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24 Page Supplement Inside

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tial change in the collector current of Q2. Q1 and Q2 are in thermal contact and so any temperature change will effect both equally. Thus the −1.9 mV/°C factor is cancelled out by Q1 acting as a compensating thermometer for Q2. The slope change is removed by using a temperature sensitive resistance (Q81 — Tel Labs) which has an equal but opposite temperature coefficient to the diode junction. This resistor is often in thermal contact with the matched transistors. If this circuit is connected to a linear current controlled oscillator, a musical VCO is produced.

VCO Circuits

Figure 12 is the circuit for an exponential VCO using an exponential current source. The oscillator is a standard triangle-square wave device. IC2 is a current-controlled integrator; the slow rate at its output is equal to I_{ABC}/C. This voltage is buffered by IC3 which drives a Schmitt trigger IC4. The output of IC2 ramps up and down between the two hysteresis levels which are determined by the two clamping diodes connected to the output of IC4. Any stray capacitance on the output of IC4 will slow down the Schmitt trigger and this will make the VCO go flat at high frequencies. Also the propagation time delay around the oscillator will cause a flattening out ot the response at high frequencies. These effects can be nulled out but they may not even affect things if the VCO frequency is kept relatively low.

A very good VCO is shown in Fig. 13. It is a monolithic device, the CEM3340 from Curtis Electromusic Specialities Inc who make a range of electronic music devices. As can be seen, very few external parts are needed to implement the VCO. All the temperature compensation is performed inside the chip. Triangle, sawtooth and variable mark/space square wave outputs are simultaneously available. The mark/space ratio is a voltage controlled parameter. A sync input is also provided so that the VCO can be slaved to another oscillator.

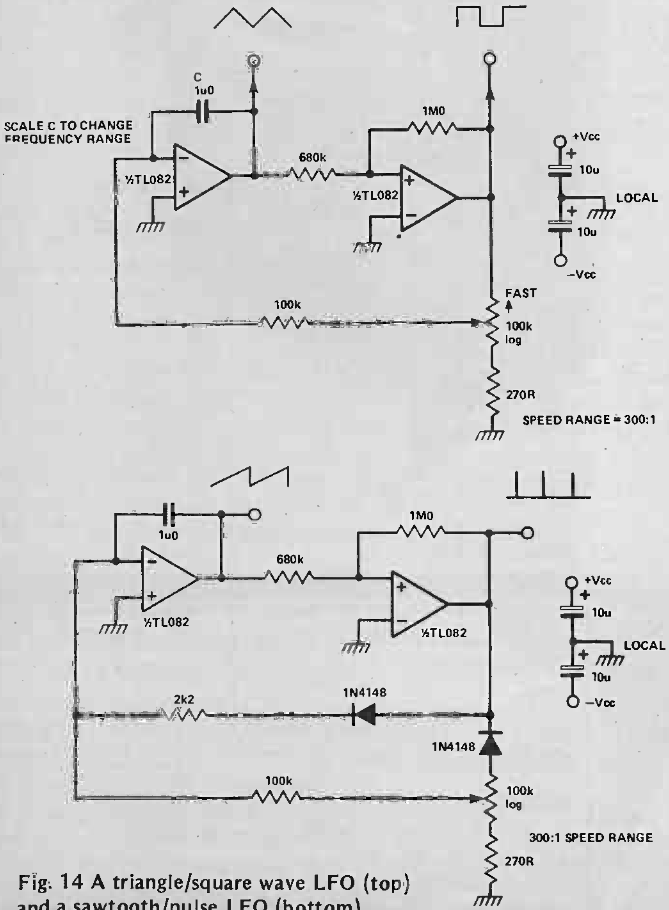


Fig. 14 A triangle/square wave LFO (top) and a sawtooth/pulse LFO (bottom).

LFO Circuits

A couple of LFO units are shown in Fig. 14. All four output waveforms can be usefully employed to sweep VCOs and VCFs. Often the waveforms are mixed together to produce strange frequency modulations. When the sawtooth is fed into one side of a ring modulator and noise into the other, a beat track can be generated; it sounds a bit like a cymbal being hit.

Noise Generators

In 'the old days' noise sources were made by amplifying the noise current of a diode junction that was zenering. These were a bit unreliable, and always involved selecting the device. However, noise can be generated digitally with a maximum length pseudorandom sequence generator (Fig. 15). The noise spectrum is relatively flat and always the same. If you slow down the clock rate you can get some interesting sounds; I think that this is used on some TV games. If a longer shift register is used, say 30 or 40 stages (the 4006 is 18 stages long), and the noise source is turned on, a tone is initially heard which gradually changes into noise as the sequence becomes more scrambled up. You can purchase a monolithic noise generator (pseudorandom); it is the MM5837 made by National Semiconductor, also sold by AMI with the part number S2688.

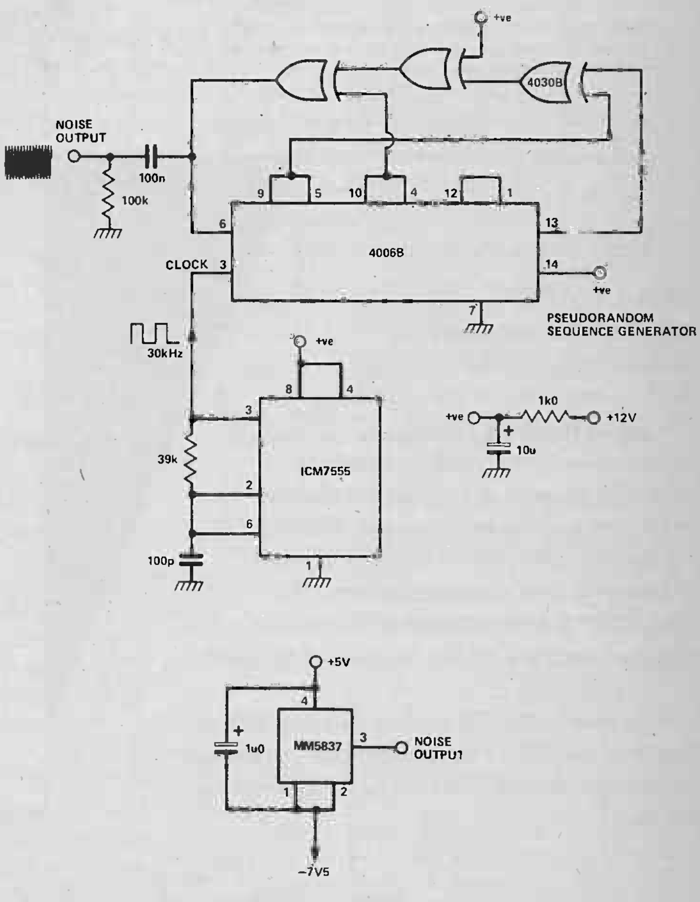
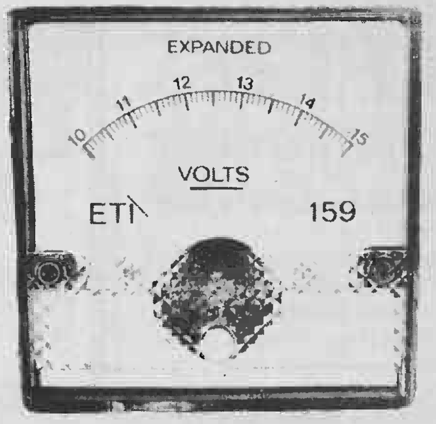


Fig. 15 A digital noise source (top) and a noise generator chip (bottom).

Five pages gone already, and we've still only scratched the surface of this fascinating subject. In part two next month, Tim Orr will continue his discussion of electro-music techniques with yet more circuit building blocks.

ETI

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.

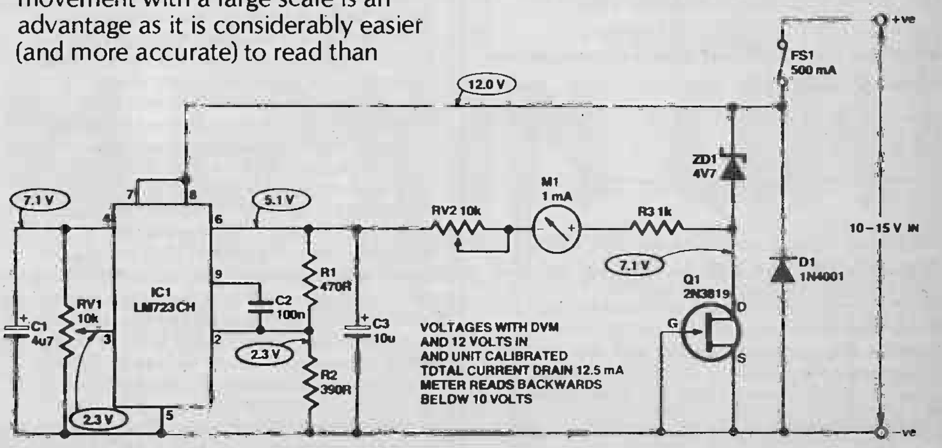


Fig. 1 Circuit diagram for the Voltage Monitor.

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 44.

PARTS LIST

Resistors (all ¼ W, 5% metal film)

R1	470R
R2	390R
R3	1k0

Potentiometers

RV1,2	10k cermet multiturn horizontal trimpot

Capacitors

C1	4u7 10 V tantalum
C2	100n ceramic
C3	10u 10 V tantalum

Semiconductors

IC1	LM723 (see Buylines)
Q1	2N3819
D1	1N4002 or similar
ZD1	4V7 400 mW or 1 W zener

Miscellaneous

M1	1 mA meter (see text)
FS1	500 mA fuse and in-line fuse holder

PCB (see Buylines); meter scale to suit meter; red and black cable, etc.

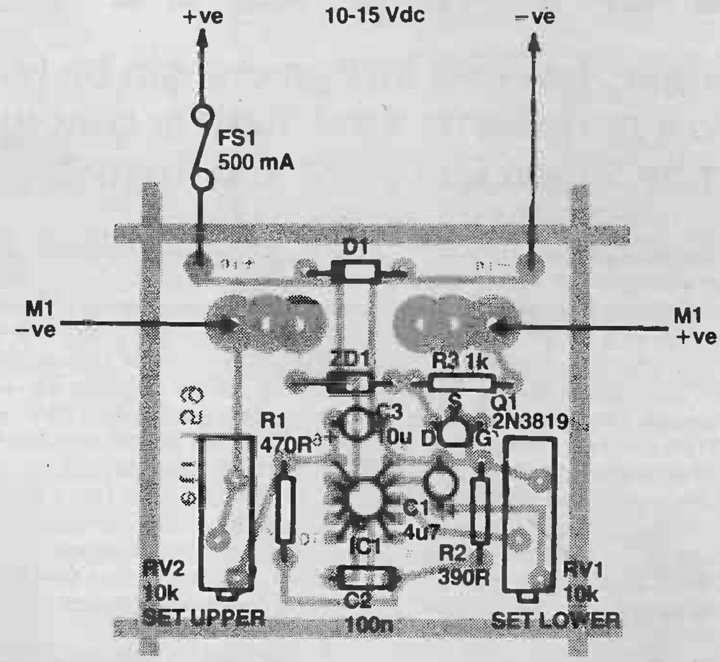


Fig. 2 Component overlay for the Voltage monitor. Note that IC1 is in a 10-pin T099 case.

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!

The charge and discharge characteristics of typical lead-acid and sealed NiCad batteries are given in the accompanying figures.

ETI

COMPUTER EXPANSION SYSTEM

How's your memory? If you're lacking EPROM and the ability to program it, the fourth of our expansion cards is just what you need. Design by Watford Electronics.

This month we present an EPROM programmer and associated EPROM cards suitable for the machine code freak to store away those beloved extra routines or the space invaders freak to capture his aliens in 0's and 1's for life.

The first major consideration when designing an EPROM programmer is just what EPROMs should it be capable of blowing. There is more than just a little confusion here. There are two basic types of EPROM currently available — those that run off a three rail supply and those that run from a single +5 V rail. The two sizes of PROM most popular at the moment are 2K x 8 and 4K x 8. Aha! here manufacturers have had some fun. Intersil and others like calling their triple rail PROMs 2716 and 2732 whereas Intel make their 2716 and 2732 single rail; not to be missed out Texas try to settle the balance by nominating their EPROMs 2516 and 2532; both are single rail!

To clear up the matter our programmer will program single rail EPROMs only, these being the most popular. It will program the Texas 2516 2K x 8 EPROM and Intel 2716 2K x 8 EPROM as these are pin-for-pin compatible (see Fig. 1). However, 2532 and 2732 4K x 8 EPROMs are not compatible and we have stuck to the 2532, as this then allows for use of the new 2764 8K x 8 EPROMs with the minimum alteration (see Fig. 2). If you wish to program 2764's then you must make the alterations to correct the $\overline{OE}$/V_{PP} and $\overline{CS}$ lines. A12 has been brought to pin 1 and power (V_{CC}) to pin 28.

Selection of the type of EPROM you want to program is made by means of a quad DIL switch. This switch is unusual in that each section operates two oppositely biased single pole switches — this means it can be

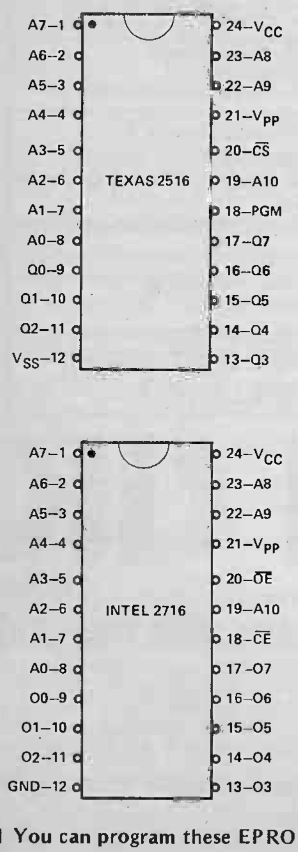


Fig. 1 You can program these EPROMs...

HOW IT WORKS

PROM PROGRAMMER

The heart of this board is two 6520 peripheral input-output chips — they serve to generate the address bus, the data and control signals for the chip being programmed.

R1 and C1 generate the power up reset; C4, 5 and 6 are included in for decoupling. The rather peculiar need of the V_{PP} pin for 0, +5 V and +25 V is met by the PSU and switching circuit. Transformer T1 supplies 30 V AC to the bridge which rectifies it and feeds it to smoothing capacitor C3. IC3 and ZD1 regulate this to +25 V DC. C2 is included in the interests of stability. Transistors Q1 and Q2 handle the switching of V_{PP} between 0, 5 and 25 V. This output is then fed to the DIL switch and then to the V_{PP} pin of the EPROM to be programmed. Ports A and B of IC2 are used to generate the address bus — note A12 is connected to pin 1 of the EPROM (on a 28 pin basis) for use later with 2764 EPROMS. The data bus is generated by port A of IC1, while port B of IC1 generates the control for V_{PP} and the $\overline{CS}$ and PGM lines which are switched with A11 to the correct pins of the EPROM by the DIL switch.

Inputs to the 6520s are straight from the expansion sockets — $\phi 2$ being used to enable the chips to reduce power consumption.

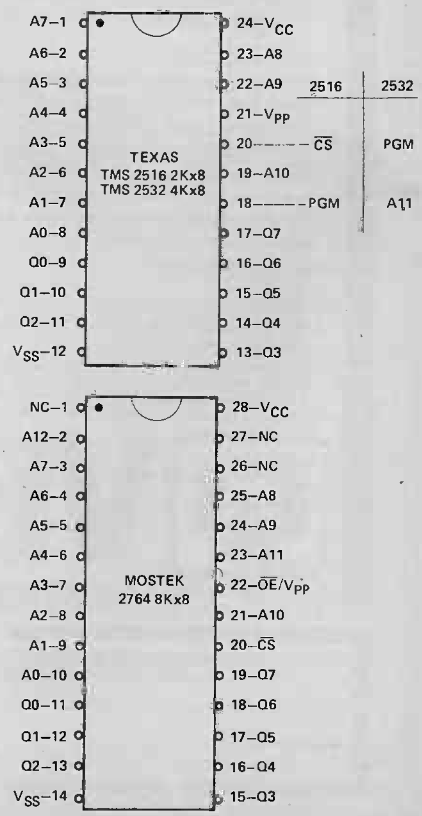


Fig. 2 ...or these ones.

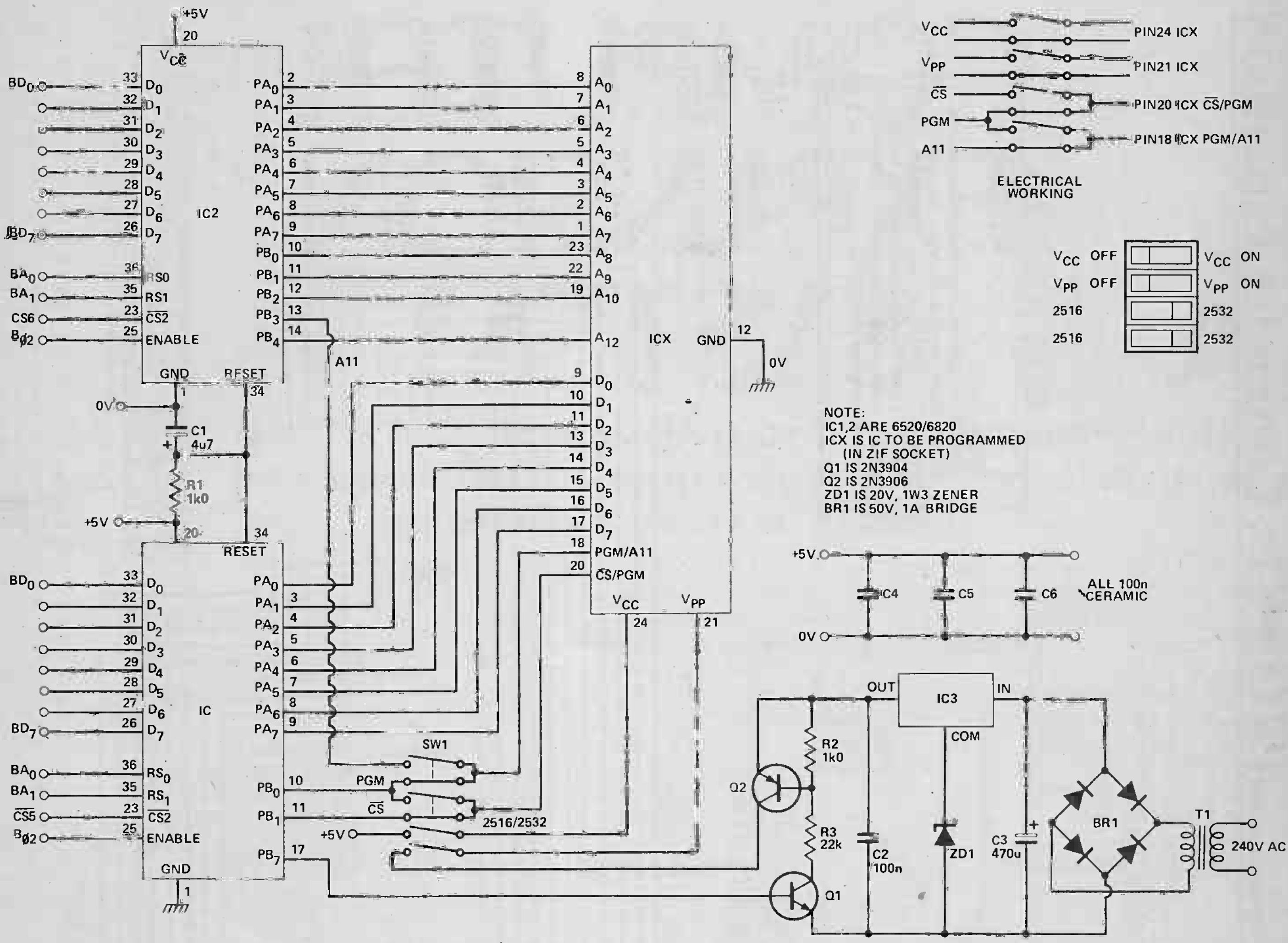


Fig. 3 Circuit diagram of the EPROM programmer, with details of SW1. ICX is the EPROM to be programmed.

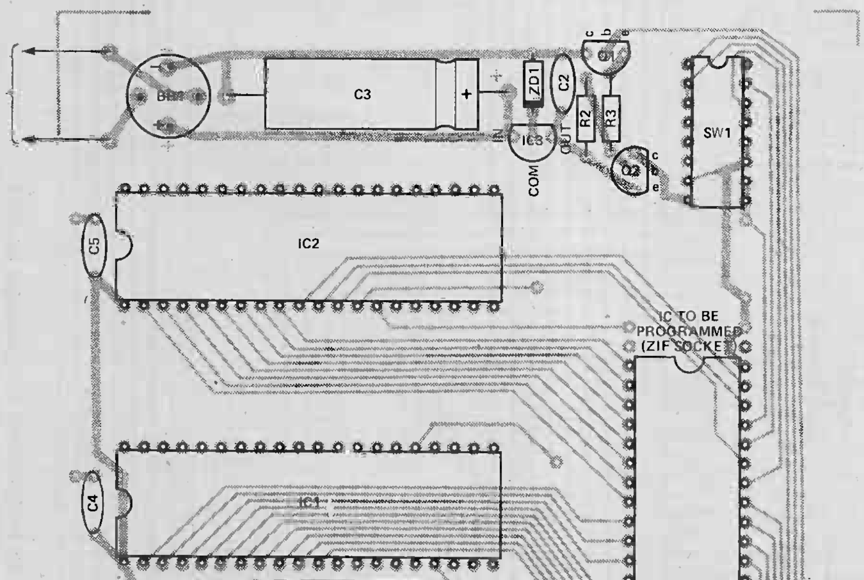


Fig. 4 Overlay for the EPROM programmer. The zero insertion force socket position has extra holes to allow for 2764s.

PARTS LIST

PROM PROGRAMMER	
Resistors (all ¼ W, 5%)	
R1,2	1k0
R3	22k
Capacitors	
C1	4u7 25 V axial electrolytic
C2,4,5,6	100n ceramic
C3	470u axial electrolytic
Semiconductors	
IC1,2	6520/6820
IC3	78L05
Q1	2N3904
Q2	2N3906
ZD1	20 V, 1W3 zener diode
BR1	1 A, 50 V bridge rectifier
Miscellaneous	
SW1	Quad DPST DIL switch
PCB (see Buylines); DIL sockets; transformer (6 VA, 0-15-0-15)	

used as a 4 pole changeover switch and makes it ideal for the job. Two of the four sections are used for chip power (+5 V) and the programming can be destroyed if V_{pp} is applied with V_{CC} disconnected. The other two sections are used to switch $\overline{CS}$, PGM and A11 to the correct pins of the ZIF socket according to whether a 2516 or 2532 is to be used.

PROJECT : Computer Expansion

HOW IT WORKS

PROM CARD
The PROM card is very straightforward as all the hard work is done by the motherboard. ICs 5 and 6 are used to generate A10 and A11 and the $\overline{CS}$ lines. The DIL header plug is used to set jumpers J1-J6 according to requirements (see text).

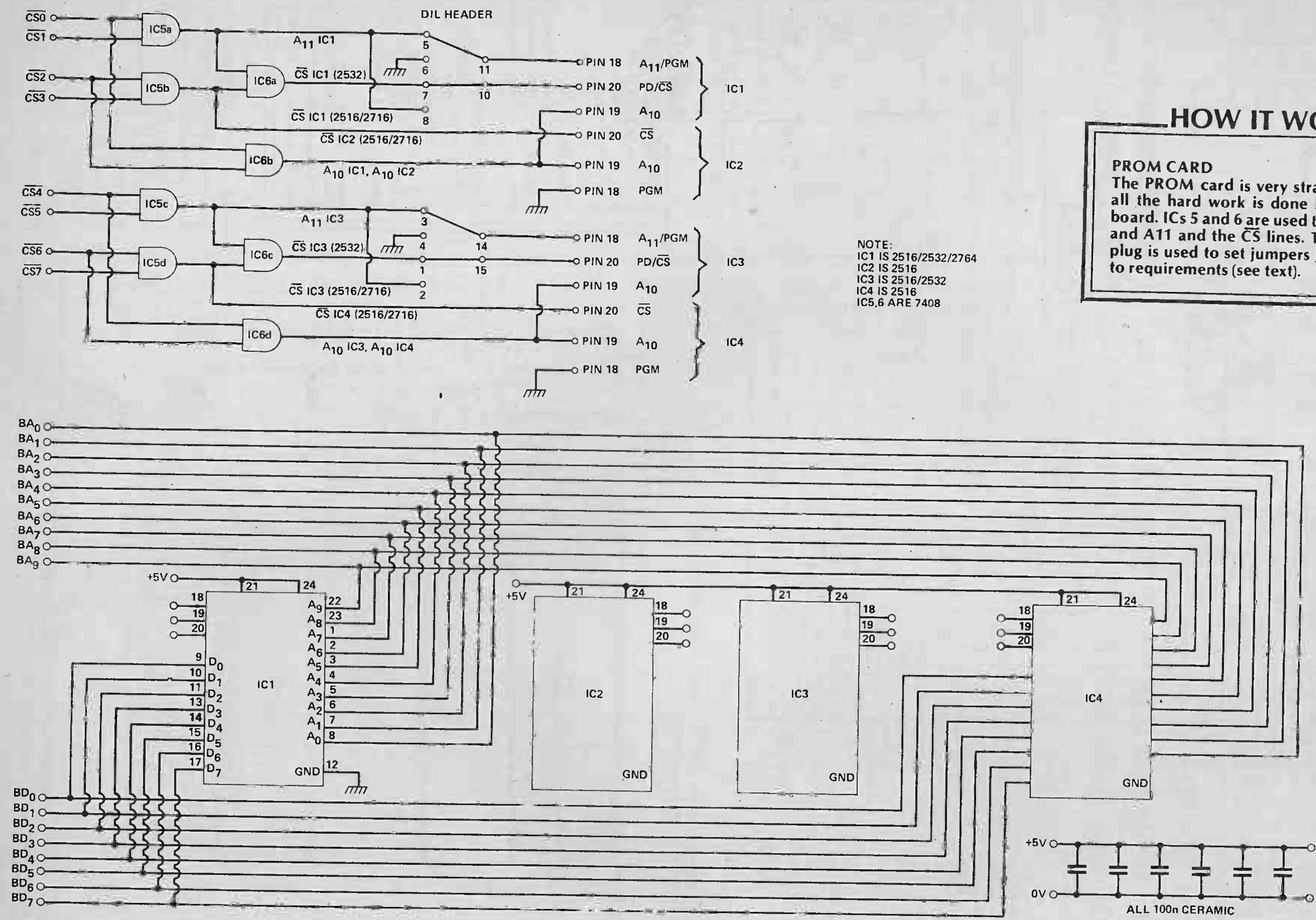


Fig. 5 Circuit diagram for the EPROM card. Links soldered to a DIL header select the correct signals for the various combinations of EPROMs — see Fig. 7 for details.

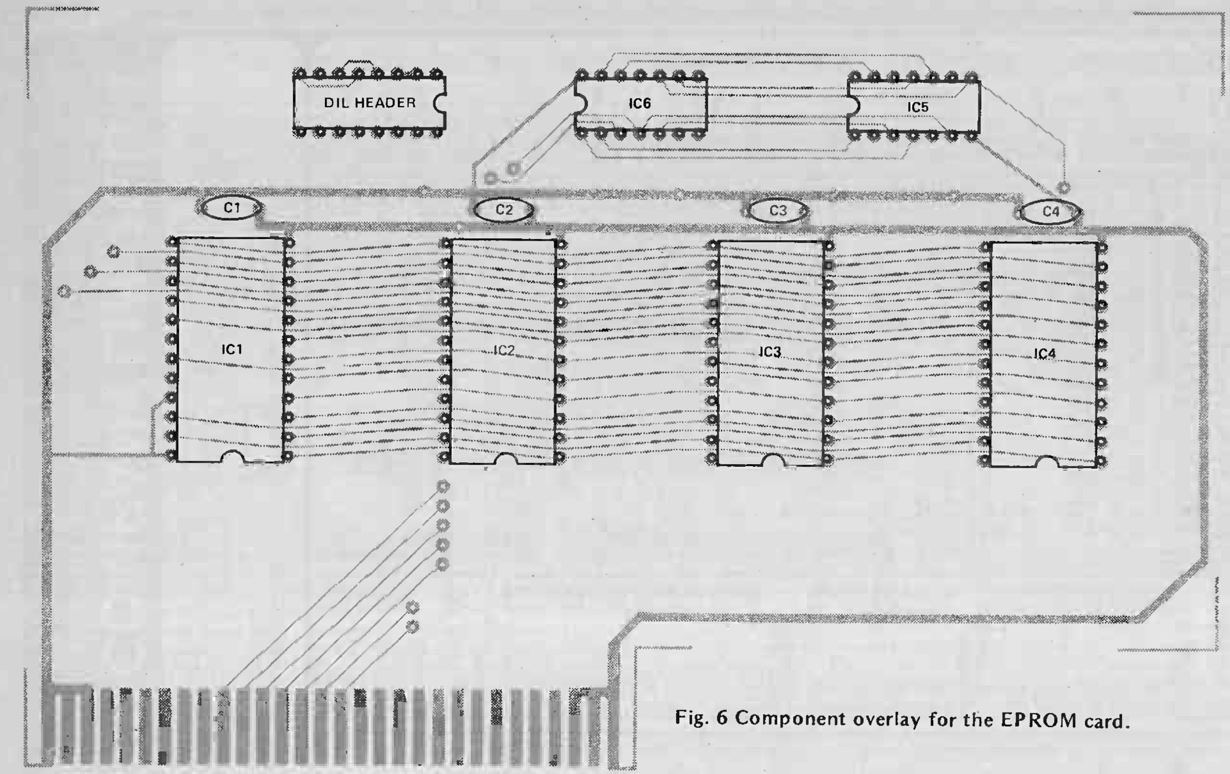


Fig. 6 Component overlay for the EPROM card.

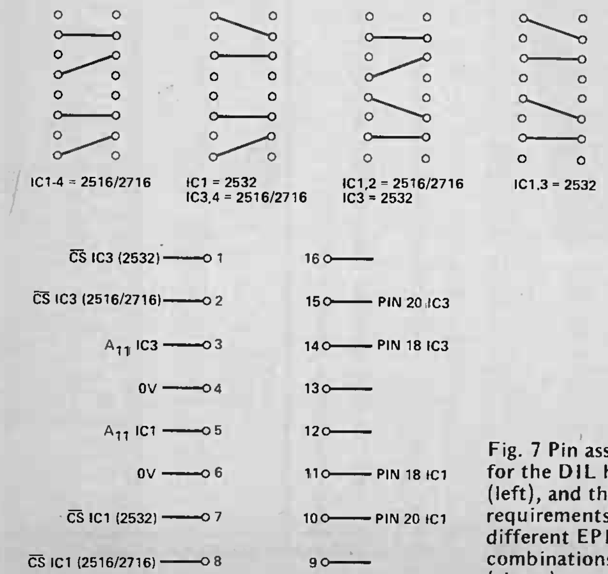


Fig. 7 Pin assignments for the DIL header (left), and the link requirements for the different EPROM combinations (above).

PARTS LIST

PROM CARD

Capacitors

C1-6	100n ceramic

Semiconductors

IC1-4	2716, 2516, 2532 — PROMs as desired
IC5,6	7408

Miscellaneous

PCB (see Buylines); DIL sockets as required; header plug (see text).

BUYLINES

A complete kit of parts for these expansion cards are available from Watford Electronics. The PROM programmer kit costs £26.95 (PCB only £9.75); the PROM card kit costs £11.95 (PCB only £9.75). All prices are subject to VAT at 15%.

A similar method has been used on the EPROM card. As there are four sets of switches needed for four EPROMs a 16 pin header plug and socket have been used. You can make up a header for four 2516s and two 2532s and easily change the role of the board by simply exchanging header plugs. This retains better flexibility than jumper links and is cheaper than the method previously considered. The DIL header plug can be wired as in Fig. 3.

Refer to Fig. 7 for an explanation of how the header plug is wired.

Construction

Construction of the two boards is very straightforward — follow the overlays given here. Note that if you want to move the card around in memory then simply break the connections $\overline{CS5}$, $\overline{CS6}$ to $\overline{CS2}$ of the 6520 and re-make to the $\overline{CS}$ line you desire.

Use two Veropins or similar to bring the 30 V AC from the transformer to the board — unfortunate as it is using a transformer mounted off the PCB to generate the V_{pp} voltage, it is about the only practical way from a computer that has supply rails of 0 and 5 V.

Fit the 28 pin DIL socket at the IC1 position on the EPROM board. This is to allow experimenters to fit a 2764 8K chip at a later date.

When you have finished you will have a very powerful means of customising your system to your own specifications. To mention one use: you could burn a renumber routine into ROM and then while writing a BASIC program simply renumber by calling the routine through the USR(X) function.

