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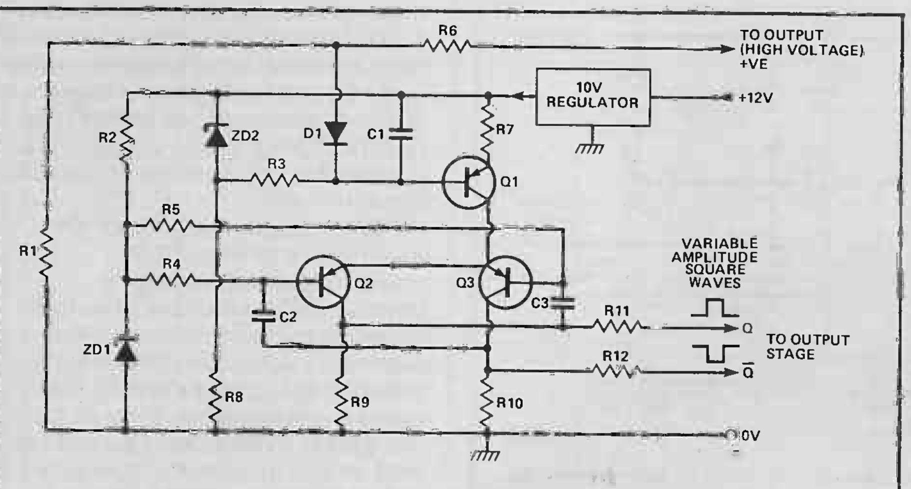


Fig. 4 A multivibrator arrangement with controlled output drive voltage.

either fully on or fully off. In any intermediate state they will dissipate unwanted amounts of power, which will reduce the low output power efficiency of the system.

What is required for such a system is a pulse-width modulator drive circuit, in which the output devices are either hard on or hard off but where the duration of the on pulse to each switching transistor could be varied from a 50/50 duty cycle (for maximum output power/voltage) to a 1/99 duty cycle (for minimum output power/voltage) as shown in Fig. 5.

Looking through my notebooks, I found I had built a circuit of this type for another application using the layout I have sketched in Fig. 6. In this, a square-wave generator feeds a sawtooth generator and a divide-by-two circuit. The output of the sawtooth generator is then sliced by a circuit having variable 'slice' level, to give a repetitive rectangular pulse waveform output whose duration can be altered from 0% to 100%. The Q and $\overline{Q}$ outputs of the divide-by-two stage can be used to control a pair of AND gates which feed the output pulse to either one or other of the output switching devices via a suitable buffer.

This circuit works very well, and could be implemented using standard CMOS logic, but it looked a bit over-elaborate for what I had in mind.

My thoughts then turned to the simple astable multivibrator, using a pair of junction FETs (Fig. 7). This would give the output waveforms a and b at the points x and y in the circuit, and these could be sliced, squared, and inverted to give an alternate drive pulse to each

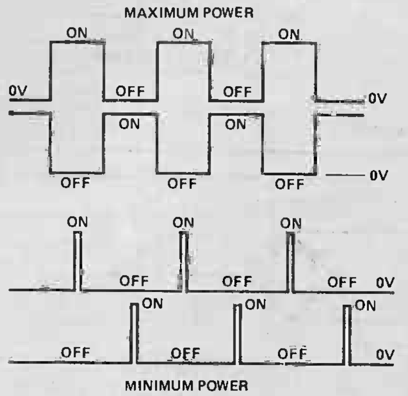


Fig. 5 Controlling the output power by pulse width modulation.

output device.

This does work, but cheap junction FETs vary enormously in their turn-on threshold voltage, and in order to get an equal mark-to-space ratio the FETs would have to be selected or R1 and R5 would have to be made adjustable. Also, the use of a Zener diode to provide the source reference voltage makes the circuit too sensitive to changes in the supply voltage.

This can be resolved by replacing the FETs with junction transistors, which are more consistent in their turn-on voltage. For silicon devices, this will always lie between 0.5 and 0.6V in the forward bias sense. Unfortunately, silicon transistors are not ideal for multivibrators because, if their bases are driven more than about 5V negative, the base-emitter junction acts as a Zener diode and can cause collector current to flow, even when the transistor is supposed to be turned hard off.

This problem can be avoided by adding protection diodes, D1 and D2 to stop Q1 and Q2, from being reverse biased (Fig. 8). Then, when the multivibrator action drives the lower ends of R1 and R6 negative, the diodes disconnect and the transistors are left with zero forward bias but with bases returned to emitters through

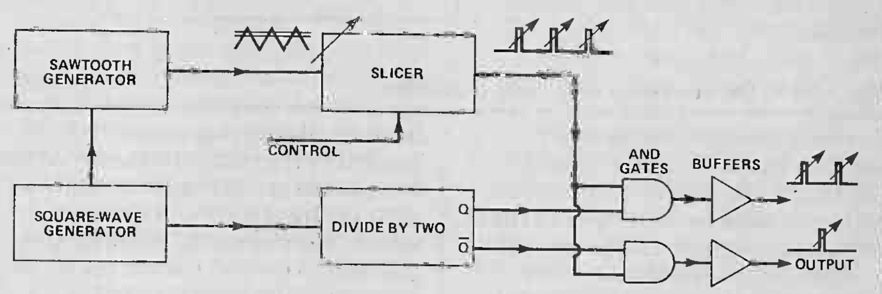


Fig. 6 Block diagram of a pulse width modulation drive circuit.

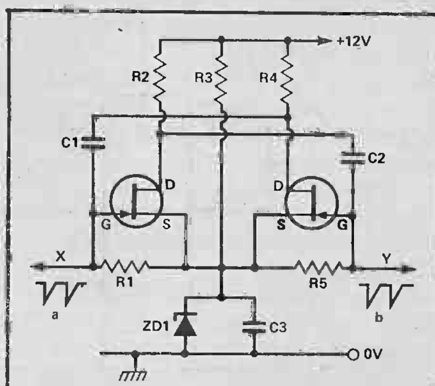


Fig. 7 Simple astable multivibrator using junction FETs.

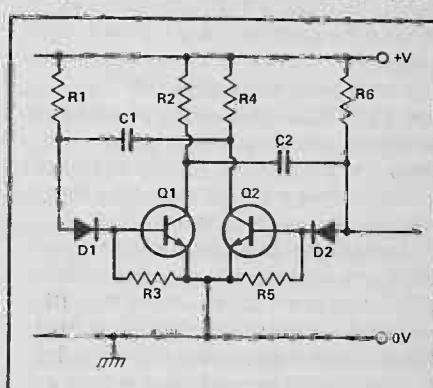


Fig. 8 An improved bipolar multivibrator.

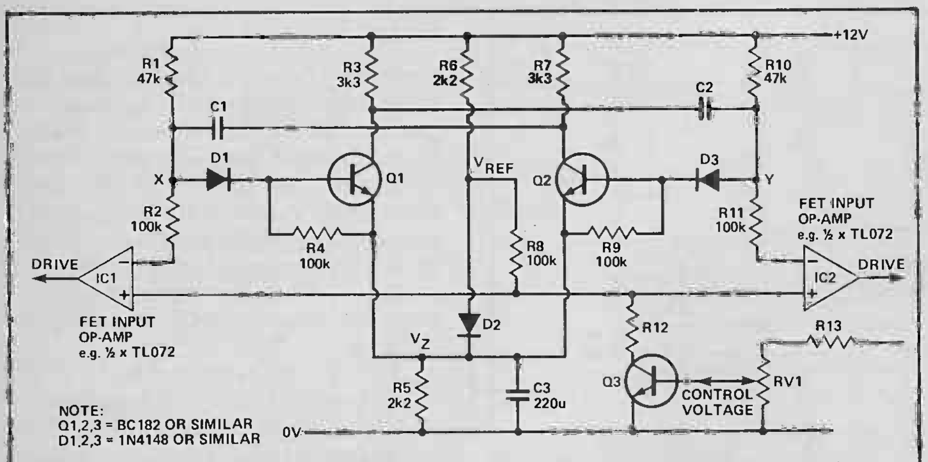


Fig. 9 A bipolar multivibrator with pulse width control of the output waveform.

R3 and R5 to prevent unwanted leakage currents.

A final version of the driver circuit based on this layout is shown in Fig. 9, in which a pair of FET input op-amps is used to sample the outputs at x and y, and deliver an alternate output rectangle drive pulse to the output switching devices.

If the output voltage from the inverter rises too high, it can be sampled by R13 and RV1 which will turn on to lower the input reference voltage applied to the op-amp. This will reduce the on time as the switch point slides down the slope of the waveforms a and b in Fig. 6.

By taking the op-amp reference voltage from the top end of D2, the maximum on time is made nearly 50%, but cannot exceed this. This is because the turn-on voltage, measured at x and y, is Vz plus two forward silicon PN junction potentials (D1 and Q1 b-e, or D3 and Q2 b-e), whereas Vref is Vz plus only one forward junction potential, (D2).

This circuit works as anticipated, but is not yet in a final form simply because it is not fail safe. If the oscillator were to stop oscillating for some reason — device failure, perhaps, or too low a supply voltage — the outputs of IC1 and IC2 would be left permanently high and the drive transistors would both be turned hard on.

This snag can be avoided by the adoption of the coupling circuit of Fig. 10. This is AC coupled and the signal waveform is DC restored positive by the networks D1R1 and D2R2. If the input drive waveform disappears, the output drive voltage will collapse to zero and both the output devices will be turned off.

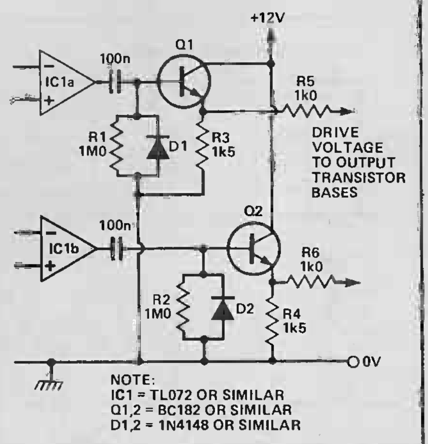


Fig. 10 An output coupling arrangement.

Other Snags

I had hoped to use a standard mains transformer as the step-up unit, which would save a lot of trouble. However, having got the prototype to work quite well on low power outputs, I started to do some serious power output and efficiency measurements at the projected output power levels and using a car battery instead of a stabilised DC power supply. The results were very disappointing unless I used a really large, high-power (and high cost) toroidal unit. For the record, the ILP 62033 (240V/50-0-50) was not too bad, but still not as good as I had hoped in terms of output power regulation.

The difficulty is that the effective secondary leakage inductance of a standard 240V unit is too high. The 'leakage' inductance is the value seen at the low voltage secondary terminals when the primary (240V) winding is short-circuited. This depends on the total number of turns in the windings, the type of core, and the way it is wound. Toroidal cores are generally better than E/I cores, and bifilar windings (side by side and wound together at the same time) are better than separate ones. However, the main problem is the number of turns in total.

With too high a leakage inductance as seen at the collectors of the switching transistors, the current at switch-on takes too long to rise to the necessary value and this means that the secondary current also takes a considerable time to build up. This is exaggerated by the fact that, in the interests of low power losses in the switching devices, we are using a square wave input drive to the transformer rather than a sinusoidal drive.

In spite of my best intentions, it looked as though I should have to wind a transformer. The choices here are whether to use a laminated iron core or a ferrite core. Ferrite-cored transformers only work well above about 15KHz where the efficiencies of power transistors, op-ams, and standard rectifier diodes are beginning to deteriorate. Laminated iron cored transformers are at their best from about 50Hz to 1000Hz, which is not too high a

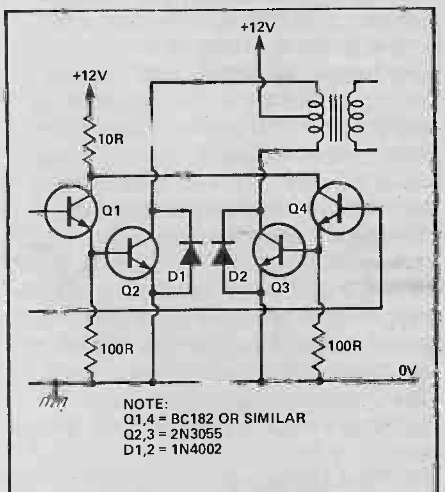


Fig. 11 The output transistor arrangement.

frequency for the other components.

If one were to operate at 300-500Hz, the number of turns needed wouldn't be very large and the transformer shouldn't be too difficult to wind — given a suitable core. I therefore used as the starting point a commercial 100W transformer kit which has a 28mm square core. On this I formed two 10T windings of 1mm copper wire wound as a doubled strand to

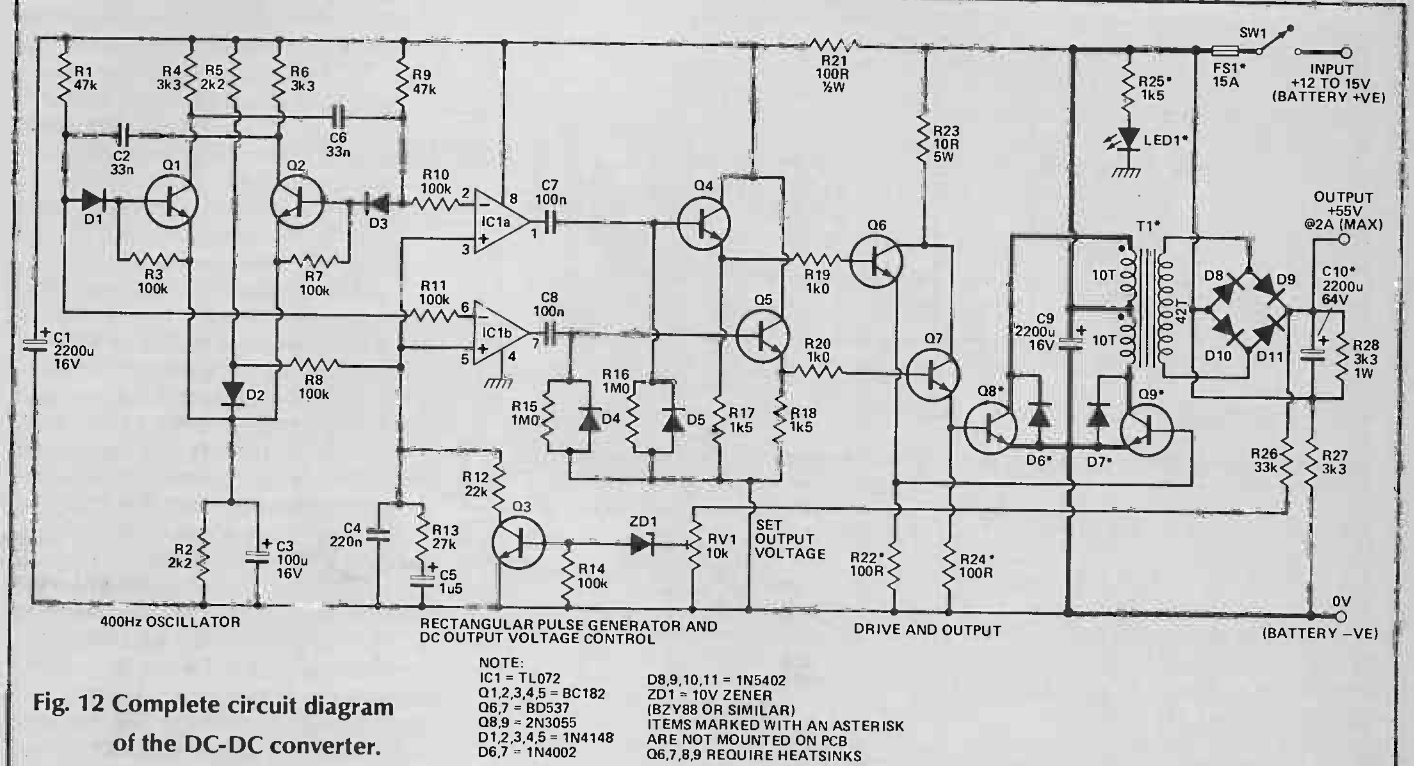


Fig. 12 Complete circuit diagram of the DC-DC converter.

lower the DC resistance, and a secondary of 42T also of 1mm wire insulated from the primary by a layer of PVC tape.

The ratio of 1:1:4.2 is dictated by the fact that, with substantially square voltage waveforms, the peak rectified output voltage will be the same as the mean. If we are using a single +ve rectified output rather than a dual-rail supply, we can sit the rectified output on top of the 12V DC input. We then only need to provide some 50V DC rectified output — giving a nominal 62V total — to be sure of obtaining the +55V supply specified earlier. It is necessary to have a little in excess to allow for poor load regulation, and to allow the control loop to have some surplus voltage upon which to work.

This transformer was much more satisfactory for the purpose, and also much less bulky. I tried varying the drive frequency a bit and found that higher frequencies made more core noise on load while lower frequencies increased the quiescent load current as the primary inductive impedance decreased. I eventrually settled for 400 Hz as a compromise.

For the output power levels required, and with the output efficiencies eventually obtained (70-80%), the peak switching currents will be about 11A. This is too high for the MJ3001 or TIP141 Darlington transistors which have a 10A current limit. To double-up these devices would be needlessly costly, as would the use of a heavier current Darlington type, so I finally opted for a pair of 2N3055s which are easy to come by and have a 15A peak current capability. These I used in the circuit layout shown in Fig. 11.

The driver transistors (Q1 and Q4 in Fig. 11), are not critical as to type. Any small power NPN device capable of handling 1A peak collector current will serve. The 10 ohm resistor in their collector circuit limits the peak currents which can flow in these devices, and also limits their dissipation to 1W or so.

Output Voltage Stabilisation

Any system which makes some measurement and then uses the result to effect a control function is a servo-mechanism, and as such may suffer problems of loop stablility. The servo loop consisting of the output DC voltage across the reservoir capacitor, sensed by ZD1 and Q3 which control the reference voltage for ICs 1 and 2, is just such a case, and the loop stabilisation components are C4, C5 and R13. They hold the output voltage constant within its working load range, with very little bounce on step changes in loading. This is helped by the 3k3 resistor which is connected as a permanent output load and which also prevents the reservoir capacitors retaining a charge on switch-off.

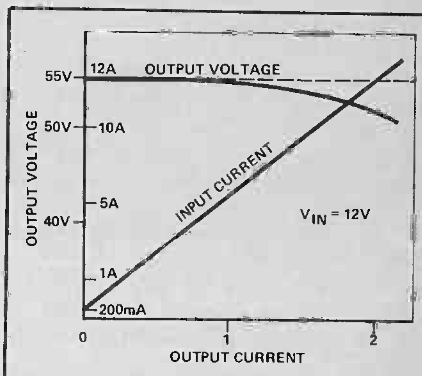


Fig. 13 Measured output voltage regulation of the prototype.

