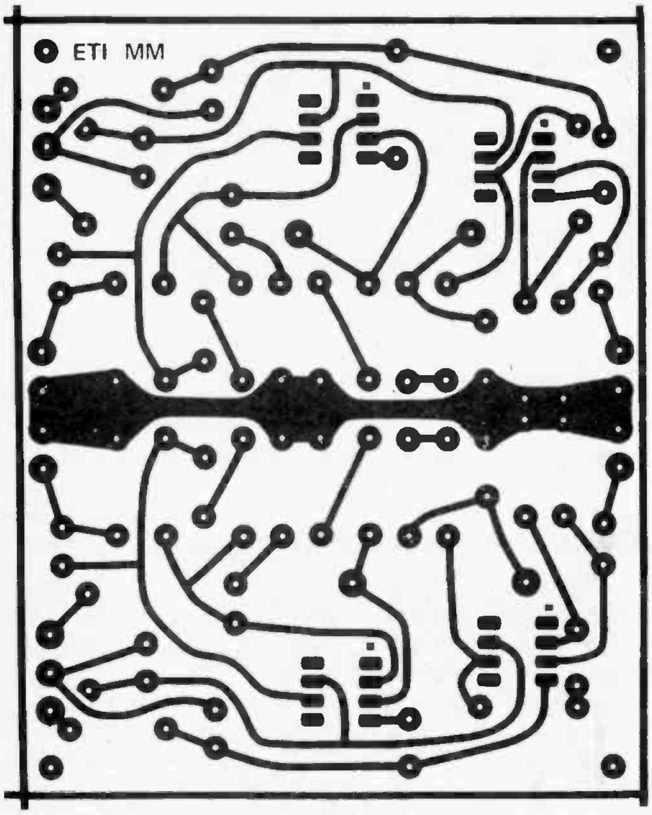
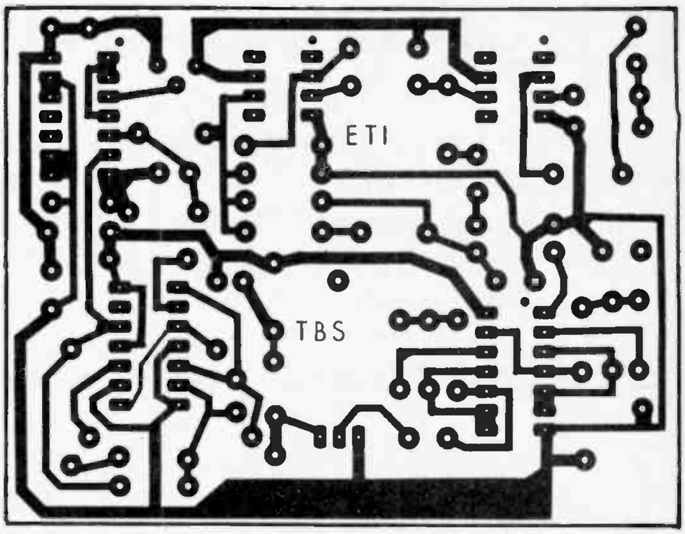
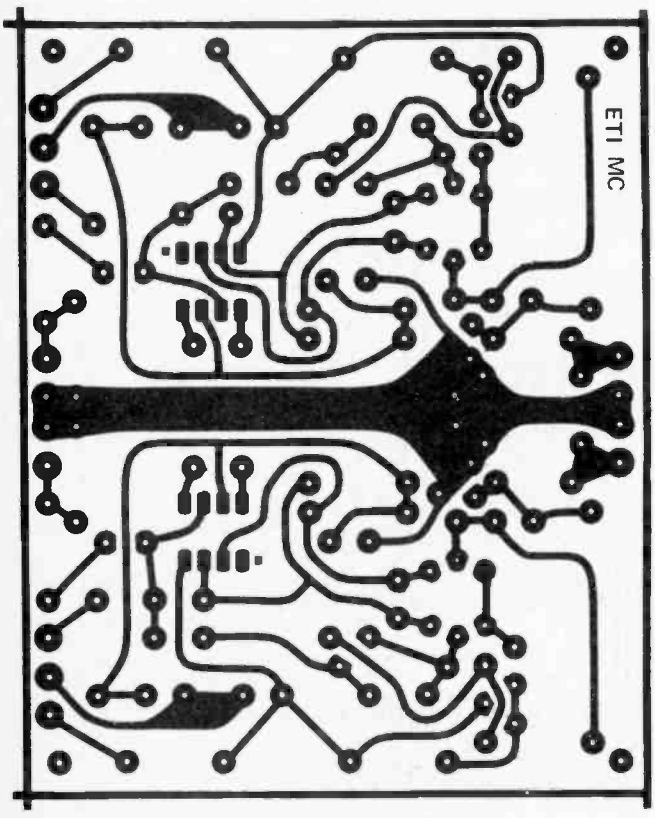
PCB FOIL PATTERNS



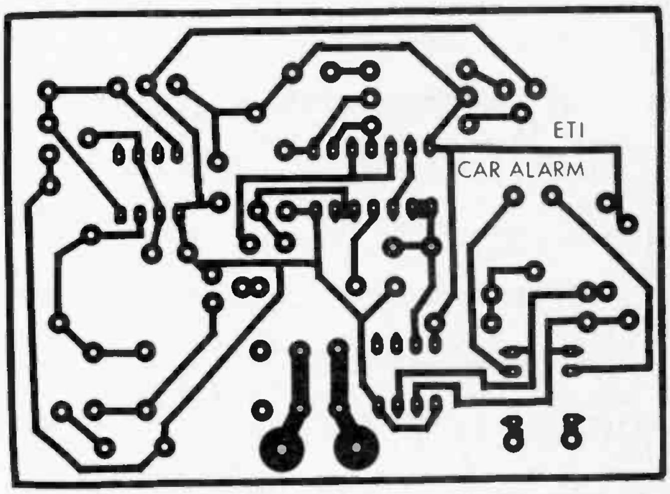
The moving magnet PCB of Phono Amplifier.