ETI

Fancy a pair of Wharfedale E70s? Can't afford them? Then why not build 'em yourself? Peter Freebrey underwent the mystic rites of woodworking and saved himself over £100.

For many years now there have been speaker manufacturers who have marketed kits for the 'do-it-yourself' audio enthusiast. At the present time there are several well known and respected firms supplying high quality kits. One such firm is Rank Hi Fi who manufacture the Wharfedale range. Their approach to this market is the Wharfedale Speakercraft series of drive units and crossovers, together with the constructional information necessary to duplicate their ready-built units using these same components. If the demand is there someone will supply that demand. . . such is the case with Wilmslow Audio who sell kits of the cabinets to suit the Wharfedale units. This review follows the construction of the E70 system using the WE70 flat-pack cabinet kit.

Why build loudspeaker kits? Well, one obvious answer is to save money; often the cost of a kit is very much less than buying the completed unit. If you are reasonably competent at woodwork, it is perfectly feasible to start from scratch with just a large sheet of flooring grade ¾" chipboard. An electric power saw makes the job much easier and can also give a better edge to the cut. It is often the edges which concern people as they are going to be visible somewhere around the loudspeaker cabinet and it is easy to think that to get rid of the ugly sight of these will be difficult. This is not necessarily true; there are several ways in which unsightly edges may be hidden from view. The simplest answer is not only to buy a kit of speakers, crossovers, and so on, but perhaps to buy a ready-cut cabinet kit as well — this does not rid you of dealing with edges, but at least they are all cleanly cut!

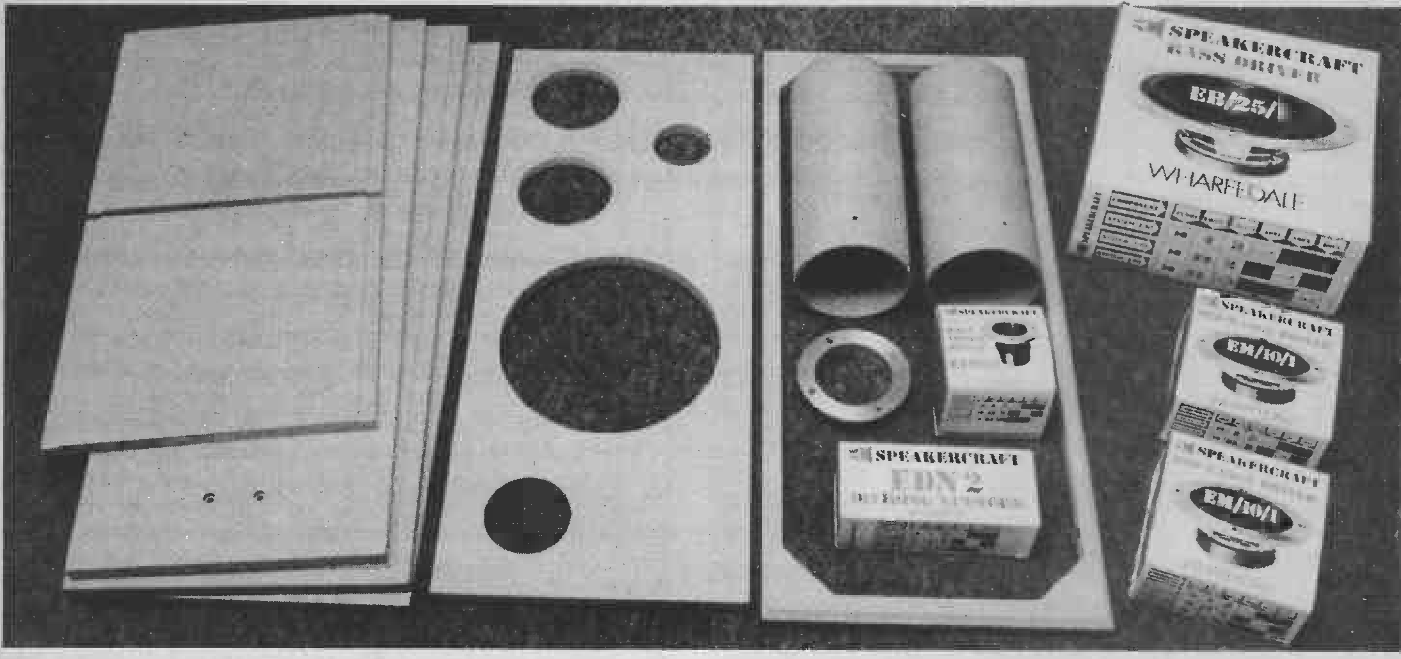
I had heard that Wilmslow kits were of a very high standard — several people having commented upon the ease with which they went together. That sort of build-up sometimes takes a bit of living up to and I waited for the delivery of the WE70 kit with some uncertainty. When they arrived my initial reaction was favourable; all cuts were clean and the method of construction looked simple and sensible. The sides, top and base are rebated by about ⅛". This not only gives you a better mechanical joint, but also makes it almost impossible to get any voids or gaps — which is good, acoustically speaking. It also means that with the minimum of care the cabinet will slot together into its correct shape with no unsquare corners or leaning sides. Included with the kit were two cardboard transmission tubes for the mid-range units, acoustic damping material, grille material (both black plastic foam for the reflex port and cloth for the front), nylon grille plugs and sockets, 3 mm wander plugs and sockets for loudspeaker lead connections, and the screws to fix the speaker units themselves. Last but not least there are written instructions on how to assemble the kit.

16 Steps To Heaven

Step one in the instructions is to examine the panels for transit damage. Presumably if any damage is noticed, Wilmslow Audio should be contacted as soon as possible. Step two is to remove all dust, etc from the panels. Any excess of wood dust from the sawing operation can only do harm so vacuum all surfaces. If there were any build-up of sawdust at the surfaces to be glued that sawdust could conceivably impair any glue joints and also cause the fit of the joints to be out of true.

Step three is to assemble the cabinet without gluing to check the fit. It is also suggested that panels be swapped around to find the optimum results. This step proved to be most encouraging. . . I assembled one unit (panels only) and held it together with just one turn of linen tape (no string please — it can bite into the corners of the chipboard and cause you extra work later). The cabinet felt as firm as a rock. No glue, just well-fitting joints. Thus encouraged I rapidly got on to step four, which was to paint the face of the baffle board matt black. I gave it a couple of coats of sanding sealer — not so much to get a 'de luxe' finish but to seal the wood surface. Chipboard is pretty thirsty stuff and you can use up a lot of paint if you do not seal the surface first. Just be careful not to get any of the sealer or paint on the edges, as this may affect the glue joint you have to make later.

Step five is to glue the midrange enclosures (transmission tubes) to the baffle boards, using plenty of glue to ensure an air-tight seal. The baffle boards are recessed to take the cardboard tubes so it is easy to line up for position. I used Evostik Resin W, which is a PVA wood-working adhesive for all glue joints. It is easy to apply and may be cleaned off the hands/clothes as it is water soluble. Just don't put your speakers out in the rain! Light pressure to a PVA glued joint gives a better joint so I placed one of the side panels across the top of the four tubes to ensure a light even pressure. Rather than apply liberal amounts of glue in one dose I used sufficient so that a *small* bead of glue was squeezed out all around the tube. This was smoothed around with a handy finger and when dry a further fillet of glue was applied all round the tube/baffle joint. Four pieces of approximately 1" thick polyurethane foam are supplied which must be



glued to the rear (outside) end of the baffle tubes. Wharfedale recommend a hard rubber pad at this position but as this 1" foam is to be compressed to about 3/16" it probably is just as good.

Step six is probably the most critical point in the whole construction procedure, for at this point the cabinet panels are glued together. This entails gluing five of the six panels; the sixth (the side furthest from the mid-range enclosures) is placed in its position while the glue is setting but is not glued. This enables you to work inside the cabinet; fitting the crossover, acoustic wadding etc.

Wharfedale suggest that the acoustic wadding be attached to the inside of the panels before you reach this step. Wilmslow Audio suggest that the wadding be fixed *after* the panels have been glued. Although I only learnt of Wharfedales' suggestion after I had completed step six, I favour the Wilmslow approach for several reasons.

If the wadding is stuck/tacked or stapled to the panels before they are fitted together two things may happen: 1) some of the wadding may inadvertantly get caught between the panels and cause either an air gap or 2) force the cabinet to go together 'out of true'. Also, with the wadding in place you cannot inspect the inside corners to check that there is a continuous fillet of glue all along the joint.

If you choose the Wilmslow way you will have to cut the wadding to fit around the mid-range enclosures but in practice this proved to be a very simple task.

Getting A Grip

Holding the whole thing together while the glue sets is quite a teaser. I was fortunate to have a set of excellent clamps known as Jet System Clamps made by TMT Design Ltd of Leamington Spa. They cost about £10 per clamp but are worth their weight in gold for this type of job. The problem comes from the 1" thick foam stuck to the rear of the mid-range enclosures; this tends to force the back panel out of position. Wilmslow suggest either that clamps be used or that the joints be held *firmly* together with masking tape. It *is* possible with masking tape but only just; remember that unlike your trial fitting in step three, the foam pads are being compressed to about 3/16" and all but one panel has glue all along the edges and is quite capable of sliding all over the place! I bought a wide webbing strap from a camping shop to assist the initial stages of holding the four vertical panels approximately in place while I set up the clamps. The cost of the strap was wasted as I could not get enough tension in it to overcome the spring in the foam . . . a linen tape would have done just as well! If you are going to use masking tape then get someone to help apply the pressure to hold the front and back panels in position while you apply the tape. Lastly, cut up a thin polythene bag and place four pieces inside each corner of the panel that is not to be glued; it would be a shame if this stuck firmly to the rest of the panels by accident!

It is useful to have a rubber-faced hammer at this stage as, having clamped or taped the cabinet firmly together, you may wish to tap the panels firmly but lightly into position. A hammer *and* a block of wood do the trick just as well, but try not to mark or dent any edges. The places to look for out of true joints are the corners . . . remember once the glue has set there is nothing you can do, so a few light taps now can save the day. Wipe off excess glue with a damp cloth. Wipe from the centre of each panel *out* towards the edge; try not to get any glue smeared over the panels.

Having completed step six the rest of the construction is plain sailing. Step seven is simply to remove the loose side when the glue has set (leave for at least 24 hours). I then put a small fillet of glue all around the inside of all joints BUT not up to the edges where the last panel is to fit . . . we want it to go back from whence it came!

Step eight is to place the drive units and reflex port trims in the baffle board and mark accurately where pilot holes for the fixing screws are to be drilled. Although the chipboard is high density it has a fairly soft texture so it is well worth buying a new ⅛" drill bit. This ensures the pilot holes are clean and in the right place . . . worn bits tend to wander! Although I'm sure it is unnecessary I drilled all my pilot holes just deep enough for the screws by slipping a small rubber sleeve over the drill bit at the right depth. No-one could accuse me of having any extra holes or air gaps here!

Step nine is to position the grille frame on the front of the cabinet with the cabinet lying on its back. Use masking tape to hold it in position and carefully drill a pilot hole through the grille and into the baffle board. I used a 1/16" drill bit and drilled four holes, one in each corner section of the grille frame. These holes can now be drilled out to the correct size to accept the nylon plugs and sockets that hold the grille in place. Wilmslow supply eight plugs/sockets for each grille but as Wharfedale suggested that four would be sufficient I chose the latter. It is far easier to line up four holes than eight! For the socket in the baffle board I used a 7/16" bit and for the grille a 7/32" bit. Don't forget to drill only from the rear of the grille and only to a depth of ¼-5/16". The 1/16" pilot hole may be filled with wood filler but when the grille material is fitted I doubt that these holes can be seen. If you are happy with the finish on the baffle board then glue the sockets in now; if not, then wait until you have quite finished before fixing them in position. Do not stick the plugs in the grille until you have fixed the material in place. I used a quickset epoxy glue for these fittings.

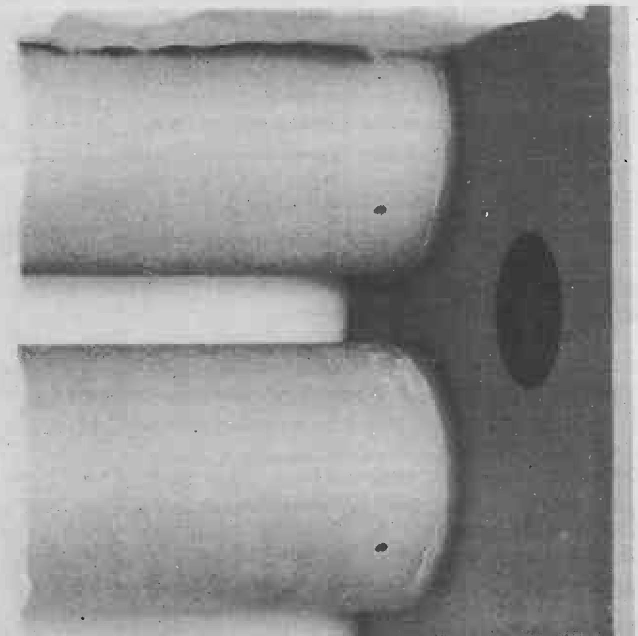
Step 10 is to glue the black, acoustically transparent foam over the inside of the reflex port aperture. You can use either PVA glue or quickset epoxy, just be careful not to get any of the adhesive on the foam where it is over the port.

Step 11 is to position the crossover network inside the cabinet on the rear panel opposite the bass unit aperture. Before you screw it into position check that the leads from the drive units can reach their appropriate tags! Wharfedale recommend that the crossover has a piece of felt or foam between it and the panel to prevent any vibration rattles. Also in step 11 is the fitting of the input terminals through the rear panel. I smeared the threads on these sockets with some latex glue, again to ensure that there would be no air gaps. Solder the leads from the crossover to these terminals . . . make sure they are connected correctly, red to red and black to black!

Step 12 is to cut three 5" discs of wadding and place these in the mid-range tubes. The Wharfedale instructions that come with every Speakercraft unit specify that the packing density of this wadding should increase towards the back of the tube and that the tube should be completely filled with wadding. In view of this I cut two extra discs and fluffed out those towards the front of the tube.

It's In The Bag

Step 13 is to line the inside of the cabinet with the acoustic wadding and glue the remaining side into place. Now comes the tricky bit — how do you slide the wadding up behind the mid-range tubes? The wadding catches on the side panel and snags up behind the tubes! Easy — get a large polythene bag 12" or more wide and about 15" to 18" long, slide the wadding into the bag, slide the bag plus the wadding up behind the tubes and, lightly holding the wadding in place, pull out the bag. Cutting the wadding to fit round the tubes sounds fiddly but turned out to be quite easy. Cut the holes for the tubes smaller rather than larger as the wadding will easily stretch to fit comfortably in place. No wadding is required on the baffle board but don't forget to put wadding on the loose side panel before you glue it into place! The wadding may be tacked or stapled into place.



The wadding is tacked or stapled in place.

Step 14 is to attach the wires to the drive units — *observing the correct polarity* (if in doubt refer to the Speakercraft instructions and double-check every connection), and screw all units and ports to the cabinet. Wire up and fit the bass unit last as the bass aperture gives you ample room to work inside the cabinet connecting wires to the crossover. The wires from the mid-range units come through small holes in the tubes and these holes should be sealed after you have connected the wires to the crossover. The fitting of the drive units should only be started after the glue joints of the final side have thoroughly set and any glue fumes have completely cleared. The comment regarding fumes is highly pertinent if you are not using a water-based adhesive. There is a possibility that the fumes could affect certain plastics used in the construction of the drive units.

Step 15: You have two working loudspeaker systems, so connect them to your amplifier and sit back and enjoy your favourite record.

Step 16: The cabinets are now ready for their final cosmetic treatment. There are a number of options open to you: they may be:

- veneered either by you or a local cabinet-maker.
- covered in iron-on veneer or plastic laminate.
- sealed and then painted (preferably sprayed) in colour of your choice.
- Wilmslow Audio also suggest the use of a 'Contact' type covering as these can be obtained in very realistic wood-grain finishes.

Whichever method you opt for you will probably have to attend to the cabinet edges/joints before you can proceed. Due to the small but noticeable tolerances in the cutting of the panels, the amount of glue and the pressure used during the construction, there are likely to be a few panels that are slightly proud of the edges that butt up to them. There are several ways to solve these problems but the simplest is to use one of the proprietary wood fillers. Which choice depends upon your choice of finish.

If the cabinets are to be covered in plastic laminate you can afford to use one of the more easily worked fillers such as Fine Surface Polyfilla, Alabastine or Plaster of Paris. If, on the other hand, you are going to cover them with 'Contact' or simply spray-paint them then I would suggest a tougher type of filler that is less likely to crack or crumble. My choice here would be one of the car body fillers — they are easier to sand than some of the loaded general-purpose fillers from the DIY shop. So you are less likely to sand away the wood from the cabinet instead of the filler!

The grille material must be stretched over the grille frames and either tacked/stapled or glued (or both) to the inside of the frame. The material supplied by Wilmslow Audio stretched easily and evenly; I smeared PVA glue over the rear faces of the frames (having first painted them black) and stapled the material in place while the glue set. When set I trimmed off the excess material (having removed the 50-odd staples) and ran another bead of the adhesive over the edge of the material.

Looking back on the construction of this E70 loudspeaker system using the WE70 flat-packs, I can only say that I am very satisfied with the way they went together. There were one or two instructions that could have been a little clearer but they have been covered in this article. Common sense would probably have solved any uncertainties but I chose to phone Rank Hi Fi to confirm my conclusions. The people I spoke to did not know that I was writing this review and so it is a pleasure to say the they could not have been more helpful. This entire project has been enjoyable from first to last.

BUYLINES

Wilmslow Audio sell the complete WE70 package (flat pack, drivers and all components for two speakers) for £220 plus £8 carriage. Wilmslow Audio, 35/39 Church Street, Wilmslow, Cheshire SK9 1AS.

ETI

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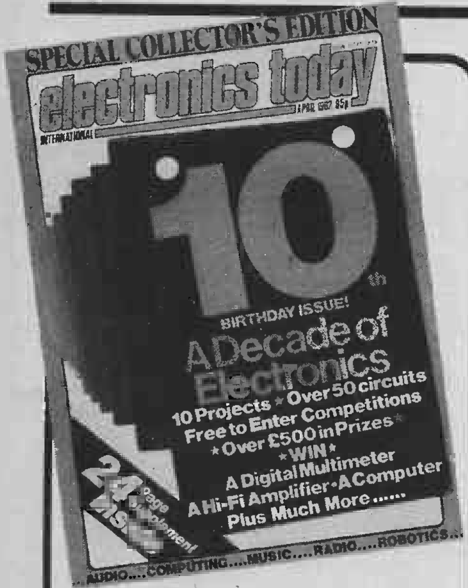

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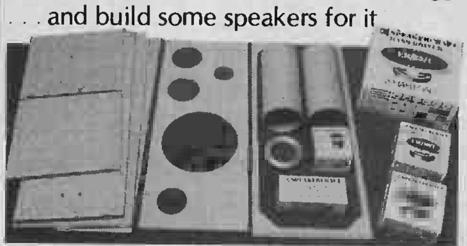

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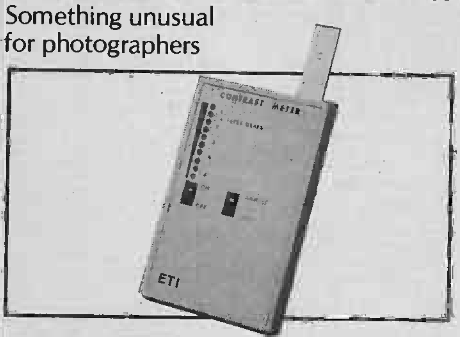
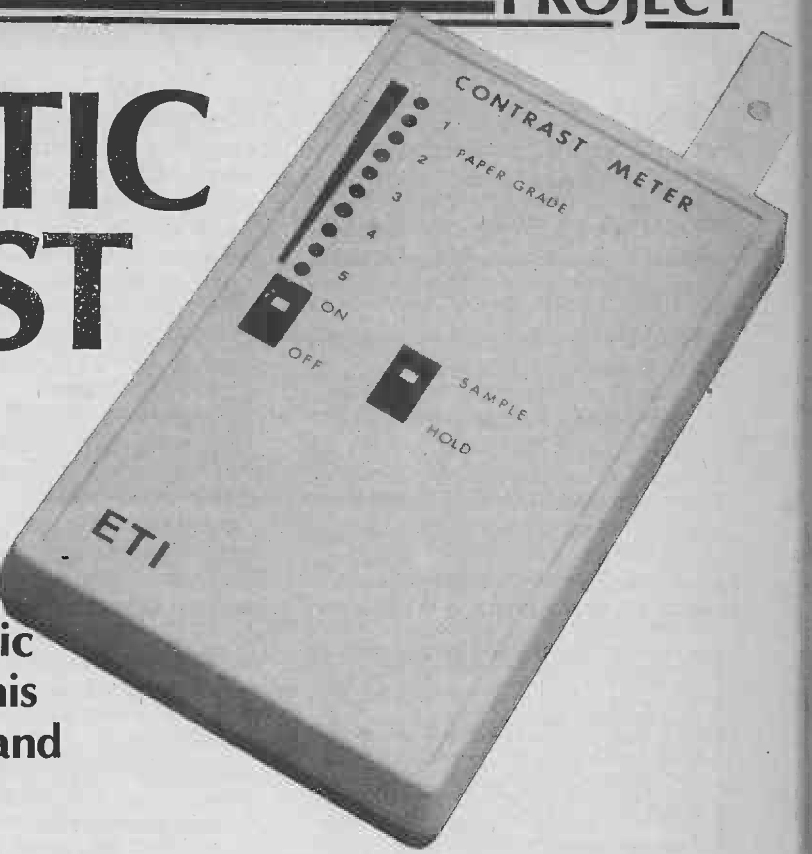


INFORMATION

AUTOMATIC CONTRAST METER



What's black and white and read all over? Answer — a photographic negative, providing you've built this simple and useful device. Design and development by Rory Holmes.

Contrast ratio is a very important quality of photographic negatives that must be assessed during the printing process, in order to select the correct grade of photographic paper. The contrast of negatives depends on the type of film used, the lighting conditions and the developing process; consequently five grades of printing paper are available to enable the full range of tones from black to white to be reproduced from any negative. Grade 1 is termed the softest and it is used with the highest contrast negatives. At the other end of the scale, grade 5 is the hardest paper, which will enhance the tonal variations of poor contrast negatives.

During the design stage of this project we experimented initially with two separate photodetectors which measured the instantaneous light difference between two points. There are a number of problems with this approach, as the photodiodes and their associated amplifiers must be carefully matched in light sensitivity.

Secondly, the lightest and darkest points of the image must be known exactly, and the two photodetectors need to be simultaneously positioned on these points while the reading is taken. This is an awkward business at the best of times, but especially so in a darkroom!

We considered that a different approach was required and developed the circuit of Fig. 1 to overcome some of these difficulties. Only one photodetector is used and the peak positive and negative voltages obtained from different light levels are followed and stored independently by sample and hold circuits.

Now, as long as the photodiode is scanned at some time through the lightest and darkest points of the image, the peak detectors will memorize the maximum and minimum voltages, and thus provide a contrast measurement.

The photodetector input stage of our meter is rather unusual in its configuration. Photodiodes are usually

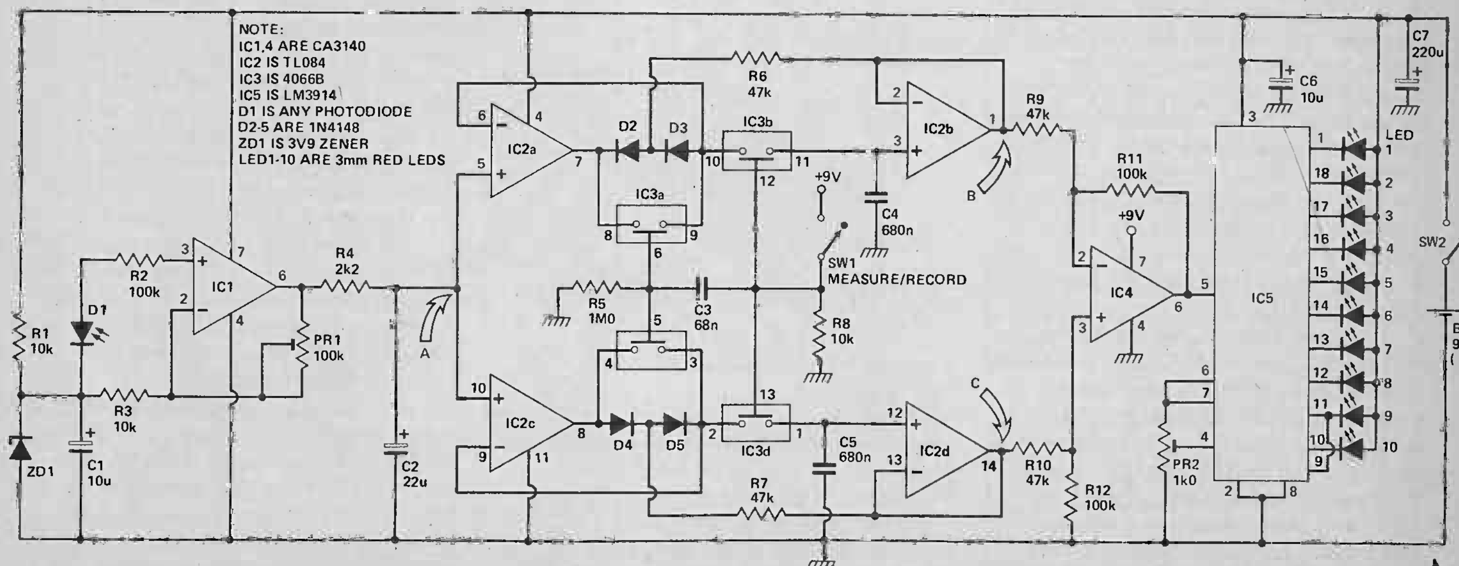


Fig. 1 Circuit diagram of the Contrast Meter.

used in the 'photovoltaic mode' where the photocurrent developed and measured is linearly proportional to the light intensity. Our input amplifier has an extremely high input impedance and thus measures the open circuit voltage generated by the photodiode. This voltage is logarithmically proportional to irradiance as the graph of Fig. 2 illustrates. This is a very convenient property since the sampling circuitry can now work on the log of the light level to provide maximum and minimum values. By simply subtracting these two values with a differential amplifier we obtain a voltage that is logarithmically proportional to the ratio of the maximum and minimum light levels, ie the contrast.

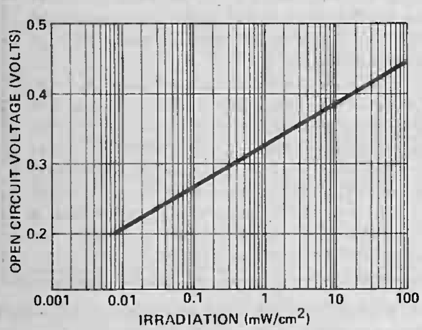


Fig. 2 Response of the photodiode used in this project.

Meter Made

The ETI contrast meter was intended primarily to determine the paper grade for a well balanced print; consequently a 10 LED bargraph type meter is sufficiently accurate for calibrating the five grades of paper. At today's prices this also works out somewhat cheaper than a moving coil meter and is less prone to damage. After calibration, the meter will be found very easy to use. It is switched on with the 'sample/hold' switch in the 'hold' position and placed down flat on the enlarger base with the photodetector probe anywhere in the image area. (The photodiode has been mounted in a separate probe with its amplifier in order to keep it as close to the focused image plane as possible. If it were much higher than this the detecting element would pass through an unfocused image, giving a false contrast reading).

Any red safety lights should be switched off before the reading is taken to avoid error since the photodiode is responsive at this wavelength. The sample/hold switch should now be moved to the sample position; this will clear any previous reading and start measuring light variations. Now the photodiode may be moved across the image and through the areas that look the brightest and darkest. This can be done quite slowly thanks to the peak detectors' long memory time; however, several areas should be scanned to ensure the recording of the true maximum and minimum. The eye can be deceived quite easily by those cunning optical illusions lurking among the shades of grey!

During the scanning process the reading on the LED scale will increase and finally level-off at the true contrast ratio when the black and white peaks have been covered. Before removing the meter from the image area the sample/hold switch should be set to 'hold'. The meter will now be immune to further light variations and will continue to display the contrast reading for a considerable time, thanks to the even longer memory of the sample/hold circuitry!

A true ratio is provided by the meter and thus the contrast reading for a given negative will be independent of the light source intensity and enlargement size (photographic aberrations known as "circles of confusion" may produce sources of error under certain conditions). Negatives may thus be compared or matched for contrast.

Construction

The meter is built into a slim style plastic enclosure produced by OK Machine and Tool company. This houses the battery and main PCB on which all the parts are mounted. Since the light sensing element must be as close to the enlarger base plane as possible, we have mounted it externally on a separate small PCB with its associated amplifier. A probe to house the external sensor is made from a short length of aluminium channel extrusion. Figure 3 shows the

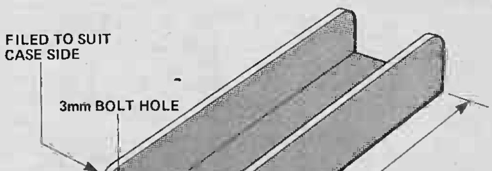


Fig. 3 Details for the aluminium extrusion that houses the photoprobe.

dimensions for the probe; if the aluminium channel proves difficult to obtain, a piece of the slotted aluminium extrusion used for commercial shelf-racking systems is ideal. This is available from most DIY stores in short lengths with the required internal width. After filing or cutting to the right size, a piece of insulating tape should be stuck down on the inside to prevent shorting out the PCB. As shown in the diagram, a hole is drilled on the end for bolting it to the bottom of the case. This bolt should eventually be connected to circuit ground, thus providing screening for the photo-amplifier. The two PCBs for probe and main meter circuits are laid out as one board, and should be sawn apart along the lines shown on the foil patterns.

For other construction arrangements, the circuit can be left as a single board, since the interconnections are already made.

Three wires are used to connect the two boards together as indicated on the overlay; these should pass through a small hole drilled in the case side where the metal probe case is bolted on. When the probe board is mounted and stuck down in its channel, a piece of thin aluminium sheet is cut to form a lid with appropriate holes for the photodiode and preset. (The photodiode case is internally connected to the cathode, so it must not short against the lid).

Calibration

Start with preset PR1 fully clockwise to set a gain of 1; also set PR2 fully anticlockwise, setting the voltage required to illuminate the lower end of the bargraph at zero. First, measure a high contrast negative that is known to require grade 1 paper for a good average contrast after developing. Initially a low contrast reading will be obtained, say about grade 4 or 5. Now, adjust PR1 anticlockwise to increase the gain of the photoamplifier. Take another measurement, when the contrast reading should be greater. Repeat this process until a grade 1 is consistently recorded.

Now select a negative with very poor contrast ratio, one known to require paper grade 5 for bringing out the contrast. Take measurements several times while adjusting only PR2 clockwise, until the bottom end of the scale illuminates at grade 5. The other contrast grades should now fall linearly between these points and can be checked for accuracy.

Although the bargraph display has a low resolution and accuracy, the rest of the metering circuit is obviously much better than this; consequently a moving coil meter could easily be added to measure the contrast voltage for those who may desire greater resolution.

HOW IT WORKS

The general circuit arrangement consists of a photo-amplifier which feeds a voltage derived from varying light levels in an enlarger, to a pair of peak detectors. One follows the peak positive voltage and the other the peak negative voltage. The capacitors used for storing the voltage peaks in the followers also form part of sample and hold circuits which are then switched to 'hold' after measurement. Their outputs represent the maximum and minimum values of light intensity. A differential amplifier then computes the ratio of these values and the result is displayed on an LED bargraph meter.

IC1, a CA3140 CMOS op-amp, is used as the photodetector amplifier. It is configured as a non-inverting DC amplifier with a gain variable from unity to about 10, set by PR1. Although IC1 can have input and output voltages all the way to ground, this facility is not used owing to the driving requirement of the TL084 quad op-amp. This requires inputs at least 1 V above ground, and thus IC1's output is offset by a reference voltage of 3V9 provided by R1, ZD1 and C1. The anode of the photodiode is connected via R2 to the non-inverting terminal of IC1 which has an effectively infinite input impedance. Thus the open circuit voltage generated by the photodiode is amplified according to the gain set around IC1 and appears at the output on pin 6 added to the reference voltage.

The voltage at point A (ignoring the reference offset) will be logarithmically proportional to the intensity of incident light, owing to the properties of the photodiode (see Fig. 2) R4 and C2 form a simple filter to remove 100 Hz ripple caused by AC mains bulbs. This voltage is fed directly to the peak detectors. These circuits are essentially the same, the difference being the polarity of the rectifier diodes. They operate in exactly the same way, and we shall deal only with the peak positive voltage follower.