The voltage regulator circuit holds the output voltage constant over an input supply range from +12V to +15.5V, which should cope with the expected variation across a car battery off load or on charge. Also, on low load (that is under stand-by conditions) the duration of the on pulses is very brief and the consequent quiescent current drain only of the order of 200-300mA, which is a trifling load on the average car battery.

The final circuit of the DC converter is shown in Fig. 12 and a graph of the output voltage/current regulation characteristic shown in Fig. 13.

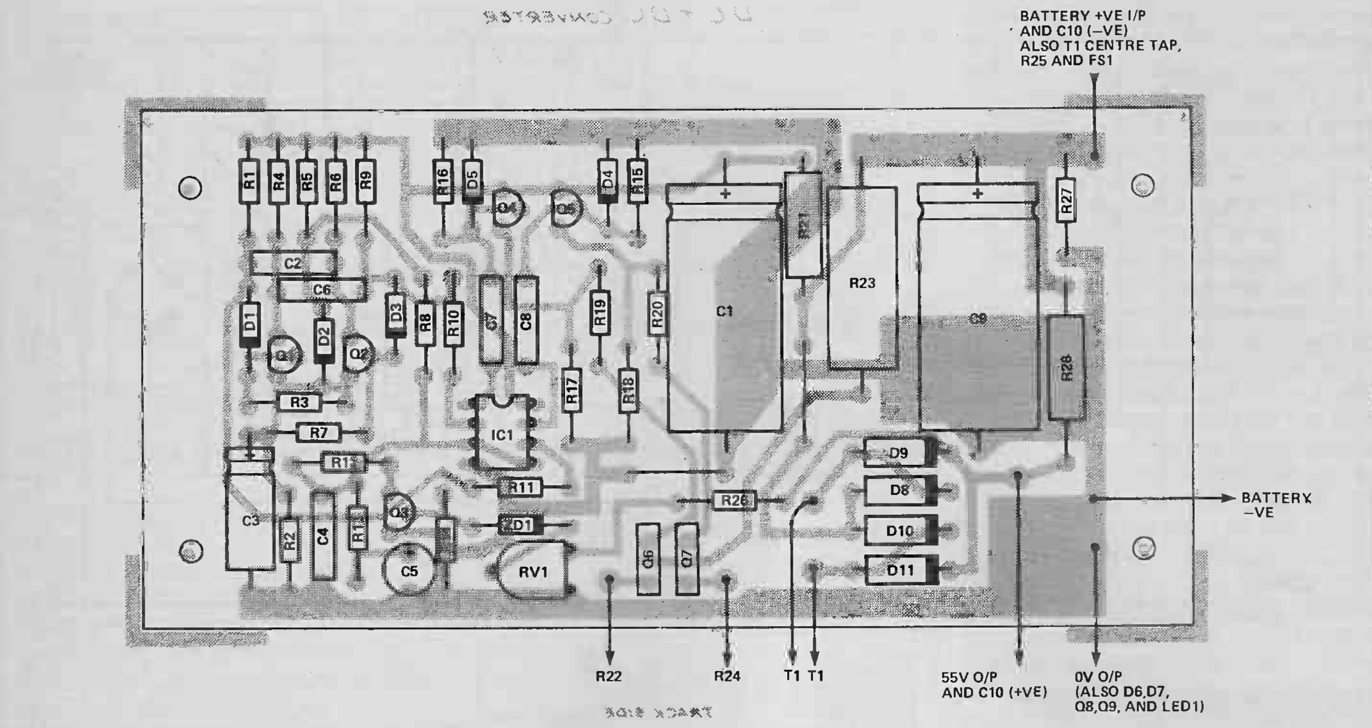


Fig. 14 Component overlay for the printed circuit board.

Construction

The component overlay for the converter is shown in Fig. 14. Assembling the board should present no problems but take care that the electrolytic capacitors, diodes and transistors are installed the right way around. The op-amp should be installed last of all and again, care should be taken that it is the right way around. A socket may be used for this component if desired.

The transformer winding details are given in Fig. 15, and the wire used should be 1mm enamelled copper. Most of the transformer kits available are supplied with a coil former which already has a 120/240V mains winding. This must be discarded and a new former made up from stiff cardboard. The E and I laminations are then fitted into the former and the completed core secured as recommended by the kit supplier, usually by means of end-cheeks which are bolted together through the laminations.

The DC converter is built into its own diecast metal case rather than being assembled in the same enclosure as the amplifier. This is advisable for safety reasons and

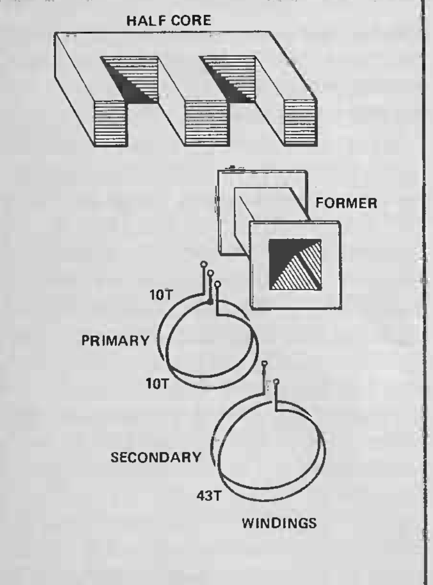


Fig. 15 Construction of the step-up transformer.

PARTS LIST

RESISTORS (¼W, 5% Unless otherwise stated)	
R1, 9	47k
R2, 5	2k2
R3, 7, 8, 10, 11, 14	100k
R4, 6, 27, 28	3k3
R12	22k
R13	27k
R15, R16	1M0
R17, R18, R25	1k5
R19, 20	1k0
R21	100R ½W
R22, 24	100 R
R23	10R 5W
R26	33k
RV1	10k horizontal skeleton preset
CAPACITORS	
C1, 9	2200u 16V axial electrolytic
C2, 6	33n
C3	100u axial electrolytic
C4	220n
C5	1u5 tantalum or radial electrolytic
C7, 8	100n
C10	2200u 64V axial electrolytic
SEMICONDUCTORS	
IC1	TL072
Q1-5	BC182
Q6, 7	BD537
Q8, 9	2N3055
D1-5	1N4148
D6, 7	1N4002
D8-11	1N5402
ZD1	10V 400mW zener diode, BZY88C10 or similar
LED1	panel-mounting LED
MISCELLANEOUS	
FS1	panel-mounting fuseholder and 15A fuse
SK1	red 4mm terminal
SK2	black 4mm terminal
SK3	2 (or 3) pole polarised connector of 2A rating or higher (eg., Bulgin P650 and P649 plug to suit)
SW1	12V 15A toggle switch
T1	see text

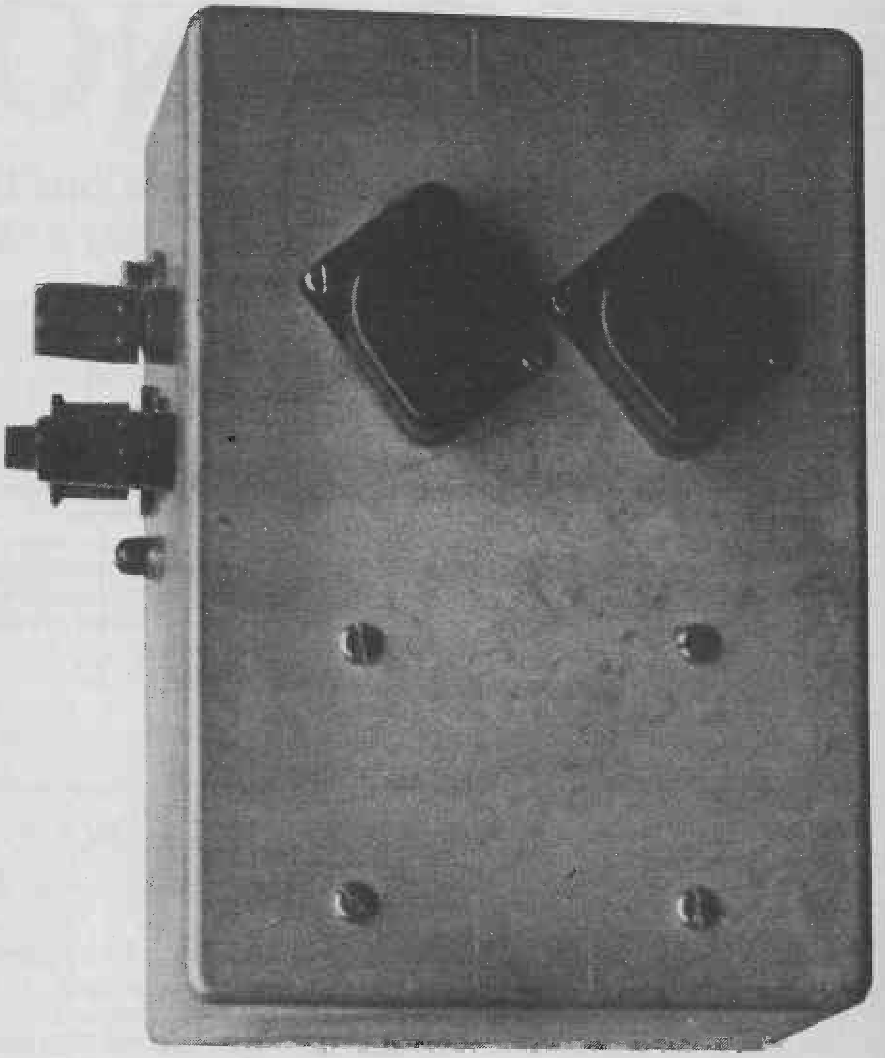
PCB; diecast box; capacitor mounting clamp (for C10); 8-pin DIL socket; clip-on heatsinksz for Q6 and Q7; plastic insulating covers for Q8 and Q9; PCB mounting pillars; nuts, bolts, etc.

also helps reduce the audibility of any transformer core noise. The power transistors are bolted to the case to ensure adequate heatsinking and a mica washer must be placed under each to provide electrical insulation. A plastic cover should be placed over each transistor to remove the risk of a short circuit should the transistor cases accidentally come into contact with any external metal object.

The positioning of the other major components within the case can be deduced from the external view in the photograph. When wiring up, bear in mind that quite heavy currents will be flowing and that too much resistance in the wiring will quickly produce an unnaceptably high voltage drop. The critical areas are represented by the heavy lines on the circuit diagram (Fig. 12). These connections should all be made using reasonably heavy-gauge wire and should not be any longer than is strictly necessary. The same goes for the wiring used to connect the converter to the car battery when in use.

Note that R22, R24, D6 and D7 are among the components which are not mounted on the board, and don't forget to attach clip-on heatsinks to Q6 and Q7.

Testing is a simple matter of connecting the unit up to a car battery and checking that an output of approximately the right voltage is present. If there is no output or the power transistors rapidly get hot, disconnect the leads from the battery immediately and check carefully for the fault. If all is well, adjust RV1 to set the output at exactly 55V and the converter is ready for use.



BUYLINES

The majority of the components for this project should be very easy to find. 100VA transformer kits are available from Maplin among others, and those who supply the kits can usually supply suitable winding wire. The PCB will be available from our PCB Service, see page 59.

Next month we will be presenting a 50W amplifier design produced by John Linsley Hood specifically for use with this converter and developed with outdoor PA systems very much in mind.

ETI

TV OR NOT TV

A 14-pin IC promises to change our television and video viewing habits utterly. Chris Giles examines it and its implications.

How many times have you been watching a videotaped TV programme and cursed as yet another string of adverts interrupts the action? You have three choices: sit through the messages (SUBSCRIBETO ETI); fast forward to the next bit of action (SBSCBTI); or watch only BBC. In view of the current debate about the BBC showing ads, the third solution may turn out to be not only restrictive but short-lived, too. There is now, however, an elegant answer to the question of how to avoid the ads — thanks to an ingenious piece of electronics.

This absurd situation has happened because technological developments have outstripped our legal, economic and social practices. . .

The Adzap was launched quietly in January this year by a formerly little-known West Country company, Specialist Semiconductors. In a moment, we'll take a look at how it works, concentrating on the remarkable new IC at its heart, but the controversy surrounding this device is well-worth a brief digression. For some time now, television companies have been disturbed at a growing tendency to use VCRs as a means of 'time shifting' — that is, watching programmes after their initial broadcast. In itself, this reduces the efficacy of TV as an advertising medium, since ads are carefully placed by agencies to attract the right audience at precisely the right time. Worse still, time shifters are known to avoid watching ads at all, if they possibly can. For the independent companies, a device like the Adzap devalues their major source of revenue. Even the BBC, whose present output includes numerous trailers, licence fee reminders and the like, are worried.

It's not just time shifting that produces the problem. The Adzap itself is a single-function device, intended only for use as a VCR add on. With a recommended retail price of £69.95, it is a piece of equipment many VCR owners would find attractive. But with a little extra circuitry and for a little extra cost, a device like the Adzap could be installed *inside* television sets to remove adverts or other broadcast material from the screen. The simplest sort of device could switch channels as soon as an ad came on. A more complicated arrangement might use a framestore to produce, say, a two-minute delay in screening received signals and edit out ads practically in real-time. A television set with that sort of facility would probably cost upwards of £2000 — even with memory as cheap as it is — but it is technically feasible, as we shall see. Television programme makers are so worried by these possible developments that they have formed a defensive organisation to fight for changes in the law to protect what they describe in publicity leaflets as 'the integrity of broadcast material".

This association, the Joint Organisation of Small-screen Production Houses (JOSEPH), has over the last few weeks been lobbying MPs in order, as they say, 'to obtain legislative clarification of the status of broadcast material'. Their aim — as stated in their leaflet 'Joseph's Coat — The Need For Protective Measures In the Television Industry' — is to introduce an Act of Parliament creating a class of licensed programme producers and expressly forbidding the unlicensed editing of broadcast material. Of course, such a law runs the risk of overkill — preventing users of VCRs from recording parts of a programme or even threatening our right to switch off the television.

This absurd situation has happened because technological developments have outstripped our legal, economic and social practices. The Adzap itself, currently unavailable in this country due to an injunction granted to the members of JOSEPH in the high court in February, can only work because the TV companies themselves introduced Teletext systems in an effort to boost their profitability. So how *does* it work?

How It Works

Happily, ETI received an Adzap shortly after its launch and before the device was withdrawn from the market. Our technical wizards have taken it apart and put it back together again. Although details of the new chip at its heart are being kept under wraps, we've been able to fill in some of the blanks. The secret is in the Teletext format.

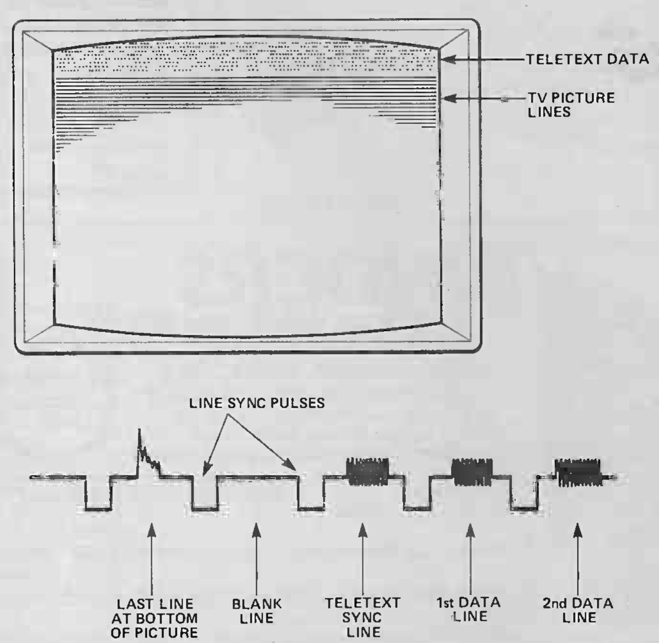


Fig. 1 Teletext data as seen on an ill-adjusted TV (a) and in relation to sync pulses and picture lines (b).

The Teletext system sends out all sorts of information on top of the ordinary television signal by utilising the normally unused top few lines of the frame sync and blanking regions of the signal, normally invisible on correctly adjusted TV receivers. The position of the Teletext data in a TV field (one picture) is shown in Fig. 1.

Thanks to extensive advertising, most viewers are familiar with the normal Teletext service of Ceefax, Oracle and 4-Tel, but it is not widely known that along with these magazine services, all the channels carry a lot of engineering and associated information, inaccessible to the domestic Teletext decoder because of its high transmission rate and its position in the sync.

For the independent companies, a device like the Adzap devalues their major source of revenue. . .

Most of this information is pretty uninteresting to us, being mostly digital transmitter keys, status information and channel locking codes. But some of the information can be really rather useful to the ordinary viewer — which is where the Adzap comes in.