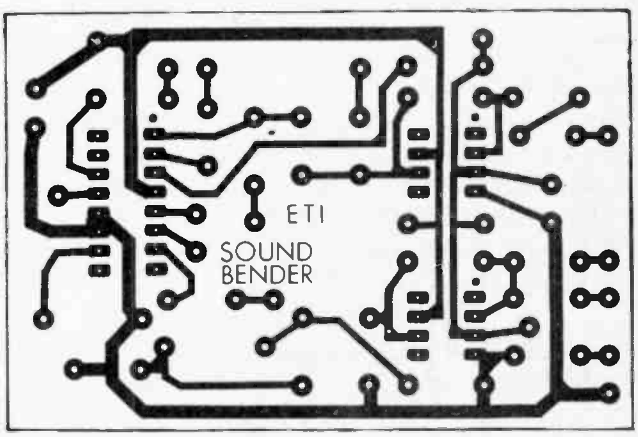
The PCB for the Phone Bell Shifter.



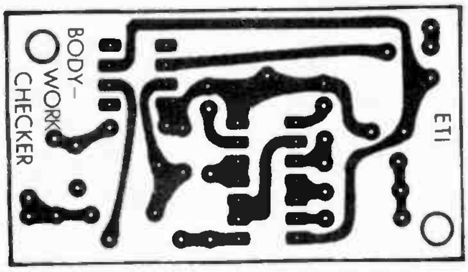
The moving coil PCB of Phono Amplifier.



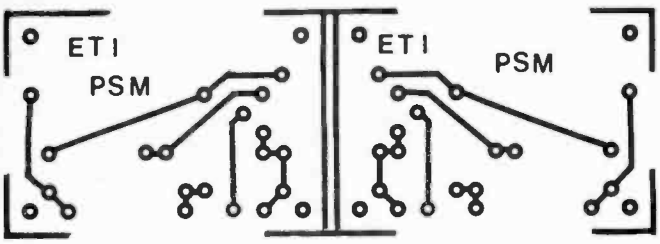
The foil pattern for the Car Alarm.



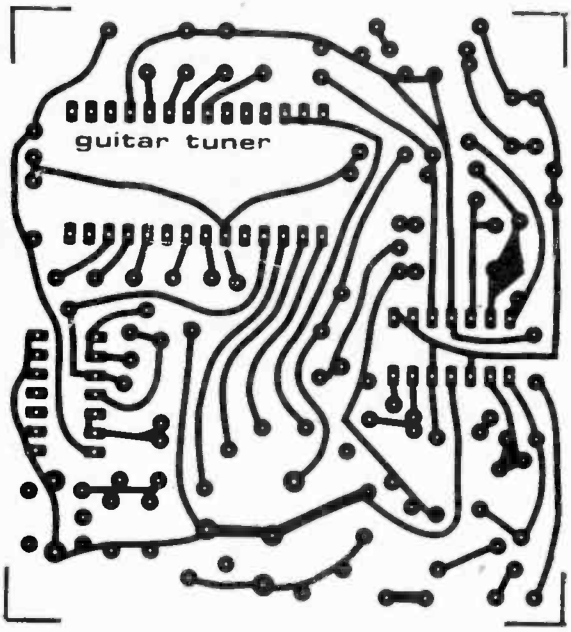
The Sound Bender PCB.



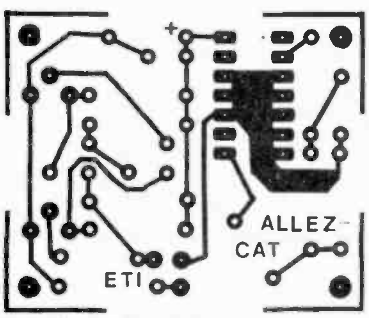
Bobywork Checker foil.



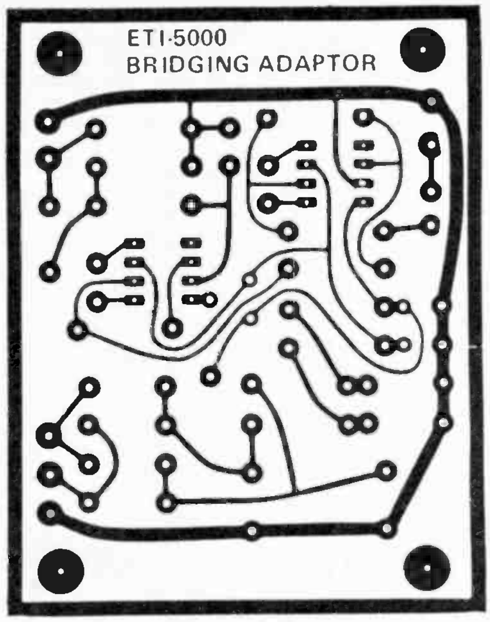
Upgarding Amplifier PSUs PCB.