Assume initially that the CMOS analogue switch IC3c is open and IC3d is closed. C5 will be connected to the output of op-amp IC2c via the rectifiers D4 and 5 (we can ignore the action of R7 for the moment). C5 will charge up via the rectifiers to the most positive voltage peak when the voltage at point A on the non-inverting terminal is greater than the capacitor voltage applied to the inverting terminal. The voltage held on C5 will droop over a period of time due to leakage current through the rectifiers D4 and 5 and the input bias current of IC2c. IC2c was chosen as a FET op-amp with a low input bias current and R7 is included to reduce the diode leakage current.

IC2d is connected to C5 as a straightforward high impedance voltage follower to buffer the stored voltage. When the input voltage to IC2c at point A drops below the peak value, IC2c's output will go negative, reverse biasing D4. However, IC2d applies the capacitor voltage via R7 to the anode of D5, effectively removing leakage current through D5.

The peak positive value of the signal at A thus appears at point C, and likewise the peak negative value at point B. When the analogue switch IC3d is now opened, C5 is disconnected from the peak detector and acts in conjunction with IC2d as a sample and hold circuit thus isolating the measured values from further light variations.

When SW1 is open, R8 and R5 hold the control pins 13 and 5 of IC3 low, opening both analogue switches. This is the 'hold' mode. When SW1 is now closed, the control pin 13 is taken high, switching to the 'sample' mode. C3 and R5 produce a positive pulse (about 50 mS) on control pin 5 to briefly short out D4 and D5, so resetting the peak detector to the current voltage at point A. When C3 has charged the IC3c switch will open again, allowing the peak detector to function.

IC4 is wired as a differential amplifier with a gain of 2, to subtract the voltage at point C from point B. Since these voltages are the log of the light levels, the output on pin 6 will represent the contrast ratio of these light values.

IC5 is a standard LED bargraph driver, the LM3914. The input voltage on pin 5 is converted linearly to illuminate one LED on a scale of 10. Full scale deflection (LED 10) is set internally at 1V2; the zero scale deflection is set by PR2 anywhere between 0V and 1V2 during the calibration process. C6, a 10 uF tantalum, is required for IC5 to ensure stability from oscillation.

NOTE: k = CATHODE

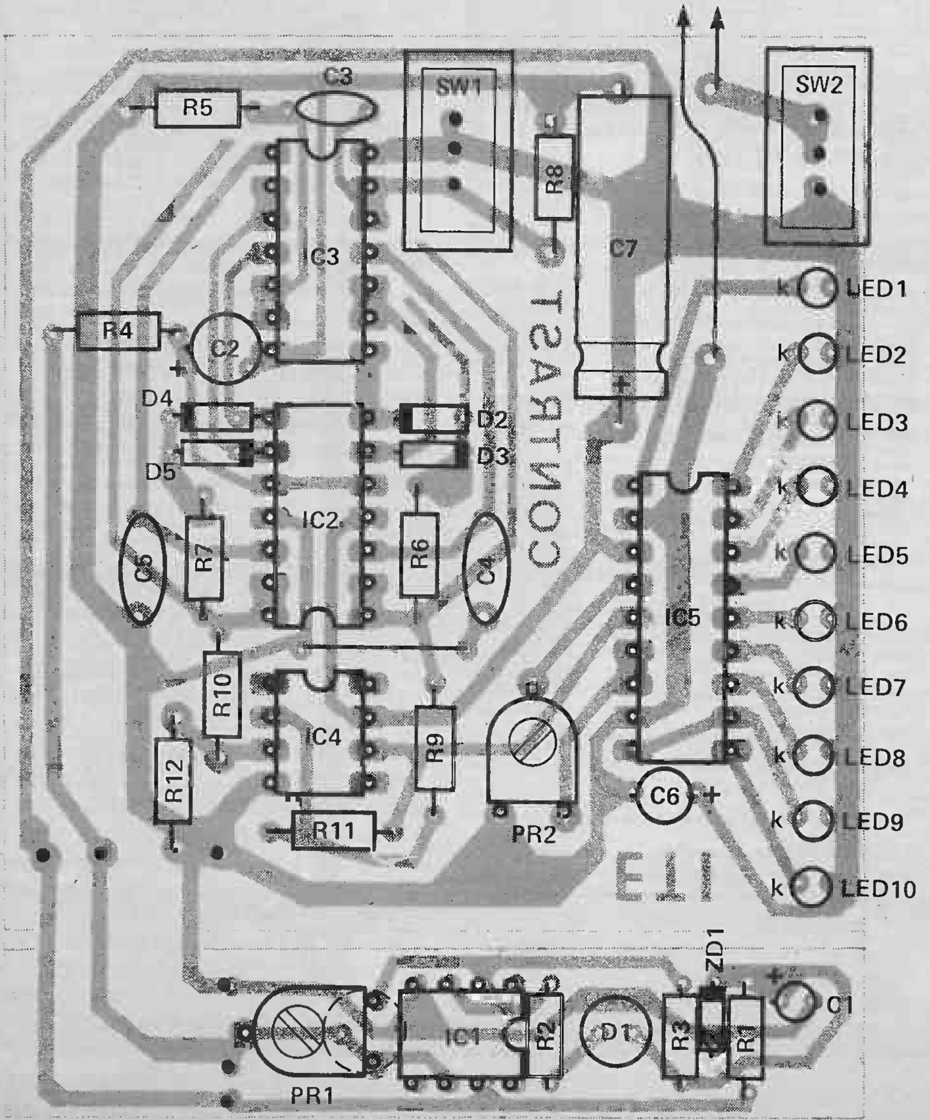


Fig. 4 (Left) Component overlay for the meter (showing the board uncut).

PARTS LIST

Resistors (all ¼ W, 5%)	
R1, 3, 8	10k
R2, 11, 12	100k
R4	2k2
R5	1M0
R6, 7, 9, 10	47R
Presets	
PR1	100k subminiature horizontal preset
PR2	1k0 miniature horizontal preset
Capacitors	
C1	10u 35 V tantalum
C2	22u 25 V tantalum
C3	220u 16 V electrolytic
C4, 6	82n polycarbonate
C5	68 n ceramic
Semiconductors	
IC 1,4	CA3140
IC2	TL084
IC3	4066B
IC5	LM3914
D1	BPX65
D2, 3, 4, 5	1N4148
LED1-10	3 mm red LED
Miscellaneous	
SW1, 2	miniature slide switches

Case (see Buylines); PCB (see Buylines); B1 PP3 9 V battery (preferably alkaline type).

BUYLINES

The photodiode specified in the Parts List is the one used in our prototype, but any general purpose type should do. The case we used is a Pactec type HP, size 146 x 91 x 28 mm. The PCB is available from us using the order form on page 44 — price is £2.12.

DESIGNER'S NOTEBOOK

Five into one does go. This month Don Keighley explains all about sampling and time-division multiplex systems, and looks closely at the advantages of pulse-width modulated telecommunications networks.

Sampling is a process we can undertake if we want to combine many different signals on to a single transmission line. The transmission line can be of any type such as wire, radio, or optical. Combining several signals into one is called 'multiplexing' and can save the expense of having many separate lines. Sampling is used in a specific type of multiplexing called time-division multiplexing (TDM) which I'll explain later. The other form of multiplexing — frequency-division multiplexing (FDM) — is the basis of all standard radio transmissions. Each signal to be transmitted is mixed with a carrier wave (or radio frequency) on to a set frequency within the radio spectrum. Thus many signals can be transmitted and received by radio link — one on each defined frequency of the radio spectrum.

Figure 1 shows an illustration of sampling. In the figure, a sinusoidal signal (known as the message signal) has a series of values taken at regular intervals. These sample values can be used to represent the message signal. For instance, we can pass the actual DC values of the samples, ie their voltages, along the line. At the other end of the line the sample values, or pulses as they are usually called, are converted back into the message signal, simply by passing them through a lowpass filter. The filter removes the high frequency pulses and thus re-creates the envelope of the original message signal — as shown by the sinewave of Fig. 2.

One of the most important questions arising is — How often do we need to sample the message signal? It is obvious that if the signal is sampled too few times we won't be able to reconvert the pulses into the message signal at the receiving end of the transmission line.

The minimum number of samples is given by the sampling theorem, which states that a message signal of bandwidth B Hz can be represented by a set of sample values taken at a frequency of 2B Hz. For example, an audio system has a frequency response of 20 Hz to 20 kHz. Its bandwidth is thus 20,000 − 20 = 19,980 Hz. The audio signal of the system can thus be represented if samples are taken at 2 x 19,980 Hz = 39,960 Hz.

But the *minimum* number of representative samples (2B Hz) isn't the *easiest* number of samples to convert back into the message signal. It's usual to take a greater number of samples because doing so makes the reconversion easier. To see why this is so we've got to take a look at the spectra of the transmitted samples and see how they differ when different sample frequencies are used. Figure 3 shows the possible spectrum of a message signal such as an audio signal. It's the sort of result you would see on the screen of a spectrum analyser. Frequency f_m is the maximum frequency contained in the signal. The lowest frequency contained is 0 Hz (the signal extends down to DC); so the bandwidth of the message signal is $f_m - 0 = f_m$ Hz.

When the message signal is sampled at a frequency f_s the overall spectrum looks something like that shown in Fig. 4 and consists of components at harmonics of the sampling frequency, with upper and lower sidebands around them, as well as the original spectrum of the message signal. In Fig. 4 you can see the sampling frequency, f_s, is more than twice f_m — hence there is a gap between the highest frequency of the higher sideband of a

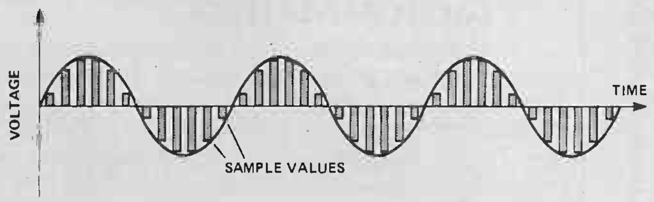


Fig. 1 A message signal can be represented by a series of sample values of the signal.

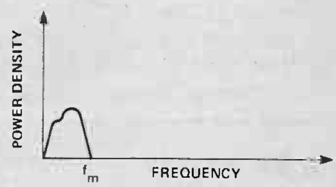


Fig. 3 Power density spectrum of typical audio signal. The higher frequency component in the signal is f_m. The signal extends down to 0 Hz, so the bandwidth of the signal is f_m Hz.

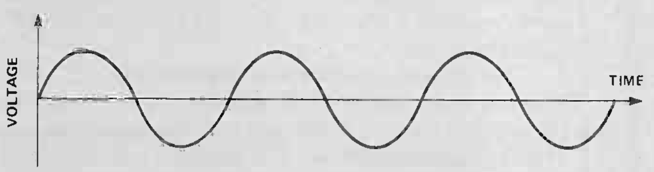


Fig. 2 If the series of sample values is passed through a lowpass filter the original message signal is recreated.

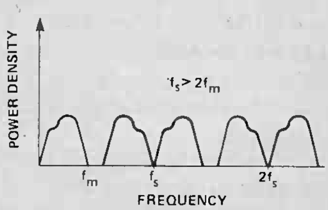


Fig. 4 Power density spectrum of an audio signal, sampled at a frequency of f_s. In this example, f_s is greater than $2f_m$.

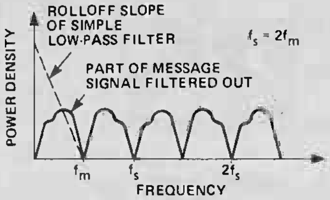


Fig. 5 Sampling frequency f_s equals $2f_m$. A simple lowpass filter may filter out some of the wanted message signal.

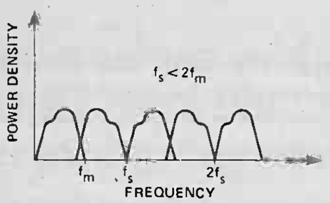


Fig. 6 Sampling frequency less than $2f_m$. A lowpass filter cannot be used to recreate the original message signal.

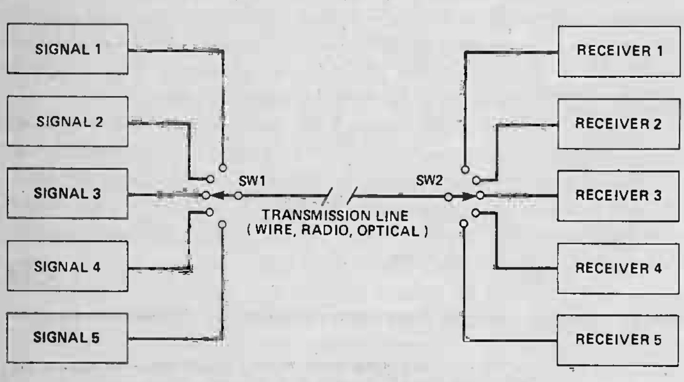


Fig. 7 A simple time-division multiplex (TDM) system.

component and the lowest frequency in the lower sideband of the next component. This gap between bands means that a simple lowpass filter can be used at the receiver to pass only the message signal and not the higher components: so the message signal is recreated.

With a sampling frequency of only $2f_m$ (Fig. 5) the highest frequency of one band and the lowest frequency of the next occur at the same point. A simple lowpass filter would filter out some of the message signal, as shown in the figure. A more complex lowpass filter (with a steeper roll-off slope) could be used to correctly recreate the message signal.

In Fig. 6, f_s is less than $2f_m$ and, as you would expect, the spectrum shows how message signal and sidebands overlap. A lowpass filter *cannot* be used to recover the whole of the message signal without letting through part of the next sideband.

TDM Tricks

A simple TDM system is shown in Fig. 7, in block diagram form. Each signal to be transmitted is connected to an input of switch SW1. This switch, although shown in the diagram as a mechanical-type switch, will be of electronic construction in a real TDM system, so that a high switching speed can be obtained. The output signal from the switch is transmitted along the transmission line to switch SW2, which connects each receiver, in turn, to the line. Providing the switches are operating fast enough so that the sampling theorem is fulfilled ($f_s \geq 2f_m$) for *all* the message signals, everything is fine and we have five signals passing down one line.

The whole process of sampling and TDM is a form of modulation because only a representation of the message signal is transmitted, not the actual signal. And because pulsed samples of the message signal are transmitted, we call the process pulse modulation.

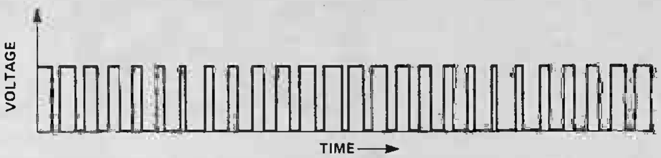


Fig. 8 Pulse-width modulation. The width of each pulse varies in accordance with the amplitude of the message signal.

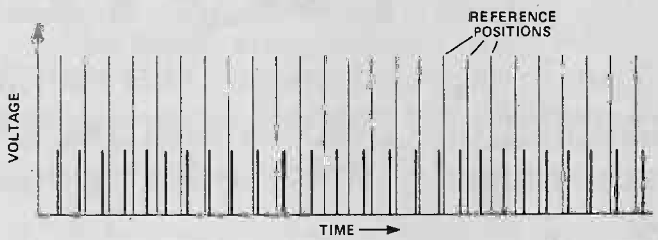


Fig. 9 Pulse-position modulation. Each pulse's position, with respect to a reference point, varies in accordance with the message signal amplitude.

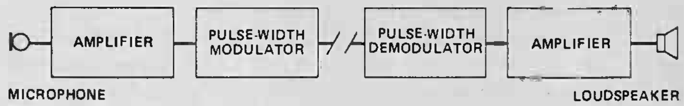


Fig. 10 A pulse-width modulation microphone/loudspeaker system.

There are various forms of pulse modulation which can be used in a TDM system, all relying on the fact that the original sample values control some property of corresponding pulses. The one just described uses the DC value (ie amplitude) of the pulses and is therefore known as pulse-amplitude modulation. Other forms of pulse modulation are: pulse-width modulation (where the width of the pulses is varied according to the sampled value) and pulse-position modulation (the position of the pulse, relative to a reference position, is proportional to the sample value). Figures 8 and 9 show examples of these pulse modulation systems and the sampling frequencies of both must follow the sampling theorem — the sampling frequency must be at least twice that of the message signal bandwidth. There is a final pulsed system, in which each sampled value is converted into a train of binary digits. This is, strictly speaking, a digital system and doesn't concern us here; however the system must still follow the sampling theorem.

Practical Matters

With careful design all the pulse modulation systems can give good results in TDM but perhaps the best — because it's easy to use, has a high immunity to interference and yet needs a minimum of component hardware — is pulse-width modulation (PWM). Figure 10 shows a block diagram of a PWM microphone/loudspeaker set-up — such as you might have in a multi-station intercom system or similar.

We can investigate the modulation and demodulation blocks in more detail, as in Fig. 11 and 12. Figure 11 shows a simplified pulse-width modulator. It consists of an oscillator to provide sampling pulses at a rate of over $2f_m$, so that the sampling theorem is fulfilled. In a good quality audio modulator, the sampling rate is therefore over 40 kHz and the time between pulses must be $1/f_s$ = 25 uS.

The pulse duration is less than this, say 1 uS, and each pulse charges the capacitor C1 to full voltage. After charging, the capacitor is linearly discharged via the constant current source. The cycle repeats itself at every pulse. The capacitor's discharge rate is a product of the capacitor/constant current time constant, which should be about 2 uS. Comparator IC1 compares the ramp discharge with the incoming audio signal — when the non-inverting input voltage is above that of the inverting input

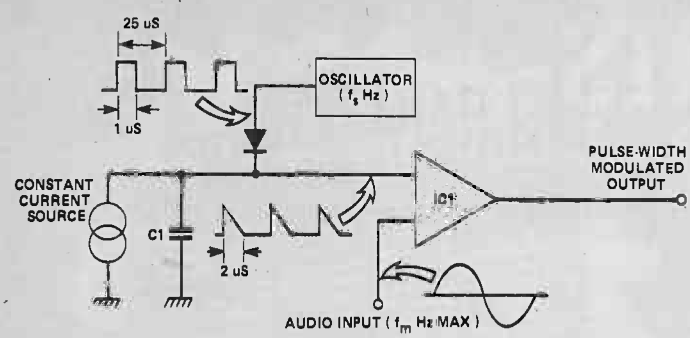


Fig. 11 A pulse-width modulator in detail.

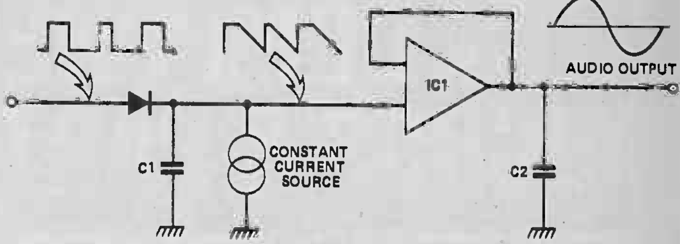


Fig. 12 A pulse-width demodulator can be built using the same basic components used in a pulse-width modulator.

the comparator output is high; when the non-inverting input is below the inverting input the output is low. Thus the output is high the instant of every sampling pulse, but falls low again after a time which is linearly related to the amplitude of the audio signal. In other words, the width of the pulse is modulated by the audio signal.

A pulse-width demodulator is shown in Fig. 12. A capacitor with a parallel constant current source is again used and the incoming width-modulated pulses cause a charge/discharge cycle similar to that in the modulator. The average DC level of charge across the capacitor is dependent on the width of the pulses — the wider the pulse, the higher the DC level. Buffer IC1 prevents loading of the voltage across the capacitor and the output is lowpass filtered by capacitor C2 to remove the sharp spikes of the sampling pulses, thus re-creating the original audio message signal.

The advantages of such a system aren't always immediately obvious, but you must remember that the audio signal is being represented by a pulse of nominal width 2 uS in a cycling time of 25 uS. This means that 12 different, high-quality audio signals can be time-division multiplexed down that transmission line simultaneously and without interference — and this is just a simple system. With a shorter nominal pulse width and more accurate modulators and demodulators, many more signals can be multiplexed on to a single transmission line.

It's all down to economics really. When you look at a large telecommunications system like the telephone network, there are literally thousands upon thousands of miles of expensive copper cable. By putting 100 telephone conversations down one line the overall cable cost is only 1/100th of that of a non-multiplexed system. Makes sense, doesn't it!

ETI

SOUND EFFECTS 1: BOMB DROP

One of the attractions of the more sophisticated video games seen in 'fun' arcades these days is the realistic array of sound effects that go with the action — gunshots, bomb whistles and explosions, etc. Make some yourself with just one IC. Design by Phil Wait.

Those 'cannon shots' and explosions that go with the popular 'Space Invaders' video games and its variants add a measure of interest, feedback and stimulation to the action in which you participate on screen. Those sounds are electronically synthesised — that is, they consist of a complex mixture of waveforms that make up the required sound.

A 'bomb drop and explosion' is a remarkably complex sound when analysed carefuly. Looking at it simply, there is a descending tone followed by a burst of noise that dies away in intensity. The descending tone starts at quite a high pitch and is not a 'pure' tone (ie a sine wave). The explosion is a burst of noise that commences suddenly and dies away slowly in a recognisable way (usually exponentially). While it is possible to electronically produce very nearly an exact replica of a bomb drop and explosion, some compromises are acceptable to reduce the complexity and cost of the task and yet produce a recognisable replica of the sound.

To produce such sound using conventional components — transistors, diodes, op-amps, resistors and capacitors — would require a whole legion of components. Fortunately, the IC maufacturers can come to our rescue here and much of the circuitry can be incorporated into a complex integrated circuit requiring the addition of a minimum of external components and the appropriate interconnections to synthesise the required sound. Generating a wide variety of sounds fortunately requires only a limited number of functional blocks, such as: a noise generator, voltage controlled oscillators, multivibrators, envelope generators (a sort of modulator), mixers and amplifiers. Tim Orr discusses such circuitry elsewhere in this issue.

Texas Instruments, the giant US-based component and equipment manufacturer, have designed a series of complex function ICs for various applications and among them is the SN76488 Complex Sound Generator. This chip contains both linear and digital circuitry and is intended for use in applications requiring audio feed-back to the user — video games, pinball, alarms, toys, etc, or industrial indicators, feedback controls and the like. Power consumption is quite low, allowing battery operation, and only a single supply rail is required.

The SN76488 is contained in a 28-pin package and can be purchased for less than £5. It is quite a versatile chip, but we have chosen to describe how to obtain only two sound effects, these being a bomb drop and explosion, and a steam train and whistle. The former is described here; the latter appears on page 118.

Construction

Both the projects described use the one PCB design. Only the required components are assembled into the board according to each overlay diagram to obtain the required sound generator. Naturally enough, the polarity of the IC should be noted as well as the polarity of electrolytic and tantalum capacitors used. Commence construction by assembling the passive components, followed by the IC. This is not a CMOS device and no special care is required, apart from being careful not to bend any pins under the device when inserting it. If you wish, a socket may be used for the IC. This way, you can assemble both projects and purchase only one IC, swapping between the boards as you need to use them!

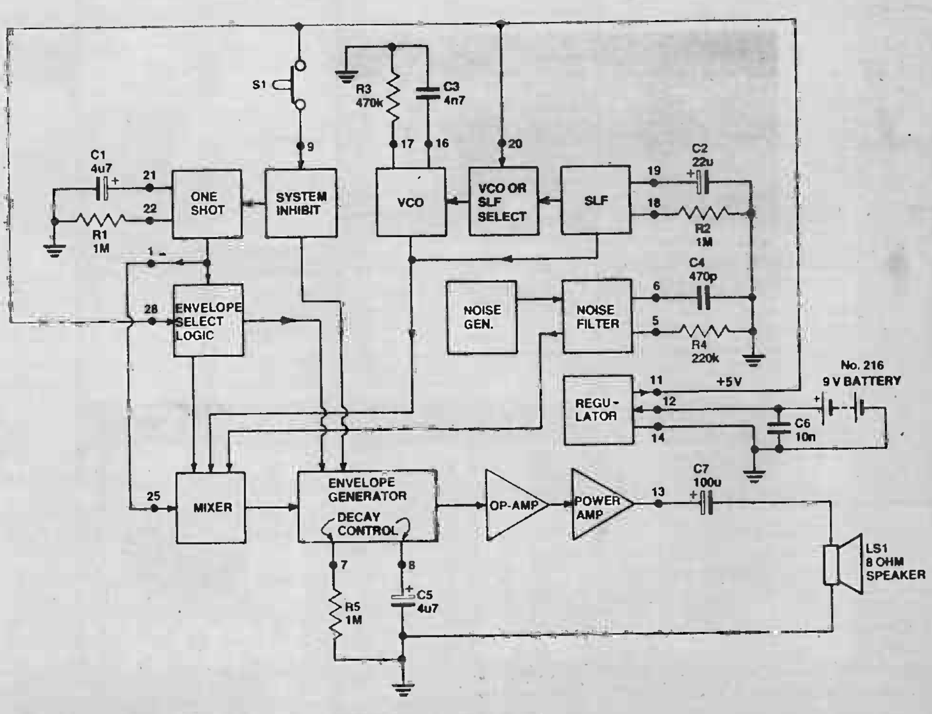


Fig. 1 Circuit diagram of the Bomb Drop and Explosion sound effects board.

Wiring to the switches, the speaker and the supply should be attached last.

The unit may be mounted in any convenient-sized box and the speaker mounted on the front. Alternatively, it may be wired into an existing piece of equipment. We'll have to leave these arrangements up to you.

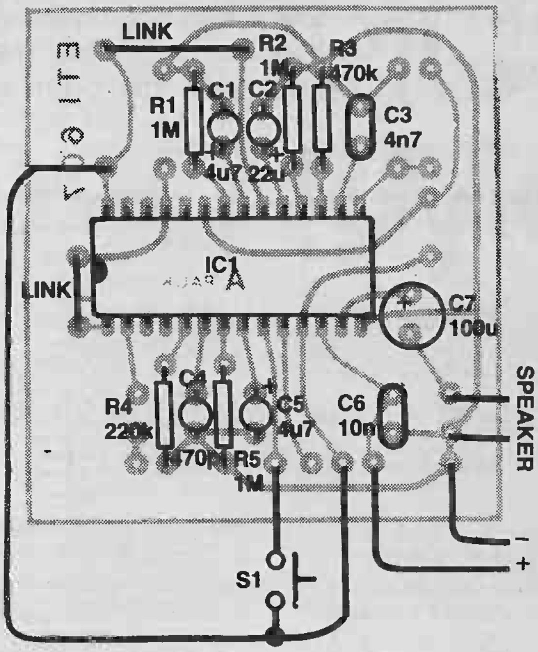


Fig. 2 Component overlay for the Bomb Drop board.

Projectile Project

This produces a 'bomb drop and explosion' sound at the press of a button. Alternatively, the push-button PB1 could be replaced by a pair of relay contacts operated by a piece of equipment or a transistor (emitter to pin 9, collector to other side of PB1) that is turned on by a logic high applied to its base via a resistor.

This project is one of the most complex, using almost every functional block within the SN76488. Varying R3 and C3 a little will vary the pitch range of the 'bomb drop' (desending whistle), while varying R4 or C4 a little will alter the characteristics of the explosion. Note that it is generally easier to 'fine tune' things by varying the resistor values. The duration of the event can be varied by changing the value of either C1 or R1 and the decay of the explosion can be changed by varying R5 (varying C5 produces quite gross changes in the decay period).

Watch that you insert the link on the PCB in this one, located at the 'notch' end of the IC.

PARTS LIST

Resistors (all ¼ W, 5%)

R1,2,5	1M0
R3	470k
R4	20k

Capacitors

C1,5	4u7 16 V PCB electrolytic
C2	22u 16 V tantalum
C3	4n7 ceramic
C4	470p ceramic
C6	10n ceramic
C7	100u 16 V PCB electrolytic

Semiconductors

IC1	SN76488 (see Buylines)

Miscellaneous

PB1	SPST push-button switch

PCB (see Buylines); 50 mm diameter 8 ohm speaker; PP3 battery and clip.

BUYLINES

Very few components and very few supply problems with this one. The SN76488 is an improved version of the Texas SN76477 and can be obtained from Technomatic. The PCB will cost you £1.80 from our PCB Service; see page 44 for details.

HOW IT WORKS

This unit employs most of the function blocks in the SN76488. The SLF provides a linearly increasing voltage waveform, or ramp, to the VCO, taking several seconds for the ramp voltage to rise from zero to maximum value. The causes the VCO to produce a tone which 'glides' down in pitch, making the 'bomb drop' effect. The explosion is generated by the Noise Generator/Filter and the Envelope Generator. It starts with a burst of noise, which dies away in intensity exponentially in a few seconds.

The whole sequence is triggered by operating the pushbutton, PB1. This applies a high (+5 V) to the input of the System Inhibit block, pin 9. This in turn triggers the One Shot and the Envelope Generator. At the commencement of the One Shot timing period, the One Shot triggers the SLF HI/LO Sync, starting the SLF, and the VCO does its things. At the end of the One Shot timing period the Envelope Select Logic becomes operative, the SLF is disabled and the Envelope Generator commences to do *its* thing. The Mixer selects the VCO output at the start of the One Shot timing period and the Noise Generator/Filter output at the end of the One Shot timing period. Thus the two sounds are switched through to the audio output stage in sequence, the Envelope Generator modifying the noise so that it dies away, the time it takes to do so being controlled by the time constant of R5, C5.

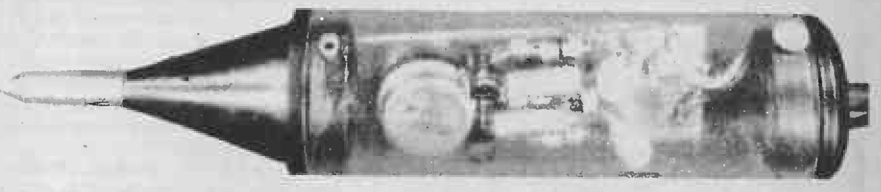
The starting pitch of the VCO is determined by R3 and C3, the rate of rise of the voltage ramp produced by the SLF is determined by C2 and R2, while the One Shot timing period is determined by the time constant of C1 and R1. The frequency characteristics of the broad-band noise produced by the Noise Generator are modified by R4 and C4 connected to the noise filter control pins (5 and 6).

Audio output is coupled to the loudspeaker via C7, a 100uF electrolytic capacitor.

ET

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 accurary 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.

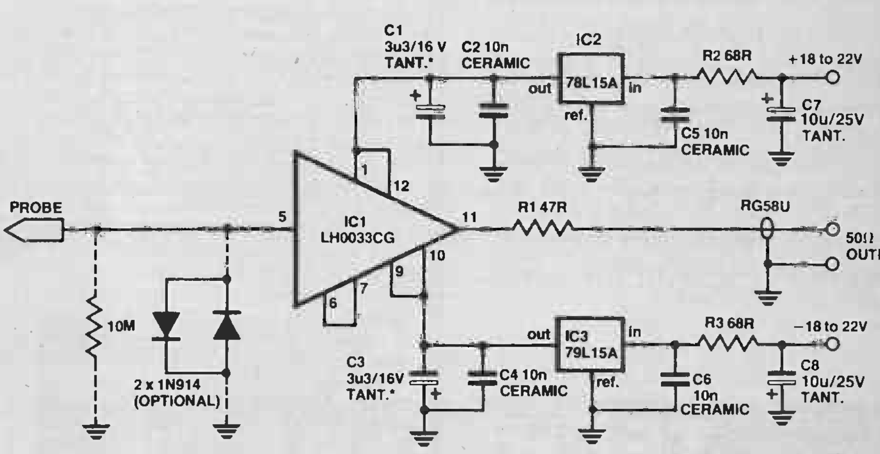
The input impedance of an oscilloscope can be effectively raised, and the capacitance decreased, by using a 'stepdown' probe. For example, a 'x10' probe will generally have an input impedance of 10M and a parallel capacitance of between 5pF and 15pF. While this improves the input impedance there are 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

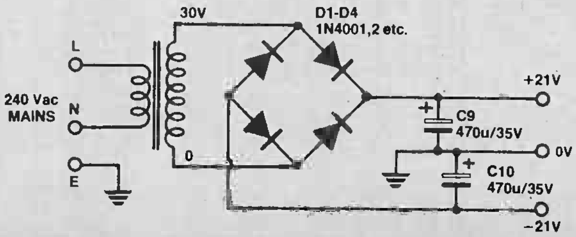


Fig. 1 Circuit diagram for the probe. C2 and C4 need to be ceramic
Fig. 1 Circuit diagram for the probe. C2 and C4 need to be 10n ceramic chip or 1n0 ceramic disc or plate types. C5 and C6 need only be disc or plate ceramic. See 'Bypassing' over the page.