Data Format

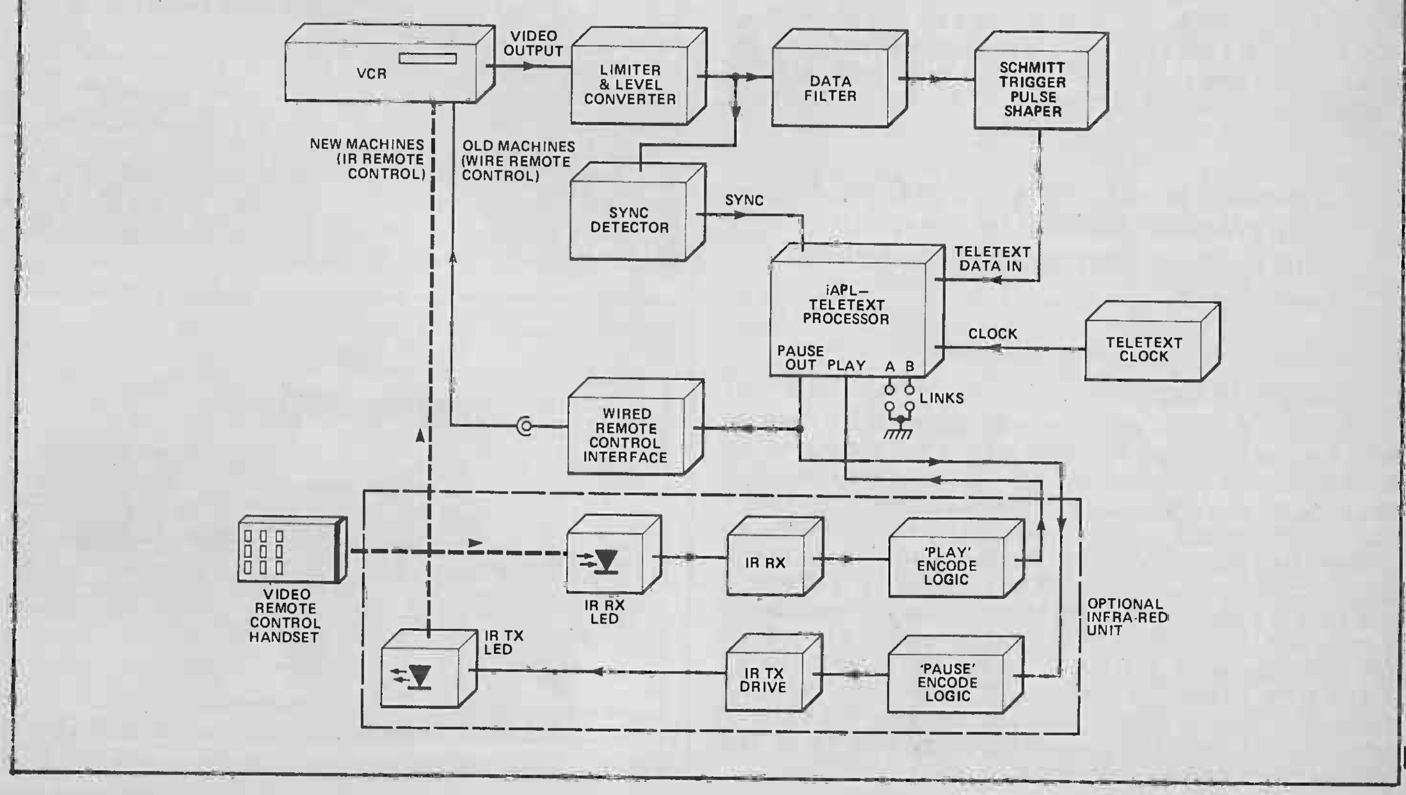
The second line of Teletext information carries the various system codes (the first is for synchronisation purposes only), and the fourteenth byte on that line is the Transmitted Signal Status Byte (TSSB). Even remembering its difficult location, the TSSB is a very underrated piece of information. Bits 1 to 4 contain the Material Content Register (MCR) which indicates what type of programme material is being broadcast. If it is set, then ordinary domestic material is on the air. If bits 1 and 2 are both zero, then engineering material is on the air. So-called 'intermediary material' (station ID, links, trailers, public information announcements and party politicals) is indicated by bits 1 and 2 being both low. Bit 1 information is used by automatic polling devices for audience research purposes, since bit 1 low material is specifically excluded from the viewer ratings.

MCR bit 1	2	3	4	Programme Type
0	0	0	0	close down — no broadcast
0	0	0	1	colour bars
0	0	1	0	black screen
0	0	1	1	full test card
0	1	0	0	future expansion — unused
0	1	0	1	station identification
0	1	1	0	party political broadcasts
0	1	1	1	links, trailers and public information
1	0	0	0	news, religion and current affairs
1	0	0	1	educational and scientific
1	0	1	0	children's programmes
1	0	1	1	chat shows
1	1	0	0	variety and music
1	1	0	1	drama
1	1	1	0	films
1	1	1	1	advertisements

Table 1 Material Content Register Bit Patterns.

Bits 3 and 4 are used to indicate the exact type of material being broadcast. All four bits set means a commercial break. Because of the lack of suitable decoders, all the information contained in the MCR has been of limited use. TV companies can easily find room and money for large boards of microprocessors, PALs, TTL devices and other chippery and, as noted above, MCR information has been quite liberally used for statistical investigations. The domestic user has not, so far, had access to the technology, largely for economic reasons. Conventional MCR decoders are bulky and expensive and, because of the high speeds and bandwidths

Fig. 2 Adzap block diagram.



involved in television signal processing, of very limited application.

Fast Byte

Now all that's changed, thanks to new developments in Indium Gallium Arsenside manufacture. At least three major manufacturers are offering custom chips in this material, making possible the design of very fast VLSI devices. Specialist Semiconductors of Gwent (with the aid of a Welsh Development Agency grant) have grabbed the bull by the horns with their design for the iAPL series of custom chips. This is the chip at the heart of the Adzap, a multi-standard bit-slice Teletext processor with on-board cache memory, provisionally type-coded the LF00 (R). The Adzap uses this chip to detect the status of the MCR byte and, with the help of some external circuitry, to pause a VCR during commercial breaks.

A block diagram of the Adzap system is shown in Fig. 2. (Under the terms of the injunction on the Adzap, we are not allowed to show a circuit diagram). As can be seen, the video output from the VCR is first limited and level-matched to the rest of the circuitry. Next, it is fed to the Very Fast Acting Sync Detector, another novel design using GaAs. The sync detection threshold needs to be about 10dB better than the average domestic Teletext decoder, in view of the increased speed and bandwidth demanded of this application. The output of the VFASD is converted to TTL levels and fed into the Sync Input of the LF00. Meanwhile, the output of the video processor block is fed through a high pass filter to remove the picture information, leaving just the Teletext data stream.

The edges of this signal are squared up by a high bandwidth Schmitt trigger and then presented to the iAPL device. The Teletext Processor examines the incoming Teletext signal, compares it with the clock input and locates the TSSB byte. The MCR bits are stored in cache memory and another part of the chip reads the data. If all four bits are in the correct logic state (see below), the internal MCR flag is set. This is detected and the Pause Out pin goes high. A high on Pause Out is fed back to the video recorder via one of several possible control routes detailed below.

Specialist Semiconductors of Gwent have grabbed the bull by the horns with their design for the iAPL series of custom chips...

Complete Control

If the VCR uses a wired remote controller, a suitable lead can be supplied with the Adzap (or, rather, could have been). An infra-red LED is also built-in to the front of the case for use with modern remote control machines. This will transmit a pause command to the VCR without cables. This IR facility allows for one simple refinement. If the Adzap is equipped with an optional IR receiver, it will detect a play signal from a tape being played back. If a tape recorded without the Adzap system is played back including ads (HAVE YOU ORDERED NEXT MONTH'S ETI?), the Adzap will detect the ads and transmit a fast forward command through the IR LED, returning to the play command when the MCR flag returns from ADVERT state.

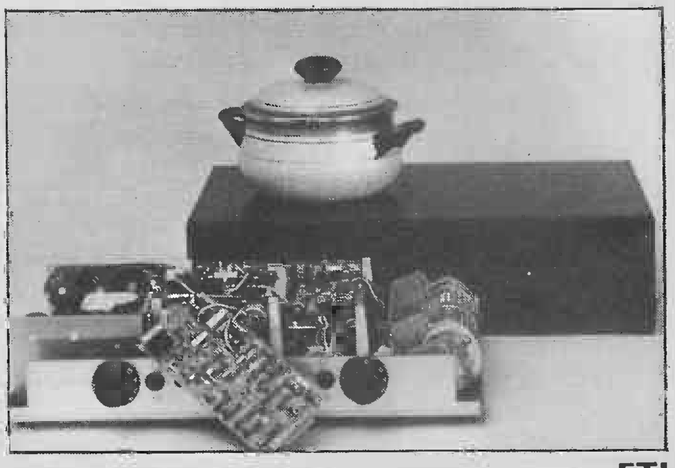
At present, the LF00 chip is a fairly simple 14-pin device — a requirement demanded by the economics of producing a domestic unit. But it is still versatile enough to be used with most TV systems.

Two inputs to the chip are used to determine the transmission standard in use. These wire links are set at manufacture of the Adzap (most countries don't change their TV system very often). Input A determines whether the negative modulation system (as used in the UK and most of Europe) or the positive modulation (French) system is in use, while input B determines 625/525 line compatibility.

The importance of knowing whether 525 or 625 lines are being used is fairly obvious, but the need for setting the moduation system is not so clear. When you consider the nature of the Teletext data it should become immediately apparent. With the wrong modulation sense selected, the MCR flag bit would be in the wrong logic state. If the modulation sense input was not provided, this inversion would have the unfortunate effect in France of preventing the VCR from taping any programmes and recording all the adverts (ACHETE ETI). Although anyone who has ever seen French TV would probably not consider this to be a fault.

A top-of-the range version may even be able to reset MCR bits, thus altering the character of broadcast material...

Future developments of the iAPL series of chips (already under way at Specialist's Gwent laboratories) should be able to automatically detect line and modulation standards. An on-chip microprocessor and ROM version is being considered with which it will be possible to program an evening's recording and, ultimately, viewing. By detecting MCR bit-patterns other than 'all set' one could, for example, avoid chat-shows or soaps. Specialist are, apparently, considering an iAPL chip with signal processing facilities able to discriminate programme material using combined analogue and digital techniques down to the level of Wogan or Dynasty. A top-of-the-range version may even be able to reset MCR bits during broadcast thus altering the character of the programme material itself. In this way, Benny Hill could be made funny. No wonder, then, that television programme producers are worried.



A prototype top-of-the-range Adzap.

ETI

MICRO-EXPANSION

Mike Barwise continues his series on up-grading 8-bit micros with a look at factors affecting reliability and the efficient use of hardware.

Reliability and efficiency are rarely spoken of in the context of the home micro. They are, of course, very subjective concepts. The only absolute is that no system is perfectly reliable and that all systems are more or less inefficient in some respect or other. The degree of reliability and efficiency expected depends on the nature of the application. In the design of the home micro, which is viewed by its producers essentially as a toy, little attention is usually paid to either.

Call Me Unreliable

The main source of unreliability is marginal hardware design. Apparent software unreliability is most often the indirect result of marginal hardware and, in any case, incorrect software (a bug) normally does the wrong thing very reliably!

Marginal hardware design is any configuration which causes your circuit to behave in a manner other than that expected. The most likely cause of such behaviour is a design element which fails to stick within the acceptable limits of some microchip parameter or other. The maximum drive capability of outputs, the voltage of logic levels at inputs, and out-of-spec or excessively noisy power supplies are examples of so-called DC parameters which can cause problems.

These are often glossed over by the 'string-it-together' brigade, but it is dangerous to look on microchips as uniform bricks that can be cemented end-to-end or side-by-side without any restrictions.

When adapting or enhancing a system the slowest component must be taken as a yardstick. For example, there has been a recent fashion for running Acorn Atoms at 2 MHz (double speed). It does just about work, but the OS ROMs in the Atom have a 450ns access time and the chances of misread data resulting in incorrect program execution are greatly increased (see Acorn User, December 1985).

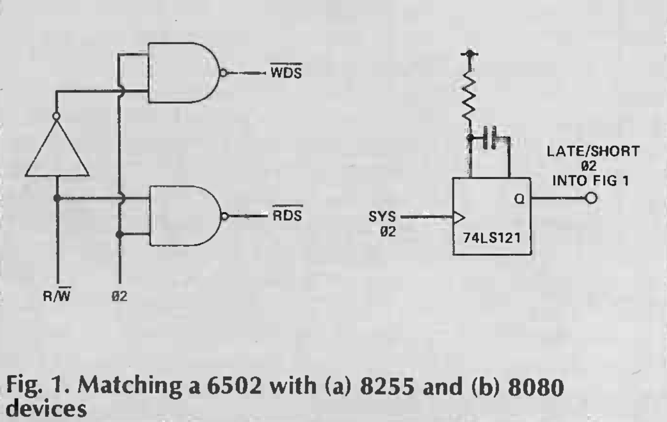


Fig. 1. Matching a 6502 with (a) 8255 and (b) 8080 devices

The real problem is that modern microchips are so forgiving. Very few examples of commercial LSI chips come anywhere near worst case parameters, so an out-of-spec circuit will often work — more or less. Only exhaustive testing or knockabout service life will show the weak spots but the urge to get on to the market before the competition makes this an un-economic proposition for most commercial companies.

If you are going to add your own expansion to a micro, you must really take note of reliability. It is not difficult. Just make sure that the worse case value for every specified parameter is at least 10% better than that specified and you won't go far wrong. This may mean some crafty support circuits if you are mixing chip families, but it's all good instructive fun.

Two fairly simple examples of mixed family device matching are shown in Fig. 1. The first is the standard way of providing a 6502 with a read strobe and write strobe for use with the Intel 8255 and similar devices (Fig. 1a). The second enables the 6502 to be used with certain 8080 devices, which will work with a short write strobe but need a data hold time about three times as long as the 6502 provides (Fig. 1b).

Watchdogs

A watchdog is a subsystem which determines whether or not the main system is functioning as expected, and aborts or corrects departures from the norm as soon as possible after they occur. Some larger systems use software checksums and the like, while advanced hardware watchdogs can actually keep track of the required operations and correct errors. The latter tend to be complex and need to be included in the original design of the system or, as is the current trend, into the CPU chip itself.

For post-implementation, the watchdog is an external discrete hardware solution. The simplest one is a counter which is reset by an event occurring periodically in a properly functioning system, and clocked by a regular event which occurs whatever the active system is doing. If the counter overflows — that is, it has not been reset for a period N — it is assumed that the system has gone wild. The resultant output is normally used as a general reset to the micro.

This type of watchdog can be implemented in a variety of ways. One method uses an address mapped write port which is periodically written to by software to prevent counter overflow (Fig. 2a). This demands that the watchdog be implemented before the software is written, and it also takes some processing time, although not a lot. Another approach is to attach a permanently enabled comparator to the CPU address bus, so that every time an address within a given range is generated, the watchdog is reset (Fig. 2b). The corollary of this is to

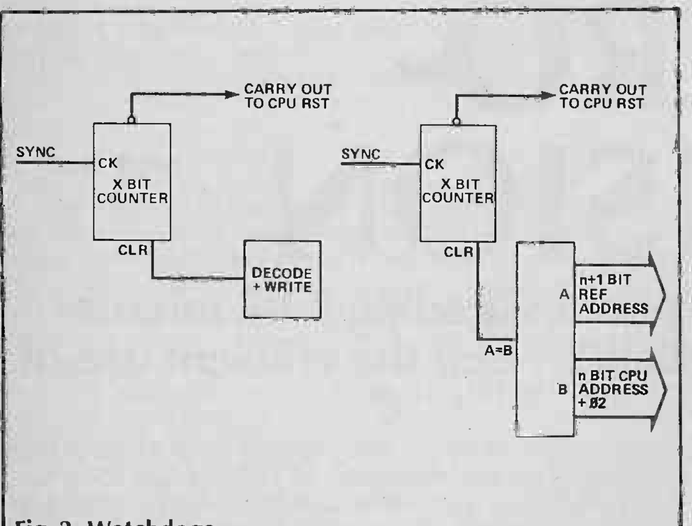


Fig. 2. Watchdogs.

look for addresses which should not be generated, and allow N of them to cause a system reset. This second approach is definitely the more flexible and is totally transparent to software. If you are really crafty, for a small penalty in terms of reliability, you could use write protected registers to hold the comparator settings so that individual applications packages could set the watchdog to suit themselves.

It is possible, with yet more ingenuity, to include a wrap-around RAM which stores the last couple of K of addresses presented to the bus so that the error point can be examined. The device is then, in essence, a logic analyser. A general schematic of such a system is shown in Fig. 3.

Assuming a 6500 based micro, the analyser address counters are advanced by the leading edge of the SYNC signal, which indicates an OP code fetch cycle. While SYNC is active, the Phase 2 clock is used to store the 16-bit CPU address in the analyser RAM. The resulting contents of the RAM amount to an execution address trace of the program. As the counters wrap around, only the last N-1 addresses are stored at any time, where N is the analyser RAM size. In this application, an overflow-type watchdog inhibits the address counters of the analyser either when or until the bus error occurs, depending on the mode of the operation.

In the HALT ON ERROR mode, which is the most useful but needs a little more logic, the address on the counters can be read to determine a 'cursor position' reference within the RAM. If you are very clever indeed, you will add a delay of several counts to the address clock inhibit, so that the start of the actual error is trapped in addition to the valid events immediately before it.

How Does Your Micro Go!

Even if you have the most reliable system you can to start with and your add-ons maintain the standards, you can still get very frustrated by inefficient operation.

We touched on the question of efficiency last month in discussing the modern massive business micro (ETI, March 1986). Efficiency is not the same as speed of operation. A lot of popular fast micros could be many times faster if their operation were tidied up and rendered more efficient.

Efficiency may be roughly defined as the achievement of the best compromise between the amount of code used and the amount of time taken in the execution of a particular task. I stick my neck out here, but I reckon most commercial applications packages are grossly inefficient. It is noticeable that the package size doing any given job has grown and grown over the last few years. The extensive use of high level compilers and the almost entirely software-orientated application has caused manufacturers to reach for 16 and 32 bit CPUs running in the 10MHz region to compensate for the inefficiency of the results.

Inevitably, the code used for driving hardware must be really efficient, particularly where bulk data transfer deivces such as disk controllers, data loggers, etc, are concerned. Often you have only a few microseconds to perform an operation if the system is to stay running, so well written assembler and the intelligent sharing of tasks between software and hardware are the order of the day.

There are, of course, times when the choice between hardware and software is obvious. Nobody in their right mind would control a floppy disk drive enitrely by host software when they could use an FDC interface of some sort. (Oops! Sorry. I think someone did!). The necessary functions were itemised in part one, and it is obvious that the CPU would get little else done. On the other hand, a low cost universal EPROM programmer is better off using software and passive control latches, as the number of permutations of mode would make a hardware solution pretty complex and the speed required is very low (20Hz).

There is a much less clearly defined area where the use of either hardware or software could appear to be the obvious solution, depending on your personal bias. A lot of designs that look neat at first sight can be improved upon. The real answer is to consider all the alternative combinations of hardware and software that present themselves, and to avoid the temptation to use something because it's available or to be satisfied with an answer just because it works.

A real-life example of this from my own experience concerns the WD1770 Floppy Disk Controller. My own board was to be used in either an interrupt or polled interface configuration. The interrupt outputs are always operational, so the interrupt driving solution is obvious. The WD1770 status register contains bits which indicate data requests (comparable with the DRQ interrupt) and FDC busy (approximating to the INT terminal interrupt).