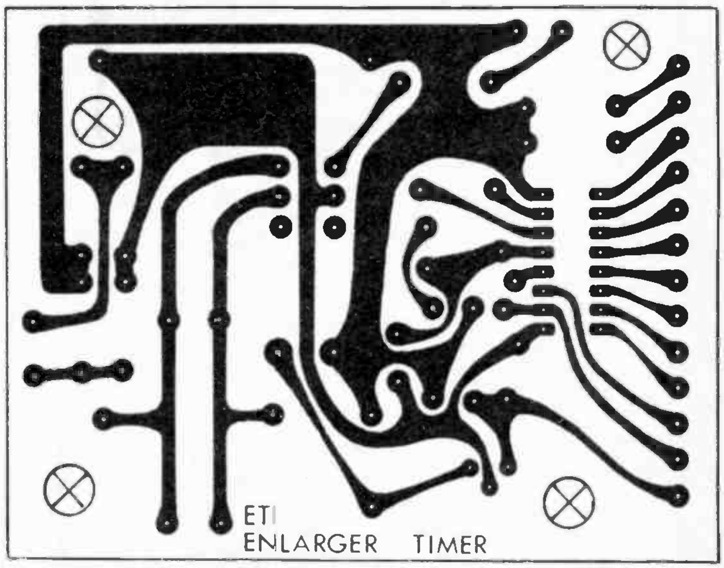
The foil for Guitar Tuner.



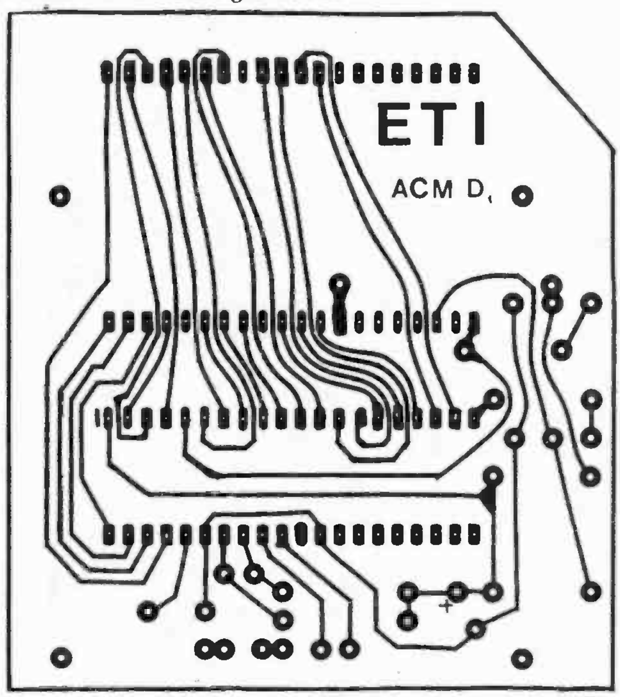
Pest Controller foil.



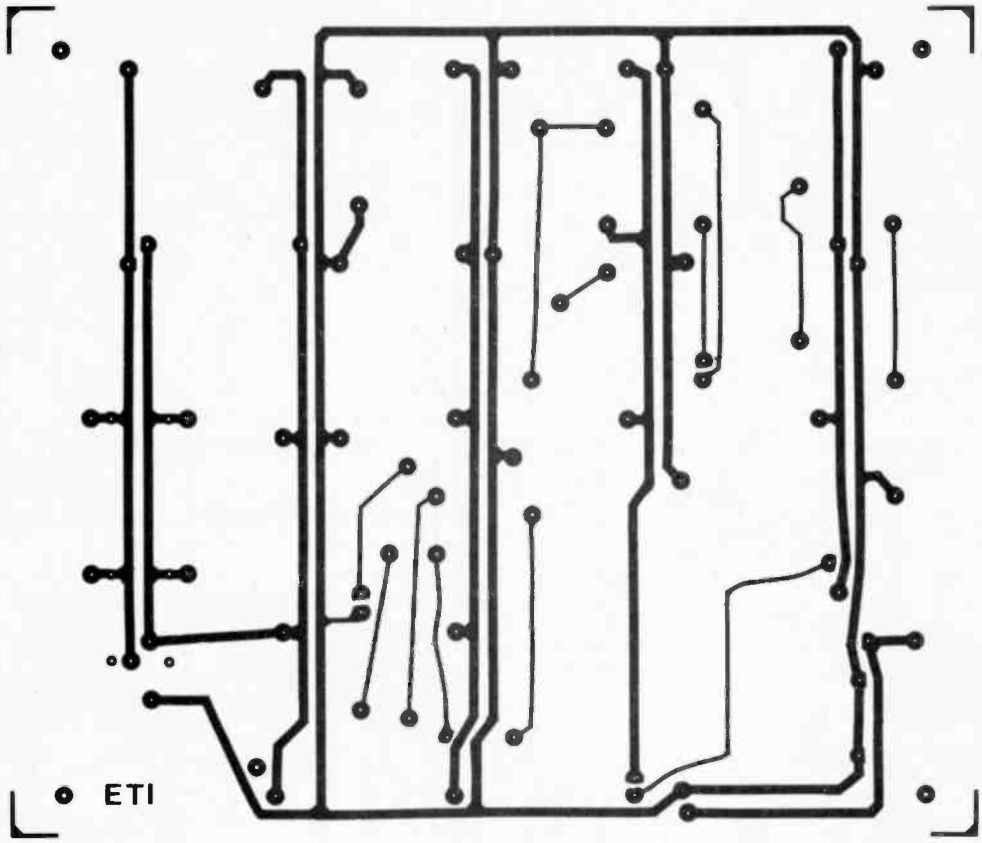
The Bridging Adaptor foil.