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.*

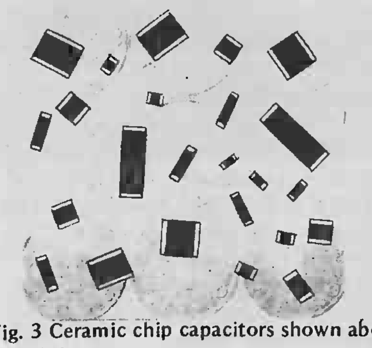


Fig. 3 Ceramic chip capacitors shown about actual size.

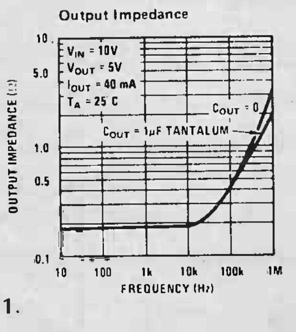


Fig. 1.

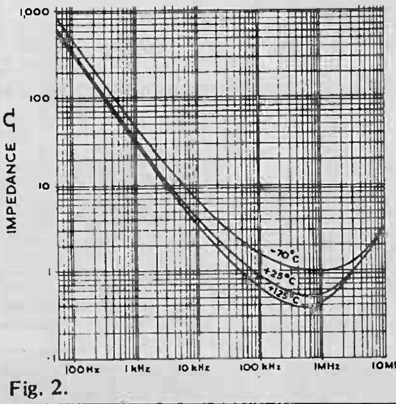


Fig. 2.

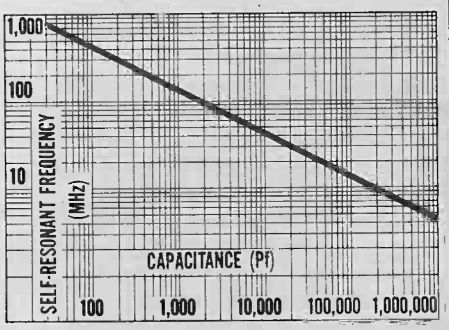


Fig. 4.

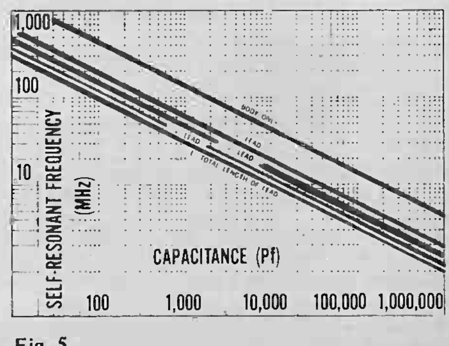
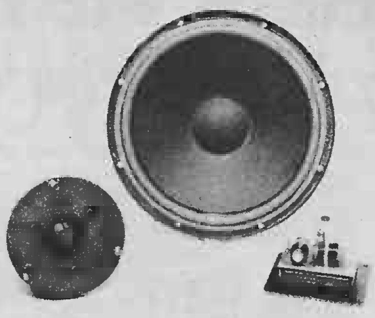


Fig. 5.

DIGEST



ETI PRICE DECREASE

Readers will have no doubt noticed (painfully!) the cover price increase on this issue of ETI. We apologise for this, but are happy to say it is ONLY FOR THIS ISSUE and the price returns to 75p with the May issue.

The one-month jump was made necessary by the sheer size of this special issue. We hope you will agree it is worth it. If you could see the price of paper these days...(moan, moan).

Thank you for sticking with us through thick and thin... (and 10p!)

Tempus Fugit

It's felt a little uncomfortable working in the ETI office this month; must be something to do with the sackcloth and ashes we're wearing. During the last few issues several of our reviews have featured Casio products, but we have consistently failed to credit the company which lent us the review models. The kindly folk in question are Tempus of 38 Burleigh Street, Cambridge CB1 1DG and we'd like to thank them for all the help they've been giving us. Tempus are leading Casio specialists and if there's something from Casio you're having problems obtaining, they will doubtless be as nice to customers as they are to us.

Sun-Day Driving

A Volkswagon Dasher car is presently being tested carrying a roof-rack of AEG-Telefunken solar modules which convert solar energy directly into electric current. The small 160 W 'solar power plant' of the test car complements the dynamo and charges the battery. This means that fuel consumption can be reduced by approximately five percent. As yet the cost of manufacturing these solar panels makes them uneconomical to use, but with the rising prices of fuel, it is foreseeable that low-priced solar generators will enter the market. Not only that, future car generations will make increased use of electricity, for example with automatic start-stop devices and pollution-free electrical energy for air conditioning in cars in warm countries. Great idea — but where will you put the luggage?

Tweeters That Go Cheap

Well, not just the tweeters, in fact. Mullard have a 40 W speaker system consisting of an 8" woofer as well as a high-power textile dome tweeter. They form part of a new low-price, two-way, self-build audio kit (whew!) being marketed by BK Electronics. The BK Electronics crossover unit have been combined with spring-loaded terminals and recessed mounting panel. The complete system, when built into the 23 litre enclosure, is capable of handling 40 W comfortably. All this for the small outlay of £13.90 plus VAT and £1.50 carriage per kit! Get yours now from BK Electronics Ltd, 37 Whitehouse Meadows, Eastwood, Leigh-On-Sea, Essex SS9 5TY.

Heading For The Top

Headphones seem to be getting lighter and smaller these days, so Sennheiser, that well-known manufacturer of headphones, has decided to launch a pair of their own lightweight 'phones. The new model HD40 is soon to be released in the UK and weighs only 60 grammes with extremely light contact pressure. They can be supplied with either a three or seven metre lead, the seven metre variety incorporating a volume control in the lead so that you don't have to march all that way back to the amp if it's too loud. Another feature is that each ear-piece can be revolved on the headband by 90 degrees if you have a funny shaped head or if you want to store them compactly (!). The Sennheiser HD40 will be launched in the UK with a suggested selling price, including VAT, of £16.55. For those of you interested in technical specs; frequency response is 22 to 18,000 Hz, impedance is 600 ohms, characteristic SPL is 90 dB and distortion factor <1.2%.

Electroware, OK?

OK Machine and Tool (UK) Ltd have launched a new division aimed at providing the electronics user with a really wide range of electronic hardware. All the products in the range will be available to everyone involved in building electronic equipment — that includes engineers, students, teaching staff, laboratory technicians and, not least, the hobbyist. The 40-page catalogue contains various products selected from OK's bench tool range — plus some new items — and includes soldering irons, wire-wrapping kits, IC tools, PCBs, cases, enclosures, connectors, sockets and test instruments to name just a few. Electroware is distributed throughout the UK by leading electronic and computer stores. Catalogues are free, but send 30p for postage and packing. If you want any further information or one of their catalogues contact OK Machine & Tool (UK) Ltd, Dutton Lane, Eastleigh, Hants SO5 4AA.

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.

PARTS LIST

Resistors (all ¼ W, 5%)

R1	47R
R2, R3	68R

Capacitors

C1, C3	3u3 16 V solid tantalum axial leads
C2, 4, 5, 6	10n ceramic block
C7, C8	10u 25 V tantalum
C9, C10	470u 35 V electrolytic (if required)

Semiconductors

IC1	LH0033CG
IC2	78L15A
IC3	79L15A
D1-D4	1N4001,2,etc. (if required)

Miscellaneous

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.

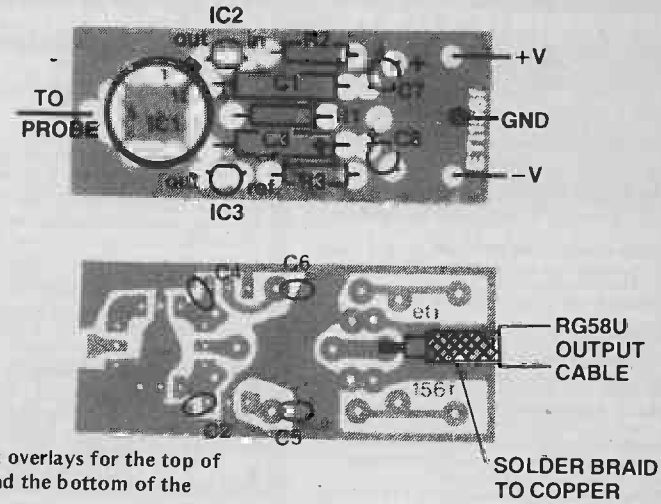


Fig. 2 Component overlays for the top of the board (top) and the bottom of the board (bottom!).

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.

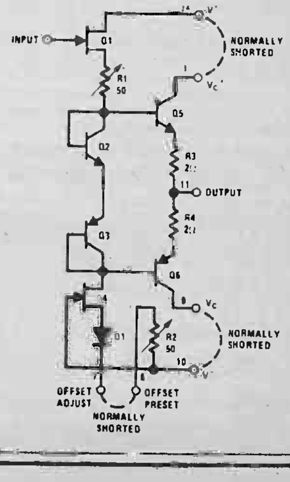
The internal circuit of the LH0033 is shown below. Basically, it consists of a FET input stage (Q1), operated as a source follower. The other FET, Q4, provides a constant current source for the source bias of Q1, while Q2 and Q3 are connected as diodes and provide bias for the bases of Q5 and Q6. Resistors 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.



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 44.

AUDIOPHILE

Soon burglars won't be bothering to nick your whole hi-fi; they'll just take the cartridge. This month Ron Harris reviews two new pickups, one with a gemstone cantilever and the other a work of modern art.

News just in of a new piece of British circuitry genius. This is a new protection circuit, soon to be added to a famous manufacturer's product, which is claimed to make an amplifier totally invulnerable electrically.

Totally in this case means "even from 240 V mains at input or output". Ultra-fast relays are set at the output and on the supply lines to the PCB. These are driven from the new circuit, which has as its final stage a voltage amp with an incredibly high slew rate. This ensures a high speed of operation for the relays.

Out Of Phase

The protection circuit operates like this: if an amplifier is suddenly faced with a massive input signal, the ratio of the feedback signal to input will drop dramatically. A comparator senses the change and a 'low-feedback' signal is generated. This by itself is sufficient to trip the supply relays, so that the overload cannot be passed on to the output stages, thus destroying them — and probably the speakers.

A second block within the circuitry watches the supply rails and any surges which are outside the requirements of normal drive will trip the protection circuit, since this is a "low-feedback likelihood situation" as the designer puts it. Great play is made of the fact that the music signal and the feedback voltage are in anti-phase at the point of comparison, so no interaction within the buffer is likely. 'Anti-phase reset', as it is called, thus introduces no colouration. Hence the protection reset of the relays can occur either in the case of low feedback-to-signal ratio, or in event of an "overload likelihood". I suppose this is where the somewhat pompous title of the circuit is derived — **A**nti-**P**hase **R**eset **I**n **L**ow **F**eedback (**O**r **O**verload) **L**ikelihood.

Shure MV30HE

A dedicated offshoot of the renowned V15 IV design, the MV30HE is for use in the SME Series III or IIIS only. The cartridge is built into a SME carryarm such that no headshell is used, or needed.

The moving components are those of the V15, save that no damper is provided. The cartridge body is all new, however, and quite a few problems it must have given them getting the coils and poles into a body as slim as this. The design is so arranged that the point of bearing intersection and the stylus line up parallel to the record. This will tend to aid stability in the replay of warped records.

As in the V15 a hyperelliptical stylus is used, which will give lower distortion results than either a spherical or elliptical tip. Tip mass is commendably low and output level is on a par with the V15 IV.

Once fitted into the SME the MV30HE looks very smart indeed is and visually extremely classy!

Testing an Armful

In the lab the MV30HE had an easy time passing just about every test. It tracks as well as the V15 IV and measures slightly better. There is no higher technical accolade than that. The LF resonance came out — surprisingly — at around 16 Hz, a little higher than optimum in my opinion. Best values are somewhere around 10-12 Hz so as not to affect extreme LF reproduction. Best tracking was obtained at around 1.0 g, and no improvement was forthcoming for increased force.

Frequency response was boringly perfect at 20 Hz — 20 kHz ± 1.3 dB with a separation figure of 27 dB at 1 kHz. Compliance measured very high at 34 cu, so only the smallest damping paddle is required. It *is* required however — see later.

Instructive Stuff

The instruction booklet is worth a special mention. It is a straight 'copy' of the SME style, right down to the little diagrams with ticks and crosses for right and wrong answers. Some sort of deal has been struck here, methinks!

One point that I just *have* to mention here; I could not,

At long last Quad have released their new tuner, the FM 4. It was shown for the first time at the Audio 82 exhibition in Swiss Cottage recently. Designed to match the Quad 44 control unit (preamp to the rest of us) the FM - only unit has digital tuning and seven pre-set stations. Programme locations are stored in memory.

A tuning knob has been retained in preference to a set of push-buttons, since Quad say it is easier to use.

Brief Specification:

Full limiting	1 V	IF Rejection	100 dB
S/N (1 V input)	7 dB (stereo)	AM Supression	60 dB
Distortion (1 KHz)	0.15%	Image Rejection	80 dB
Capture Ratio	2.5 dB	Crosstalk (1 KHz)	40 dB

The MV30HE set up in a Series III. About the best looking piece of of hi-fi you'll ever see. As the compliance is very high only the smallest damper paddle is required, despite the lack of dynamic stabiliser (as fitted to V15 IV).

in any circumstances, get the arm to balance out using the recommended mixture of weights and spacers in the SME. Despite Shure patiently showing me the same arm set up perfectly in *their* Series III, would it balance once I left the factory? It would not!

I would be most interested to hear from any MV30HE purchaser as to balance requirements. I agree the total mass is 11.8 g but I'll be damned to an eternity of music centres playing Barry Manilow before I'll agree that an MV30HE balances in an SME Series III with four weights and six spacers.

Down To Sound

I suppose the obvious comparison with the MV30HE must be the V15 IV. So, as I'm getting old and predictable (I *still* hunger after Felicity Kendal . . .) that was the first pickup against which I auditioned the unit. Frankly, I had expected to discern no difference, and initial tests confirmed this to some degree. However, having settled in at home with my own system around me and the reassuring brandy in hand, subtle differences began to manifest themselves. The MV30HE has a more coherent sound — the mid-range is more open under close examination and the bass is 'cleaned-up' and tighter, if a little more prominent. These are *exactly* the changes to be expected from a unit which simply matches the arm better, but has a higher resonance. Which, of course, if I'd thought about, is *exactly* the MV30HE/V15 IV relationship. Serves me right for being so sure of myself!

In comparison to the market as a whole, the MV30HE/SME pickup stands well-up with the best. The damper is not significantly missed, provided the SME paddle is employed. Leave it off and boom is liable to result, as is a certain lack of stability on warped records.

A limited application, then, but a very creditable performance and one which will compete with Shure's own V15 IV. After all, if you've got an SME and were contemplating a V15 IV, the MV30HE is a better bet all around. It is no more expensive than the V15 IV with a CA1 arm to hold it and it provides a cleaner, more refined performance. All in all, a nice touch Shure. Whither goest thou now?

Dynavector Karat Ruby

Both this month's cartridges are unusual in their own way; Dynavector's Karat is notable for its gemstone cantilever. This 2.5 mm long piece of single-crystal ruby is cut with a laser to accept the stylus (diamond) and then allowed to cool, thus fixing the stylus in place. The length is remarkably short, since Dynavector say that the less material the stylus information has to pass through, the higher will be the fidelity of the output.

Wave propagation through a medium is something not many of us take up as a hobby, but someone down at Dynavector must have it all well sussed! Apparently this equation:-

$$\frac{EI}{m}\frac{\partial^4 y}{\partial x^4}+\frac{\partial^2 y}{\partial t^2}-\rho\frac{EI}{m}\left(\frac{1}{E}+\frac{\gamma}{G}\right)\frac{\partial^4 y}{\partial x^2\partial t^2}+\frac{\rho^2\gamma}{mG}\frac{\partial^4 y}{\partial t^4}=0$$

$$C_B=\alpha\sqrt{2\pi f}\left[1-\frac{1}{4}\beta\frac{2\pi f}{\alpha^2}+\frac{1}{4}\delta(2\pi f)^2+\cdots\cdots\right]$$

where E = Young's modulus; I = secondary moment of section area; G = shear modulus; m = mass per unit length of a cantilever; ρ = density of the cantilever material; x = distance from the end of the cantilever; y = flexural displacement of the cantilever; r = constant; t = time.

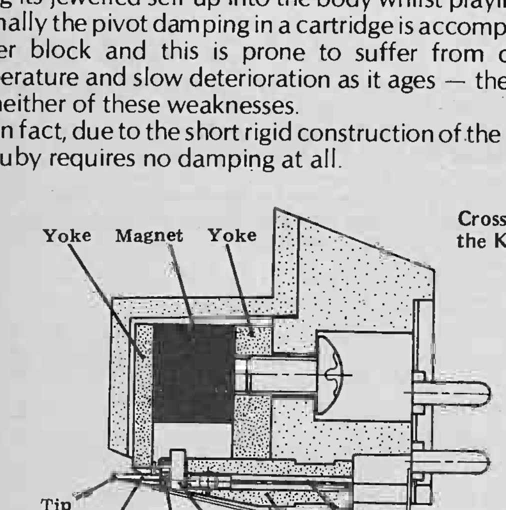
sums up the vibrational behaviour of a cantilever under dynamic conditions. It can also be used to prove that rigid materials, such as ruby and diamond, make for better cantilevers than boron, berylium and the rest.

(There is a 'big brother' to the Ruby, which has a diamond cantilever and costs around £450 as opposed to the Ruby's £100. If I can persuade the ever-helpful Dynavector into lending one I hope to report on the differences soon. Maybe if I say "please"... ?)

Temperate Zones of Test

Another piece of original thinking has gone into solving the problem of temperature dependence and damping material. The only rubber used in the Karat is to prevent the cantilever taking its jewelled self up into the body whilst playing records. Normally the pivot damping in a cartridge is accomplished by a rubber block and this is prone to suffer from changes in temperature and slow deterioration as it ages — the Karat suffers neither of these weaknesses.

In fact, due to the short rigid construction of the cantilever, the Ruby requires no damping at all.



Cross-section of the Karat Ruby.

Under test the Karat showed a ruler flat response from 100 Hz to 30 kHz of under ± 0.5 dB! It was only 1 dB down at 30 Hz and separation measured an excellent 24 dB at 1 kHz and a more than adequate 18 dB at 20 kHz. Stylus resonance fell at 49 kHz and in the SME Series III (what else?). LF resonance was well placed at 12 Hz, below audibility and above warps.

Tracking was exemplary for a moving-coil unit — at 1.75 g it tracked all my test bands perfectly; the first moving coil to do so. Bias was set for 2.0 g, a high value, but one that worked well. In actual use the Karat was never caught out by any recorded information.

If at this point you're looking around the pages in search of the usual response graphs, don't bother — I haven't included any. If you really want to see a straight line, go buy a ruler. Dishearteningly disappointing for us cynics.

Listening Out

As the Karat Ruby matches the SME Series III so well, it was left in that arm all through the listening test. One brief excursion into a Linn Itokk showed the two to be completely incompatable in my opinion, as the sound stage broke up and the bass became so loose as to be positively flapping! Strange that, as both are capable of much better and there is little on paper to point to such obvious mutual abhoration.

The loudspeakers used were my trusty KEF 105 II's fed by a variety of amplification from Crimson, Monogram and Trio. Source equipment remained at Thorens 160S/SME III throughout.

On the very first LP side I played with the Ruby it was obvious that here was something special. The sound is so detailed and open, with such tight control of the bass that it makes you sit up and take notice of the music. This is a cartridge that will be much appreciated by reviewers, as it is so easy to listen *through* for long periods.

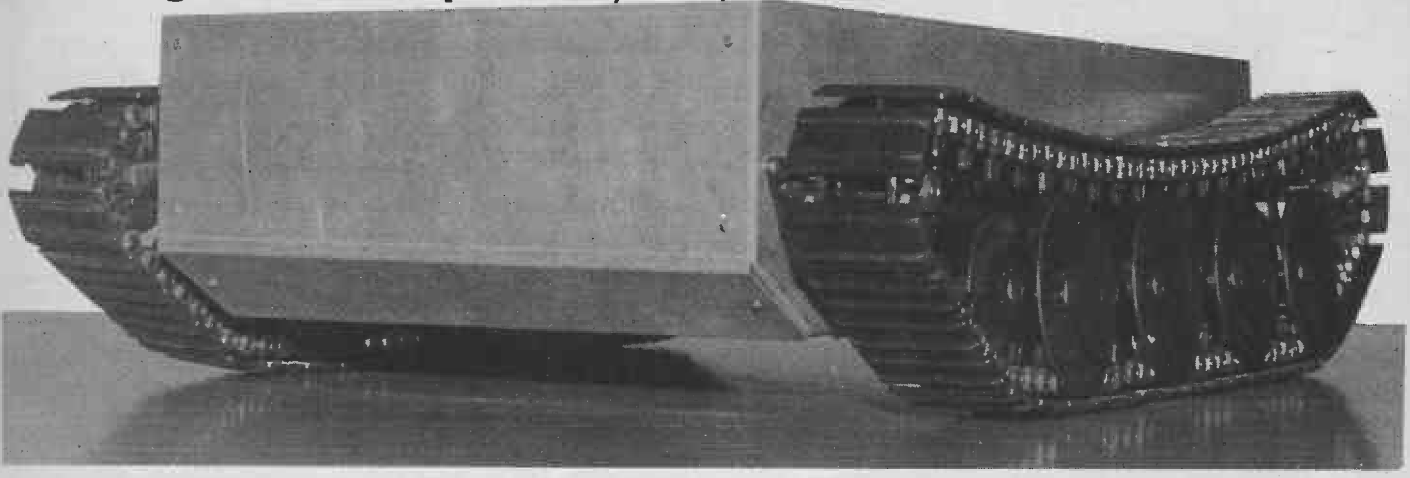
In fact there is little I can say against the Karat. It is a trifle recessed — I cannot account for this impression from the lab results, however, but it remains a definite impression — but is so relaxed and balanced a sound that none but the most obnoxious could find aught to quibble with. The sound quality reminded me greatly of the Ortofon MC30, but with greater resolution of complex passages and a more extended bass end.

At around £100 the Karat Ruby is an excellent bargain. Even accounting for the required step-up device, this pickup is required listening for anyone in the market. I have no hesitation in saying that it out-performs many units costing much, much more and will give more musical pleasure than just about any other cartridge I know.

Mind you, I haven't heard the Karat Diamond yet... but can it *really* be worth £350 more? On this evidence I would doubt it! (Pause while Dynavector work out whether this is a compliment or an insult...)

ROBOT MOTOR CONTROL

This month we feature a control board for last month's motor driving board. This is part 2 in a series of DIY robot modules — collect them all! Design and development by Rory Holmes.



In this second part of the series on the ETI intelligent programmable mobile we shall describe the design of an analogue pulse width modulator for controlling the motor driver stage featured last month. We shall also take a brief look at some of the modules being offered later in the series which can be added in stages to enhance the motorised vehicle. The intention is to build up to a complete computerised mobile.

A lot of flexibility has been allowed for in the actual use and configuration of the modules, as we are well aware that constructors interested in this type of project have firm ideas of their own on the final form and capabilities of their mobile. Construction and interconnection details for all the modules we are presenting will be given along with guidelines to a range of applications.

The facilities we have planned for the mobile will continue with the digital motor control and an on-board programmable computer for overall control of other modules. A light-weight manipulator arm complete with teaching arm has also been designed, for mounting on the front of the mobile. It is powered by four radio control servo motors and the electronics interface between the servos and computer will be described along with details of the arm mechanics. Optical proximity detectors for object sensing, and infra-red tachogenerators for speed sensing will also be featured on the ETI mobile.

It is hoped that the designs will also prove useful as stand-alone modules for individual use in other applications. Optical proximity detectors, for example, have numerous applications in batch counting, limit sensing, detection, alarms and so on.

The digital pulse width modulator in next month's issue will find many uses in the control of analogue functions; how about a computer interfaced to a pulse width modulated optical data link, for analogue information transmission? Our version will control two pulse width modulated channels, with a resolution of one part in 256, via an eight bit data port; modulation being achieved solely by logic to satisfy the all-digital purists.

Optical Proximity Detectors

These have been designed as small independent units with as much in-built versatility as possible. The circuitry is housed in a short length of aluminium tube axially aligned in the detector direction, with three external connecting points; ground, positive supply, and an open collector digital output. A number of detectors can thus be easily mounted in strategic locations. All circuit operating parameters are independent of the supply voltage, which can be anywhere between 5 and 35 V at a current of 20 mA.

The proximity switch works on the principle of transmitting and detecting a modulated infra-red beam. The infra-red transmitter receives 1 A peak current pulses, of 10 uS duration, with a modulation frequency of 1 kHz. The 100:1 duty-factor thus achieved allows high currents to be used to increase the detection range, while reducing the average supply current to only 10 mA.

The sensor can be set by a preset pot, accessible through a small hole, to detect an object at any distance in the range 1 cm to 35 cm.

A small amount of hysteresis is introduced into this switching distance to ensure clean switching thresholds and stability of the output signal. The use of tuned detector amplifiers provides excellent infra-red interference rejection.

Analogue Speed Control

The analogue speed control has

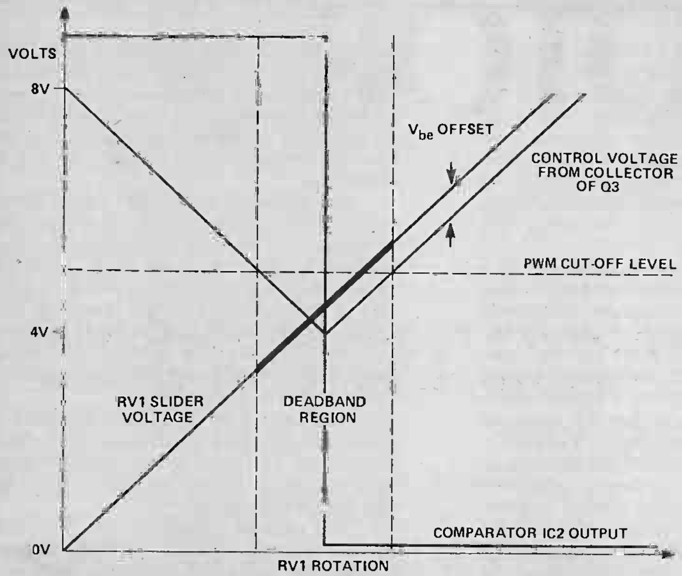


Fig. 1 Various voltages associated with the circuitry around Q3. The control voltage is measured at point A in Fig. 5.

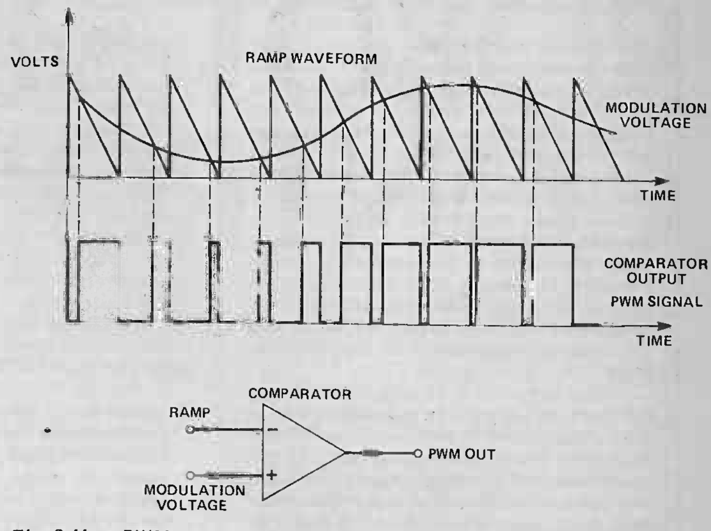


Fig. 3 How PWM waveforms may be generated using a comparator.

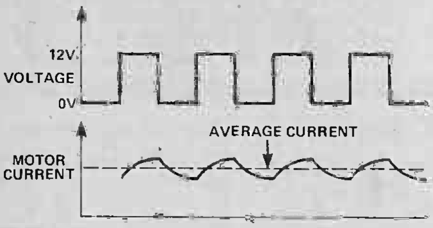


Fig. 2 PWM motor driving waveforms for last month's circuit.

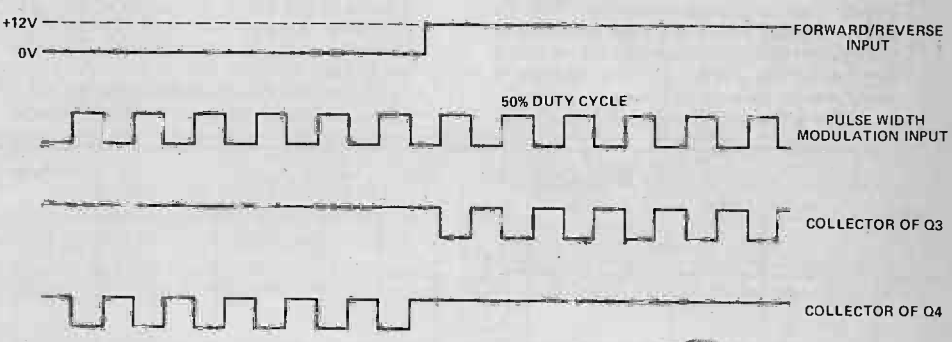


Fig. 4 The waveforms needed by our motor driver board, published last month. (Q3 and Q4 refer to last month's circuit.)

been devised for manual control of the main traction motors; it provides two pulse width modulated signals suitable for the motor driver amplifier.

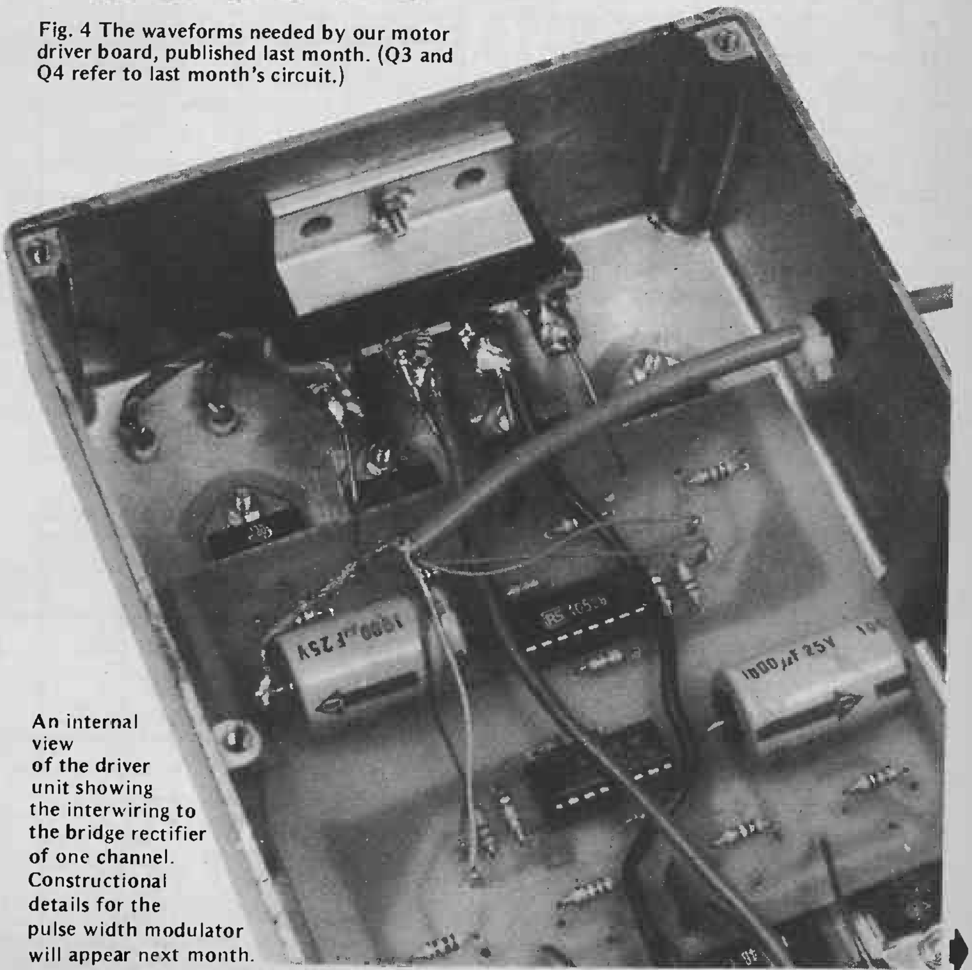
The circuit is designed to provide a linear control-voltage-to-pulse-width relationship for greater flexibility in application, and to simplify the addition of speed feedback velocity control.

The modulator can be built either single or dual, and the manual control section, if not required, is easily omitted. Speed control is achieved via two remote potentiometers, allowing speed to be set in either forward or reverse directions independently for each traction drive.

Since both motors are controlled via switching amplifiers from the same battery supply, it is important to reduce the peak currents that are drawn. This can be achieved by offsetting the phase of the switching waveforms relative to each other, such that at 50% duty cycle modulation, power

BUYLINES

No problems here with any of the components specified — most mail order companies who advertise in the magazine will be able to supply everything. We can supply the PCB — see page 44 for details.