At first sight this presents no problem, but the design was for 6500 series processors, and the bit positions (0

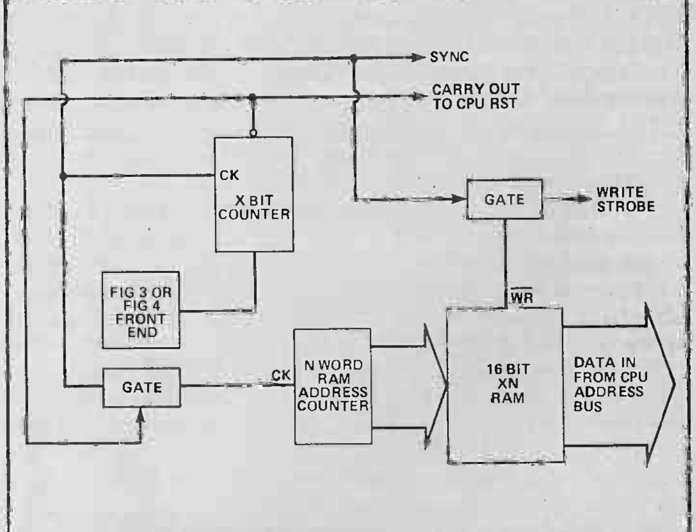


Fig. 3. Logic analyser schematic.

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FEATURES

PROJECTS

Due to illness, a number of items scheduled for this month's ETI have had to be held over. We especially regret not being able to run the constructional details for the digital sound sampler digital board which should now appear next month. Our apologies to our readers.

ETCETERA

INFORMATION

and 1) are exactly the wrong bits for convenient fast polling during data trasnfer. Furthermore, the data request bit doubles as an index pulse indicator for non-data commands.

Worse is to come, as supplementary information showed that 64 microseconds must be allowed between writing a command and reading the DRQ bit in the status register. The FM byte time is 64 microseconds, so on read and write operations you can't poll the DRQ bit for the first byte to be transferred.

The solution was to tie the real interrupts to the inputs of an addressed external buffer register, put them in bits 6 and 7 (which suits 6500 series devices just fine) and poll that! The register was needed anyway, to hold drive and side parameters.

Examples of this type are endless, and I won't bury you in details, but I think this shows how a little rethinking can not only improve performance, but also make the software writer's task easier.

Interrupts

When you get around to the multi-processor configuration outlined in part one, each intelligent peripheral or module will have to be able to talk to any other as well as to the real world. Keeping control in the time domain is then your problem. Yes, I can hear you muttering "interrupts; interrupts ...". Are interrupts the universal best answer, though?

A superficial look will suggest that you can do anything with interrupts, but you can't. They are probably the most over-used and misused features of the eight-bit generation of processors.

When you need occasional asynchronous transfers of a few bytes between the micro and the outside world, there is nothing more useful than the interrupt. There are a few rules to observe, but in general the maskable type of interrupt is very easy to use for this kind of work.

Interrupts cause most problems when they are used to handle large volume transfers at high speed, or when used to drive multiple asynchronous events simultaneously without proper prioritising. The two main points to remember are that:

(1) their response is not instantaneous, even in their most efficient implementation;

(2) they are demands, not requests. You may be able to inhibit service, but you cannot abort a recognised interrupt except by returning as normal.

The underlying point is that interrupts absorb time. In 6500 terms, an interupt which points directly to an RTI (return instruction) absorbs around 20 clock cycles.

Suppose for convenience that we are running a 6502 at 1MHz (BBC 1MHz bus, Acorn Atom, Tangerine, etc.). An MFM (double density) disk interface requires data transfers every 32 microseconds, and each transfer must be completed within the first 28 microseconds.

Of the 20 cycles required by the interrupt 14 occur before the address bus points to the first byte of user code, and six after the last user instruction to return to the point where the interrupt occurred. Half of the allowable byte transfer time is thus taken up before any useful work is done if interrupts are used to synchronise the data transfer. I guarantee that a 1MHz 6502 cannot pass a byte and increment its data vector within the remaining time. On the other hand, polling by use of the BIT instruction and a relative branch takes only seven clock cycles, so you might just get away with it. Even more cunning would be the use of the 6502 SO (Set Overflow) pin, which would allow the polling loop to take over three cycles (see Acorn User September 1985).

Using our floppy disk controller example, suppose we find a way to drive data transfers under interrupt. Even in FM (single density) we must transfer a byte every 64 microseconds, and the best code I have come up with takes about 28 clock cycles to perform the transfer. At 1MHz this is 28 microseconds, and we must add to this the 20 microseconds required to fetch the interrupt vector and return. The 58 microsecond result is, of course, three quarters of the total CPU time, so any foreground task might as well just stop for the time being. The moral of all this is: keep your interrupt service time short in proportion to total CPU time, ideally never more than ten per cent.

Multiple Interrupts

So far we have assumed only one device as a source of interrupts. This is a very severe limitation. If we require a real-time plot of a graph of selected data from a data logger, we would need interrupts to drive keyboard input, input from the logger and output to a pixel plotter simultaneously. Each of these devices has an infrequent but time-limited data transfer window caused by an asynchronous outside world event.

This introduces a problem. How do we control the various interrupts to ensure that they all get services whenever they need it?

As soon as we have more than one source of interrupt, there must be a very clearly defined method of identification of the device currently in need of service. The very simplest answer is to provide all the interrupting devices with a register bit which acts as an interrupt flag. On interrupt, the device registers are each read in turn, and the first device with set flag is serviced. As soon as an interrupt is serviced, the device flag is cleared by the CPU, so another interrupt can be identified by continuing the flag register scan.

The big catch to this simple approach is that it is terribly slow. The more complex the system and hence the more devices it contains which can cause interrupts, the longer the time taken to establish which device needs service. You will rapidly reach a point where, by the time the correct device has been identified, its need for data is long gone and the system fails.

We can solve this by providing a separate interrupt for each device. One option is to design the hardware so that each module puts a unique address into the register at a fixed point in the address map. A set of bus lines would have to be dedicated to this, but as you would rarely need more than 16 interrupt driven devices in an 8-bit system, only four lines would need to be allocated. Now the CPU has only to read the contents of a fixed single register to determine where to go for the device driver code. The time overhead is short and constant, regardless of the number of devices generating interrupts.

But what happens if two devices generate interrupts simultaneously? Obviously, there will be contention between their output codes, and the register contents will be invalid.

There is a very simple hardware solution to this problem. A Priority Encoder (for example, a 74LS147 and 74LS148) provides a binary coded output which depends on the most significant active input at any time. If, for example, you drive inputs 0, 3 and 6 active, the outputs read six and the state of inputs 0 or 3 will have no effect until input 6 has been de-activated. Thus no contention can occur in our interrupt register. The priority encoder will need a bus line allocated to each interrupt input so it will use up more space, but it prevents so many problems that it is well worthwhile.

ETI

CONSTANT CARE

Audio designer Graham Nalty has been investigating constant current sources. This is his report.

Constant current sources, like individual components, are often taken for granted in electronic design. If we wish to keep the current in part of a circuit constant, we simply make up a constant current source to a known recipe and assume it does its job. But the constant current source is a complex circuit in its own right, and requires quite detailed specification to describe its performance.

Constant current circuits are used in a number of situations:

1) On equipment subject to wide fluctations of supply voltage, where we wish to maintain adequate current for operation of the circuit at the lowest suppply voltage, but must keep heat dissipation within limits at the highest;

2) In amplification circuitry, a constant current load increases the gain of an amplification stage and enables high open loop gain to be achieved in a feedback amplifier;

3) In instrumentation amplifiers, the use of a constant current source to supply the input differential pair decreases the error at the output due to a common mode input voltage. When a sensor with a balanced output is used, the interference picked up by the cables is a common mode signal and will be rejected by an amplifier with a high common mode rejection ratio;

4) A constant current source in an audio amplifier reduces the effect of power supply ripple on the audio signal. In practice, the ripple we wish to eliminate is not the AC ripple from rectification circuits, which is not really a problem but the ripple caused by other parts of the amplifier drawing current from the power supply due to the signal;

5) Constant current circuits are used in discrete regulator circuits and their use improves the line regulation, reducing the change in output volgage due to a change in input voltage.

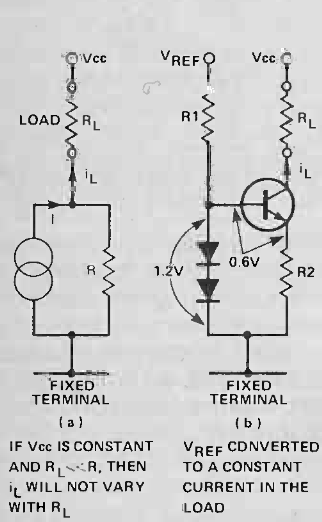


Fig. 1 Two and three terminal constant current sources.

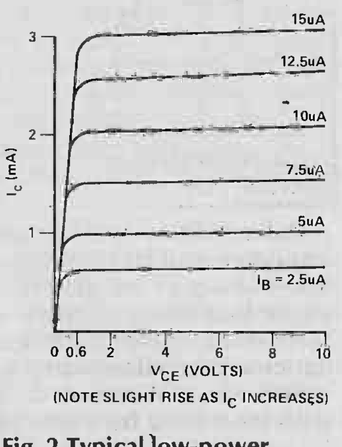


Fig. 2 Typical low-power transistor collector characteristics for a common emitter circuit.

Terminal Zone

Most constant current sources are either two or three terminal devices. The two terminal device may be regarded as a resistor substitute comprising a current generator and high value resistor in parallel (Fig. 1a). The three terminal device has a third pin which needs to be held at a reference voltage in relation to the fixed terminal (Fig. 1b shows a typical example).

To assess performance, we need to measure how constant the current delivered is in relation to the applied voltage and the other variables. For a two terminal device, we can measure the change in current in relation to the change in voltage across it which gives a figure for the source's dynamic impedance, Z_D. The dynamic impedance, which can be considered a measure of the way in which a device's impedance changes under different operating conditions, is equal to the change in voltage across a source, dv, divided by the associated change in current, dI. A good constant current source will offer substantially the same current over wide range of applied voltages. So, dI will approach zero over the device's useful range and the dynamic impedance will approach infinity. In fact, an ideal constant current source would appear as an infinite impedance delivering a finite current.

The collector characteristic of an ordinary transistor in common emitter mode (Fig. 2) shows that it approximates an ideal constant current source. In fact, the common emitter circuit appears as an impedance of something in the order of 10k — a figure derived from the slope of the characteristic curve which is the common-emitter output conductance, h_{oe}:

$$h_{oe} = dI_C/dV_{CE}$$

and

$$Z_D = 1/h_{oe}.$$

The common emitter circuit is a three terminal constant current source, the base having to be held at a constant potential in order to achieve a constant current through the collector-emitter circuit. The series diode arrangement of Fig. 1b is designed to hold Q1 base at a suitable constant potential. This sort of arrangement is fairly complex — certainly when you consider that some designs use nothing more than a high value resistor as a constant current source. It is, however, to be preferred over simpler arrangements if only because a considerably lower voltage is required to deliver a useful current. In the tables that follow, the dynamic impedance of a constant current source is compared to the equivalent resistor value. This latter is the resistance which will deliver a current equal to that produced by the constant current source when connected across the DC Power rails. As will be seen, an active circuit displays a far higher impedance at a given current and voltage than would be associated with a simple resistor.

The major problem with the Fig. 1b circuit is to make sure that the transistor base voltage actually is constant. If the reference voltage is taken from power rails, as it usually is in audio designs, there is almost certain to be variation. Real, rather than ideal, diodes will respond to variations. To judge the performance of a constant current circuit, then, it is important to see how it responds to changes in reference voltage. I have done this with the aid of the arrangement shown in Fig. 3.

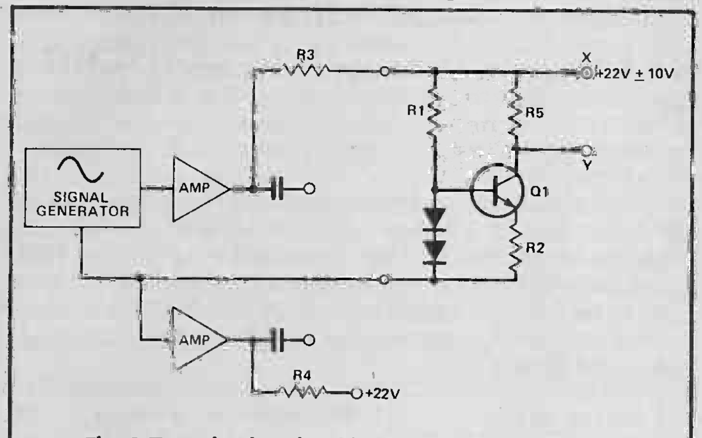


Fig. 3 Test rig showing Fig. 1b circuit in position.

Testing Times

The circuit of Fig 1b is typical of the simplest and most widely used constant current sources. Resistor R1 and diodes D1 and D2 hold the voltage at a the base of Q1 at 1.2V above the fixed terminal. Allowing for V_{be} of 0.6V at Q1, we have 0.6V across resistor R2. This sets the current through Q1 collector at (0.6/R2) A.

To test the circuit we need a stereo power amplifier with a very clean ripple-free power supply. The amplifier used in actual tests was a Classic series 2 A25 which uses a single rail supply and has a DC output level before the output capacitor of 22V. By feeding a sine wave of 7V rms (20V peak-to-peak) to the constant current source under test, the voltage across it is made to vary from 12V to 32V. This variation also affects the reference terminal.

Most constant current sources deliver something in the order of 1mA, and this was my target value for tests and comparisons.

R3 and R4 are low value resistors (20R) to protect the amplifier — and the constant current circuits — from damage due to accidental shorts whilst making measurements. R5 is the test resistor which was arbitrarily set at 4k7. The current was measured with a digital voltmeter across R5. Change in current could also be observed using an oscilloscope in the AC position across R5.

The test configuration is not a practical constant current arrangement. No practical circuit would actually look like Fig 3. However, it serves to gauge the performance of the constant current source under widely differing conditions. The results of the tests should be considered comparatively rather than as absolute measurements of practical performance.

Table 1 shows the results of measurements taken on the Fig. 1b circuit. The first point to note is that variations in the bias resistor, R1, have hardly any effect on overall dynamic impedance. This is because the dynamic impedance of the diodes is inversely proportional to the current through them — so that increasing R1 has the effect of increasing diode dynamic impedance and holding the base bias voltage constant.

R1	R2	R5 V DC	Q1 mA	R5 V AC	Z_D	R Equ.
220k	470R	4.2	0.9	1 p-p	94k	24k
100k	470R	5	1.06	1.1	85k	21k
47k	470R	5.8	1.23	1.1	85k	17k8
22k	470R	6.4	1.36	1.1	85k	16k3
J503	470R	5.6	1.19	0.033	2M7	18k5
47k	1k0	2.5	0.53	0.48	196k	40k
47k	220R	11.8	2.5	2.1	42k7	8k8

Table 1 Dynamic impedance (Fig. 1b) — AC signal, 1kHz.

The dynamic impedance figures in the table were calculated by dividing the load resistance, R5 (=4k7), by the proportion of AC voltage across the source as a whole appearing across R5. Or, to put it more simply, by dividing the AC voltage across the constant current source by the AC current in the load. While this cannot be considered to be 100 per cent theoretically sound (for one thing, the figures invovled are averages rather than instantaneous values), assuming a linear response in the circuit and a stable and regular AC signal it does give a realistic estimate. The equivalent resistor value was calculated, as mentioned above, by dividing the DC voltage across the source by the resultant collector current.

R6	R8	R7 V DC	Q3 mA	R7 V AC	Z_C	R Equ.
47k	470	6.2	1.3	0.28 p-p	330k	17k
22k	470	6.3	1.34	0.28	330k	16k4
220k	470	5.7	1.22	0.26	360k	18k4
J503*	470	6.1	1.3	0.004	23M5	16k9
74k	470	12.9	2.7	0.52	170k	8k5

*Compare with Table 3.

Table 2 Dynamic impedance (Fig. 4) — AC signal at 1kHz.

Having noted the effect of changing R1's value, we should observe that the fluctuating voltage across R1 due to the signal injected by the sine-wave generator has considerably more impact. Since the emitter resistance of Q1 is small compared to R2, the AC voltage seen at the base of Q1 appears across R2, affecting the AC voltage across R5 and, therefore, the dynamic impedance. Replacing R1 with a constant current diode, J503 (rated at 0.56mA±20%), had the effect of considerably reducing the AC voltage across R5 and therefore giving something very much better in the way of dynamic impedance.

Another common constant current circuit is the two transistor configuration of Fig. 4. The current is equal to V_{be} (Q2)/R8. The collector of Q2 is held at 0.6V above its base by the base-emitter voltage of Q3. Table 2 shows that this circuit will have a higher dynamic impedance than the circuit of Fig. 1b, because feedback action

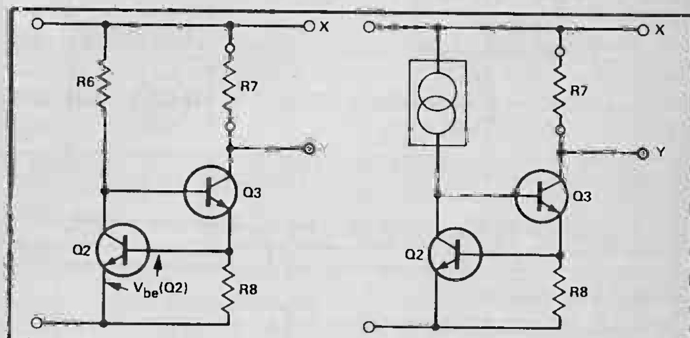


Fig. 4 Two transistor constant current source. **Fig. 5 Using a constant current diode improves performance.**

stabilises Q3's base more effectively. The dynamic impedance of the two transistor circuit is 20 times the impedance of its equivalent resistor compared with about four times for the single transistor circuit. Further tests showed that with resistor R6 in circuit the impedance was constant up to 100kHz. If R6 was replaced by a J503 constant current diode the impedance decreased with increasing frequency (Table 3).

Constanter Still

A common economy when using several constant current sources is to share the base bias voltage between them. The two circuits in Fig. 6 show ways in which a shared bias might be employed with discrete op-amps in, say, an audio power-amp. We shall ignore the effects of

AC kHz	R7 mV AC	Z_D
(1)*	(4)	(23M5)
(2)	(5)	(18M8)
5	8	11M7
10	13	7M2
20	25	3M7
40	53	1M9

*Compare with Table 2.

Table 3 Dynamic impedance against frequency (Fig. 5).

power supply ripple in the circuit and consider the effect of large output voltage swings. We can use the circuit of Fig. 7 to make the measurements we require. R9 and R11 are the load resistors, used for measuring.

To represent the circuit of Fig. 6a, we supply points A and B with 22V from the unused channel of the power amplifier and point C with 22V ± 10V p-p. An oscilloscope will show a voltage of 3mV p-p at the base of Q6 and also across R12. A 20V p-p voltage at the output will change the current in the long tail pair by 3mV/R12(=6.4μA p-p). If we supply the sine wave voltage to point A, while keeping B and C at 22V, we measure a voltage of 3mV p-p across R13 but the AC waveform at Q4 base or across R11 is not visible on the 'scope — it is well below 1mV. The point of this test is that the circuit of Fig. 6a does allow a small measure of feedback from the output to the input of the long tailed pair, but the feedback is considerably reduced in the circuit of Fig. 6b.