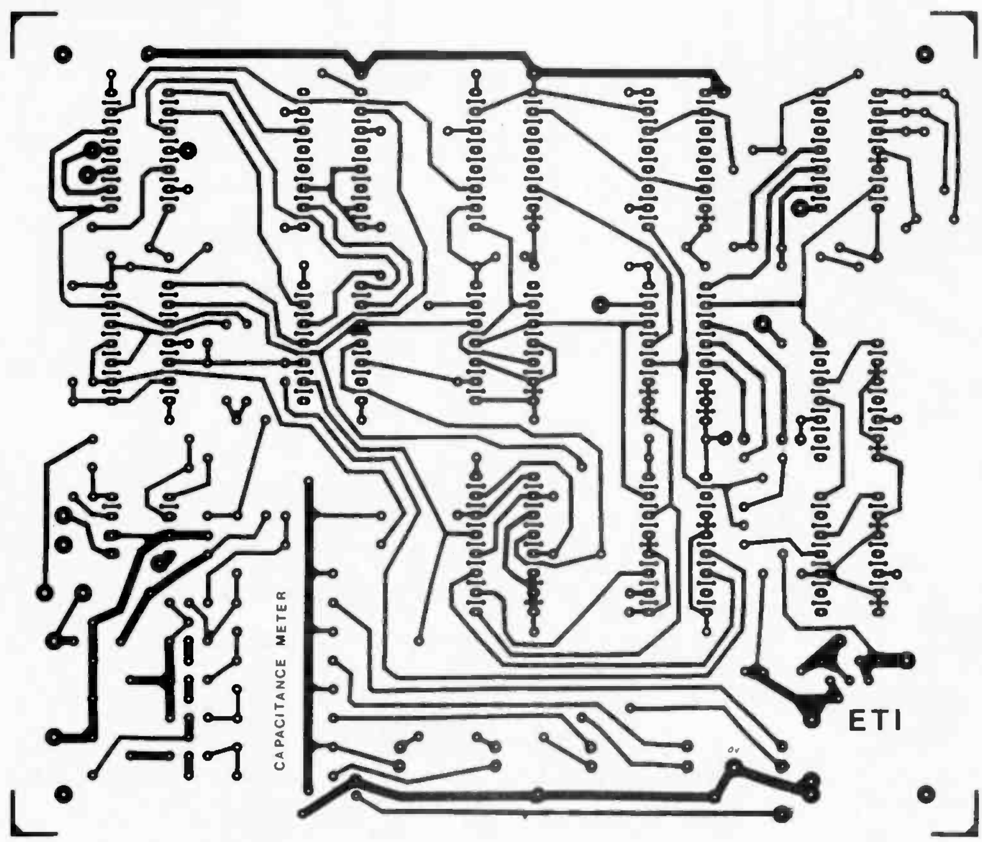
The PCB for the Enlarger Timer.

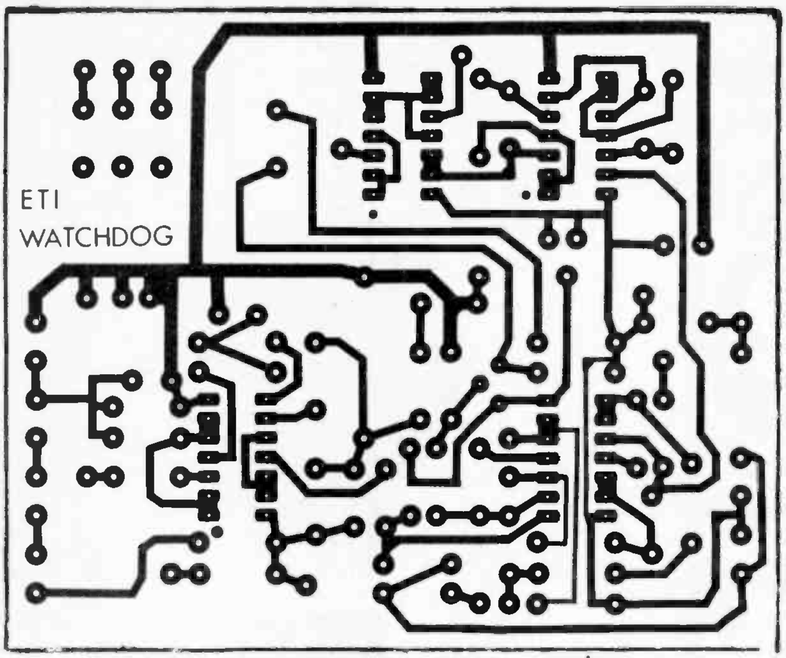


The Autoranging Capacitance Meter display board.

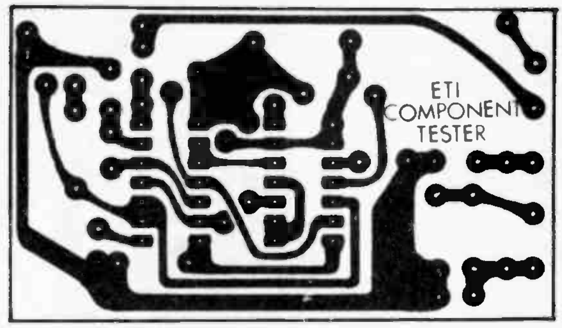


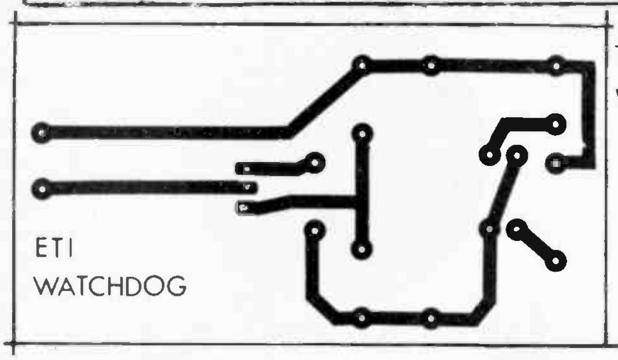
Above: the foil pattern for the top side of the Auto-ranging Capacitance Meter main board, which is double-sided.
Below: the foil for the bottom side of the board.