An internal view of the driver unit showing the interwiring to the bridge rectifier of one channel. Constructional details for the pulse width modulator will appear next month.

HOW IT WORKS

The circuit for the dual analogue pulse width modulator is shown in Fig. 5; it will be seen that each channel is identical with the exception of the circuitry around the CMOS gates IC1 and IC4. As described earlier the two switching waveforms must be the same frequency and synchronized 180° out of phase, to distribute the motor current peaks more evenly through the cycle. This is achieved by synchronizing both pulse generators to a master clock based around IC1a and b. A 20 kHz square wave is generated by this conventional astable arrangement and its frequency, set by R1 and C1, is fairly independent of supply variations.

The output of IC1d at pin 6 provides a buffered square wave in the same phase as the output on pin 10 of IC1b. C2 and R3 differentiate the positive-going edge of the square wave to produce a very short logic low pulse at the output of Schmitt inverter gate IC1c. In similar fashion C9 and R16 produce a logic high pulse coinciding with the negative-going square wave edge. IC4b further inverts this signal to a logic low pulse. Two separate trains of 500 nS negative-going pulses are thus provided in the correct phase relationship for resetting the charging cycle of two sawtooth oscillators as described below.

The pulse width modulators are identical from here on and we shall refer to the topmost circuit for description. Voltage controlled pulse width modulation is, in principle, very simple; a ramp waveform (sawtooth) is applied to one input of a comparator and the modulation voltage to be encoded is applied to the other, producing the required PWM squarewave at the comparator output. Figure 3 illustrates this operation.

Due to the design requirement of a linear relationship between control voltage and pulse width, a constant current source formed from Q2 is used to generate the linear ramp waveform. LED1 and the base-emitter junction of Q2 are forward biased by R6 and together define a temperature-compensated voltage across R7 which in turn defines a constant emitter and collector current of about 1 mA. C3 is charged up negatively from this current, until the negative-going reset pulse arrives from inverter IC1c. This pulse turns Q1 hard on for a very short period (500 nS), during which C3 is completely discharged, taking the ramp voltage back to +8 V. This process repeats at the clock frequency of 20 kHz, providing a negative-going sawtooth of about 3 V peak-to-peak referenced to the +8 V rail.

IC3b, the comparator used to perform the modulation, is an LF353 dual op-amp, chosen for its large bandwidth and high slew-rate. The inverting terminal on pin 2 is fed from the ramp waveform, while the non-inverting terminal is fed from op-amp IC3a, an inverting amplifier configured to sum control voltage inputs relative to a 4 V reference.

The potential divider R11 and R12 provides the 4 V reference to the non-inverting terminal of IC3a, and the control voltage applied to R13 at point A is summed relative to the 4 V. An offset voltage set by PR1 is also summed at the inverting terminal of IC3a, and is used to bring the control voltage into the correct operating range and for setting a deadband region on the manual control pot RV1.

The output of op-amp IC3b (and indeed most others) will not swing to the full supply rail voltages, so the inverter gate IC1e is used to buffer the square wave to full CMOS logic levels.

The manual control system included in this circuit enables a single potentiometer to control the speed in both forward and reverse directions. When the pot is at centre travel, and for a certain deadband around this point, the motor must be stopped and no switching pulses should occur (ie the PWM signal is continuously low). As the pot is turned in either direction from its midpoint, the pulse width should in-

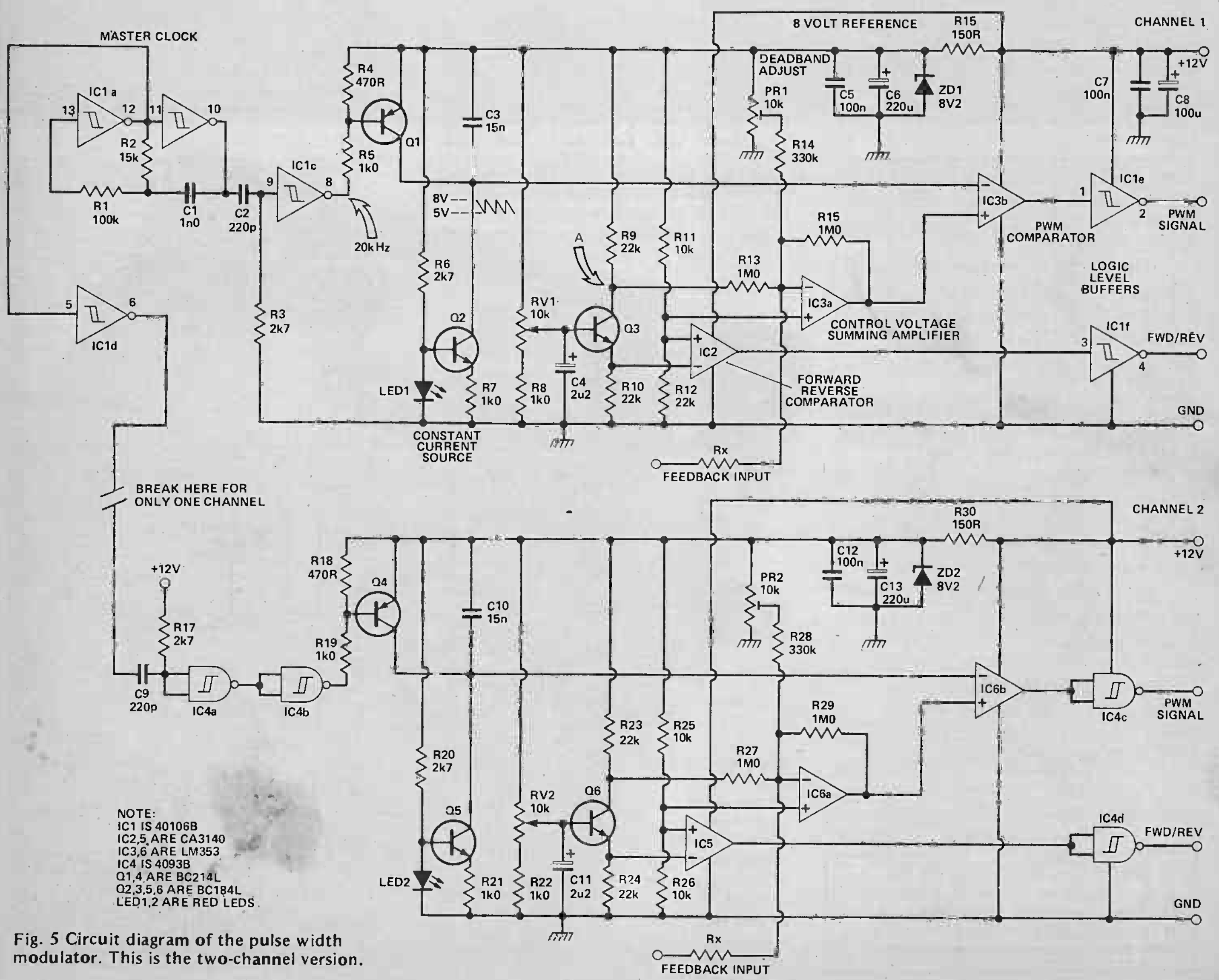


Fig. 5 Circuit diagram of the pulse width modulator. This is the two-channel version.

crease and this requires a positive-going input voltage to the summing amplifier IC3a. The forward/reverse logic level should also change state as the pot moves through its midpoint. Q3 provides the necessary voltage transfer function from the pot RV1 to the control voltage summing amplifier, as explained graphically in Fig. 1.

The emitter and collector resistors of Q3 are both equal and the base voltage is taken directly from the slider of the manual control pot RV1. The output voltage is taken from the collector of Q3 to feed the summing amplifier, and will be held at +8 V via R9 when Q3 is switched off. As the slider of RV1 moves toward the centre of travel, the base voltage rises, slowly turning on Q3 and lowering the collector voltage.

When Q3 is turned hard on as RV1 reaches its mid-point, R9 and 10 will form a potential divider giving 4 V as the minimum control voltage. Further increase of base voltage can now only increase the emitter and collector voltages back up to the positive rail, reaching a maximum at one V_{be} drop from the +8 V rail.

During the above process the voltage on the emitter of Q3 rises from zero to the same maximum voltage, and is fed to the inverting terminal of IC2, a CA3140 used as a comparator. The other comparator input receives 4 V derived from the potential divider R11 and R12. This provides the required forward/reverse signal that corresponds to each half of the control pot. Inverter gate IC1f buffers the output of IC2.

C7 and C8 provide supply decoupling for both channels, while C5 and C6 provide further smoothing for the 8 V zener regulator formed by R16 and ZD1. This 8 V reference rail is used for two reasons; firstly to allow for fluctuation in the 12 V battery power supply that would otherwise affect the output pulse width, and secondly to ensure that the op-amp supply voltage is well above the maximum input voltage.

The resistor marked as Rx in the circuit shows where a speed feedback voltage will be added to the controller to close the velocity control loop. An infra-red tachometer module to directly sense the traction speed will be described later in the series.

If the manual control input is not required, the components associated with this can be simply omitted (ie RV1, R8, R9, R10, C4, Q3, IC2 and their equivalents in the other channel). Control voltages may now be fed to the unconnected end of R13, where a variation of 3 V, set by PR1 to be anywhere in the range 0 V to 8 V, will provide 100% control of the output pulse width. Forward/reverse switching must also be applied to the input of IC1f on pin 3.

will be switched alternately to each motor. This spreads the current peaks more evenly over the switching cycle.

Construction and setting up with interconnection details for the motor driver will be described next month.

PARTS LIST

Resistors (all ¼ W, 5%)

R1	100k
R2	15k
R3,6,17,20	2k7
R4,18	470R
R5,7,8,19, 21,22	1k0
R9,10,23,24	22k
R11,12,25,26	10k
R13,15,27,29	1M0
R14,28	330k
R16,29	150R

Potentiometers

RV1,2	10k linear
PR1,2	10k linear miniature horizontal preset

Capacitors

C1	1n0 ceramic
C2,9	220p ceramic
C3, 10	15n polycarbonate
C4, 11	2u2 35 V tantalum
C5, 7, 12	100n ceramic
C6, 13	220u 16 V axial electrolytic
C8	100u 25 V axial electrolytic

Semiconductors

IC1	40106B
IC2,5	CA3140
IC3,6	LF353
IC4	4093B
Q1,4	BC214L
Q2,3,5,6	BC184L
LED1,2	red LED
ZD1,2	8V2 400 mW zener diode

Miscellanous

PCB (see Buylines)

ETI

SOLID STATE REVERB UNIT

Where have all the spring lines gone? Gone to lesser projects in other magazines, that's where. Meanwhile we present this cheap, simple, but high-quality unit using solid state technology. Design by Charles Blakey.

At last — a reverberation unit which is not a pseudo echo effect and does not suffer from the defects of spring line devices. The unit described below will interface with virtually any preamplified signal and is ideal for direct use with most musical instruments or for incorporating in the 'echo-send' line of mixers. The design has been made possible by a new 3328-stage bucket brigade device having six tapped delays and capable of producing a useful reverberation time of about three seconds.

Sound emitted in an enclosed space will be subjected to both simple and multiple reflections from internal surfaces. Since these surfaces are at varying distances, the time for these reflections to occur and then decay by absorption will vary. The effect is a build-up of sound known as reverberation. When playing a musical instrument in the home, small studio or some other venue, the decay time can be very small coupled with a high absorption loss; the result is a weak sound when compared to recorded music or to live music played in a large hall.

Until now the only low-cost method of simulating acoustic reverberation has been the use of spring lines. These units, however, are prone to vibration, require a high power consumption for effective driving and are prone to producing distorted resonant peaks. Furthermore it is not possible to adjust the reverberation time and in many instances a short reverberation can be very effective. Another option has been available for some years, namely, the use of bucket brigade devices to electronically delay signals. While claims have been made for reverberation effects based on these products, a realistic unit would require at least three dual 512-stage BBDs, such as the Reticon SAD1024A. The cost and complexity of the latter approach puts it beyond the reach of the average constructor.

Beyond The Pail

The reverberation unit utilises the MN3011, which is the latest in a series of bucket brigade devices for audio applications to come from National Panasonic. They are all fabricated in PMOS and for a start you can forget most of what you may have read about the disadvantages of PMOS BBDs. It is a fact that they are somewhat limited in clocking speed (10 kHz to 100 kHz) and also have a limited bandwidth, typically 10 to 12 kHz. The latter, however, is not usually a limitation since the bandwidth is often restricted by the desire for long delay times. What makes the series ideal for audio applications is their low insertion loss, low distortion and excellent signal-to-noise ratio and for the MN3011 the specified values are 0 dB, 0.4% and 76 dB respectively.

The IC is unusual in that it has 12 pins but is the length of a normal 18-pin package; the functional block diagram and pinout for the MN3011 is shown in Fig. 1. As is normal with such devices it requires two power supplies, V_{DD} and V_{GG}; the former may be up to -18 V with respect to ground while V_{GG} should be $+1$ V higher than V_{DD}. Bucket brigade, or charge coupled, devices are analogue shift registers which operate by sampling the input signal at a rate determined by an external clock. The signal level at the time of sampling is stored on an internal capacitor; this charge is then clocked down a series of capacitors by means of internal switches. The transfer process is accomplished by a dual clock whose outputs are in antiphase and so are alternately opening and closing adjacent switches. It will be apparent that the slower the clock speed the longer the delay. Since the devices operate at high clocking speeds the input signals are faithfully reproduced at the output.

The most interesting feature of the

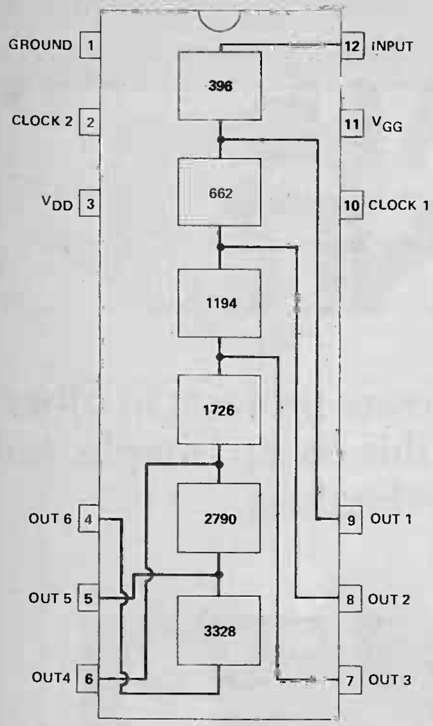
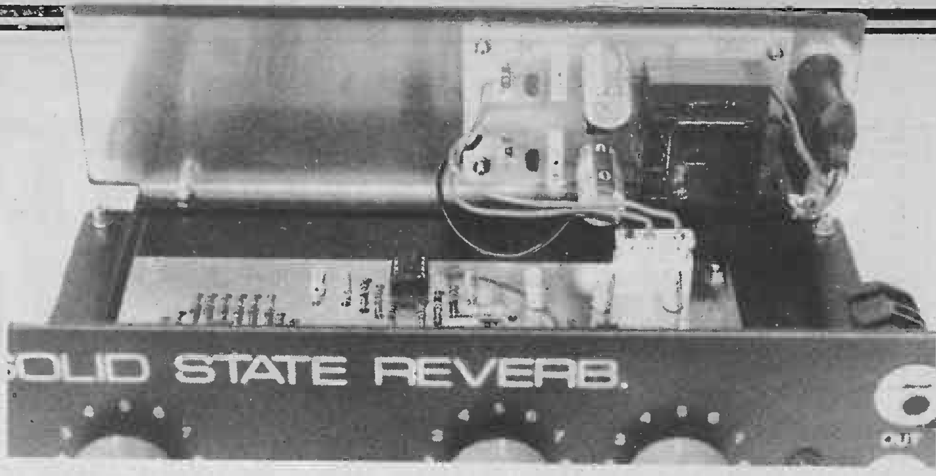


Fig. 1 Pinout and internal layout of the MN3011. The centre three pins on each side of this 18 pin package are absent.



MN3011 is that it has six tapped delays and Fig. 1 shows the number of stages for each tapping. The tappings are not evenly spaced since otherwise the reverberant sound would have a distinct flutter. If the device was being clocked at 10 kHz then the delays from outputs one to six would be 19.8, 33.1, 59.7, 86.3, 139.5 and 166.4 milliseconds respectively. If these delay times are mutiplied by 0.33 then one obtains the equivalent room path length for one trip, ie the longest delay is equal to a room length of 55 metres (181 feet). Reverberation time is usually measured as the time taken for the power to decay to one millionth of its initial level (60 dB down). For the present design the time was measured for the output level to fall to one hundredth of its initial level (−40 dB) and at the longest delay this was found to be about three seconds.

Blocks 'n Clocks

The block diagram of the circuit for the reverberation unit is shown in Fig. 2. First there is the dual clock driver, which is another National Panasonic device, the MN3101. It has an oscillator, divider and wave form shaping and produces the dual clock pulses required by the MN3011. It reduces component count and is lower in cost than other alternatives, such as a 4007. A further advantage is that it also generates the required V_{GG} voltage.

The unit will operate satisfactorily with any input signal greater than 280 mV RMS and higher input signals are attenuated by the input potentiometer. The signal is also reduced by half in amplifier A1 and inputs higher than 140 mV to the first filter are indicated by a LED peak detector circuit. Although the MN3011 will accept signal levels up to 780 mV before the distortion value stated earlier is exceeded, it will become apparent that the effect of reverberation can lead to reinforcement of signals and consequently this has to be allowed for. The only preset in the circuit is used to apply a bias voltage to the signal. The precise value of this voltage is not very critical in the current design and the object is to keep the signal at a level where it will not be distorted or clipped within the BBD.

The main problem with BBDs is the inability to completely cancel out the clock pulses and these can form audible cross products with the input signal. In order to prevent this foldover distortion, the bandwidth of the input signal should be limited to between a half and a third of the clock frequency. Filter F1 in Fig. 2 is a lowpass filter with a cut-off frequency of 3.6 kHz. This may seem rather low but in fact it is equivalent to the upper reverberation limit of most spring lines and the BBD scores in respect of low frequency responses since springs usually give rise to 'booming' below 100 Hz. The limited bandwidth is compensated by mixing the original signal with the reverberated signal at the output stage. The filtered signal goes to the MN3011 and the six output stages are summed to give a composite signal with different delay times. This signal is again filtered with a lowpass filter with a cut-off frequency of 3.6 kHz, to remove residual clock glitches, prior to mixing with the original signal at the output amplifier, A2.

The most important feature, however, is that the signal from the longest delay is returned, slightly attenuated, to the input and subjected to further delays. This is the reverberation effect and with the times given earlier the sound will simulate the effect of the first reaching a surface 55 metres away (assuming slowest clocking rate) and then being reflected back as well as being reflected from other surfaces closer than the 55 metre surface. The whole process is repeated until the original delayed signal and its reflections die away. In the meantime new signals are being recycled and the overall effect is a build-up of sound — reverberation.

Construction

The construction is very straightforward but the following precautions should be observed. First,

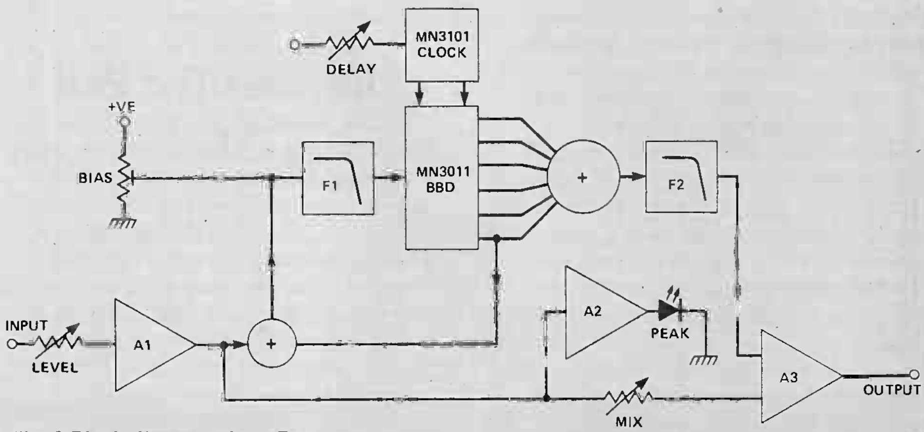
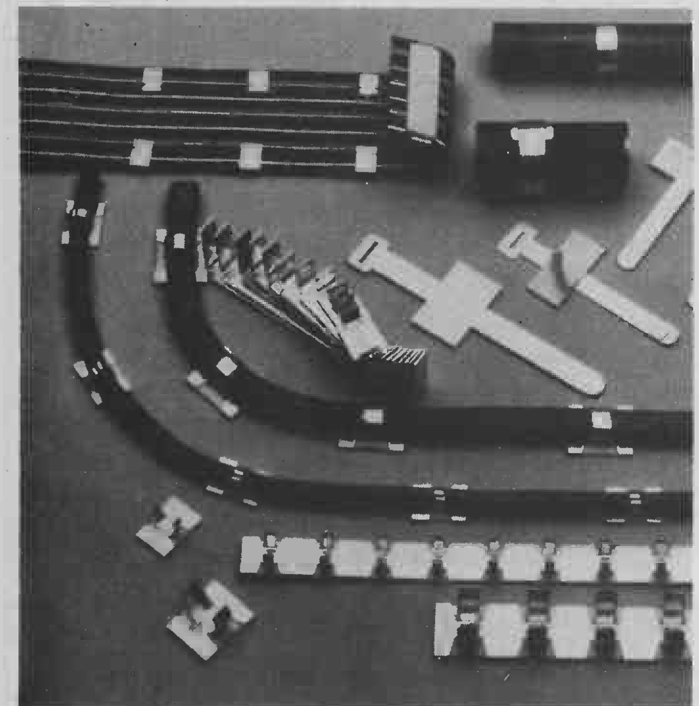


Fig. 2 Block diagram of the ETI Solid State Reverberation unit.

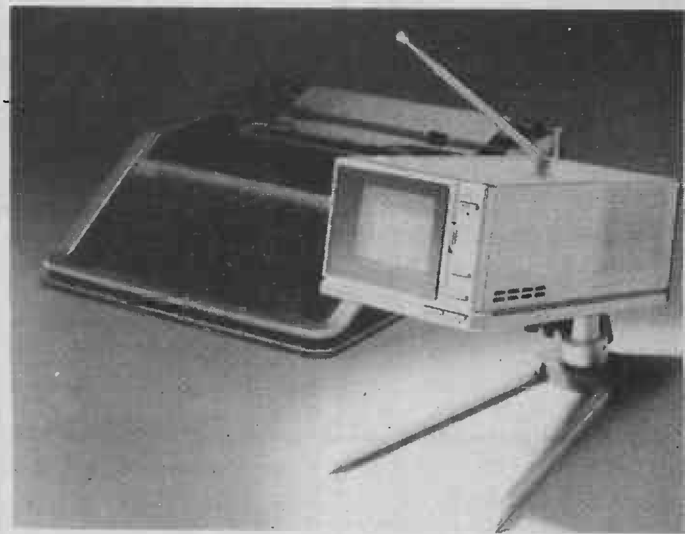
High-res Printing

New from Hi-Tek is the Facit 4542, a high-speed, high-resolution printer which combines a new type of 'Flexhammer' print-head with advanced microprocessor control to make it equally suited to text printing, label or bar code production, and graphics output. Using 260-character-per-second bidirectional two-colour printing and a 14 x 9 dot-matrix format, the 4542 can produce a virtually unlimited range of characters as well as different grey scales in graphics applications. In normal text-printing applications, the 4542 features proportional spacing, justified right-hand margin and an extensive set of up to 512 characters in 11 national repertoirs with red/black, elongated and underlining facilities. For label printing, a variable-size option is available which allows characters or bar codes to be generated in 95 different sizes from 2.52 mm up to 240 mm. Selection of size and position is easily controlled by software commands. In the graphics mode, scanning, semi-graphics and 10 levels of grey/red scale are available to illustrate reports with histograms, curves and diagrams, as well as generating half-tone illustrations in applications such as tomography, process monitoring and computer-aided design. The key to the versatility of the 4542 is the print-head, which consists of a set of nine stored-force flexible metal hammers mounted directly on a magnet armature. No adjustment or lubrication is necessary, wear is minimal, and a 'floating' mount means that the correct paper/print-head distance is always maintained irrespective of the paper thickness or number of copies. Further information is available from Hi-Tek Distribution Limited, Trafalgar Way, Bar Hill, Cambridge, CB3 8SQ.



Small And Beautiful

Hailed as 'the World's Smallest, Lightest and Lowest Power Consumption' television, the TH3-W3V from Matsushita certainly caught our editorial eye. Closer inspection revealed a colour TV set with a 3" colour picture tube, only 115 mm x 86 mm x 323.5 mm in size and 1.5 kg in weight. Power consumption is a mere 9.5 W and it operates on AC power, car batteries and on optionally available rechargeable batteries. Yet, despite its small size, it is equipped with video input/output terminals and operates as a colour monitor and a video tuner when connected to a video camera and a portable VTR, respectively. This 3" colour TV was launched on to the Japanese market in mid-December 1981 at the approximate price of £200. It is due for launch in the US in June this year and, hopefully, will be seen in this country shortly after. Further details will be supplied by National Panasonic (UK) Ltd, 300/313 Bath Road, Slough SL1 6JB.



Sticky Clips

Brandauer adhesive cable clips from Stotron provide an inexpensive method of fixing round or ribbon cables to clean, dry surfaces. The range can handle round cables from just a few millimeters up to 19 mm and flat ribbon cables from 13 mm to 75 mm can be accommodated by a selection of clips with widths in stages of 6 mm. The adhesive is instant acting and polyethylene pads provide high levels of insulation, where necessary. Further information is available from Stotron Ltd, Unit 1, Haywood Way, Ivyhouse Lane, Hastings, East Sussex.

Video Victory

Thorn EMI have just announced that agreements have been signed with Telefunken and JVC to form a holding company for the manufacture of video consumer electronics products in Europe. Thomson-Brandt was originally intended as a fourth partner, but this was not possible. However, the three other parties hope an opportunity will arise for Thomson-Brandt to join the venture.

Products manufactured by the joint venture will include VHS video cassette recorders, VHD video disc players and video cameras.

BT Bill Beater

Following the success of the Telcost TNA25 from the Ansafone Corporation, it was decided that a single line unit should be manufactured. The new machine offers a range of functions which are all designed to save money by monitoring telephone use. Ansafone's single line Telcost 1 has features including a 24-hour clock display, which instantly shows the cost of a call as soon as a user is connected with a number dialled. The unit also has a built-in printer which records details of the call including cost and number dialled. It also prints out the date, time, machine identification number and the duration of the call. Telcost 1 has a built-in memory which retains information even if the machine is disconnected from a power source. It also gives a special security midnight printout each night which frustrates any attempts to conceal the day's telephone costs by the destruction of the daily printout sheet. The machine is virtually tamper-proof as the printout will indicate if it has been disconnected from the line at any time or if any information parameters have been changed. The machine has provision for it to be reprogrammed at any time to enable the user to keep in line with British Telecom unit rate charges and the date, time and identification number can be changed for any reason if the machine is moved to a new location. This desk-top unit is no bigger than a telephone and for an investment of around £249 could help to cut out the abuse of telephones in both large and small companies.

BUYLINES

The PCB and a kit of components for the reverberation unit is available for £32.00, inclusive of postage and VAT, from Digisound Limited, 13 The Brooklands, Wrea Green, Preston, Lancs PR4 2NQ. The power supply may also be obtained for an inclusive price of £7.00. As the PCBs are copyright they will not be available from our PCB Service; however, the foil patterns are reproduced at the back of the magazine. National Panasonic do not distribute active components in the UK and the ICs may *only* be obtained from Digisound.

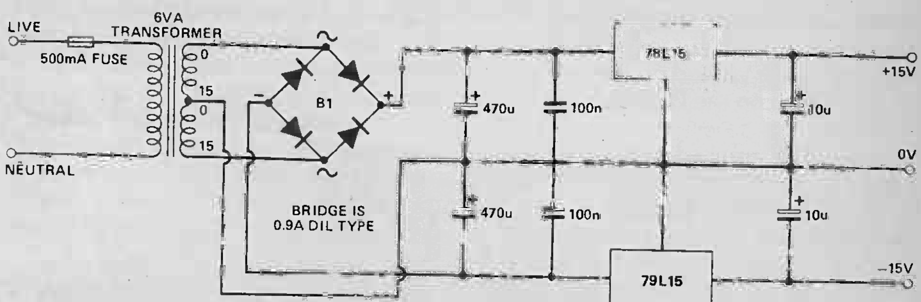


Fig. 4 Circuit diagram of a suitable PSU for this project.

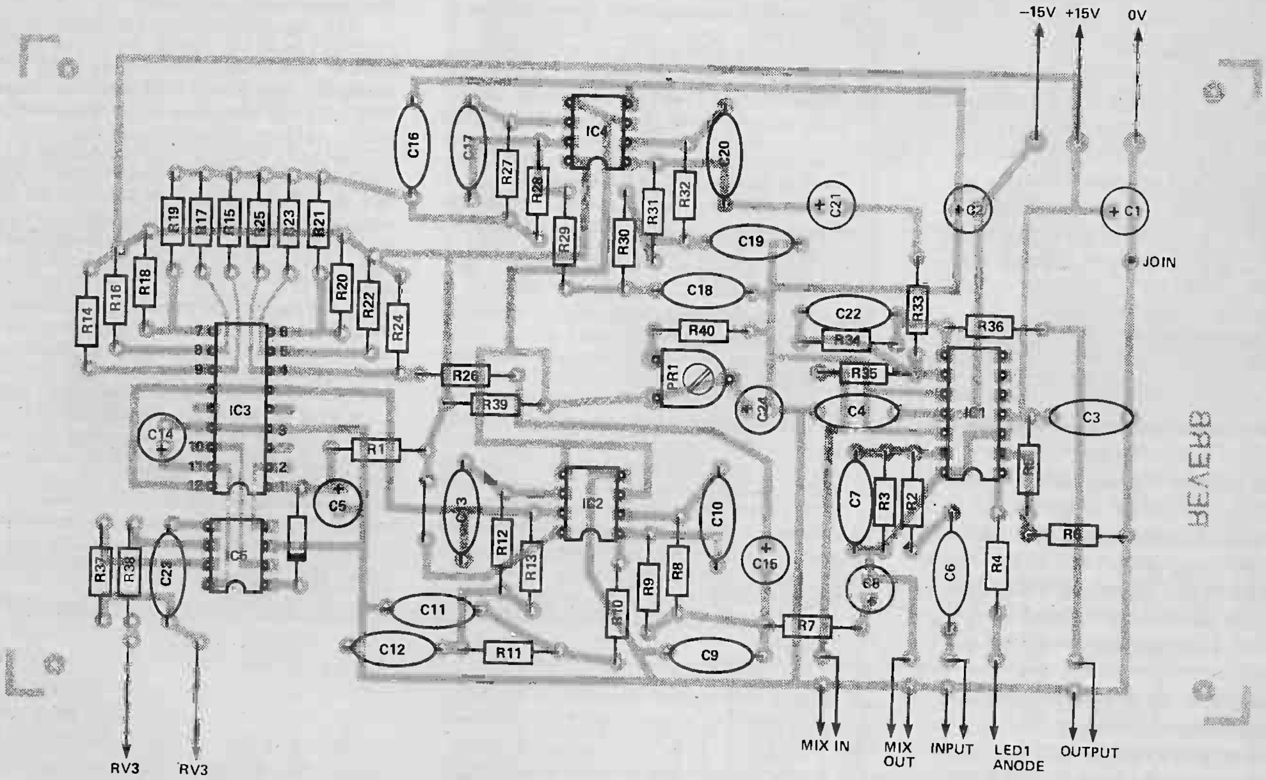


Fig. 3 Component overlay for the reverberation unit.

make sure you get the correct orientation of the ICs which are clearly shown on the component overlay. Second, the MN3011 is a CMOS device and with the advent of 'B' series devices we have all become rather careless as regards handling such ICs. For the MN3011, however, take the precaution of working on a grounded

PSU PARTS LIST

Capacitors

C1,2	470u 35 V PCB electrolytic
C3,4	100n polyester
C5,6	10u 35 V PCB electrolytic

Semiconductors

IC1	78L15
IC2	79L15
BR1	0A9 DIL type

Miscellaneous

PCB (see Buylines); PCB-mounting transformer (15-0-15, 6 VA); 500 mA mains fuse and chassis-mounting holder.