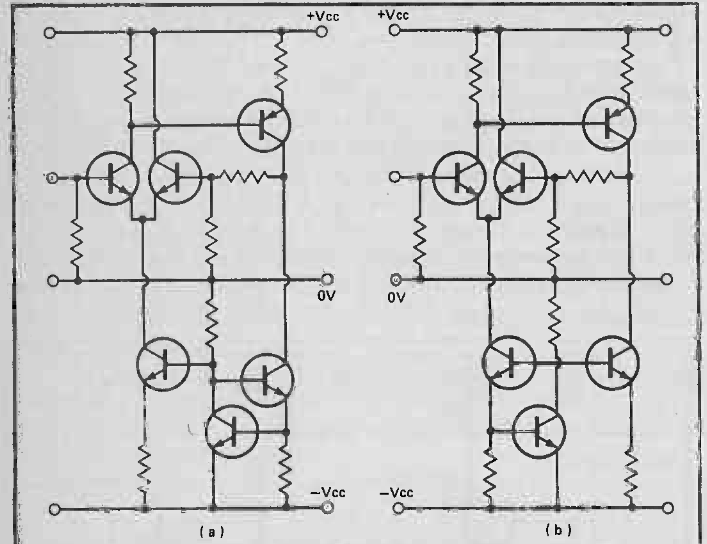
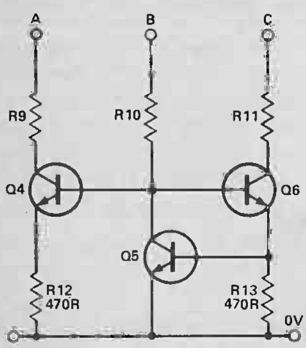


Fig. 6 Two designs for a discrete op-amp using CCS.

Fig. 7 Two sources using a single biasing arrangement.



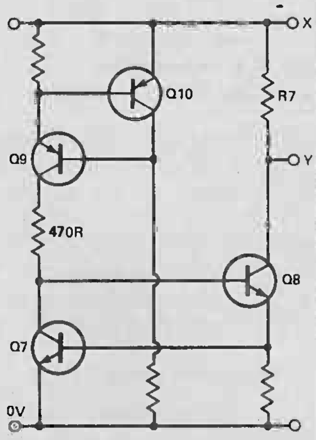


Fig. 8 You can replace the bias resistor in Fig. 4.

The circuit of Fig. 8 is a natural progression from the circuit of Fig. 4. Tests on the two transistor circuit showed that replacing the resistor R6 with a constant current source greatly increased the dynamic impedance. Constant current diodes are quite expensive and Q9 and Q10 are an alternative at a lower cost. A particular feature of this circuit and the similar circuit in Fig. 9 is the use of 470R resistors between the separate constant current sources. In the test circuits, omission of these resistors produced results which were rather doubtful. The 470R resistors were chosen because they were closest to hand and of a value which would enable the AC voltage across them to be measured on a 'scope. Their position produced credible readings, shown in Tables 4 and 5, for reasons which are unclear. A similar resistor in series with the J503 in the circuit of Fig. 5 had no apparent effect.

AC kHz	R7 mV AC	Z_D
5	16	5M8
10	18	5M3
20	25	3M7
50	52	1M8
100	100	0M9

Table 4 Dynamic impedance against frequency (Fig. 8).

FET Up

Field effect transistors can also be used to generate a constant current. In the N-channel JFET, the gate is held at a voltage negative to the soruce and the value of this voltage, V_{gs}, determines the current, I_d flowing from the drain to the source.

Figure 10 shows a JFET connected as a constant current source. The current flowing through RV1 raises the voltage of the source above the gate. The transfer characteristic of a JFET suitable for constant current generation operating with a drain-source voltage of 15V is also shown.

The cut-off voltage, $V_{gs\ (off)}$, is reached when I_d= 0 and is, in the example −4V. The saturation drain current, I_{dss}, is the drain current when V_{vs}=0 and is 8mA (In the data books, a JFET such as the J232, which is one of the best for constant current circuits, has $V_{gs\ (off)}$ specified to be between −3V and −6V and I_{dss} to be between 5mA and 10mA).

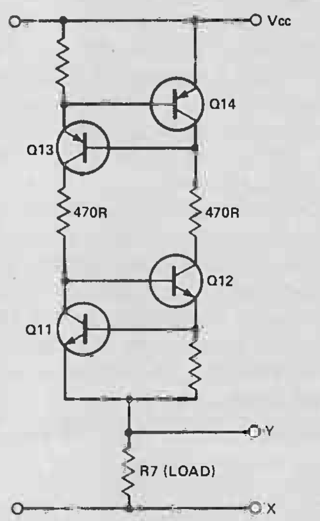


Fig. 9 Another configuration providing constant current biasing for a CCS.

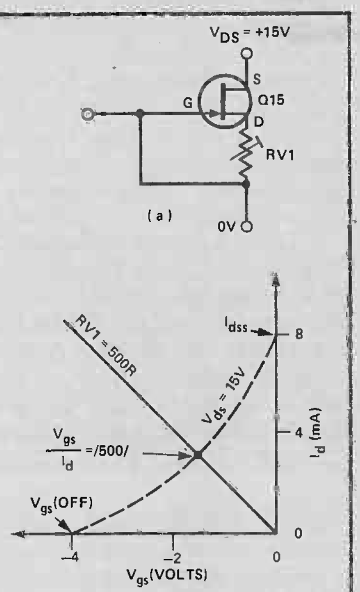


Fig. 10 Using a JFET as a constant current source.

If we look at Fig. 10b and draw a 'load line' for RV1 = 50 ohms, it intersects the transfer characteristic curve where V_{gs} is just over −2V and I_d just below 4mA. These values satisfy all the circuit parameters. The output charateristic curves in Fig. 11 show that drain current remains almost constant for large changes in V_{ds} while this voltage is above 4V. In our case, V_{ds} was 15V, so the circuit will deliver a constant 4mA.

AC kHz	R7 mV AC	Z_D
(2)	(8)	(1.2M)
5	14	6M5
10	27	3M5
20	57	1M6
50	120	0M8

Table 5 Dynamic impedance against frequency (Fig. 9).

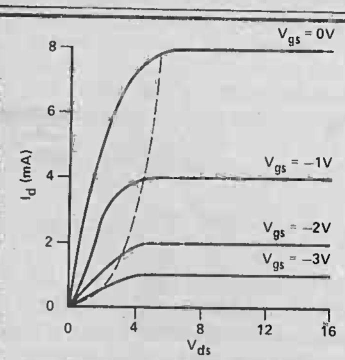


Fig. 11 N-channel JFET drain characteristics. (cf. Fig. 2).

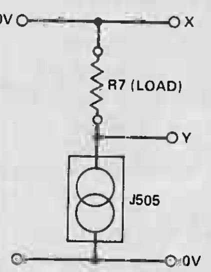


Fig. 12 A J505 constant current source.

Because of the wide spreads in V_{gs} (off) and I_{dss} for different FETs, a variable resistor is used to set the current. This is not very satisfactory as it adds extra time and costs in manufacturing. An alternative is to manufacture FETs with built in resistors to give a fixed current from a two terminal device (Fig. 12). These can be obtained from Siliconix for currents from 200mA to 5mA. Tolerances vary from 10% for the CR022-CR470 devices down to 30% for the J553-J557 devices.

Table 6 shows regulator current and dynamic impedance for five randomly selected J505 devices. Table 7 shows the variation in dynamic impedance with frequency for one sample.

R7 V DC	1 mA	R7 mV AC	Z_D
3.0	0.83	40	2M4
4.1	0.87	40	2M4
3.9	0.83	33	2M8
4.4	0.94	48	1M9
4.8	1.02	45	2M1

Table 6 Performance of five sample J505 devices — AC signal at 1kHz, circuit of Fig. 12.

The variable resistor in a FET current source could also be replaced by a constant current device (Fig. 13). Table 8 shows the dynamic impedance of a J505 with a J232 FET at various frequencies. A point to note is that in this configuration the voltage across the J505 (measured with a DVM) was 2.6V and 2.9V in two separate samples. The impedance of the J505 would be much higher at higher voltages, but the combination of, effectively, two FETs gives a very high overall dynamic impedance (Table 8).

AC kHz	R7 mV AC	Z_D
5	40	2M4
10	42	2M2
20	45	2M1
50	77	1M2
100	140	0M7

Table 7 Dynamic impedance against frequency (Fig. 12).

AC kHz	R7 mV AC	Z_D
(2)	(3)	(31M)
5	5	16M
10	11	8M5
20	23	4M0
50	60	1M6
100	130	0M7

Table 8 Dynamic impedance against frequency (Fig. 13)

The circuit of Fig. 14 uses a bipolar based constant current soruce in place of the J505. The results were not significantly different from those of the previous, simpler, circuit (Table 9).

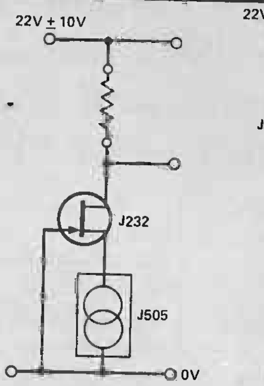


Fig. 13 Replacing the variable resistance in a JFET CCS configuration improves reliability and performance.

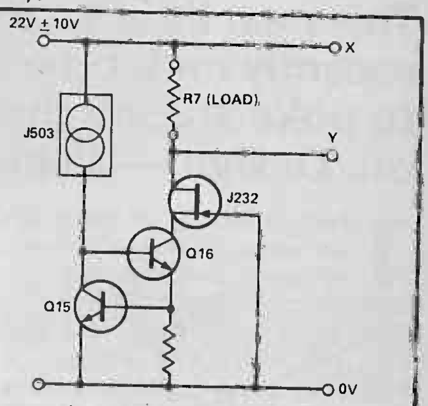


Fig. 14 Composite CCS using bipolar and FET devices.

AC kHz	R7 mV AC	Z_D
(2)	(3)	(31M)
5	7	13M
10	14	6M5
20	28	3M3
50	65	1M4
100	105	M9

Table 9 Dynamic impedance against frequency (Fig. 14)

Conclusion

A number of general trends seem to emerge from these investigations. Firstly, all the circuits revealed that dynamic impedance falls with increasing frequency — in some cases, very considerably. Secondly, the dynamic impedance of the more complex circuits is much greater than the simple one and two transistor regulators. Not unexpectedly, the performance of three terminal sources can be very ineffective when variation occurs between the reference voltage and the 'fixed' terminals of the device, and this can have significant consequences.

When designing current sensitive circuits, we must realise a further disadvantage of using three reminal regulators. The reference terminal carries a current which has to be taken somewhere. If it is take to ground, it should not share the signal earthing arrangements as it may contain ripple which will modulate the signal due to the impedance of the earth returns.

Perhaps the most important consideration is cost. It is clear that bipolar current regulators are inexpensive, but use a substantial number of components to achieve effective results. FET current regulators, on the other hand, are fairly expensive, but can achieve excellent results with simple circuitry and a low component count.

ETI

MTE CIRCUIT PROBE

This neat little board conforms to the four boards we produced recently under the title Modular Test Equipment. You can use it to poke around digital and analogue circuits to see what's going on. Design — as ever — by Mike Meakin.

The two circuits on this board are separate and only share common power supply connections. Apart from being good value-for-money, this means that they will be described individually.

Logic Probe

This circuit gives an audible and visual indication of correct TTL/CMOS logic levels. A pulse catcher circuit can capture fast pulses and either expand them or hold them in a memory circuit. This is particularly useful for detecting glitches or spikes not easily visible on an oscilloscope.

The audible indication allows the user to concentrate on holding the probe without having to glance at the LEDs. It can even provide a somewhat un-musical signature analysis!

Note that the top pin of SK1 must be connected to the positive supply of the circuit under test. This ensures that the CMOS thresholds of 30% and 70% V_{cc} are referenced to the supply of the circuit under test. When testing TTL circuitry on the standard 5V supply the thresholds are set to 0.8V lower and 2.25V upper. Again the top pin of SK1 should be connected to the 5V TTL logic supply to establish these thresholds, although if not connected the levels will be approximately correct.

In order to achieve a flexible input arrangement some compromises are necessary and the input stage is somewhat limited in speed. The response time of the LM311 is 200ns and this will be the shortest pulse that can be reliably detected. This is quite adequate for most CMOS and TTL circuitry used by the average hobbyist and represents a 5MHz signal.

Analogue Probe

This handy circuit is the equivalent of the logic probe but

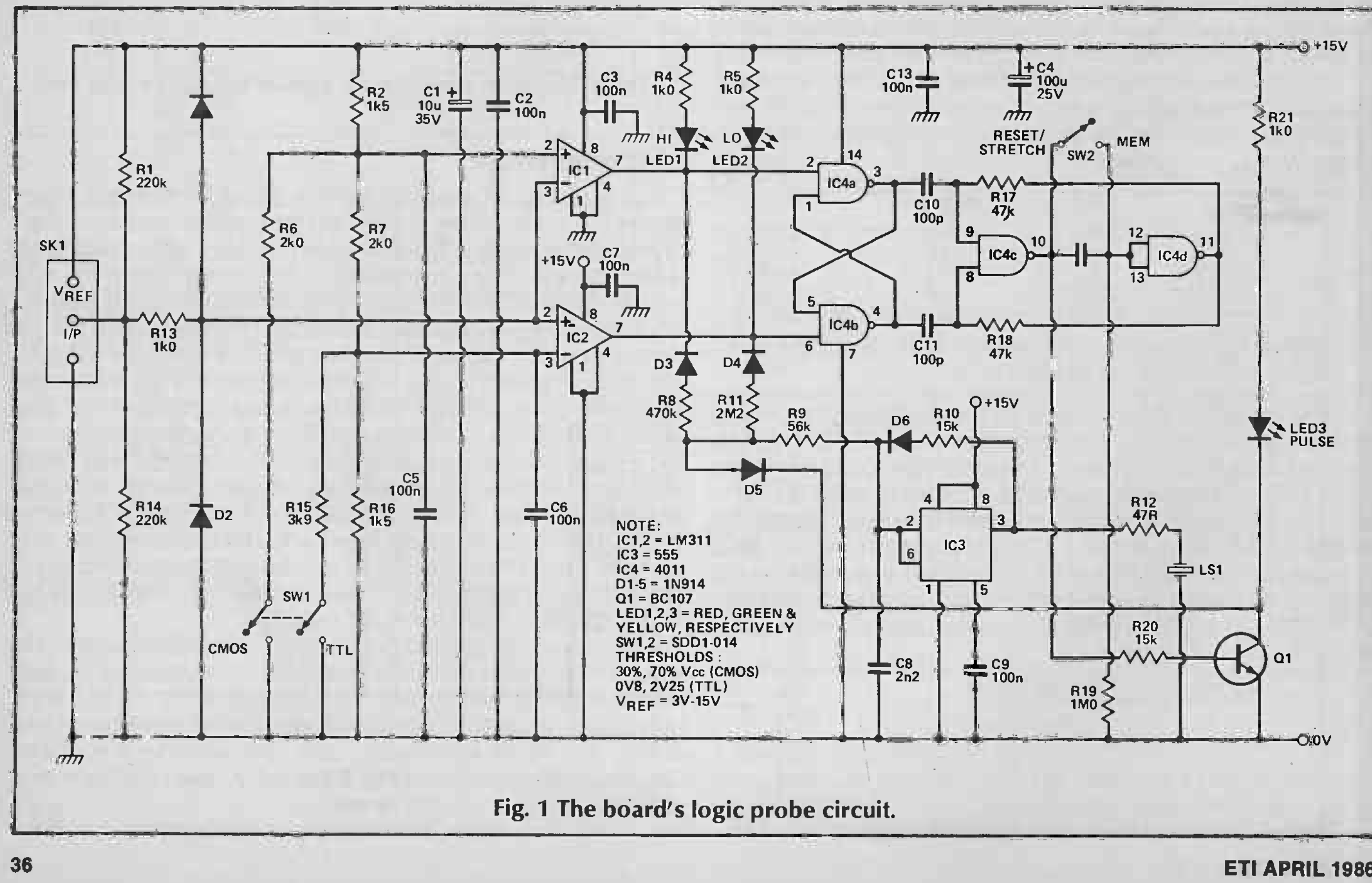
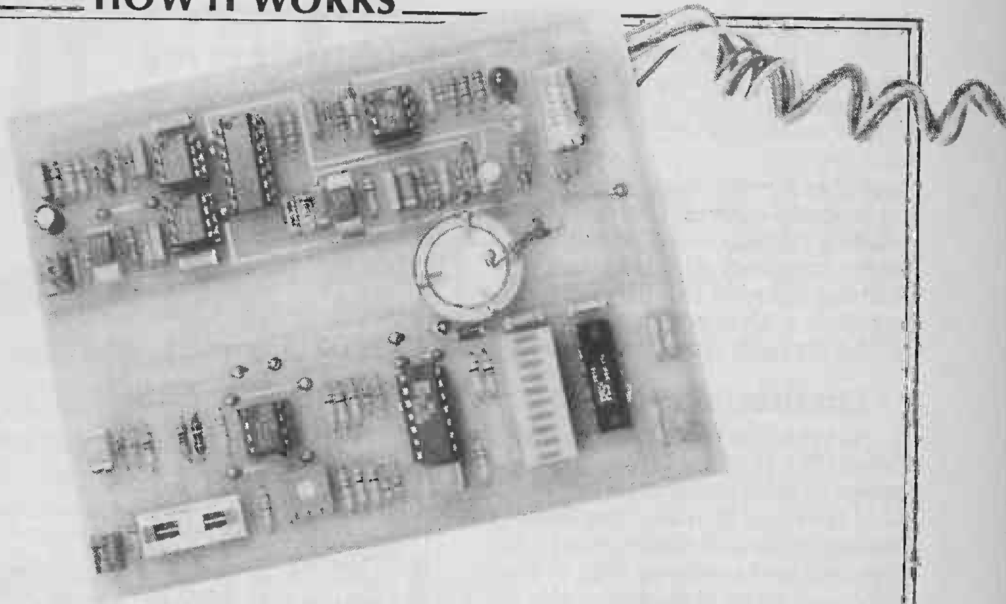


Fig. 1 The board's logic probe circuit.