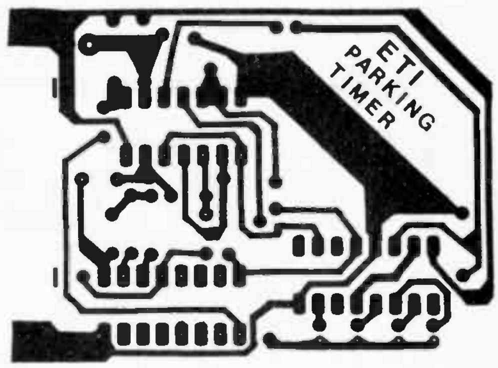


The Compact Tester PCB.

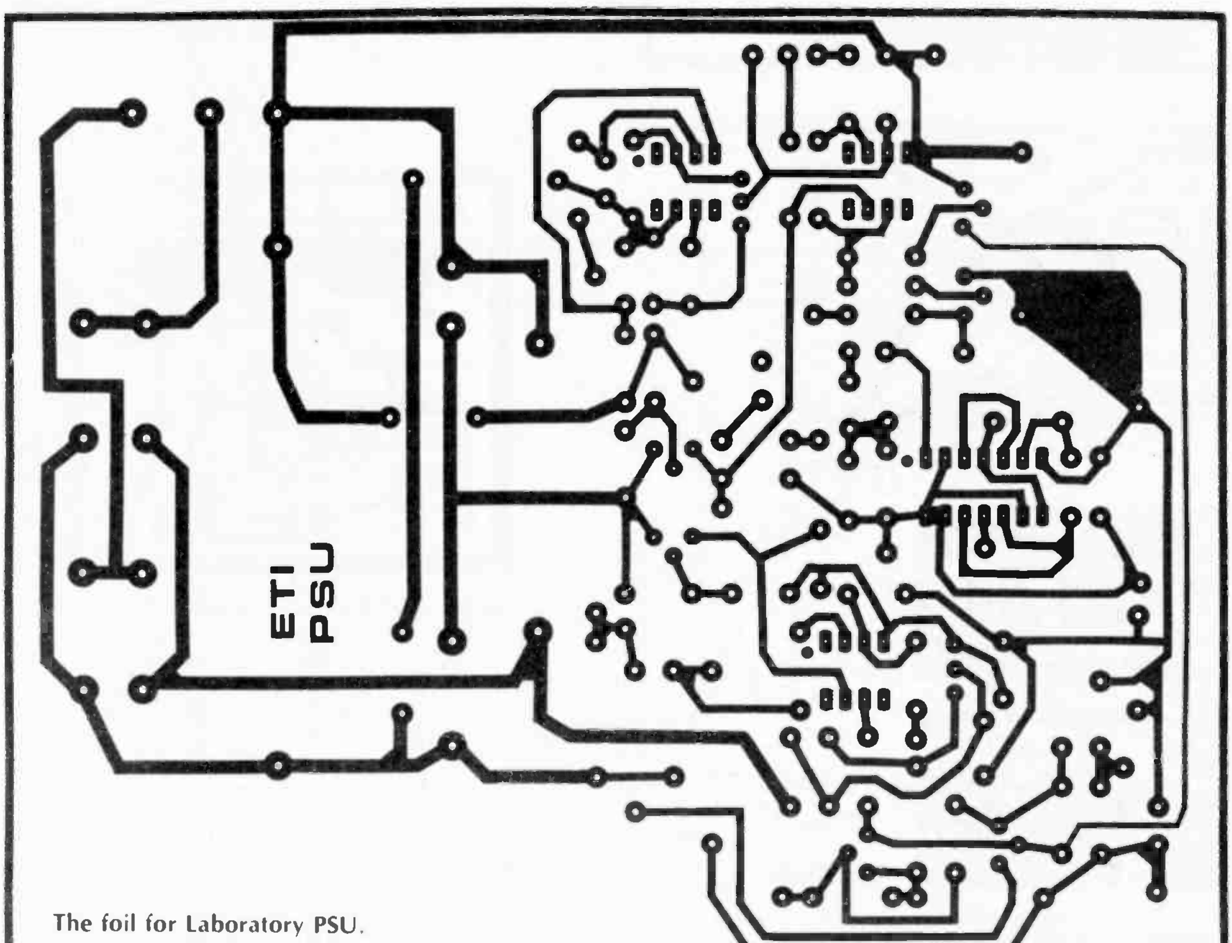




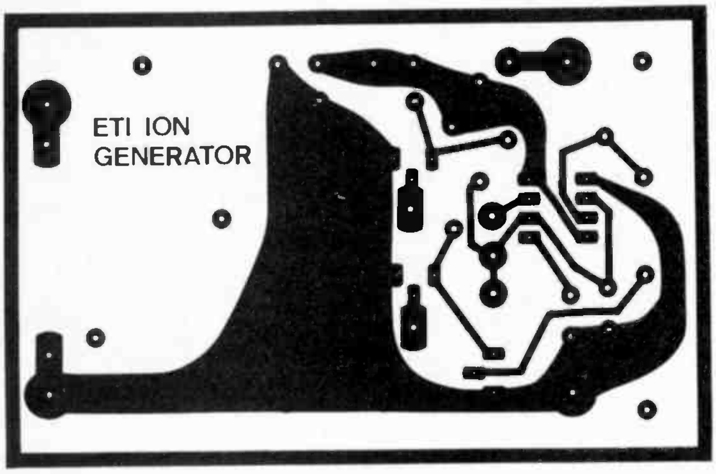
The foil for Watchdog.