PARTS LIST

Resistors (All ¼ W 5% except where stated)

R1	10R ½ W
R2,5,7,9, 13,32,33,39	100k
R3,34	51k
R4	330R
R6	1k3
R8,12,27,31	33k
R10,29,37	47k
R11, 30	56k
R14,16,18,20, 22,24	56k 1%
R15	100k 1%
R17	110k 1%
R19	120k 1%
R21	130k 1%
R23	150k 1%
R25	160k 1%
R26	200k
R28	82k
R35	18k
R36	1k0
R38	36k
R40	68k

Potentiometers

RV1	100k logarithmic
RV2	10k logarithmic
RV3	470k linear

PR1	47k miniature horizontal preset

Capacitors

C1,2	10u 35 V PCB electrolytic
C3,4	100n polyester
C5	22u 35 V PCB electrolytic
C6	220n polyester
C7,10,13, 20,22	220p polystyrene
C8,14,15, 21,24	3u3 63 V PCB electrolytic
C9,11,12, 18,19	2n7 polystyrene
C16	2n2 polystyrene
C17	270p polystyrene
C23	33p polystyrene

Semiconductors

IC1	TL074
IC2, 4	LM358
IC3	MN3011
IC5	MN3101
D1	1N4148
LED1	5 mm red LED

Miscellaneous

SK1,2	mono jack sockets

PCB (see Buylines); IC sockets; case (Vero order no. 91-2673G).

metal surface, such as a piece of aluminium foil, do not insert the IC with the power on and do not use a soldering iron on the PCB with the IC installed.

The PCB supplied with the kit has a ground plane to reduce interference from and to other electronic equipment as well as to reduce noise. This feature allows greater freedom in locating the unit, eg it does not have to be housed in a separate metal case. A ground plane comprises a metallized surface on the component side except for small areas around the holes for the components. Ensure that the component leads do not touch the ground plane — which is not difficult — and preferably solder the resistors and axial capacitors in place with a thin piece of card between the component and the board so that the former are not in physical contact with the ground plane. After soldering the card is removed. The latter step is not essential. The one wire link must be made with insulated wire. The ground plane has to be connected to the 0 V line and some 15 mm from where the latter is connected to the PCB there is a hole marked 'join'. A piece of wire should be placed through this hole and soldered on both sides of the PCB.

The PCB has been laid out such that the BBD and clock are as far away as practical from the signal input and output. This separation should be maintained if the unit is housed in a box and all wiring should be kept as short and as neat as practical, with the audio connections being made with miniature screened cable.

The unit requires a ±15 V power supply and the current consumption is a miserly 13 mA at +15 V and 9 mA on the −15 V line. If a separate power supply is required then a suitable PSU is shown in Fig. 4. A PCB-mounted transformer is preferred, and it should be mounted as far away from the BBD as practical. The photographs show the unit inside a Vero 'G' range case with internal dimensions of approximately 218 x 138 x 50 mm.

HOW IT WORKS

The input signal is attenuated by RV1 and also by the inverting amplifier built around IC1a which has a gain of about 0.5. From IC1a the signal goes three ways. A comparator built around IC1b forms a peak detector to indicate optimum signal level, while RV2 and R35 allow mixing of the original signal with the reverberated signal in the inverting amplifier configured around IC1c. The component values in this section are such that equal proportions of the two signals may be mixed. Finally the signal also passes to two active filters constructed around IC2 which have a 12 dB/octave roll-off for each stage and a cut-off frequency of 3.6 kHz.

From the above filter stages the signal passes into the MN3011 and the six delay outputs are summed by the resistor network formed by R14 to R25. Note that the shorter the delay, the less the attenuation. From the longest delay (pin 4) the signal goes via R25 back to the input of the filter and thus provides recycling of the delayed signal in order to generate a true reverberation effect. The reverberated signal is filtered by two active filters constructed around IC4 and these have the same characteristics as the input filters. Between the active filter stages some passive filters have also been added to increase the roll-off; the loss in these filters is compensated by increasing the gain of the active filters.

The dual clock for the MN3011 is provided by IC5 and with the components shown, the clock frequency may be manually varied with RV3 over the range 10 kHz to 100 kHz, allowing maximum first pass delays from 16.64 to 166.4 milliseconds. Pin 8 of IC5 provides the V_{GG} voltage for the MN3011. Since both IC3 and IC5 are P-channel CMOS it would be normal to operate them from a −15 V supply. Voltages are, however, relative and by connecting +15 V to the ground pin and ground (0 V) to the V_{DD} pin they will operate happily with positive signal inputs. R1 and C5 prevent clocking signals getting back into the power lines. The filters are also operated from a single +15 V supply and this avoids any problems which may arise from excessive bipolar signals, ie they will be clipped at +15 V or ground and not damage the BBD. The bias voltage required by the BBD and the filters is primarily to allow them to accept bipolar signals; this voltage is provided by the resistive divider using components R39, PR1 and R40 and is applied to the non-inverting input of the filter op-amps.

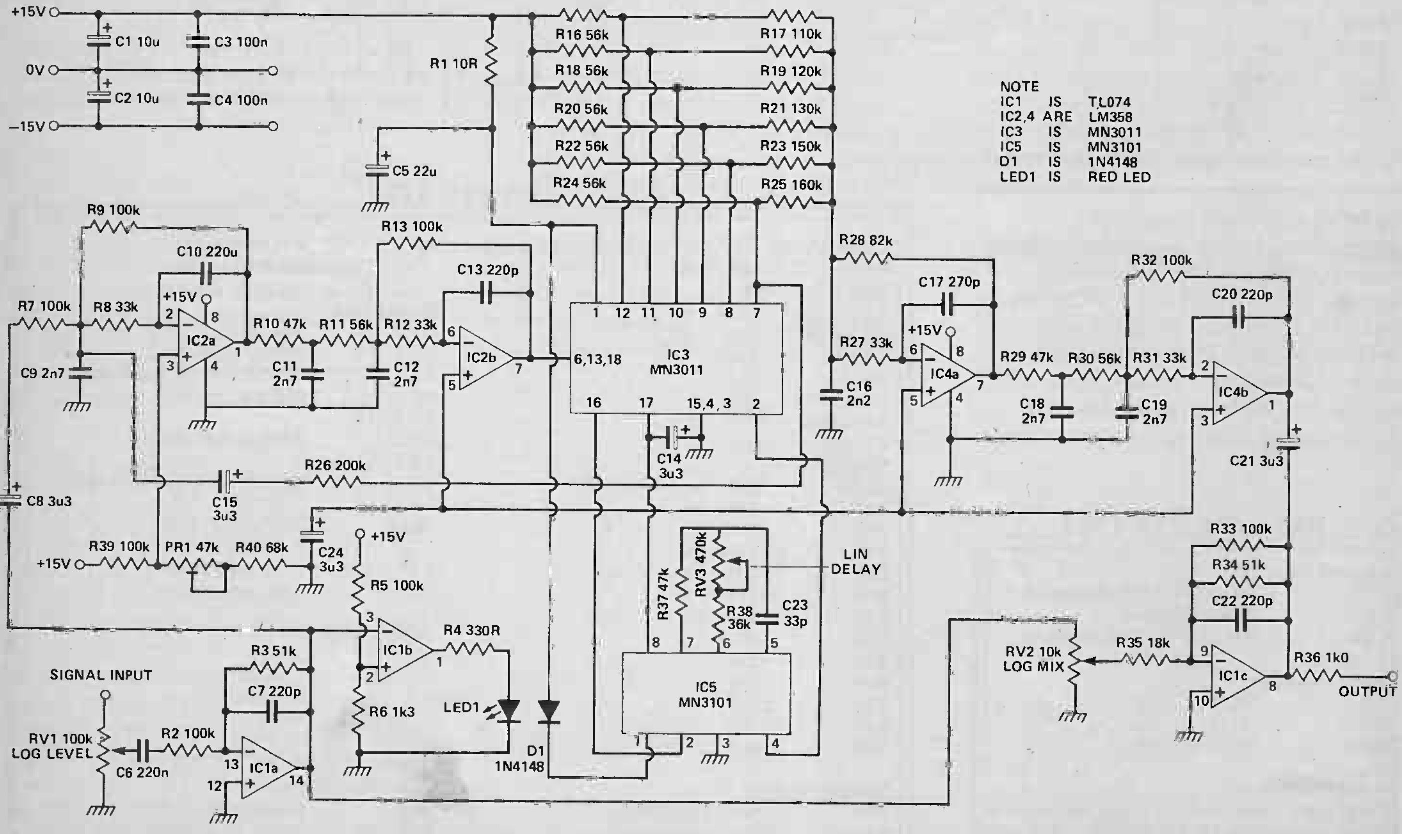
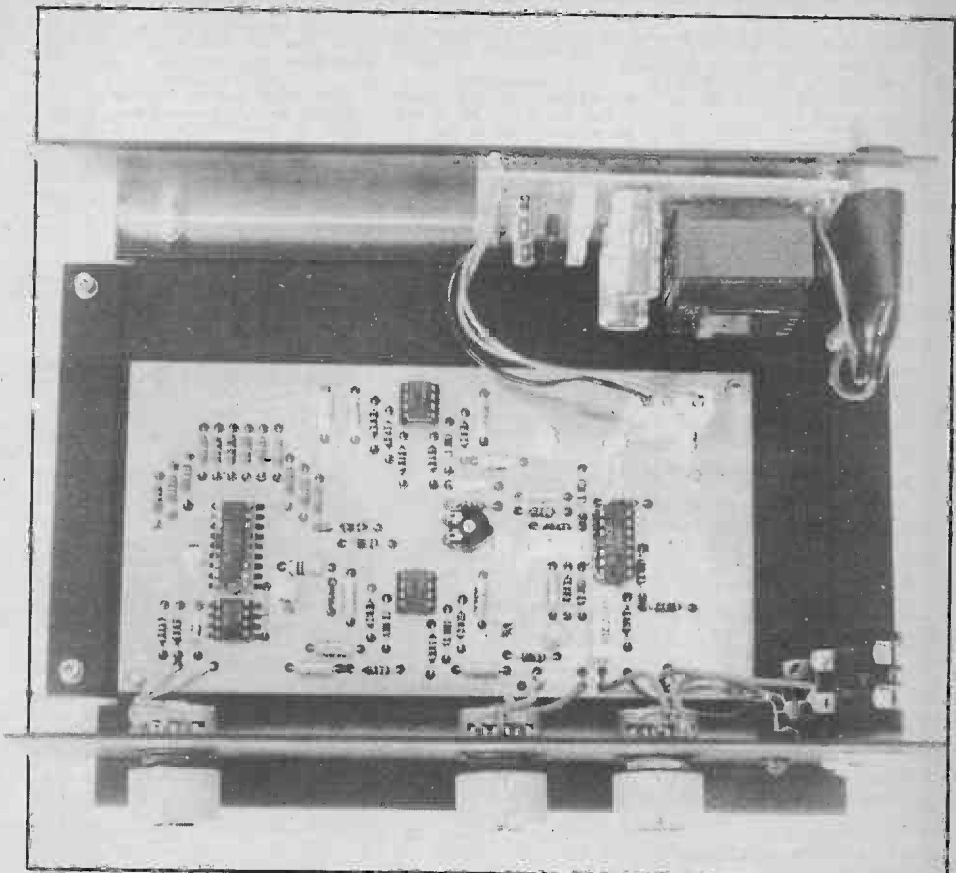


Fig. 5 Circuit diagram for the ETI Reverb.

Setting Up And Use

The only setting up required is adjustment of PR1. If a sinewave source is available then the latter may be used as the signal source and PR1 adjusted by ear, or with an oscilloscope, for minimum distortion. Alternatively measure the voltage at the junction of PR1 and R40 and adjust PR1 to give a reading of 6V2.

The unit has a signal-to-noise ratio of better than 60 dB but this requires that it is operated with the peak indicator LED just glowing or occasionally illuminating. The output level will vary from about 0V5 to 1 V RMS, depending on the amount of mixing of the original signal, and these levels should ensure adequate response from most amplifiers, mixers, and so on. In other words, by keeping input signals at maximum level the amplifier setting will be such that during periods of no signal the residual noise will not be obtrusive. This is common practice with recorders, many of which have much lower signal-to-noise ratios.



ETI

AUTORANGING CAPACITANCE METER

Look — no hands! The only control on this piece of test-gear is the on/off switch; the only connection is to the test terminals. This month — construction. Design and development by Phil Walker.

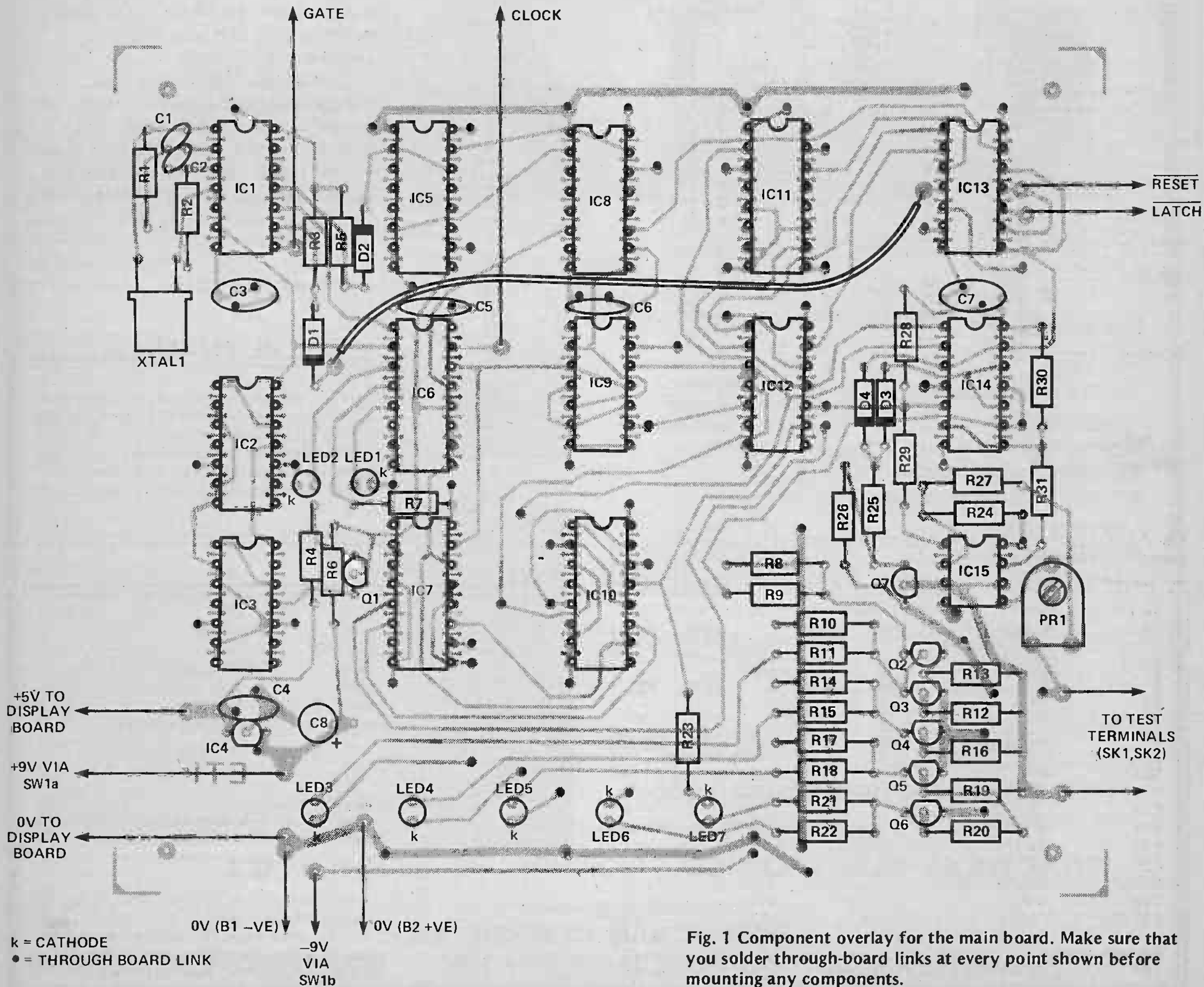


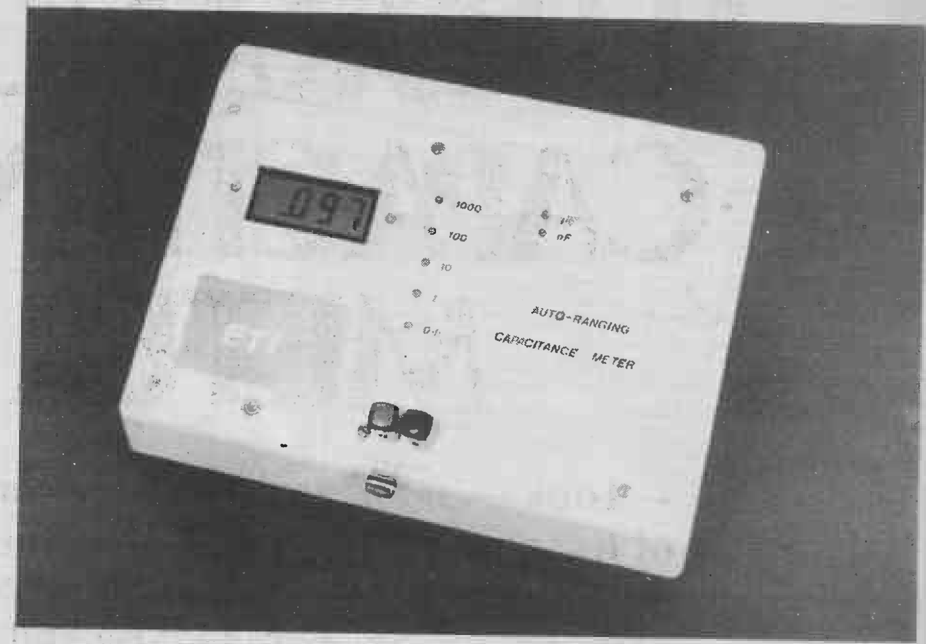
Fig. 1 Component overlay for the main board. Make sure that you solder through-board links at every point shown before mounting any components.

This is a fairly complex project and should only be attempted by those with a good deal of constructional experience. 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.

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).



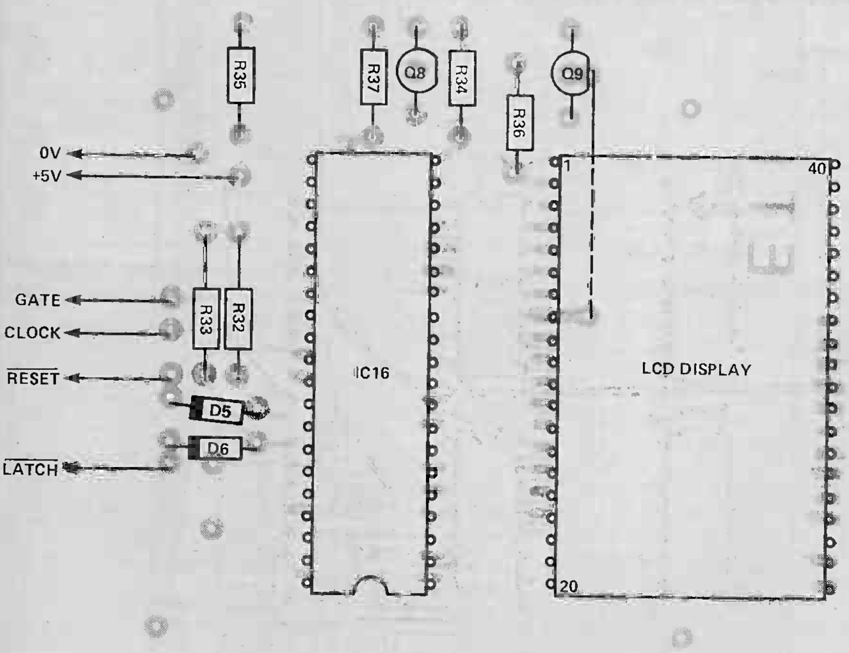


Fig. 2 Component overlay for the display board. Insert the link under the display first.

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 44.

PARTS LIST

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

R1,2	470R
R3,4,5, 28-37	15k
R6,25	4k7
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
R24,27	27k 2%

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
Q3-7	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).

TECH TIPS

Frequency-To-Phase Controlled Power Supply

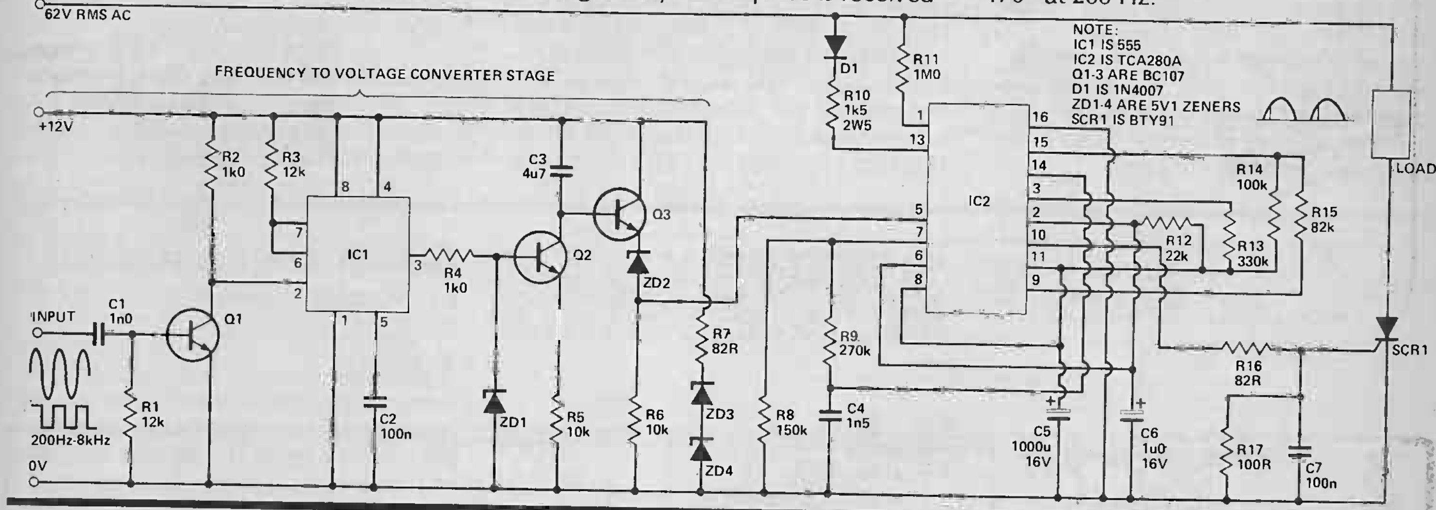
Dilbay Singh (B.Tech), Crawley

The circuit shown in the diagram was initially designed to obtain a phase-controlled power supply to use with a ¼ horsepower stepping motor. The phase angle can be varied over the complete cycle period and is dependent on the *frequency* of the input. Clearly the circuit can be used to control resistive loads such as lamps or motors.

The first stage of the circuit consists of a frequency-to-voltage converter. C1, R1, and Q1 effectively differentiate and amplify the input signal waveform to provide triggering pulses for the 555 timer, which is used in the monostable mode. The output of the monostable is used to charge C3 by a constant amount of charge every time a pulse is received at the base of Q2. The voltage across C3 acts as an input to the common collector stage formed by Q3. The voltage across C3 is DC-shifted by means of the zener diode ZD2 to a suitable value, providing the input to the trigger IC (the Mullard TCA280A). The TCA280A provides the phase control signal for the gate of the thyristor.

A triac may be used in place of the thyristor, if phase-controlled AC is required.

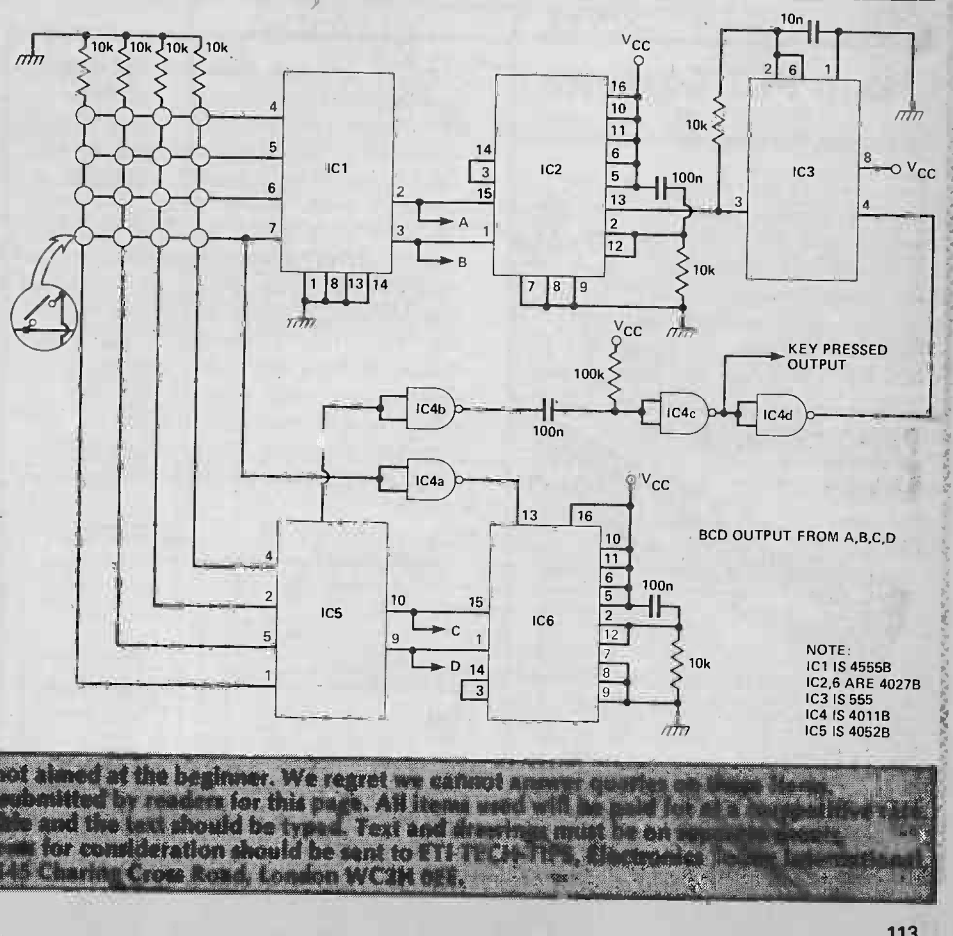
The component values shown are suitable for providing phase control using frequencies in the range 200 Hz-8 kHz on the control input. The firing angle can be varied from 0° at 8 kHz to 170° at 200 Hz.



Fully Debounced Keyboard

Graham Kyte, Bexleyheath

This circuit produces a debounced output whenever a key is pressed. Each matrix point is scanned in turn and the output of the 4052 data distributor goes high when a pressed key is detected. This stops the scanning oscillator (555) for about 10 mS and a 'key pressed' output is produced, thus enabling the BCD output to be stored in a latch or otherwise made use of. The use of CMOS ICs enables current consumption to be minimised, making the circuit suitable for operation in a car. The circuit is easily modified for a larger number of keys by using an eight-way data distributor (with relevent counter made from three J-K flip-flops rather than the two as used here).



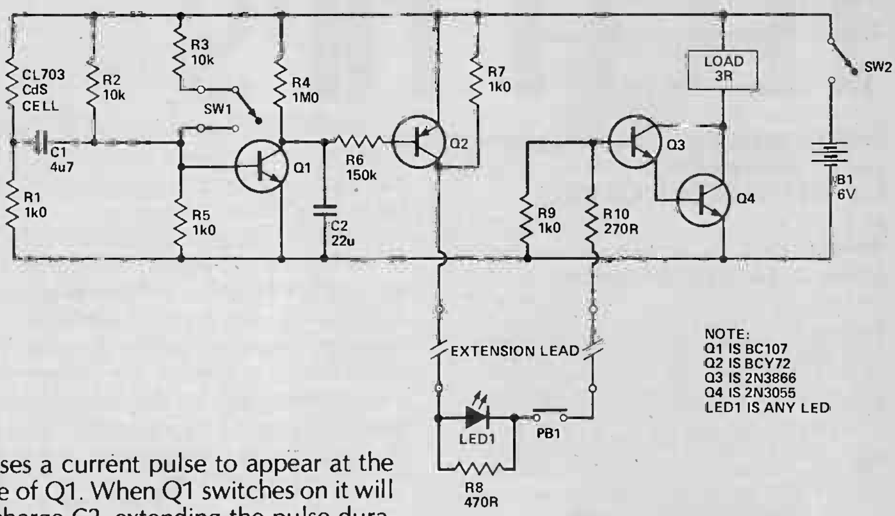
Tech-Tips is an ideas forum and is not aimed at the beginner. We regret we cannot answer queries on these items. ETI is prepared to consider circuits or ideas submitted by readers for this page. All items used will be paid for at a competitive rate. Drawings should be as clear as possible and the text should be typed. Text and drawings must be on separate sheets. Circuits must not be subject to copyright. Items for consideration should be sent to ETI-TECH-TIPS, Electronics Today International, 145 Charing Cross Road, London WC2H 0EE.

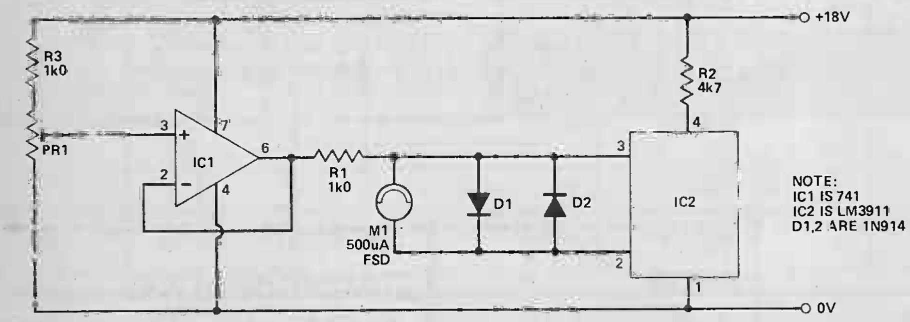
Remote Camera Release

Geoffrey Ammon, Welling

When taking photographs from a distance, a pneumatic remote release is normally used. These will only work over a limited distance and it is not always possible to tell if the camera has operated. This simple circuit uses a low current trigger circuit to operate the camera and provides a visible indication that the camera or flashgun has worked correctly.

The circuit operation is as follows. When the remote release push-button PB1 is operated, a current flows via the extension lead, which may be a 100 metres or more in length, to switch transistors Q3 and Q4. This combination provides the load current of up to 2 A for the camera release solenoid. When the flashgun fires, light falling on the CdS cell causes a current pulse to appear at the base of Q1. When Q1 switches on it will discharge C2, extending the pulse duration to about one second. While C2 is charging Q2 will be turned on, causing a large enough current to flow in the extension lead to operate LED1. If a flashgun is not used, the camera flash contacts (SW1) may be connected to bypass the CdS cell and remotely operate the indicator LED1.





Room Thermometer

J. P. Macaulay, Crawley

With the advent of the LM3911 temperature controller IC the task of measuring temperature has become simple in the extreme. The internal circuitry of this device comprises a temperature sensing element, an op-amp and a stable reference voltage. The device gives, in its simplest form, a stable 10 mV change in output for every 1° change in temperature over the range −25 to 85°C. For the application of room thermometer it is only necessary to utilise part of this range from, say, 0° − 50°C. The circuit to be described measures this range.

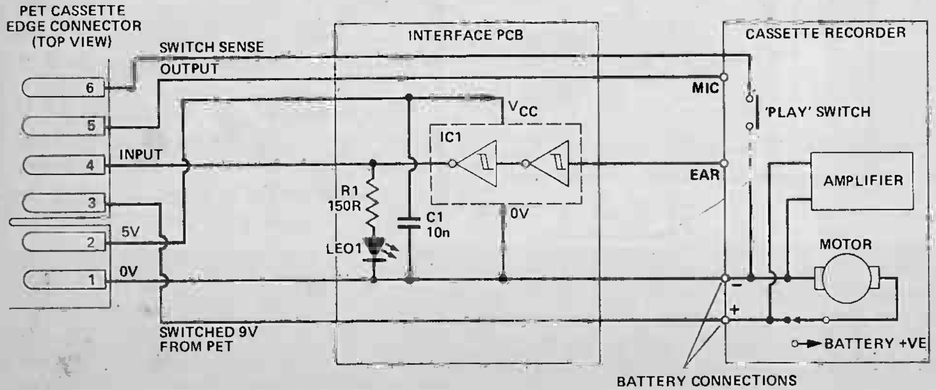
The figure shows the complete circuit of the thermometer. The meter, a 500 uA FSD type, is connected between the output and inverting input of the internal op-amp. Resistor R1 connects the inverting input to the output of the 741 op-amp. This is used with 100% AC and DC feedback to form a unity gain voltage follower with a current output capacity of several milliamps. The input of the 741 is connected to the slider of PR1 which in turn is connected across the stable supply voltage produced by the IC. D1 and D2 protect the meter from overrange temperatures and thus protect its delicate movement from harm.

Once completed, a calibration can be made with a room thermometer of known accuracy. Simply leave the equipment in the room for 10 minutes or so for its own temperature to stabilise and then adjust PR1 until both thermometers read the same; the calibration is now complete.