HOW IT WORKS



LOGIC PROBE

The input stage consists of two LM311s forming a window comparator. The non-inverting input of the lower threshold comparator and the inverting input of the upper threshold comparator are biased to mid supply voltage of the circuit under test by R1 and R14 giving an input impedance of approximately 100k. Input protection is afforded by the combination of R13 and clamp diodes D1 and D2. The upper and lower thresholds are established by the potential divider R2, R7 and R16. When SW1 is closed additional resistors are shunted across the divider to set TTL thresholds. If the input voltage is below the lower threshold then LED2 will light, if it is above the upper threshold then LED1 will light and if it is between the thresholds then neither LED will light. The set-reset latch formed by IC4a and IC4b detects positive and negative transitions that correctly cross both thresholds and triggers the monostable formed by IC4c and IC4d, stretching pulses to approximately 50ms which are then indicated by LED3. If SW2 is closed the monostable is converted into a latch which memorizes input pulses. It is reset by momentarily opening SW2.

IC3, a 555 timer, is configured as a voltage controlled oscillator whose freqency is governed by the inputs connected to D3, D4 and D5. If D4 is low, a low tone is emitted. If D3 is low, a high tone is emitted. If D5 is low, a very high tone is emitted overriding inputs to D3 and D4. If none of the inputs are low the oscillator is inhibited. The frequency of the oscillator thus indicates the input logic levels. Low for logic low, high for logic high, off for indeterminate and a pipsqueak for pulses and transitions! A piezo sounder is used as the transducer.

ANALOGUE PROBE

A non-inverting amplifier with presettable gain of eiether 1, 10 or 100 is configured around IC1. It has an input impedance of 100k and is protected from input over-voltage by the combination of R10 and D1, D2. The input may be AC coupled, DC coupled or shorted to ground. The output of this amplifier drives a precision full-wave peak detector formed by IC2a and IC2b. The output of the detector is connected to a LM3914 bar graph driver arranged to give a full scale display of plus 10V and a LED current of 25mA per segment. When all segments are lit some 250mA is required from the 5V supply. Also connected to the output of IC1 is the polarity and null detector. This circuit requires either plus or minus 100 mV on the input to R16 to light the bicolour LED. Thus at maximum sensitivity, 1mV on the input to the probe will be indicated. When an AC signal is detected the LED glows yellow.

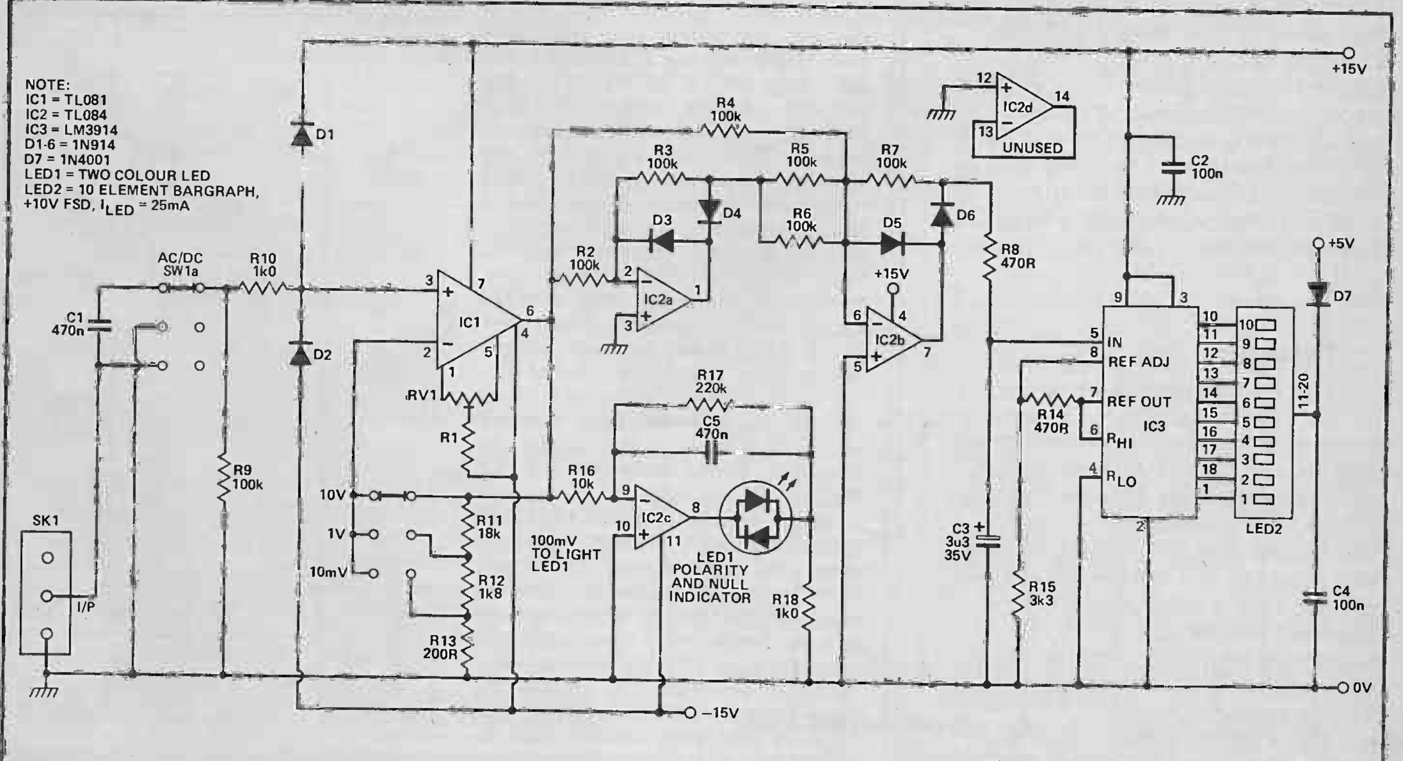


Fig. 2 The board's analogue probe circuit.

for poking around linear circuits!
It will indicate both AC and DC input voltages with a selectable sensitivity of 100mV, 1V or 10V FSD. A bi-colour LED indicates the polarity of the input signal and is useful as a null indicator for setting up op-amp offsets. The peak reading display is well suited for audio monitoring. Finally, the flashing lights of the bargraph together with those on the logic probe do look rather impressive.

Construction

As with the universal counter timer (ETI, January 1986) the board used is double sided. Track pins are used to make the through connections and these should be inserted and soldered first. It is very easy to miss soldering the pins both top and bottom and careful checking at this stage will save much time later. Check the polarity of diodes, the LEDs and electrolytic capacitors.

The bi-colour LED should be aligned so that the flat on the LED faces the bargraph display. This ensures that for positive inputs the LED glows red and for negative inputs the LED glows green. Logical, eh!

The only special point to note is the orientation of the bargraph driver IC. This is placed unusally with pin 1 at bottom, which makes for a more logical PCB layout. The bargraph display itself must also be correctly orientated with the chamfer at the top left. If the display is not marked then a little experimentation with a power supply and a suitable current limit ring resistor should be used to establish its polarity. The piezo sounder is attached to the board by a double-sided adhesive foam pad.

Testing

The board requires + and −15V supplies and a +5V supply. Note the current requirements for the 5V supply (250mA maximum). An understanding of how it works should suggest suitable tests for the board. The only adjustment is the offset of IC1 on the analogue probe. This is easily done by shorting the input with SW1, switching to maximum sensitivity and adjusting RV1 until the null LED is extinguished.

The logic thresholds can be checked with a variable input voltage and a digital voltmeter monitoring the input.

PARTS LIST

LOGIC PROBE

RESISTORS (all ¼W, ±5%)

R1, 14	220k
R2, 16	1k5
R3	10k
R4, 5, 13, 21	1k0
R6, 7	2k0
R8	470k
R9	56k
R10, 20	15k
R11	2M2
R12	47R
R15	3k9
R17, 18	47k
R19	1M0

CAPACITORS

C1	10µ 35V tantalum
C2, 3, 5, 6, 7, 9, 12	100n polyester
C4	100µ 25V electrolytic
C8	2n2 polyester
C10, 11	100p polystyrene

SEMICONDUCTORS

IC1, 2	LM311
IC3	555
IC4	4011
D1-5	IN914
Q1	BC107
LED1	Red LED
LED2	Green LED
LED3	Yellow LED

MISCELLANEOUS

SK1 3-way Molex plug; SW1, 2 ERG SDD1-014 DIL switch; PCB; IC sockets - 3 x 8 pin DIL; 1 x 14 pin DIL; PCB track pins; LS1 piezo sounder.

ANALOGUE PROBE

RESISTORS (all ¼W, ±5%)

R1	1k5
R2-7, 9	100k
R8, 14	470R
R10, 18	1k0
R11	18k
R12	1k8
R13	200R
R15	3k3
R17	220k
RV1	100k enclosed horiz. preset

CAPACITORS

C1, 5	470n polyester
C2, 4 100n polyester	
C3	3µ3 35V tantalum

SEMICONDUCTORS

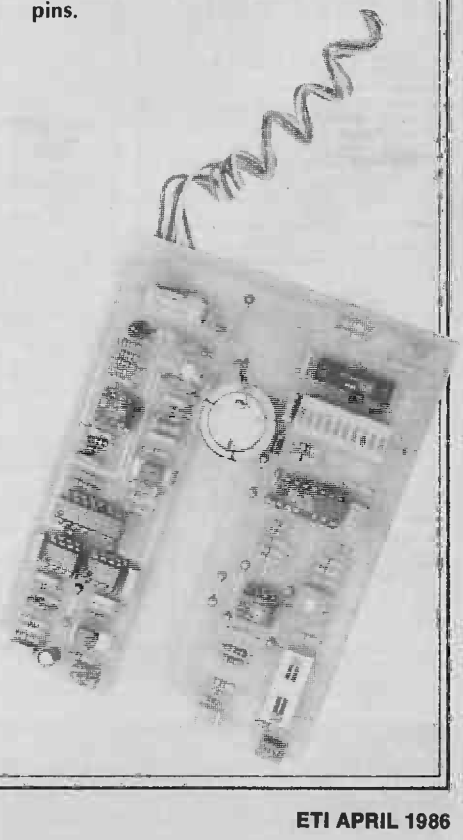
IC1	TL081
IC2	TL084
IC3	LM3914
D1-6	IN914
D7	IN4001
LED1	Red/green bi-colour LED
LED2	10-element LED bar display

MISCELLANEOUS

SK1 3-way molex plug; SW1 ERG DS16D 1-3 +1-3 DIL switch; PCB; IC sockets 1 x 8-way DIL, 1 x 14 way DIL, 1 x 18-way DIL, 1 x 20-way DIL; track pins.

BUYLINES

None of the semiconducutors or passive components should prove difficult to obtain. Bi-colour LEDs and 10 element bar-graphs are obtainable from Maplin, PO Box 3, Rayleigh, Essex SS6 8LR (tel: 0702 552911). The switches specified are available from ERG, who only deal with trade and professional customers, and we do not know of a dealer who will obtain parts from them. However, the SDD1-014 DIL switches may be replaced by 2-way DIL switches which are available from Maplin or Circkit (Park Lane, Broxbourne, Herts. tel: 0992 444111). Farnell sell a ganged 2-way DIL switch which would be more suitable (order code: SDD2 023). Like all Farnell stock, it can be obtained through Trilogic Ltd., 29 Holme Lane, Bradford BD4 0QZ. There is no very satisfactory alternative to the DS16-D 1-3 + 1-3 DIL switch as specified. The best option is to use an octal SPST DIL switch and carefully set the required range and input signal type before powering up the board. The risks of this approach are that, by setting more than one range switch (or none at all), your analogue reading will be inaccurate or that you may ground the signal taken from the circuit under test. To avoid this latter, cut the track to ground from SW1a after calibration.



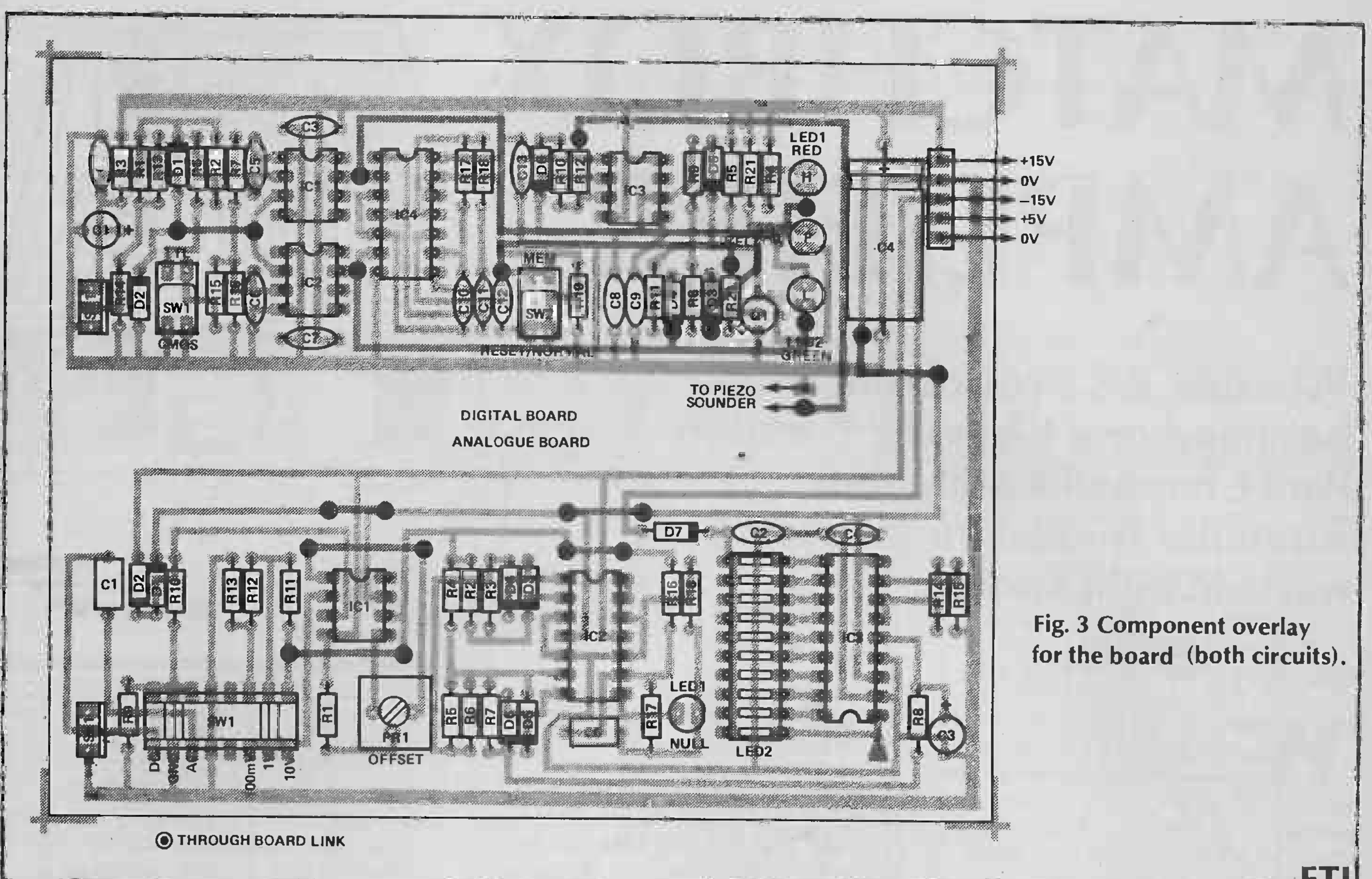
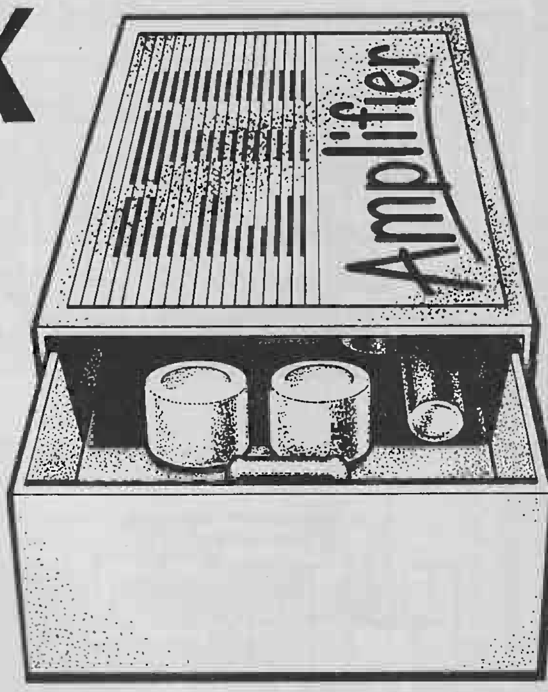


Fig. 3 Component overlay for the board (both circuits).

MATCHBOX AMPLIFIER



Whether it's a rock band in a hat-band or a PA in the Portaloo, Paul Chappell's Matchbox Amplifier is ideal for providing watts in tight spots.

What projects are on the list for April ETI? Design a general purpose amplifier for the workbench, maximum component cost £10. OK, what else? An amplifier that isn't fussy about power supplies, that will work from batteries or just about any old PSU. Fine. What next? Another amplifier, this time to boost the output of an in-car stereo, and yet another amplifier project to give 50 watts or so for heavy metal fans. And another one still — light enough to be carried around by buskers and walkperson addicts and ... what's that? ... small enough to fit in a matchbox? You must be joking!

Well, this is the result. Being much too lazy to design five separate amplifiers, I offer you two that will do the lot. And they *will* fit in a matchbox.

The circuits are based on the L165V amplifier IC, a very versatile little device indeed. The minimum rated supply voltage is 12V, although it seems quite happy to work below 9V with reduced performance. The absolute maximum supply voltage is 36V, so there's a fair old range to choose from. If you want to get the maximum possible power you will, of course, have to choose a supply voltage towards the top of the range, but if you are prepared to settle for a reduced output you can run it from just about anything.

The first of the amplifier circuits is shown in Fig. 1. If a single rail power supply is used, a large electrolytic capacitor, C6, is needed between the output and the loudspeaker. A split rail supply can also be used, with 0V connected at point G on the board, the positive rail connected to V+ and the negative rail to V−. If you do this, you can leave out R2, R3 and C6 and the speaker can be connected between the output and 0V. The only point to bear in mind is that the tab of IC1 is internally connected to pin 3 and will be at the negative rail voltage, so be careful what you bolt it to for heat sinking. Use an insulating set if necessary.

The second amplifier, Fig. 2, uses a bridge arrangement to get more power, particularly at low voltages where the output will be limited by the maximum voltage swing. This arrangement will give about four times the maximum power of the single IC version since the voltage seen by the speaker is doubled. This feature makes it useful as a booster for car stereo systems, if your speakers can handle the extra power!