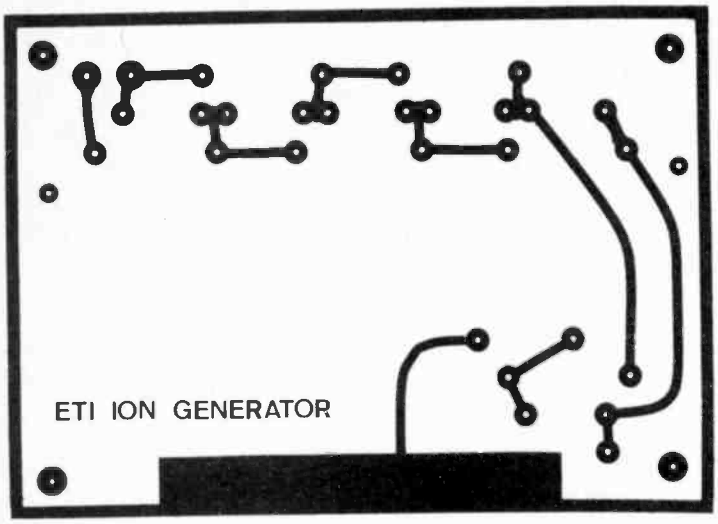


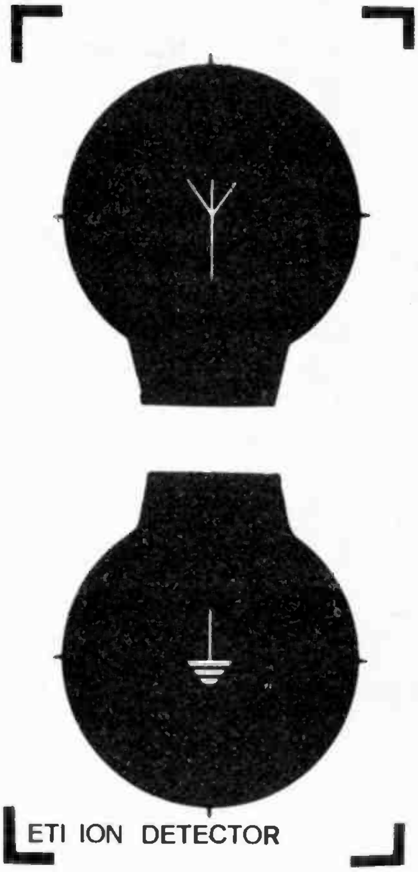
The Parking Meter Timer PCB.



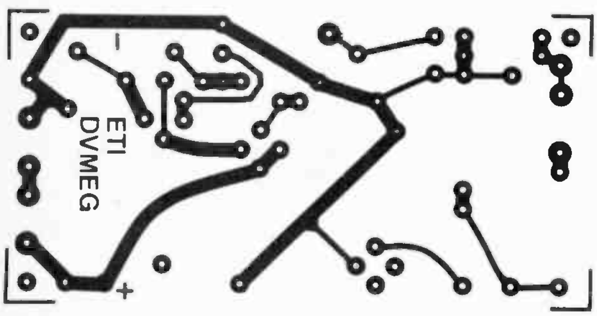
The foil for Laboratory PSU.



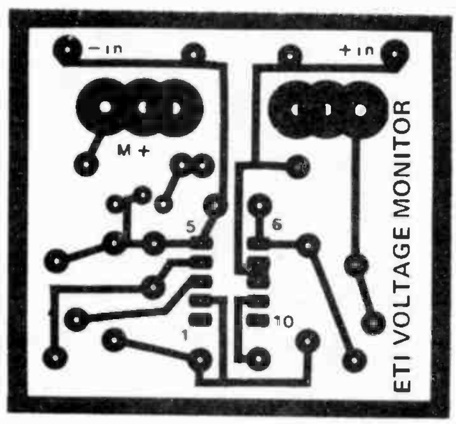




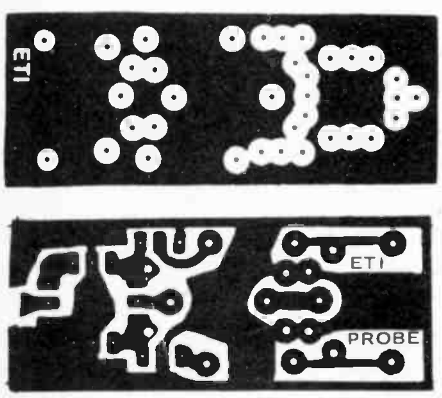
Board for the Negative Ion Generator.