Cheap PET Cassette

D.J. Cocker, Portsmouth

In view of the price of the Commodore cassette unit, the following adaptation may be of interest. I have been using this arrangement for some time and have experienced very few problems. In order to signal the PET when the PLAY key has been pressed, a switch mus be incorporated into the cassette key assembly – a small microswitch is ideal. This is an improvement on the Commodore unit, in which any key activates the switch, leading to confusion and ambiguity. The 'signal present' LED is very useful in locating the start and end of the data tape. The cassette recorder is supplied with power from the PET, batteries only being required for fast forward and rewind functions — a switch should be fitted to facilitate this. When the PLAY key is depressed, the PET has control of the tape motor. It may be found necessary to disable any tone control circuitry or AGC which may be fitted in the cassette recorder. Any suitable TTL Schmitt gate may be used as IC1.



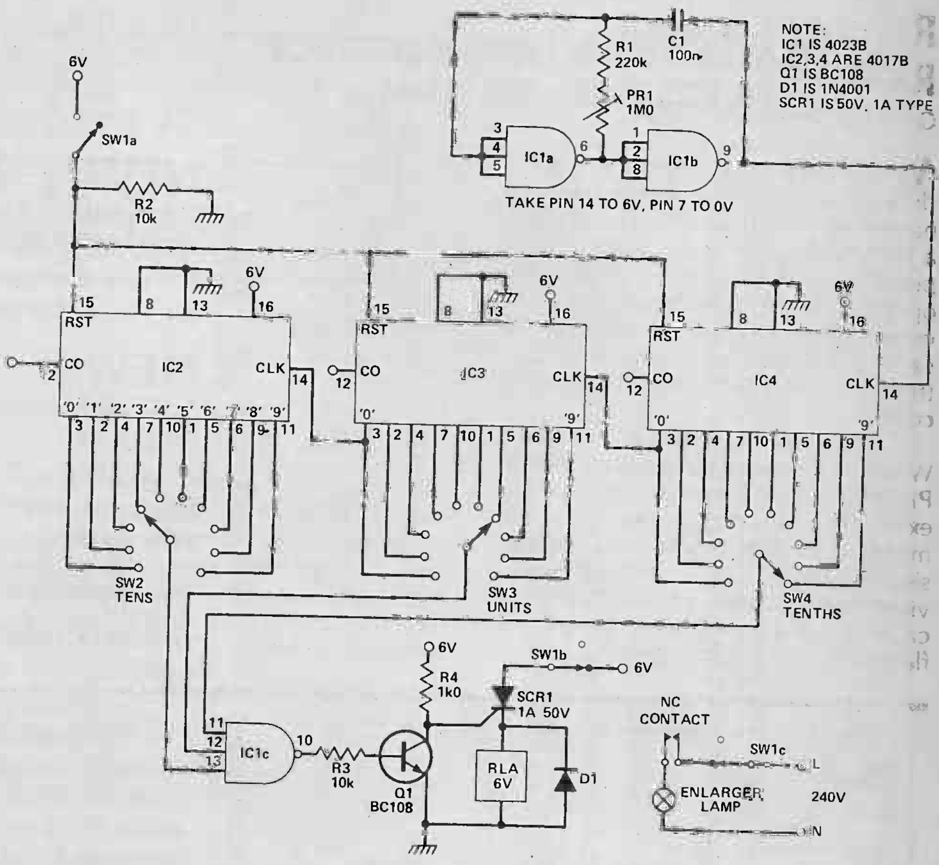
Enlarger Timer

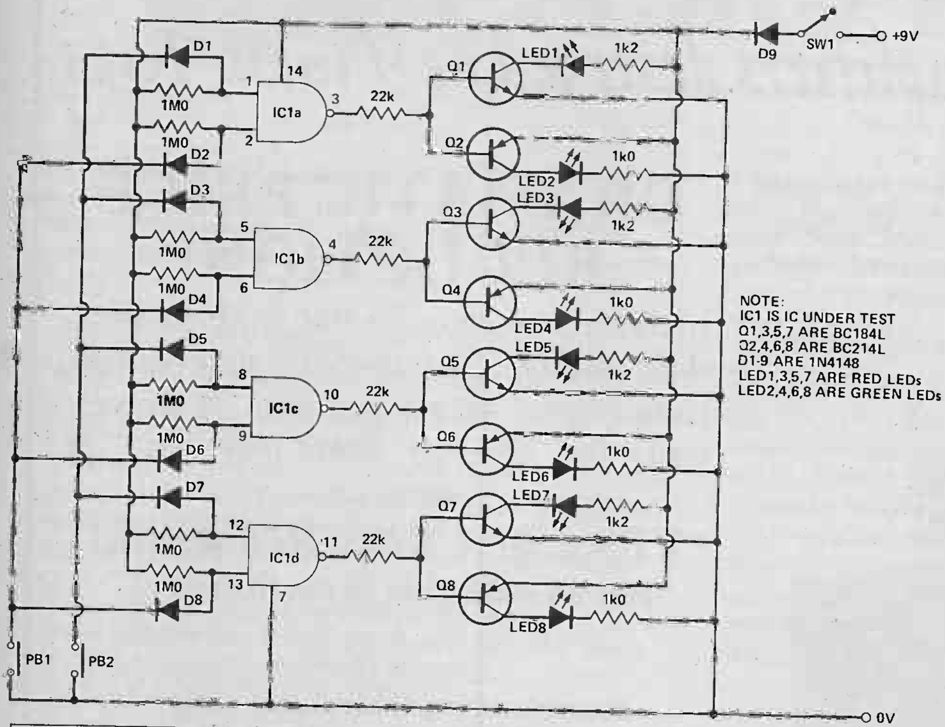
C. E. Basson, South Africa

The circuit of the enlarger timer can time periods from 0 to 99.9 seconds in 0.1 second steps. PR1, C1, R1, IC1a,b form an oscillator that feeds a 10 Hz signal to the first 4017 counter stage. Either the 'carry out' or '0' outputs of IC2 and IC3 can be used to feed the next stage, as the frequencies are the same and the positive-going edges of the pulses appear at the same time. Outputs '0' to '9' go high in sequence as the pulses are received at the 'clock in'. The desired time is selected by SW2, SW3 and SW4.

Q1 is used as an inverter and with the NAND gate it performs the same function as an AND gate. As soon as the desired time is reached, all the inputs on the gate will be high and this will trigger SCR1. The relay will be turned on and switch off the enlarger lamp. The lamp will remain off until the circuit is reset.

The circuit can be resetted by closing SW1a and opening SW1b and SW1c. SW1a will reset the 4017s and keep them in the reset condition. SW1b will remove the current from SCR1 to reset it. SW1c prevents the light from going on when in the reset condition. When SW1 is switched back to normal, the light will go on and remain on for the desired time.





4011 (NAND)	GREEN	RED
SW1	ON	OFF
PB1	OFF	ON
PB2	OFF	ON
BOTH	OFF	ON

4001 (NOR)	GREEN	RED
SW1	ON	OFF
PB1	ON	OFF
PB2	ON	OFF
BOTH	OFF	ON

4081 (AND)	GREEN	RED
SW1	OFF	ON
PB1	ON	OFF
PB2	ON	OFF
BOTH	ON	OFF

4071 (OR)	GREEN	RED
SW1	OFF	ON
PB1	OFF	ON
PB2	OFF	ON
BOTH	ON	OFF

4070 (EXOR)	GREEN	RED
SW1	ON	OFF
PB1	OFF	ON
PB2	OFF	ON
BOTH	ON	OFF

4077 (EXNOR)	GREEN	RED
SW1	OFF	ON
PB1	ON	OFF
PB2	ON	OFF
BOTH	OFF	ON

SW1 ONLY – BOTH INPUTS HIGH
PB1 PRESSED – ONE INPUT LOW
PB2 PRESSED – OTHER INPUT LOW
BOTH PRESSED – BOTH INPUTS LOW

Comprehensive CMOS Logic Gate Test Rig

David Ian, Surrey

This simple test rig will check out all possible functions of any type of dual input CMOS logic gate allowing, for example, a faulty gate to be pinpointed so that the rest of the IC may still be used.

Each gate is provided with a green LED to indicate a high output and a red LED to show a low output.

Use a 14-pin holder for the IC and orientate the LEDs to relate to their appropriate gate.

SW1 connects power and a logic 1 to all inputs: press A to put a 0 onto one input of each gate; B puts 0 onto the other inputs; A and B togther force all inputs low.

A milliammeter in series with a 9 V supply should only indicate the current drawn by the LEDs, ie about 7 mA per LED. An appreciably higher reading indicates a completely faulty IC.

SOUND EFFECTS 2: STEAM TRAIN

Railway modellers looking for something special to improve their layout need look no further. Our second sound effect project simulates a steam train and whistle. Design by Phil Wait.

Aahh, the nostalgia! If you're young at heart, old in years, or both, then this is for you — a steam train (chuff-chuff) and whistle. The *electronic* construction details are given on page 50 in the bomb drop project; but for that *authentic* touch, deft constructors can also fashion a cow-catcher out of tinned copper wire to attach to the unit!

The chuff-chuff runs continuously once power is applied and the whistle sounds when the push-button is pressed. The VCO is used to provide the whistle while the SLF modulates the noise generator/filter output to produce the steam train's chuff-chuff sound. The chuff-chuff rate may be varied by changing the values of R1 and C1, while the chuff-chuff sound may be varied by changing the values of R2 and C2. For a special effect, you can control the chuff-chuff rate manually by replacing R1 with a 1M0 potentiometer.

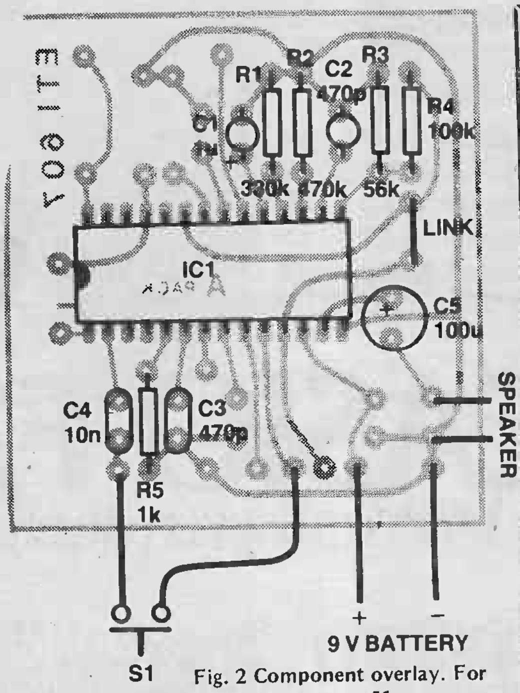


Fig. 2 Component overlay. For Buylines, see page 51.

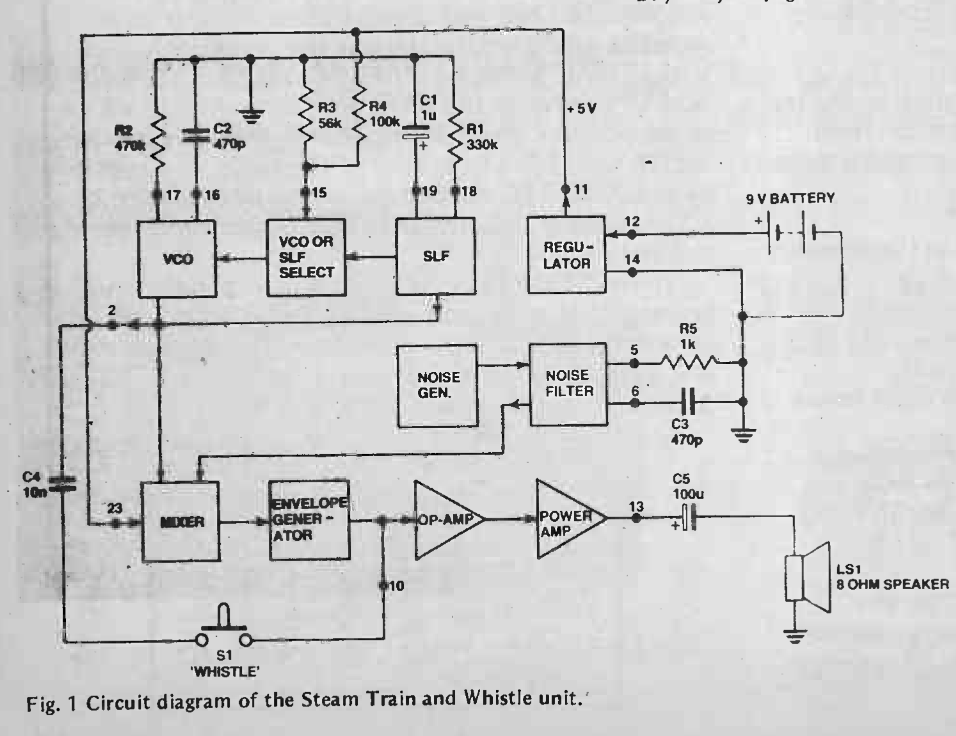


Fig. 1 Circuit diagram of the Steam Train and Whistle unit.

HOW IT WORKS

In this unit the Noise Generator/Filter is employed to produce the basic 'steam engine' sound, this being modulated by the SLF to produce the 'chuff-chuff' so characteristic of steam locomotives. The whistle is produced by the VCO, which is set to a particular non-varying pitch, and the output is switched into the audio input pin to produce the whistle.

The broadband noise from the Noise Generator is modified by the Noise Filter, the frequency characteristics being determined by R5 and C3 connected to the Noise Filter Control pins (5 and 6). The Noise Filter Output is fed via the Mixer and the Envelope Generator (which doesn't function here) to the audio output stages. The SLF square wave output effectively modulates the noise to produce a noise burst followed by a silent period, then another noise burst. Thus the chuff-chuff sound is produced. This sound is continuous whilst power is applied to the unit.

A resistive divider, R3/R4, provides about 1V8 at the VCO frequency to a convenient pitch within its range, providing a suitable pitch for the whistle. The VCO output is coupled to the audio input (pin 10) via C4 and the push-button, PB1. When PB1 is pressed, the whistle is heard over the chuff-chuff sound.

The SLF frequency is determined by C1 and R1, while the combination of R2/C2 and the voltage on pin 15 determines the VCO frequency. Output to the loudspeaker is coupled via C5, a 100uF electrolytic capacitor.

PARTS LIST

Resistors (all ¼ W, 5%)	
R1	330k
R2	470k
R3	56k
R4	100k
R5	1k0
Capacitors	
C1	1u0 16 V tantalum
C2, 3	470p ceramic
C4	10n ceramic
C5	100u 16 V PCB electrolytic
Semiconductors	
IC1	SN76488 (see Buylines)
Miscellanous	
PB1	SPST push-button switch

PCB (see Buylines); 50 mm diameter 8 ohm speaker; PP3 battery and clip.

GUITAR PRACTICE AMPLIFIER

Simple construction, low cost, good performance and super neighbour relations are the features of this project! Design and development by David Tilbrook.

This project has been designed to enable guitarists to put in long hours of practice and still keep that high power amp in the cupboard, where it belongs! It is a compact amp capable of about 7 W into a 4 ohm load. This is enough power for practice purposes and just think of the greatly improved relations you will have with your neighbours.

We were in a considerable quandary as to how to present the project, whether it should be done as a complete practice unit with inbuilt speaker or simply as an amplifier to be connected to an external speaker. Finally we chose a compromise. The PCB has been designed in such a way that it can be used as a totally self-contained unit. The heatsinks for the output stage have been mounted on the PCB so that the only components separate to the board are the power transformer, 240 volt power switch controls, input and output jacks. We have shown the project mounted in its own box with power transformer but it should be a simple matter to construct the whole unit inside a small loudspeaker cabinet.

The unit has two inputs so that two guitars can be mixed together using the relative settings of the two input level controls. A preamp output enables your main high power amp to be driven from the guitar practice amp using the practice amp as foldback.

We provided the PCB with the necessary circuitry for a battery input but you might elect not to use this feature. If so diode D8 and the battery switch can be omitted with points 'A' and 'C' connected together by a wire link.

Construction

Construction of the project is reasonably simple since it is almost entirely devoted to construction of the PCB. Start as always by mounting the resistors and non-polarised capacitors. Mount the tantalum and electrolytic capacitors next, being careful to orient them correctly. These components could be irreparably damaged if inserted the wrong way around. Mount the LM301 IC, transistors and diodes, again being careful to insert these the correct way round.

Finally the output devices can be mounted. Although the transistors are in TO220 packages, our PCB is laid out to accept heatsinks drilled for TO3 transistors. The overlay and photograph should make the construction method clear. Cut the centre (collector) lead off. This lead is connected to the case of the transistor internally, so in this case, electrical connection is made through the mounting screw that also serves to hold the heatsink in place. Place the heatsinks on the PCB and secure with the lower nut and bolt (not used to mount the transistors). There is only one right way round. Bend the leads of the output transistors and, using a small amount of thermal compound, mount the transistors with the leads protruding through the PCB.

Secure each transistor with a nut and bolt through both the transistor 'flag' and heatsink. Use a star washer between the head of the bolt and the copper pad on the PCB to ensure good electrical contact. Now the base and emitter leads can be soldered to their pads.

The prototype unit was constructed in a steel box measuring

Grabbed By The Dooleys

Those tireless chappies down at Casio have taken time off from disguising BASIC computers and arcade games as pocket calculators and watches, and have turned their attention to the music scene. Although there is undoubtedly a market for top-flight organs and synthesisers amongst home musicians, many people wilh prefer something more modest — for financial reasons, because the living room is too small or because they can't figure out what all the knobs do. At the other end of the scale (sorry), the type of hand-held organ made notorious by Rolf Harris is a little too limiting. With the Casiotone 701, Casio have not just produced a solution to this problem but a radically new type of instrument.

The CT-701 is not just a 61-key polyphonic (eight voice) mini-synthesiser, but also contains an on-board computer that acts as a built-in sequencer; among other things. You can play along with the built-in rhythm unit, store your own music in memory and play it back automatically, or just load the machine with a Casio music score and let it get on with things by itself. The latter function is quite extraordinary — Casio supply the music scores as bar codes and you read them into the machine using a light pen (like those at supermarket check-out desks). In melody guide mode you can even teach yourself to play the instrument, as LEDs above each key light up to tell you which note to play next.

Twenty preset sounds are available, such as pipe organ, flute, piano, oboe, bassoon etc, plus the synthesised drum sounds of the rhythm unit and the 'pneooum' sound so beloved by producers of disco records. Opinions of the preset sound quality vary from "beautiful" (Casio) through "very good" (an independent reviewer) to "too sharply filtered" (another independent reviewer). Since they can't agree and we haven't heard it (though, we're trying hard to get our mucky paws on one), you'll have to listen to one yourself before parting with any cash, but professional musicians seem to like it — the Dooleys use Casiotone mini-keyboards in their stage shows (fellow headbangers may not see this as a compliment). With so much packed into such a compact case (only slightly larger than the actual keyboard) and such a low price (about £500), Casio would certainly seem to have done it again.

ZX Revamp

For those of you who are serious ZX-81 owners (is there such an animal?) or would simply like to disguise the machine, there is a professional standard keyboard and enclosure now available from Protos Computer Systems. The keyboard is the first of a range of peripherals to make the computer suitable for more heavy-duty use. The 40-key Sinclair coded board uses top quality mechanical contact type key switches with relegendable tops. A steel mounting board holds the keys firmly in position and a high quality printed circuit board completes the board's electrical circuit. Connection to the Sinclair board is made by a flexible connector which is a push fit to the sockets provided on the ZX81. Access to the edge board connector is via a side port on the Protos enclosure and tape in/out, power and UHF connections are made through the rear. To fit the Protos entails removing the Sinclair board from the black ABS case it comes in and fixing it inside the Protos enclosure with four Phillips type screws. No soldering is required and all electrical connections are plug/socket connections provided either on the Sinclair or the Protos. Further details on this and other forthcoming peripherals can be obtained from Protos Computer Systems, Frome Computing, 20 Ashtree Road, Frome, Somerset BA11 2AS. Please enclose a large SAE with any enquiries.

Thin Meters

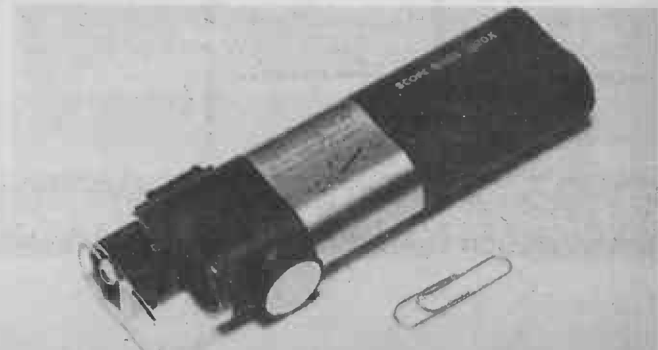
Sifam Ltd of Torquay in Devon are to market a range of very thin edgewise meters manufactured by General Electric of the USA. There are three sizes in the range with case widths of 38 mm, 63 mm and 89 mm and the units are scaled for vertical or horizontal presentation. The special feature of this design is the extreme thinness; the smallest has an overall depth of face of only 13 mm and the two larger sizes of about 17 mm. The smallest model has a rear-access zero set and a simple spring-clip method of mounting. The two larger models have front access zero set at end of scale and a slide bracket form of mounting. They incorporate jewelled pivot movements with special high-torque magnets for reliable and accurate operation. The standard meters are available ex-stock from Sifam and have a maximum sensitivity of 50 microamperes. Scale markings can be produced to suit individual requirements. Further details of these and Sifam's own range of meters are available from: Sifam Limited, Woodland Road, Torquay, Devon TQ2 7AY.

Power For Peanuts

Grenson Electronics, designers and manufacturers of power supplies for the Nuclear Research Industry have come up with a series of bench power units. The first unit in the series is priced at £59 and gives a variable stabilised output up to 30 V at 2 A in two ranges, has foldback re-entrant short circuit protection and current and voltage metering. This unit is also available in kit form at only £35 and further details are from Grenson Electronics Ltd, High March Road, Long March, Industrial Estate, Daventry, Northants NN11 4HQ.

Miniature Magnification

New from Stotron Ltd is the Scope Mark III pocket microscope with stand. Priced at under £20 it is a useful tool for laboratories, schools, workshops, service engineers and the electronics, electrical, automotive, print and graphic trades, Uncle Tom Cobbley and all! It is 125 mm long, with 20x magnification and a graticule showing linear and angular measurements. Illumination is powered by standard 1V5 'pen-light' batteries and a micro-stand (with spring clips for sample slides) is available as an option so that the device can be used like a conventional microscope. Further details on this device are available from Stotron Ltd, Unit 1, Haywood Way, Ivy House Lane, Hastings, East Sussex.



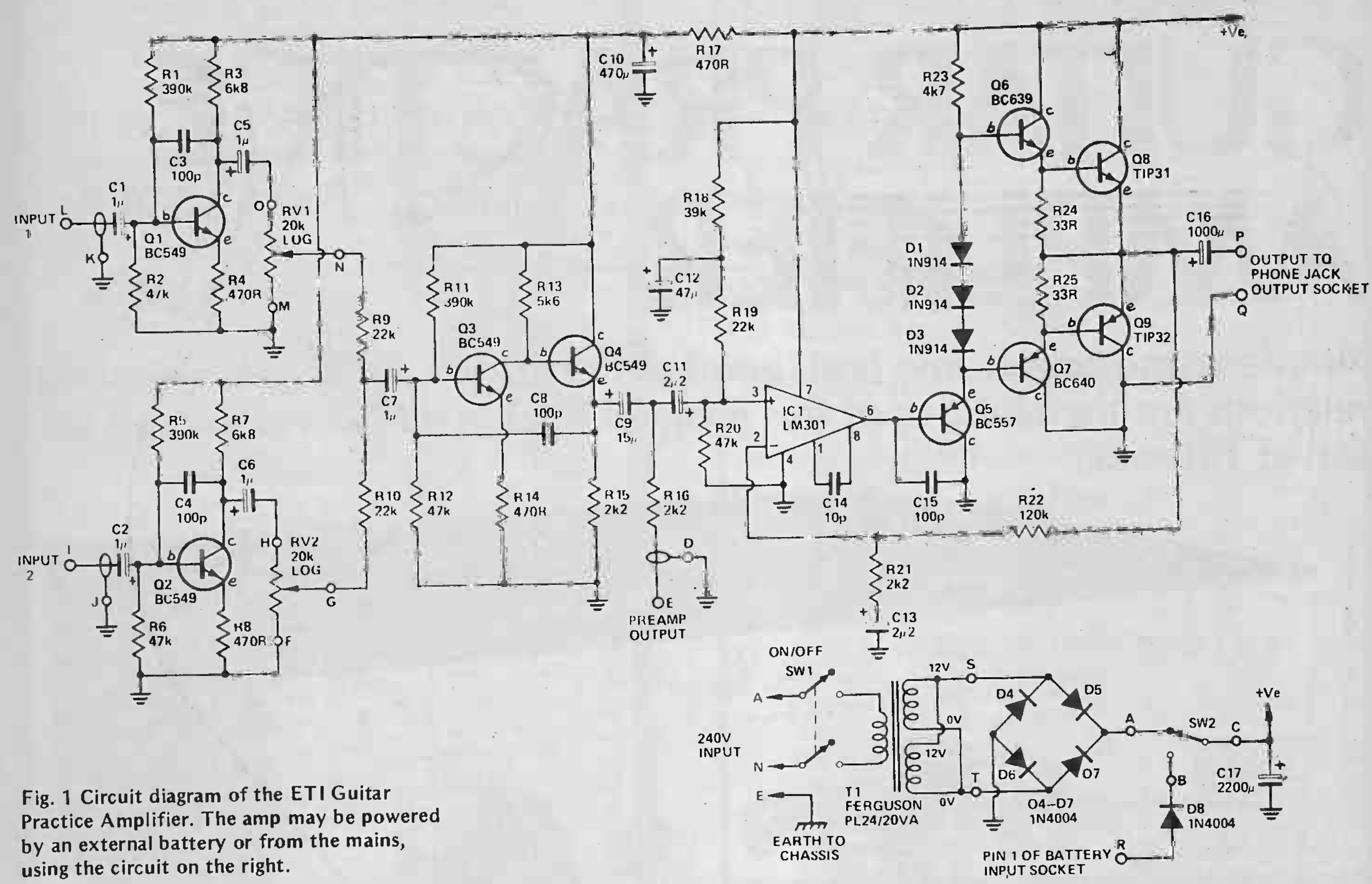


Fig. 1 Circuit diagram of the ETI Guitar Practice Amplifier. The amp may be powered by an external battery or from the mains, using the circuit on the right.

HOW IT WORKS

The two input stages formed around Q1 and Q2 are identical. Resistors R1, R2 and R4 form a very stable biasing configuration around Q1. The gain of this type of circuit is determined by the values of R3 and R4 (specifically, the gain is R3/R4). The load impedance on the output of the input stages is in parallel with R3, effectively decreasing the total value of impedance from collector to ground. Remember that, as far as signal is concerned, the positive supply rail is a short circuit to ground, since it is connected to ground through C17, a 2200uF capacitor. When all these factors are taken into account the gain of the first stage is about 10 since the impedance from collector to ground is about 4k7.

The signal, which should now be around 200 mV, is then applied to the input of the second stage through potentiometers RV1 and RV2. The 22k resistors R9 and R10 prevent the output of one of the stages being shorted to ground when the other is turned right down.

The second stage works in exactly the same manner as the input stages, resistors R11, R12 and R14 forming the bias network for Q3. The voltage present on the collector of Q3 is around 9 V which is approximately half the supply voltage. This is used to bias Q4 which is an emitter follower. This type of amplifier has no voltage gain but provides a low output impedance to drive the preamp output socket. Q3 has a gain of approximately 10. If the volume controls RV1 and RV2 are used in their middle positions, the voltage out will be around one tenth of the voltage at their inputs since these are logarithmic pots. So, the signal voltages into Q3 should be in the order of 20 mV. This will be amplified to a level of 200 mV and applied to the input of the power amp. The power amp has been designed to deliver full power with an input voltage of 300 mV, so the amp should be easily driven to full output with usable settings.

Since this is a guitar amplifier, it will spend most of its life hard into clipping. The output stage had to be robust! The basis of the output stage is the LM301 op-amp. This device gives all of the voltage gain in the power amp. The output IC1 is fed through a voltage follower Q5. This has no voltage gain and, like Q4, serves to decrease the impedance feeding the output stage. The three diodes, D1, D2 and D3, maintain 1V8 between the bases of Q6 and Q7. Each of these transistors will drop approximately 0V6 across their base-emitter junctions. This leaves a total of 0V6 to be dropped by the two 33R resistors, R24 and R25. Since these are of equal value they will each drop 0V3 and hold this voltage across the base-emitter junctions of the two output transistors Q8 and Q9. As these transistors require 0V6 to turn on they will remain off until the applied signal voltage causes the voltages on their bases to rise above 0V6. The extra 0V3 needed to turn on the output devices will be supplied by a mere 10 mA of current through the 33R resistors. Resistor R22 forms a feedback loop around the entire output stage to decrease distortion, stabilise the DC output voltage and set the overall gain of the power stage (a process too difficult to go into here).

The op-amp will at all times attempt to make the DC voltage at the output equal to that voltage set up on its positive input. This voltage is determined by the potential divider formed by R18, R19 and R20. Since this is also the main input to the power amp any noise which might be on the positive supply rail (and supplies can get very noisy sometimes!) will be communicated directly to the input of the power amp, only to be amplified and applied to the loudspeaker. Capacitor C12 prevents this from happening by bypassing to ground any noise above a frequency of around 0.1 Hz.

approximately 250 x 210 x 80 mm. Mount the pots and switches on the front panel, using the pot and switch nuts to secure the front escutcheon if you have one. Mount the output and battery input sockets on the rear panel. If you are using a battery input socket use something different to the output socket (which is usually a two-pin DIN socket or a 6.5 mm jack socket) to avoid confusion.

Mount the power transformer and make the 240 V connections. The mains lead should be terminated immediately inside the case into a terminal block and the earth lead secured firmly to the chassis by a solder lug bolted to the case using a star washer. This lead must be the longest. A length of 240 V cable should be used between the terminal block and the power switch. Wire the transformer to the power switch as shown in the circuit diagram, then wrap the whole switch with insulation tape or enclose in large diameter heat-shrink tubing so that no 240 V connection is exposed.

Finally, the fully-loaded PCB can be secured into the case using short metal spacers. If Veropins are used, all the connections to the board can be made after the board has been mounted. Connect the front panel controls, rear panel sockets and input sockets, using short lengths of shielded cable to make the connections to the two inputs and preamp output.

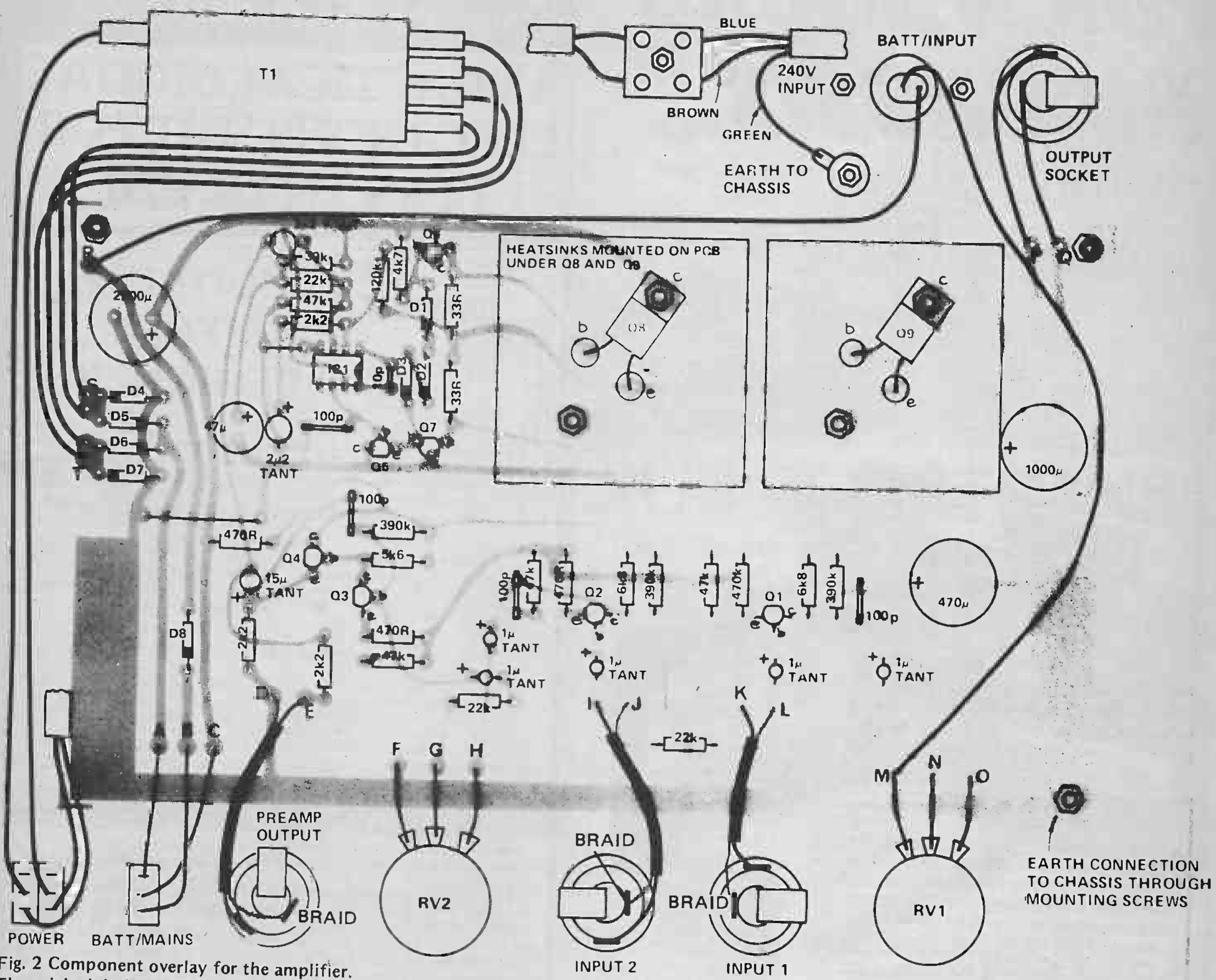


Fig. 2 Component overlay for the amplifier. The original design used a BC639/BC640 complementary pair for Q6 and Q7, and these are shown on the overlay, but they may prove hard to obtain. Consequently the PCB we will be supplying is laid out for a BC140/BC160 pair, which have different pad layouts — the b,c and e pads are etched onto the board for your guidance.