You can also use a split rail supply with this circuit if you wish, although there is not much benefit from doing so. Once again, 0V would be connected to point G, the positive rail to V+ and the negative rail to V−, and the warning about the IC heat sinking

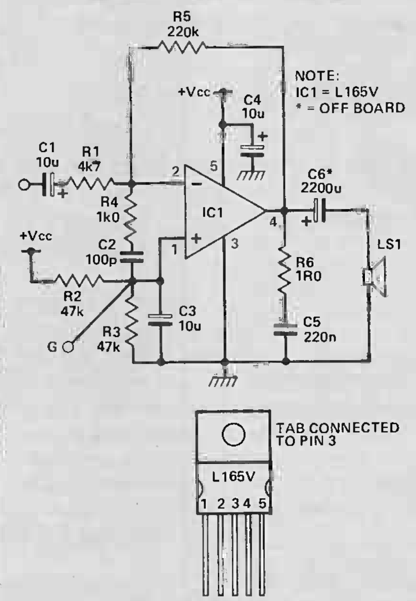


Fig. 1. Circuit diagram of single IC amplifier.

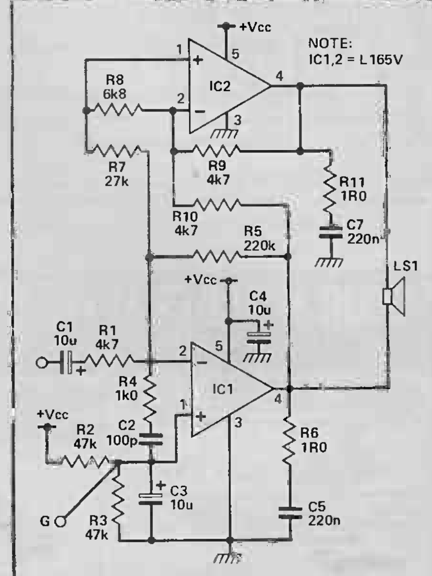
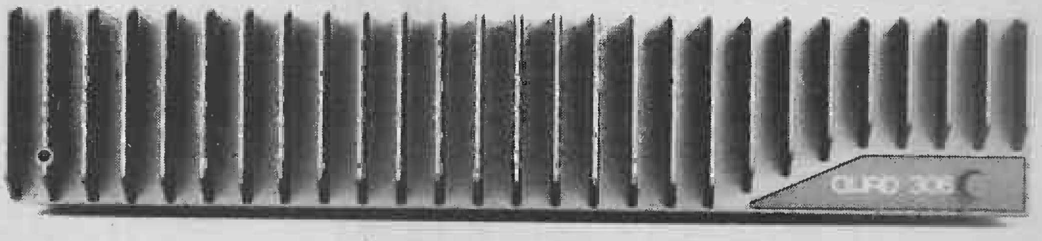


Fig. 2. Circuit diagram of the bridge amplifier.

DIGEST

Quad Celebrate With New Amplifier



The Acoustical Manufacturing Company, now known as Quad Electracoustics Ltd, was founded in 1936.

To kick off thier fiftieth anniversary year in accustomed style, they have launched a new stereo power amplifier, the Quad 306.

Designed for use in high quality domestic systems, the 306 has an output of 70 watts into eight ohms and uses a refined version of the 'current dumping' system which was pioneered in the highly-acclaimed Quad 405 amplifier. Separate power supplies are used for each channel, fed from separate windings on a common toroidal, transformer, and the amplifier has a signal-to-noise ratio of 110dB. Fuses, relays and the like have been carefully excluded from the signal path without sacrificing the standard of protection.

The Quad 306 should be available from dealers by the time you read this and will cost £229.00 including VAT.

Quad Electroacoustics Ltd, Huntingdon, Cambridgeshire PE18 7DB, tel 0480-52561.

System A Transistors

Universal Semiconductor Devices Ltd, tell us that they can supply all of the transistors required for the System A amplifier. They offer a kit of semiconductors for the System A power amplifier for £75.00 inclusive and are happy to quote for individual semiconductors for the rest of the system. Their address is 17, Granville Court Road, Hornsey, London N4 4EP, tel 01-348 9420.

Free PCB

We would like to remind readers that the multi-purpose PCB (given away free with the last issue of ETI) is available from our Readers' Services Department at the special price of £1 each (plus a stamped addressed envelope). This offer will last only as long as we have stocks of the mass-produced board — after they run out, the board will be available through the PCB service at the normal full price of £2.87 (plus 50p P&P). So if you want to save money, get your order off now!

Four-In-One Plug

Duraplug have introduced a 13 amp plug which enables up to four appliances to be connected to a single socket outlet without the need for additional plugs and adaptors.

The MultiLine plug is simple to wire and features a mains-on indicator light. It is available in black or white, is fused at 5 amps and constructed in a tough thermoplastic which is highly resistant to impact.

The plug has been especially designed for use with lighting, hi-fi units, video and television, etc. It will cost about £5.00 from most leading electrical and hardware stores.

Ah, So Old

Events at the Hitachi factory in Aberdare have angered Gwent MP Anne Clwyd. In January she told the House of Commons she wants to introduce a Bill to fight age discrimination at work. She informed MPs:

"A year ago the management invited workers aged 35 or more to take so called voluntary redundancy." This, she said, told them they were over the hill at 35. She described how the company had sent a letter to all employees saying that older workers cause problems through sickness, slower reactions, poor eyesight and resistance to change.

The letter did not mention that these problems are almost certainly the direct result of working in electronics factories.

As a recent report on electronics workers in Malaysia* points out:

"Well before they are thirty years of age women electronics workers begin to lose their value. Their productivity begins to diminish as a result of boredom, accumulated fatigue or failing eyesight caused by years of peering through a microscope at hair-breadth wire."

The report also comments:

"It costs less to hire new workers at base starting rates than to keep old workers on at their enhanced rates and fresh workers can achieve high marginal productivity within three or four months of starting a job."

In Malaysia, natural wastage runs at a staggering 6% turnover per month, substituting young workers for older ones. In Britain where the figure is nearer 6% per year, the replacement process has to be forced, but replacing older women with young workers is precisely what Hitachi is doing. The firm announced 500 redundancies when it took over the Aberdare plant from GEC two years ago. Unions at the factory, which makes televisions and videos, insisted on a 'last in first out' policy for the sackings. This left a workforce of 800, mainly older workers, 60% of them women.

Since then the EETPU has won sole negotiating rights and has made a 'no strike agreement'. The union also approved the management proposal for redundancy for older workers. Clearly many older workers felt this left them with no options. They left. Ninety-two workers took voluntary redundancy, 78 of them women. Management refused to consider younger workers for redundancy, but their motivation was hardly philanthropic.

"Redundancy for anyone under 35 is nonsense," company spokesman Tony Pegg told ETI. "Young workers can adjust to change. Older workers can't."

As an example of the intractability of older workers he points out that among the older workers: "Heavy smokers can't cope with the no smoking rule in the production area."

It is difficult to believe that such matters really form company labour policy, but Tony Pegg probably doesn't care whether we believe it or not. His firm has successfully, as he puts it, 'shed' older workers and has found suitable replacements. The company has taken on 77 school leavers.

Hitachi is not the only electroncis company which pursues an apparently ageist employment policy. Another Japanese company, The Nippon Electric Company, has made it a policy to recruit school leavers for its factory in Livingstone, Scotland. The average age of the workforce is reported to be 21.

*Electronics Development: Scotland and Malaysia in the International Electronics Industry. Published by Scottish Educational Action for Development.

tabs still applies. With a split rail supply they will be at V−, so don't bolt them to metalwork at 0V. If you are not absolutely sure, check with a voltmeter that the tabs and the metalwork are the at the same voltage.

Heat And Power

As far as heat sinking is concerned, the rule is the larger the better. By all means use finned extrusions if you wish, but otherwise just bolt the heat sinking tabs to the metalwork of the case, to a sheet of heavy gauge aluminium, or even to the loudspeaker chassis. A touch of heatsink compound and a clean mounting surface won't go amiss. Don't panic if the ICs seem to run rather hot — you can boil water on the heatsink before they get too unhappy, and an internal thermal shutdown should prevent them from coming to any harm. Having said that, they will work more efficiently if you can keep them cool. Don't run them without heatsinks, even at low power.

The power supply can be just about anything you like although if you are looking for outputs around 50W from the bridge circuit a supply of 30V or so at 3.5 amps will be needed. Avoid using supplies above 32V — although the ICs will work at up to 36V, anything above this may damage them, so leave a safety margin to allow for ripple and overvoltages of one sort or another on the supply.

HOW IT WORKS

The first amplifier circuit is much the same as the series feedback circuit commonly used with op-amps, with the gain being set by R5 and R1. R2 and R3 provide a voltage at a half of Vcc as a reference for the non-inverting input of IC1. R4 and C2 form a lag network to improve the damping and phase margin of the circuit. Without these components, the amplifier would oscillate when the input was left floating and would be prone to overshoot on transients. R6 and C5 form a Zobel network to minimise the variations in load impedance seen by the amplifier output.

With a split rail supply, the speaker can be connected between the output and 0V and C6 can be omitted. The output voltage offset, which will consist of the input offset ± the voltage across R5 developed by the bias current, will be around 2mV at the most. This is negligible, so no offset trimming is needed.

The bridge circuit is essentially the same as the single IC circuit, but has a second amplifier with a gain of −1 to provide an inverted signal for the other end of the loudspeaker. R9 and R10 give series feedback around IC2 for this purpose. The non-inverting input of the amplifier is connected to the virtual earth point at IC1 pin 2 to avoid having to adjust the output offset. R8 is included for stability.

PARTS LIST

RESISTORS	
R1, 9*, 10*	4k7
R2, 3	47k (see text)
R4	1k0
R5	220k
R6, 11*	1R0
R7*	27k
R8*	6k8

CAPACITORS	
C1, 3, 4	10u, 35V tantalum
C2	100p ceramic
C5, 7*	220n polyester
C6	2,200u 35v electrolytic (see text)

SEMICONDUCTORS	
IC1, 2*	L165V

MISCELLANEOUS
PCB (single IC or bridging version as required); heatsink (see text).

*Additional components required for bridging version. Not needed for single IC amplifier.

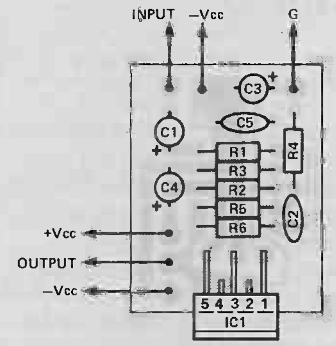


Fig. 3. Component overlay for the single IC amplifier board.

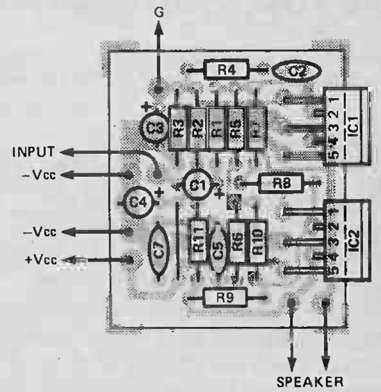


Fig. 4. Component overlay for the bridge amplifier board.

Applications

I hardly need to point out the applications of amplifier modules in general, but the small size, high power and tolerance of supply variations make these little units very versatile indeed.

The bridge amplifier can give a useful boost to car stereo systems, and will also find a home in other applications where a reasonable output is required and the supply voltage is low. Both modules are ideal for use in any project that needs a sound output. They will work from any old supply voltage, and they won't take up much space!

They won't disgrace themselves as part of a hi-fi system either. The frequency response is excellent — at quite high power levels there is no appreciable drop in output up to several hundred kHz (measured on a 'scope, since my ears and speakers give up long before that!). Noise and distortion are astonishingly low for such a simple circuit, and if your taste leans towards rock msuic, they certainly pack quite a punch. Don't underestimate them!

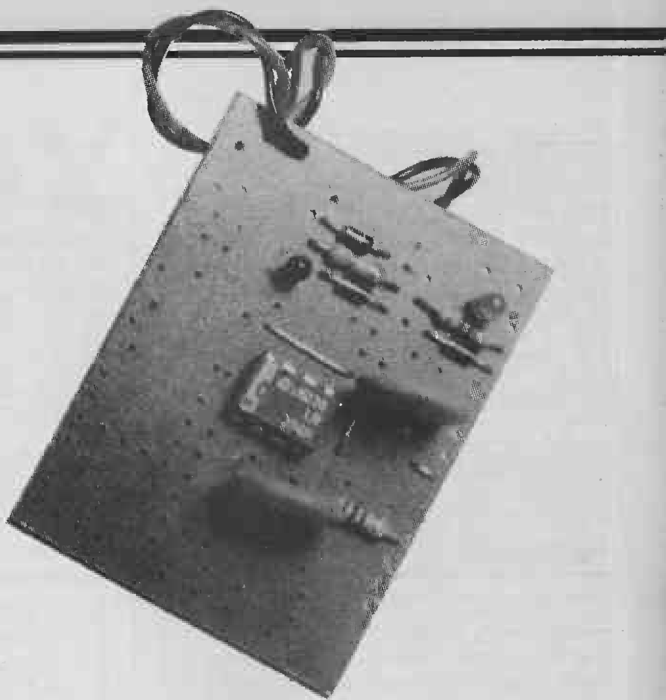
If you play in a rock band, you can sack the roadies. Who needs them when you can carry your amp about in your pocket? Of course, you'll also need matchbox-sized speakers and power supplies ....

BUYLINES

Kits for both amplifiers, including PCB and all components, are available from Specialist Semiconductors, Founders House, Redbrook, Monmouth, Gwent. The single IC amplifier kit costs £6.50 and the bridge version £8.95. The L165V ICs are available separately from the same source and cost £3.90 each. Postage on the complete order is 60p. The PCBs will also be available from our PCB Service, see page 58 for details.

ETI

TOUCH SWITCH



Paul Chappell, who designed all the free PCB circuits, goes over the top. Some say he's touched ...

Current trends in touch technology have far reaching implications for any dynamically balanced meta-system, with particular reference to their immediate spheres of utilisation and deployment. One may envisage major consequences arising from extended and/or prolonged disbursement techniques; moribund in nature, yet not without a certain attribution in the context of a pseudo-dictatorial stance — the details of which elude many who are otherwise conversant with the 'methodology' (**sic**) involved.

This project is a touch switch. You touch it. It switches. It has a delightful little LED in a tasteful shade of puce (unless that's something Alf spilled on it) and a very earnest TTL output. One of its most pleasing aspects is the little snippity-snap noises it makes when you break it in half. Oh, joy! I'd like to tell you more, but I'll quickly hand you over to Auntie Static who would like to leak some confidential information about the LM358. I'm off to get some more of my special green tablets. Over to you, Auntie.

Digression

It probably hasn't escaped the attention of our readers that all the FREE PCB projects have been based on the same IC — the LM358. The editor has expressed the opinion that a few details about the IC may be of interest, and who am I to disagree?

When we decided to make the project battery operated, our first concern was to find an op-amp that would work from a low voltage and consume very little current. The LM358 fits the bill nicely on both counts — it will run from voltages as low as 3V (and up to 30V) and draws only 500μA from the supply. The current drain is essentially independent of the supply voltage since most stages apart from the output have constant current loads, as you can see from the schematic in Fig. 1.

Split rail power supplies can be a nuisance to arrange in battery powered circuits. Two batteries can be used, of course, if there's room in the case for them, if they are evenly loaded so that they both run down at a similar rate, and if you're prepared to mess about finding out which one is flat when the circuit stops working or you don't mind throwing away the good one with the bad. In short, it's not an entirely satisfactory arrangement. The usual alternative is to arrange a suitable voltage bias for the op-amps with a pair of resistors. With the LM358, you don't even have to bother with this. The input common mode voltage range of the IC includes 0V, so for many applications it is possible just to ground one of the inputs. This was particularly useful, for instance, in the current to voltage converter used in the frequency and capacitance meters.

Have a go at designing a circuit to do the same job using a 741 (say), and you'll soon see the

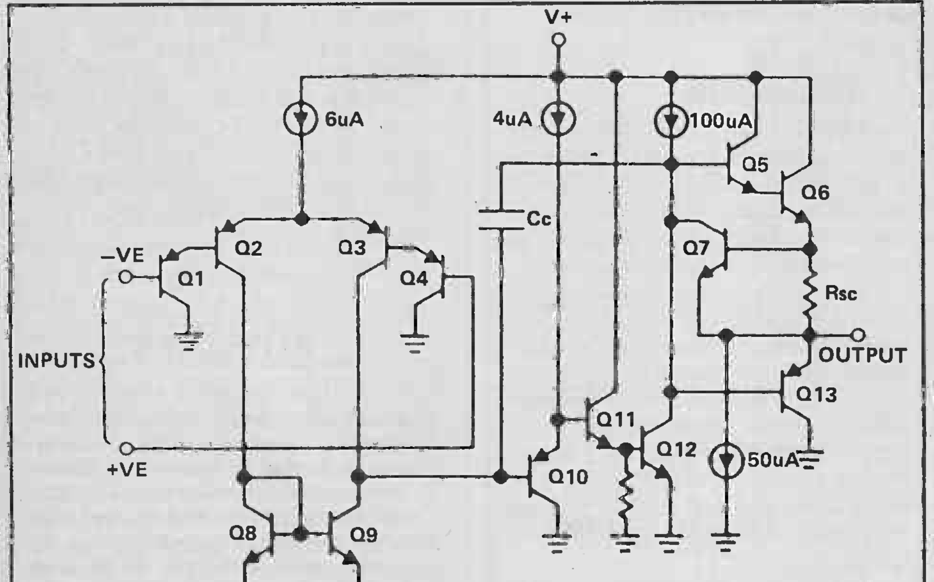


Fig. 1 Internal diagram of one half of an LM358 (courtesy of National Semiconductor).

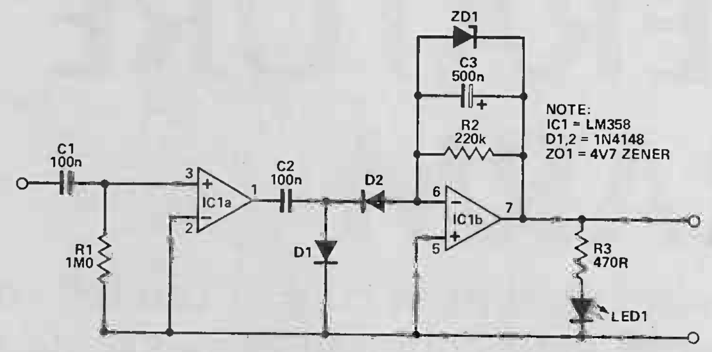


Fig. 2 Circuit diagram of the touch switch.