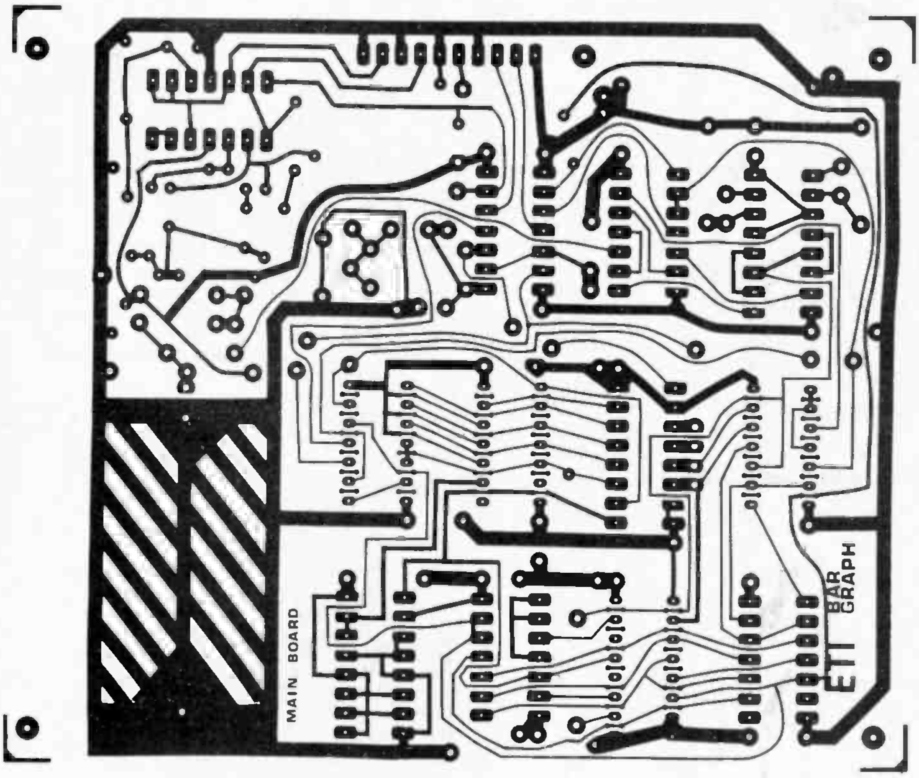
The Insulation Tester PCB.

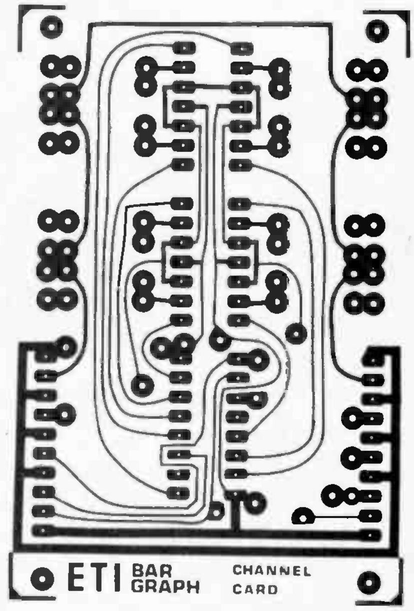


Above: the PCB for the Voltage Monitor.

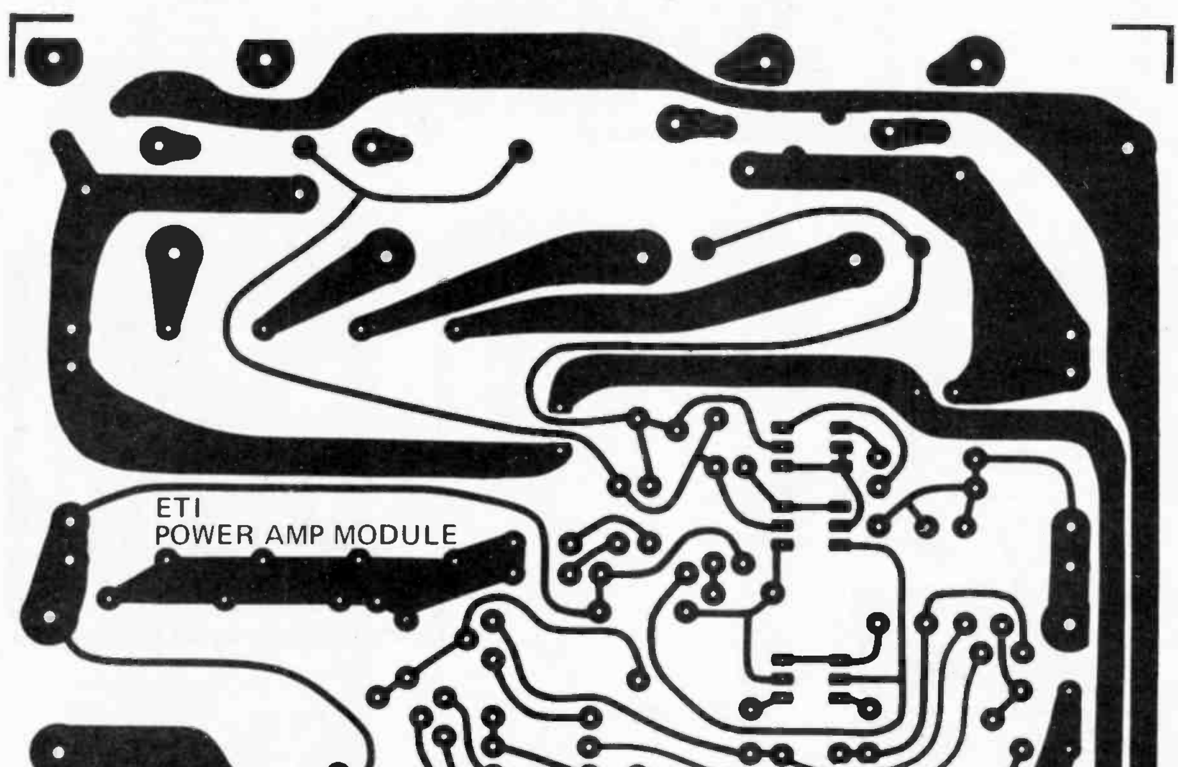


Above: the two foils for the double-sided 100 MHz High Impedance Probe PCB.