Powering Up

Make a final check of the wiring and PCB. If all is well, apply power. A slight turn-on thump should be heard at the moment of turn-on. If the 'Input 1' volume control is now wound up, some hiss should be heard from the loudspeaker. Do the same check on the other input. There is no set-up procedure since the power amp stage is operating in class B and requires no bias adjustment.

BUYLINES

Lots of nice, standard, easy-to-obtain components in this project, so you shouldn't encounter any problems with supply. The PCB will be available fro our PCB Service at the price listed on page 44.

PARTS LIST

Resistors (all ½W, 5%)

R1,5,11	390k
R2,6,12,20	47k
R3,7	6k8
R4,8,14,17	470R
R9,10,19	22k
R13	5k6
R15,16,21	2k2
R18	39k
R22	120k
R23	4k7
R24,25	33R

Potentiometers

RV1,2	22k logarithmic

Capacitors

C1,2,5,6,7	1u0 35 V tantalum
C3,4,8,15	100p disc ceramic
C9	15u 16 V tantalum
C10	470u 25 V PCB electrolytic
C11,13	2u2 35 V tantalum
C12	47u 25 V PCB electrolytic
C14	10p disc ceramic
C16	1000u 25 V PCB electrolytic
C17	2200u 25 V PCB electrolytic

Semiconductors

IC1	LM301
Q1-4	BC549 or BC109
Q5	BC557 or BC179
Q6	BC140
Q7	BC160
Q8	TIP31
Q9	TIP32
D1-3	1N914
D4-8	1N4004

Miscellaneous

SK1-4	mono jack sockets
SK5	DIN socket (or other type — see text)
SW1	DPDT toggle switch (mains rated)
SW2	DPST toggle switch

Transformer (12-0-12-0, 20 VA); TO3 type PCB-mounting heatsinks; PCB (see Buylines); case to suit; mounting hardware.

ELECTROMUSIC TECHNIQUES

Tim Orr, our tame electronic designer, emerged from his workshop this month just long enough to hand over this bundle of circuits for the ardent build-it-yourself musician.

Virtually all of the electronic music synthesisers that have been produced to date employ analogue circuits to generate the synthesised sounds. The process is known as subtractive synthesis, and operates by dynamically filtering out parts of the spectrum of a signal that is often rich in harmonics. The results are instant, easy to modify and relatively inexpensive to implement. It is not possible to produce an arbitrary output spectrum, and so it is very difficult to synthesise realistic copies of naturally generated sounds. This can be done using a digital technique known as harmonic synthesis, whereby the sound is constructed by precisely defining the amplitude and phase of each of the harmonics. These are then added together to produce the output. However, natural sounds are constantly varying and so the data defining all the harmonics must also vary. Harmonic synthesis can produce very realistic sounds and is in itself a powerful technique for generating completely new sounds, but the hardware is a combination of sophisticated microprocessor and digital technology and so is outside the scope of this article.

When we hear a sound we unconsciously analyse it for useful information; "Who wants another drink?" for example. Nobody knows how the human brain analyses incoming sounds, but it does it with incredible speed and sophistication. It can extract precise information from sounds (speech perception), it can experience pleasure from a rich harmony, or it can even learn to ignore certain sounds, such as a ticking clock. The brain is very good at perceiving pitch (or at least it thinks it is; it is also a fairly good liar); see Fig. 1. When you hear a pure tone you will get a strong impression of its pitch. You will not be able to define its frequency in Hertz, but you will be able to remember its pitch. A sawtooth has a strong harmonic structure but even so you will get the same pitch perception. The ringing tone has virtually no energy at the fundamental frequency and yet it is still possible to correctly perceive the pitch of the signal, although it is more difficult than for the pure tone.

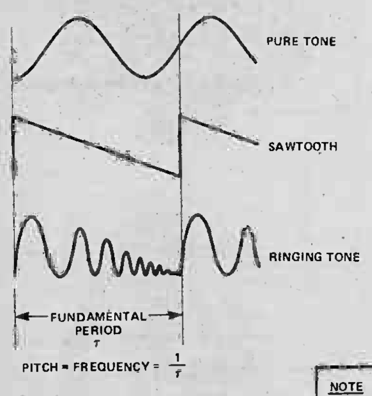


Fig. 1 Pitch perception.

Most musical instruments produce a range of notes. Some instruments, like violins, can produce a continuous range of frequencies; because, unlike the guitar, there are no frets along the neck of the instrument. Keyboard instruments have fixed tuning; the piano, for example. The keyboard is an excellent choice for controlling a synthesiser, as it is easily converted so that it generates suitable electrical signals and it is widely accepted by musicians. Equal temperament tuning is used, that is there are twelve notes per octave and they are spaced at intervals of the twelfth root of two (that is 1.0594631) along an exponential curve, as in Fig. 2.

When You Hear The Tone . . .

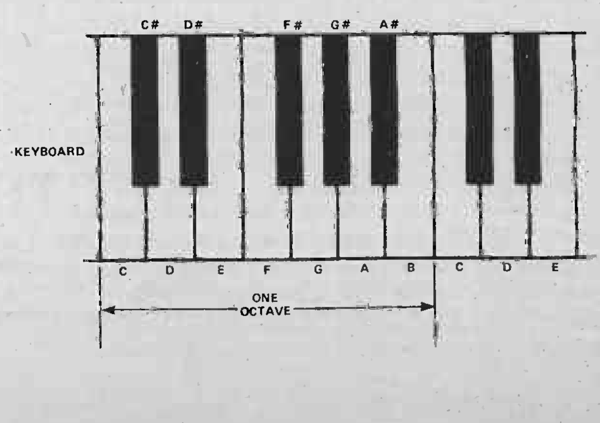
The keyboard is used to define the fundamental pitch of a sound, but the actual shape of the waveform will determine its harmonic structure (Fig. 3). A sinewave is a pure tone and has no harmonics. A ~~half~~wave-rectified sine wave contains a fundamental plus a series of *even* harmonics. A fullwave-rectified sine wave is composed entirely of *even* harmonics. The squarewave and the triangle are both composed of a series of *odd* harmonics; in fact if you lowpass filter a square wave you can produce a triangle. The triangle is a fairly pure tone, with little of the energy in the waveform contained in its harmonics. The sawtooth is a rich waveform, having both *odd* and *even* harmonics.

The harmonic structure of all these waveforms extends to infinity, but the drawings only show the first 15 harmonics. If we call the harmonic number n, then the harmonic amplitude is easy to define. The rate at which the harmonic amplitude

Fig. 2 (below) Keyboard layout with table showing equal temperament tuning.

NOTE	FREQUENCY (Hz)
A0	27.5
A1	55.0
A2	110.0
A3	220.0
A4	440.0
A5	880.0
A6	1760.0
A7	3520.0

NOTE	FREQUENCY (Hz)	RATIO
C4	261.6	1.0000
C4 #	277.2	1.0595
D4	293.7	1.1225
D4 #	311.1	1.1892
E4	329.7	1.2599
F4	349.2	1.3348
F4 #	370.0	1.4142
G4	392.0	1.4983
G4 #	415.3	1.5874
A4	440.0	1.6818
A4 #	466.1	1.7818
B4	493.9	1.8877
C5	523.2	2.0000



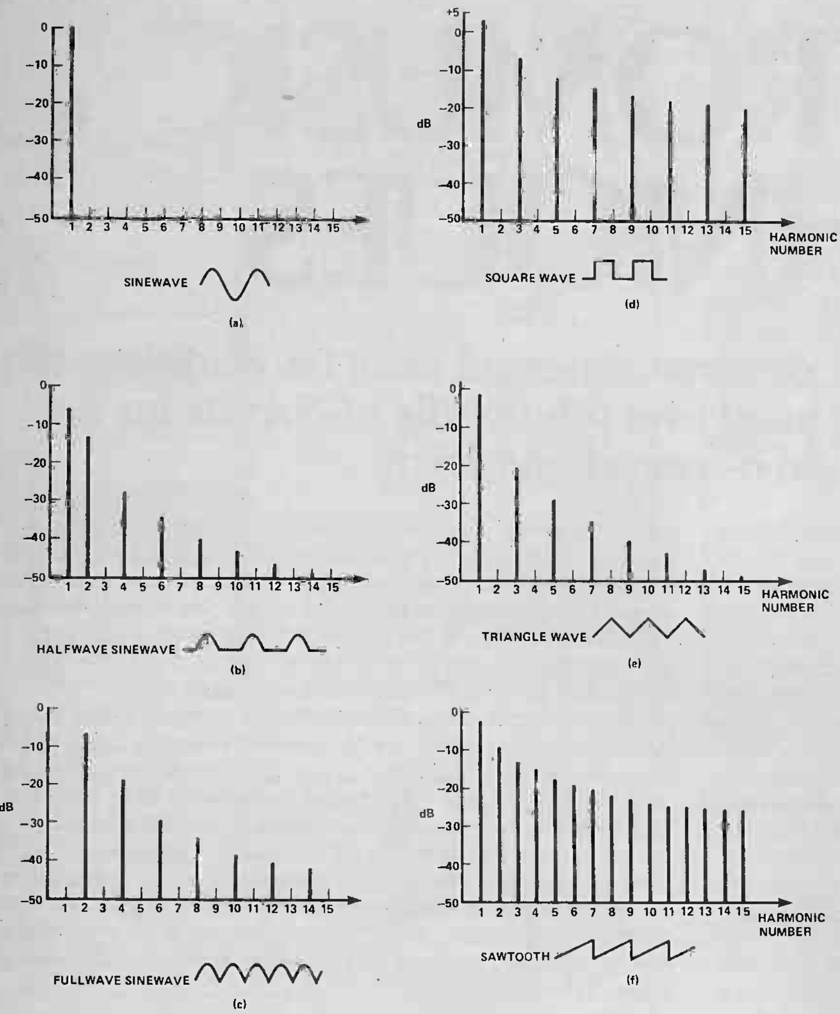


Fig. 3 Harmonic structure of various standard musical waveforms.

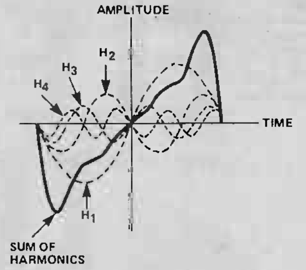


Fig. 4 Adding the first four harmonics to construct a sawtooth waveform.

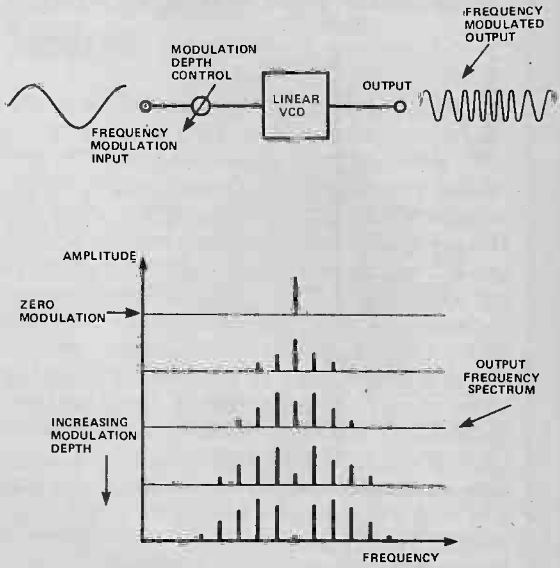


Fig. 5 The effects of frequency modulation.

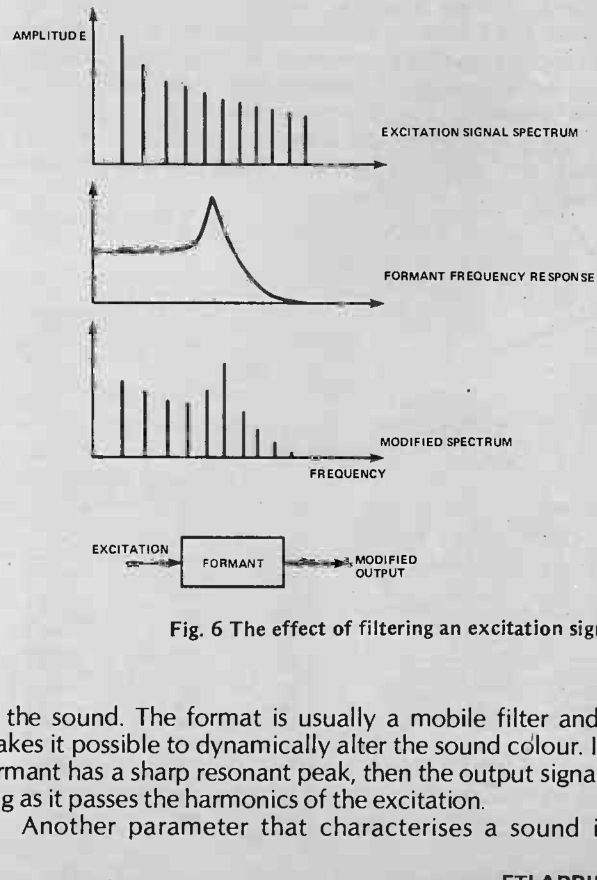


Fig. 6 The effect of filtering an excitation signal.

decreases is $1/n$ for the sawtooth and square wave and $1/n^2$ for the half and fullwave rectified sine wave and the triangle. Figure 4 shows a sawtooth being constructed from harmonics. The sum of the harmonics is beginning to look like a sawtooth. As more harmonics are added (with the correct phase and amplitude) the sum will converge upon the correct sawtooth shape. An interesting effect can be produced by changing the mark/space ratio of the square wave. This modifies the *odd* harmonic spectrum and introduces *even* harmonics. The mark/space ratio is often dynamically modified as a synthesis process.

Frequency modulation is often employed in synthesisers to produce vibrato and other dramatic pitch change effects. Figure 5 shows some of the effects of frequency modulation. As the modulation depth is increased, frequency sidebands are generated. Their spacing and amplitude are determined by the modulation depth and the modulation and carrier frequencies. To precisely calculate them involves some complex maths and Bessel functions (which I have forgotten all about). To make matters worse, synthesisers usually use voltage controlled oscillators with an exponential transfer function, which tends to exponentially distort the sideband positions. But so what! Music synthesisers are all about making music and not the calculation of sidebands. If a particular electronic device produces a useful musical effect, then use it, don't analyse it.

The output from an oscillator is known as an excitation signal. This defines the pitch of the signal, and to a certain extent the harmonic content of the final signal. It is common practice to filter the excitation signal (Fig. 6). The frequency response of the filter is referred to as a formant. The formant modifies the harmonic spectrum of the excitation, producing a colouration of the sound. The format is usually a mobile filter and this makes it possible to dynamically alter the sound colour. If the formant has a sharp resonant peak, then the output signal will ring as it passes the harmonics of the excitation.

Another parameter that characterises a sound is its amplitude contour or envelope (Fig. 7). A sound that has a sharp attack and a slow release is similar to a plucked instrument. Other envelopes will make the sound seem like something else.

Building Blocks

Most synthesisers are constructed from standard building blocks, and most of these blocks are voltage controlled. This is a very powerful concept, because it enables you to control a unit with a combination of control voltages and/or audio signals. Building blocks can be patched together in any arbitary order to produce any system that is wanted. Some standard building blocks are detailed below.

Voltage Controlled Oscillator Used to generate the pitched excitation signals. Often a VCO will generate a wide range of waveforms. The control sensitivity is usually +1 V/octave. Therefore a one twelfth of a volt change will alter the oscillator pitch by one semitone. The exponential control law is a very powerful concept. If a VCO is being driven so that it produces a melody, then adding +1 V to the control input will transpose the melody up by one octave. Thus musical transpositions are very simple to produce. Often more than one VCO will be used, so that a rich chord is obtained.

Voltage Controlled Filter This is used as a formant for the excitation signal. The VCF is generally a lowpass filter, but it can often be a multi-mode device with lowpass, highpass, bandpass and notch responses. The VCF also has a Q (resonance) control. The control sensitivity is +1 V/octave for the frequency parameter, and undefined for the Q.

Voltage Controlled Amplifier The VCA controls the level of audio signals. The control law can be linear or logarithmic. The VCA is usually controlled by an ADSR unit and is employed to generate signal envelope contours. The device is a two quadrant multiplier.

Attack, Decay, Sustain, Release unit The ADSR is used to generate the signal envelope contour and also the VCF sweep waveform.

Ring Modulator This is a four quadrant multiplier or balanced multiplier. The output voltage is the product of the two input signals. It is often used to generate discordant or clangerous sounds.

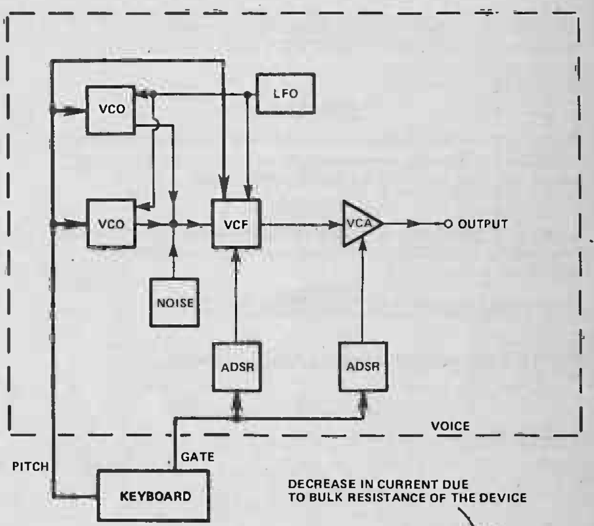
Noise source Generates random noise, which can be used in the synthesis of non-pitched sounds such as explosions. Filtered or sampled noise can be used as a random control voltage.

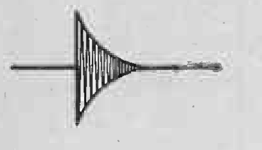
Low Frequency Oscillator These oscillators are used to generate vibrato in the VCO or a filter sweep in the VCF.

Keyboard Musical control interface, generating pitch voltages of +1 V/octave and also a gate signal to indicate that a note is pressed. A monophonic keyboard only allows one note at a time to be pressed, but if more than one can be pressed simultaneously then the system is polyphonic.

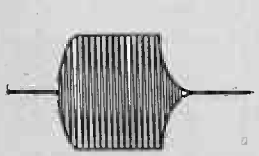
There are several other building blocks such as flangers, sequencers, frequency shifters, and pitch detectors, but there isn't enough space to deal with them.

Polyphonic synthesisers tend to be voice-based; ie all the building blocks are pre-routed to form a voice (Fig. 8). Modular systems are not pre-routed and have to be patched, either with lots of jack-to-jack patch leads or via a matrix patch board using patch pins. Patch leads are relatively inexpensive, but the leads get in the way and it is often difficult to see just what you have patched. Matrix patch boards are easy to understand, but they suffer from crosstalk and a large board (60 by 60) might cost £500!





SHARP ATTACK, SLOW DECAY – PLUCKED



SLOW ATTACK, SLOW DELAY – "PIPE ORGAN"

Fig. 7 (Above) Two typical amplitude contours, or envelopes.

Fig. 8 (Top right) The standard synthesiser voice.

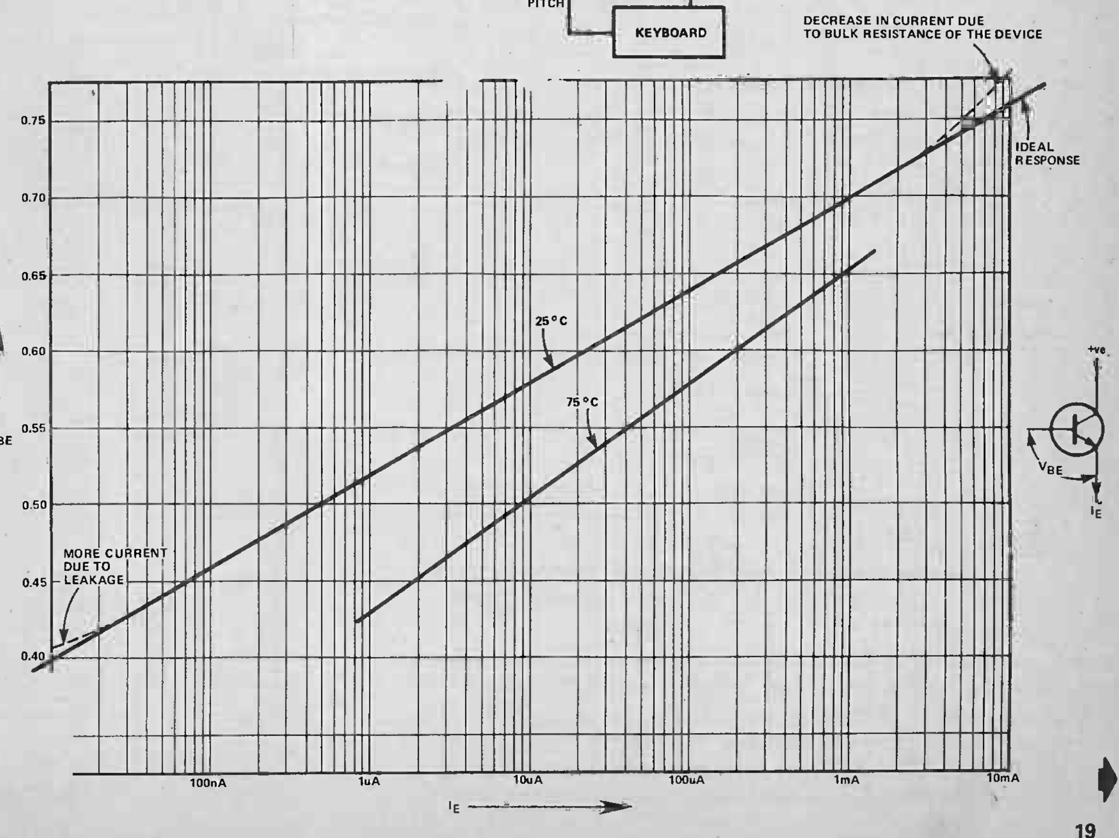
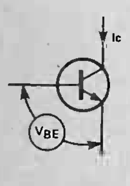


Fig. 9 (Right) Silicon diode transfer characteristics.

Diode Data

The silicon diode has an exponential transfer function, that is the diode current increases exponentially for linear increments in the diode voltage (Fig: 9). This can be used to turn linear changes from, say, a keyboard into exponential or musical intervals in a VCO. The required musical range is probably no more than 200 to 1 and so a suitable operating current would be 0.5 uA to 100 uA, thus avoiding the non-exponential parts of the curve. The silicon diode is temperature dependent (it is often used as a thermometer) and so great care must be used to avoid thermal problems. The junction voltage changes by −1.9 mV/°C, but a semitone change is equivalent to 1.5 mV, therefore a 1°C change could result in a 1.27 semitone change in pitch! Figure 9 shows two temperature effects in operation; there is a large shift and the slope of the line changes.

Figure 10 illustrates the equations that determine the diode operation. Two facts emerge from these equations. First, an 18 mV change in V_{BE} will double the current I_C, and second, this parameter has a temperature coefficient of −0.33%/°C. Both the temperature problems can be resolved by using a circuit similar to that shown in Fig. 11. Transistor Q1 is run at constant current (12 uA) by the op-amp. Q2 is used as the exponentiator transistor. The emitter of Q2 is held at a voltage of about −0V6. Any voltage change at the base of Q2 will result in an exponen-



$I_c = I_o\,(e^{\{qV_{BE}/KT\}} - 1)$

$\therefore\ I_c \simeq I_o\,e^{qV_{BE}/KT}$

WHERE
Io IS THE EMITTER SATURATION CURRENT
K IS BOLTZMANNS CONSTANT
q IS THE CHARGE ON AN ELECTRON
T IS THE TEMPERATURE IN °K

HOWEVER, $\frac{KT}{q}$ IS 26mV AT 28.58 °C (301.73 °K IS ROOM TEMPERATURE).
THEREFORE, $I_c \simeq I_o\,e^{V_{BE}/26}$
WHERE V_{BE} IS MEASURED IN mV
REARRANGING THE EQUATION,

$$26.\ln\left(\frac{I_c}{I_o}\right) = V_{BE}$$

THEREFORE, AN OCTAVE CHANGE IN Ic IS CAUSED BY A 18.021827mV CHANGE IN V_{BE} (AT 28.58 °C). HOWEVER, IF THE TEMPERATURE WERE +1 °C HIGHER, THEN V_{BE} WOULD HAVE TO BE INCREASED IN SIZE TO A NEW VALUE OF

$$26 \times \left(\frac{302.73}{301.73}\right)$$

SO, FOR AN OCTAVE CHANGE IN Ic AT THE NEW TEMPERATURE, V_{BE} MUST CHANGE BY 18.08155mV, AN INCREASE OF 0.059723mV. THIS CAN BE EXPRESSED AS A PERCENTAGE CHANGE PER °C :–

$$\text{TEMPERATURE SENSITIVITY} = \frac{0.059723 \times 100}{18.021827} = 0.33139\%/°C$$

Fig. 10 Exponential transistor characteristics.

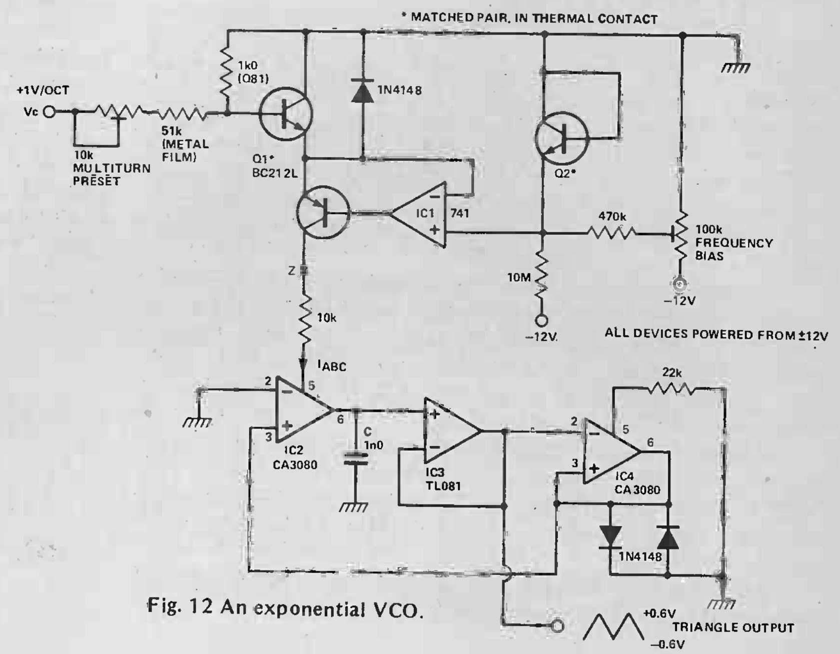
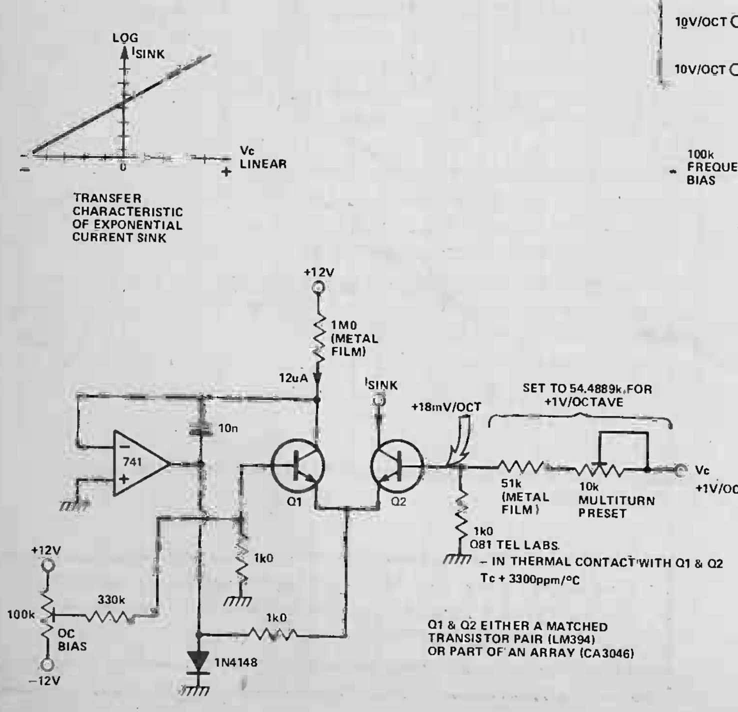


Fig. 12 An exponential VCO.

Fig. 11 (Below) An exponential current sink.



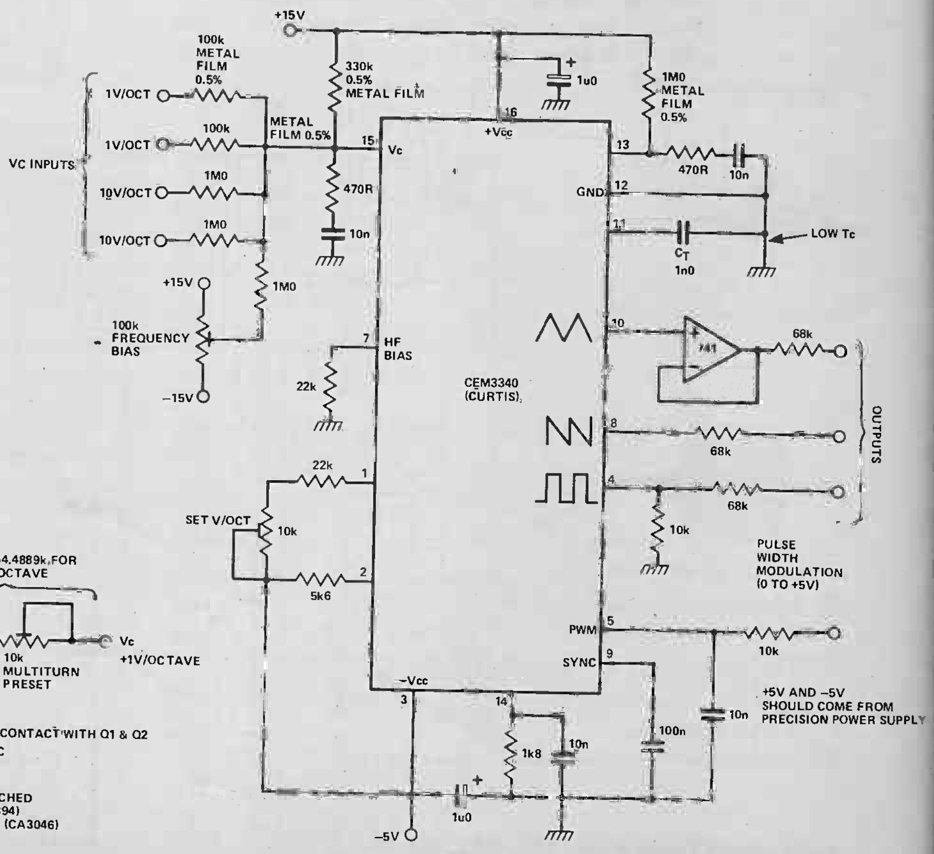


Fig. 13 A VCO using a monolithic device.