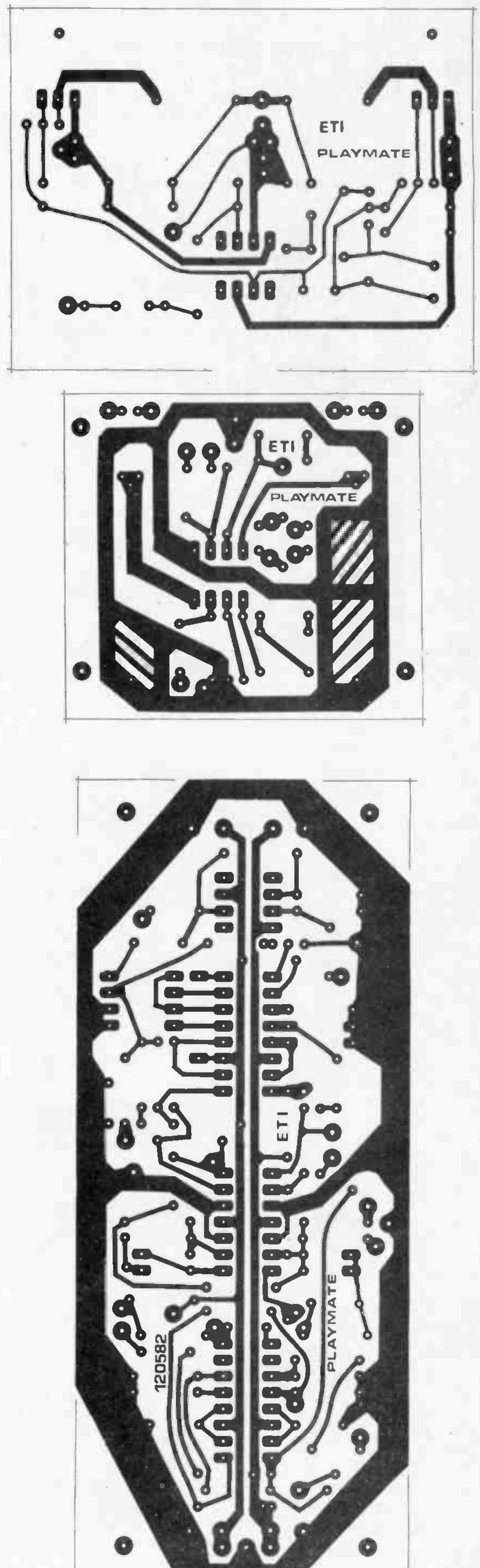


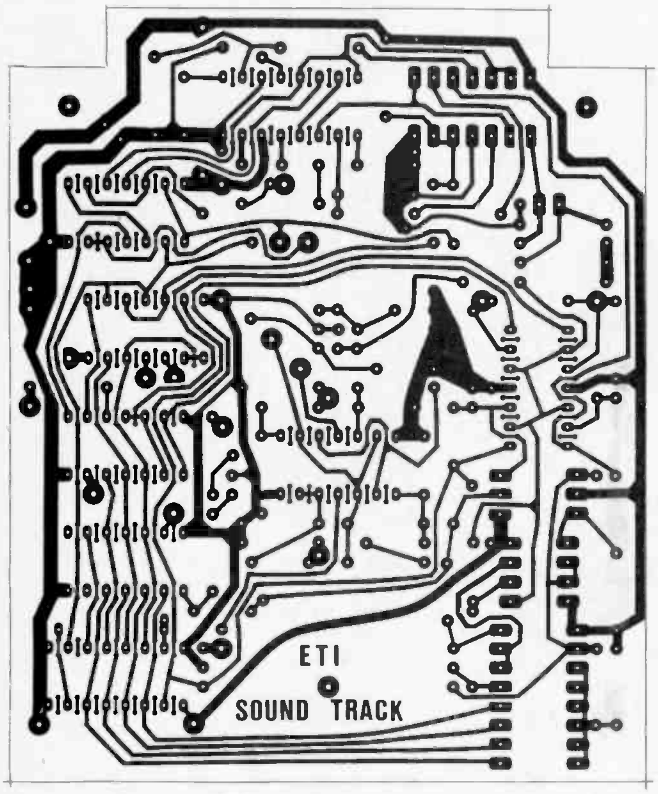
Above: The TV Bargraph Channel card.
Left: The main board of TV Bargraph.



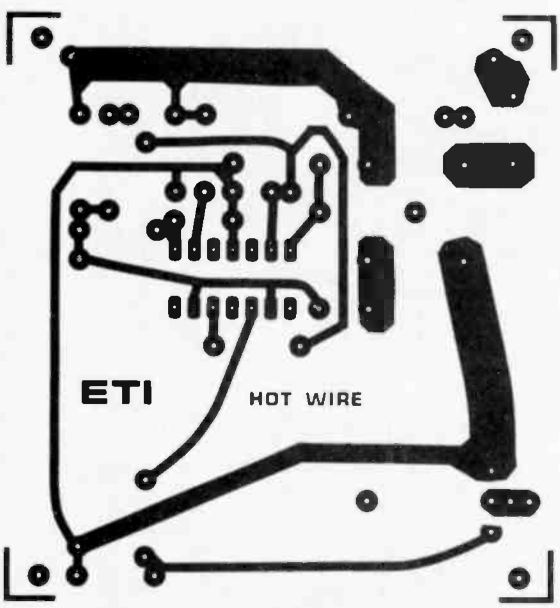
PCB Foil Patterns

Foils for the Playmate.





The Sound Track PCB.



The Polystyrene Cutter foil pattern.

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