HOW IT WORKS

IC1a amplifies the electrical grunge that is coupled to the circuit from the touchperson's body — mostly mains hum. As we are not particularly concerned with hi-fi, the amplifier uses a bare minimum of components and no feedback whatsoever. C2, D1 and D2 act as a charge pump in much the same way as the equivalent section of the frequency meter, except that C2 is much larger so that the current will be high enough to drive IC1b output hard up to the positive end of its range (or would do if not for ZD1) at low frequencies.

ZD1 is included to give a suitable positive output voltage to drive TTL, and also to make sure the output stage functions correctly. Without the zener, the output would rise to within about 1.5V of V_{cc} and no higher — limited by the operation of the output stage of the LM358.

At this point, any further current through D2 would result in a drop in voltage at pin 6 of the IC. The voltage drop would be held between cycles by C3, so the voltage at pin 6 would get progressively lower. The op-amp is being taken outside its operating limits, and sooner or later it will get upset.

What actually happens is that the op-amp has an internal hiccup, the output voltage drops by a few volts, this restores the feedback action which maintains the voltage at pin 6 at 0V, and the output begins to rise again. When it reaches V_{cc}–1.5V, the whole process begins again.

In short, without ZD1 the output would oscillate. ZD1 prevents the output from rising above 4.7V, so this problem does not arise and the output remains stable.

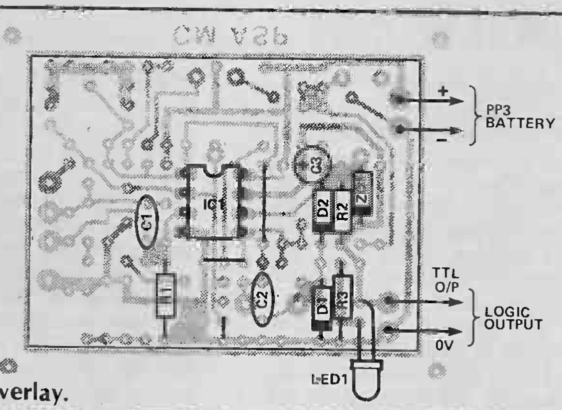


Fig. 3 Component overlay.

PARTS LIST

RESISTORS (all ¼W 5%)		SEMICONDUCTORS	
R1	1M0	IC1	LM358
R2	220k	D1, 2	1N4148
R3	470R	ZD1	4V7 zener diode
		LED1	red LED
CAPACITORS		MISCELLANEOUS	
C1, 2	100n	PP3 battery and connector	
C3	500n tantalum		

difference. In some circumstances this feature is not quite so useful, as in the input stage of the frequency meter, but nothing is perfect.

The output voltage swing of the LM358 also includes 0V, the full range being 0V to V_{cc}–1.5V. Having said that, the output won't quite pull all the way to 0V of its own accord. If you look again at the schematic in Fig. 1, you'll see that the combination of Q12 and Q13 will pull the output down to about 1V. If the output is very lightly loaded, the 50μA current sink will take it even lower, but even so the sink itself will take a certain minimum voltage to operate, so it won't quite make it all the way. A resistor from the output to ground is needed, as in the frequency meter circuit.

The difference between the LM358 and (why not?) a 741 in this respect is that if you want the output of a 741 to reach the negative supply rail, you'll have to short-circuit it! The arrangement of the output stage is such that it will continue to drive current into a load connected to the lower supply rail under all circumstances, whereas the current from the LM358 shuts off. A useful consequence of this is that the output of the IC is directly compatible with TTL logic (with a 270R resistor from the output of the amplifier to ground). Most op-amps won't pull low enough to give a good logic 0.

Other features of the IC are a fairly low input offset voltage of 2mV — not really spectacular when you consider more specialised circuits like the OP37 with around 30μV offset — but no cause for complaint. The input offset current of 5nA is quite respectable, too. The output can source and sink a reasonable current — up to 40mA, which should be more than adequate for most purposes.

BUYLINES

All the parts for this project are readily available and the LM358 will be supplied to ETI readers at the special price of 5 for £2 by Specialist Semiconductors, Founders House, Redbrook, Monmouth, Gwent. Spare printed circuit boards are also available but see the note in News Digest.

ETI

TEMPERATURE ALARM

This free PCB project knows hot stuff from cold. It can tell you when your bath water runs cold, your coffee becomes tepid or your plants are at risk from frost. Design by P. Chappell, MIPBCCMMG (Member of the Institute of Plumbers, Bathminders, Chartered Coffee Makers and Marrow Growers).

This little project will be invaluable to those of you who are in the habit of turning on the bath taps and wandering off to make a cup of coffee, build an ETI project, or whatever. Most household plumbing systems seem to have a very unfavourable hot-water tank-to-bath capacity ratio, resulting in a water temperature to volume function which declines rapidly below the Eulenschafft-Muller comfort rating as the bath fills beyond a level L=7V/8 x 1/A where V is the volume of the hot

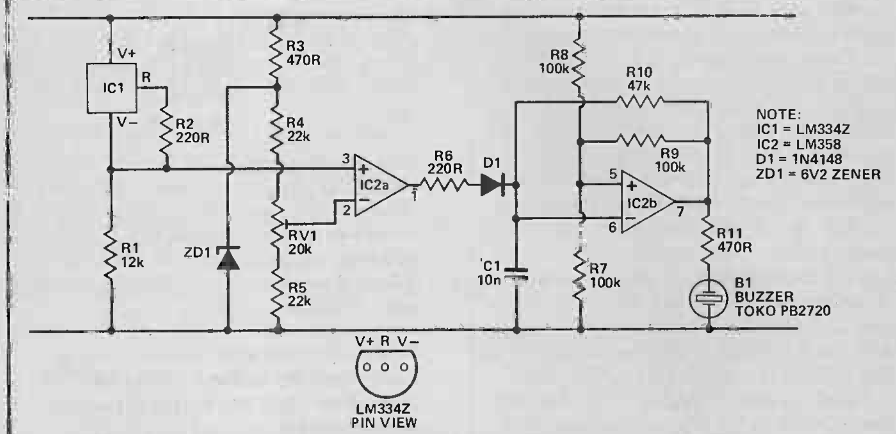


Fig. 1 Circuit diagram of the temperature alarm.

HOW IT WORKS

The LM334Z is a current source IC which has a linear variation in output with temperature. This feature can be a nuisance if you actually want to use it as a current source, but is very useful for making thermometers and suchlike. The current from the LM334 develops a voltage across R1 which is compared with a reference voltage set by RV1.

IC2b forms an oscillator. Let's suppose that the +ve input is at a higher voltage than the −ve input. The output of the op-amp will therefore be high, a little below Vcc, so the +ve input will be roughly 2/3 of Vcc by the action of R7, R8 and R9. Since the output is high, the voltage at the +ve input will rise as C1 charges up via R10, and will eventually reach 2/3 of Vcc. At this point, the output of the op-amp will go low so the +ve input will fall to 1/3 of Vcc and the capacitor C1 will begin to discharge. Eventually, the voltage at the −ve input will fall to 1/3 of Vcc, the output will go high, and the whole cycle will begin again.

Since we have received a number of Tech Tip suggestions containing rather questionable oscillator circuits, it may be in order to say a little more about why this circuit is absolutely certain to oscillate. Let's go back to the point where the −ve input of IC2b has risen to match the voltage of the +ve input. The output of the IC is just about to go low by the normal action of an op-amp. The important thing to notice is that the voltage across C1 is *still rising,* and will continue to rise even as the output of the IC falls, until it drops below (roughly) 2/3 of Vcc. On the other hand, any drop in output voltage at all will immediately pull down the voltage at the +ve input, so the cause of the switching action (the −ve input being positive with respect to the +ve input) will continue, and even increase, as the output falls. By the time the output voltage has fallen to a point where C1 can no longer charge, the +ve input will already be at a voltage well below the −ve input, and any further drop in output voltage can only increase this difference, so the switching action is hard on until the output reaches its lowest level.

Just to avoid confusion with Auntie Static's description of the LM358 (in the Touch Switch project), I should add that the upper voltage limit of the IC is, in fact, about Vcc-1.5V, so strictly speaking the input switching levels will not be quite 2/3 and 1/3 of Vcc, but the principle is still the same.

To return to the bath minder. The oscillator is allowed to run freely when the output of IC2a is low (as it will be when the temperature is low) and is held off by D1 and R6 when the temperature, and therefore the output of IC2a, is high. The transducer specified for the circuit is the one we used, mainly because there happened to be one lying around, so don't feel that you have to stick to the same type. No particular attempt was made to match the transducer frequency or drive requirements, so maybe you'd like to experiment with that for yourselves.

A modification you may like to consider is to increase C1 to 10μ or so, remove R11 and LS1, and insert a buzzer module between the output of IC2b and ground. The alarm will then be an intermittent buzzing sound. This is probably more satisfactory than the specified transducer if you intend to install the circuit permanently.

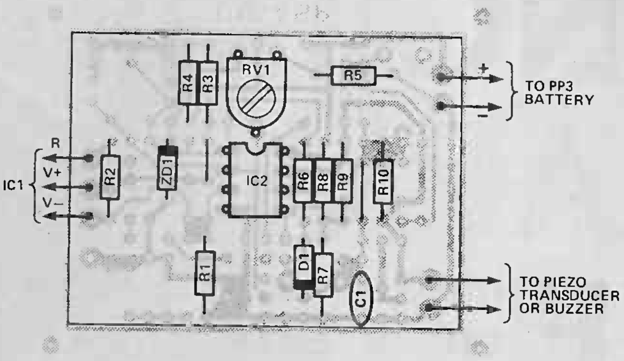


Fig. 2 Component overlay for the temperature alarm.

PARTS LIST

RESISTORS (all ¼W, 5%)		CAPACITORS	
R1	12k	C1	10n
R2	220R		
R3	470R	SEMICONDUCTORS	
R4	22k	IC1	LM334Z
R5	22k	IC2	LM358
R6	220R	D1	1N4148
R7	100k	ZD1	6V2 zener
R8	100k		
R9	100k	MISCELLANEOUS	
R10	47k		
R11	470R		
RV1	20k		

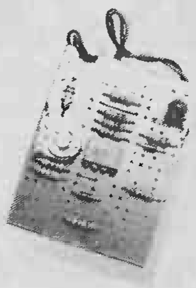
LS1 piezo sounder (PB2720); PB3 battery and connector; suitable housing for LM334; blob of Araldite; length of 3-core wire for sensor; PCB.

water tank and A is the average cross-sectional area of the bath. The eigenfactor 7V/8 is empirically determined as the breakpoint at which the ingress of cold water into the h.w. tank has reduced the temperature at the outflow, by the process of zonal intrusion, to a level at which it is $%&£@!! cold. In short, even if you return to the bath before it has overflowed, the chances are that it will 'freeze your naughty bits off' (Minsk and Kornbluth, Proceedings of the United Plumbers and Bathminders Conference, June 1983).

The answer, of course, is the auto-bathminder. With commercial models selling at upwards of £240, perhaps this little circuit will be of interest. Under the hot tap, the sensor detects the reduced temperature of the water when the tank is just about empty, and sets off an alarm. You then rush to the bathroom and turn off the tap. The circuit would also be useful as a frost warning device, although I wouldn't like to guarantee the long term stability of the temperature set point if you use the components specified for the bath minder.

Construction

The temperature sensor was housed in an old biro tube (what else?) and waterproofed with a blob of araldite. No doubt you could come up with better ideas yourselves. The circuit board is assembled according to the component overlay in Fig. 2 and once again I would urge you to double-check the layout. It's so easy to make mistakes on this board.



BUYLINES

The PB2720 piezo sounder is available from Cirkit, Park Lane, Broxbourne, Herts EN10 7NQ (Tel: 0992 444111). Otherwise the Maplin FM59P will do the trick. If you want to try the buzzer modification (see text), Maplin supply a suitable module with the catchy title of FL39N. Maplin's address and phone number can be found on their ad in this issue. The particular type of buzzer is not critical and suitable devices can no doubt be obtained from many other suppliers. The LM358 ICs are available from just about anybody, but the best price we have come across is 5 for £2 from: Specialist Semiconductors, Founders House, Redbrook, Monmouth, Gwent.

Extra PCBs can be obtained by sending £1 and a stamped, self-addressed envelope to: ASP Readers Services, PO Box 35, Wolsey House, Wolsey Road, Hemel Hempstead, Herts. HP2 4SS.

See the note in News Digest about this.

COMPETITION

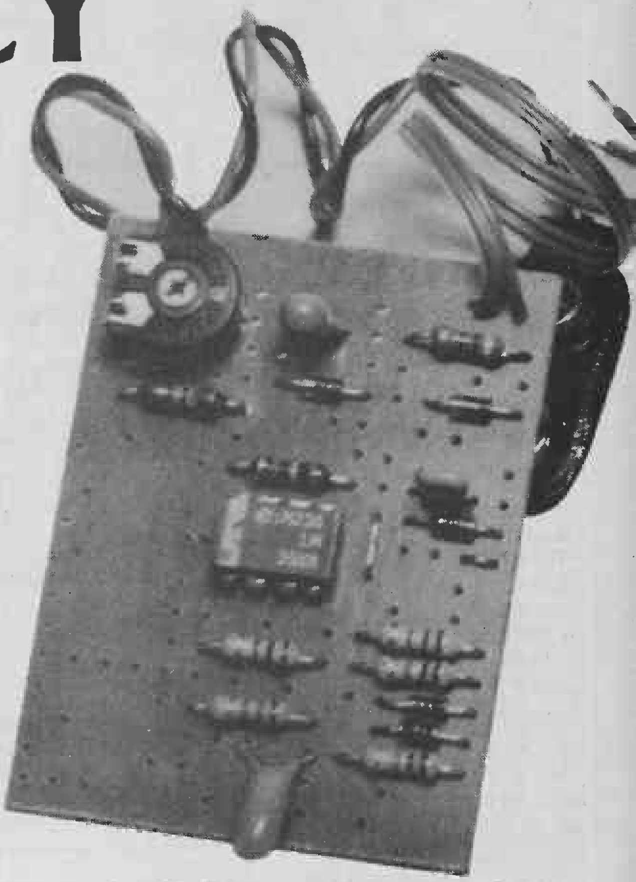
Why not use your own ingenuity to design a circuit for our free PCB? The sender of the winning entry will receive £40 and the two runners up will win £20 each. We will publish the winning circuits later in the year. The circuit should fit neatly onto the board, without components soldered to the back, and we will be looking for original, well presented and elegant designs.

Entries should include a circuit diagram, component layout, a parts list and a description of the circuit (typed, please, and not more than 750 words in length.) If you are not the kind of person who has circuits published, have a go anyway. You may find out that you are!

Send your entries to: ETI (FP), 1 Golden Square, London W1R 3AB. Please enclose a stamped, self-addressed envelope if you want your entry returned and, please, no submissions after 2 May, 1986.

ETI

FREQUENCY METER MODULE



Yes, it's the Not-the-ETI-Cycle-Minder, a 100% re-cycled free- PCB design from Paul Chappell.

A frequency meter can be an expensive luxury for many home workshops and will usually take second (or third) place when essential items like an oscilloscope or a signal generator are needed. There is no doubt that the possibility of making quick checks on frequency can be useful at times (is your ultrasonic transmitter oscillating at 40 kHz or 100kHz?) so perhaps this little circuit will plug the gap for some of you until a more expensive instrument can be bought or built.

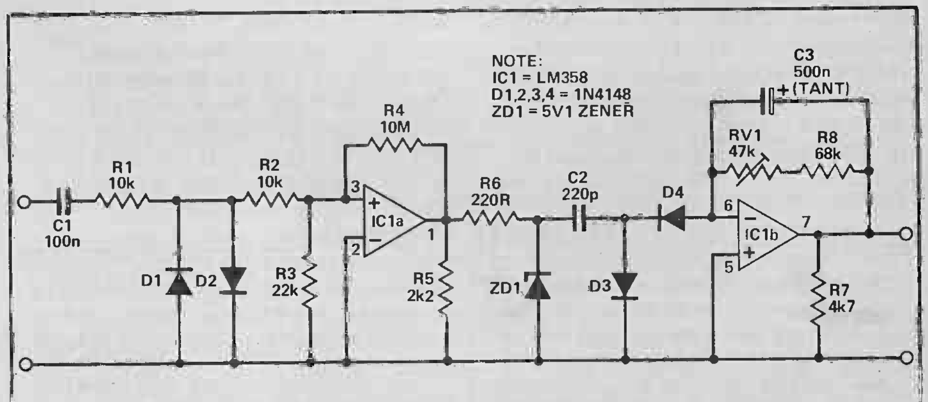


Fig. 1 Circuit diagram of the frequency meter module.

The Circuit

The circuit is essentially a linear frequency to voltage converter, the voltage being read on your bench multi-meter. The output can be adjusted so that frequency can be read directly from the meter scale — 0 to 2V could represent 0 to 10kHz, for instance. As it stands, the circuit will convert frequencies up to 20kHz, but a simple pre-scaler can be added to extend the range to 1MHz and beyond.

The principle of the frequency meter is similar to that of the capacitance meter described last month. The input waveform is squared, clipped and then applied to one plate of a capacitor. Each high to low transition of the voltage on this plate will draw a fixed amount of charge from the virtual earth at pin 6 of IC1 to charge up the other plate (Fig. 1). The current into the virtual earth comes from the output of the op-amp via RV1 and R8, so the voltage across these resistors will be proportional to the current drawn by C2 — or to the average current, since C3 will effectively smooth out the fluctuations.

The higher the input frequency, the more times each second C2 will draw its charge, so the higher will be the average current through RV1 and R8, and therefore also the voltage across them. Since pin 6 of IC1 is maintained at 0V by the action of the feedback, the output voltage of the circuit will be equal to the voltage across RV1 and R8, which is proportional to the input frequency.

RV1 is adjusted to calibrate the circuit. The component values given are for an output covering the audio frequency range 0 to 20kHz at a scaling of 1V per 10kHz, so the output will be in the range 0 to 2V. If you want a different range, the easiest way to achieve this is to alter C2. For 1V per kHz, for instance, C2 would be increased to 2n2. In this case, 1V would represent 1kHz, 2V would be 2kHz, and so on. The upper limit is determined in the first case by the fastest rate at which IC1a will switch, and in the second case by the battery voltage, so with C2=2n2 and a fresh battery, you could expect outputs up to about 8V and could therefore measure frequencies to 8kHz or so.

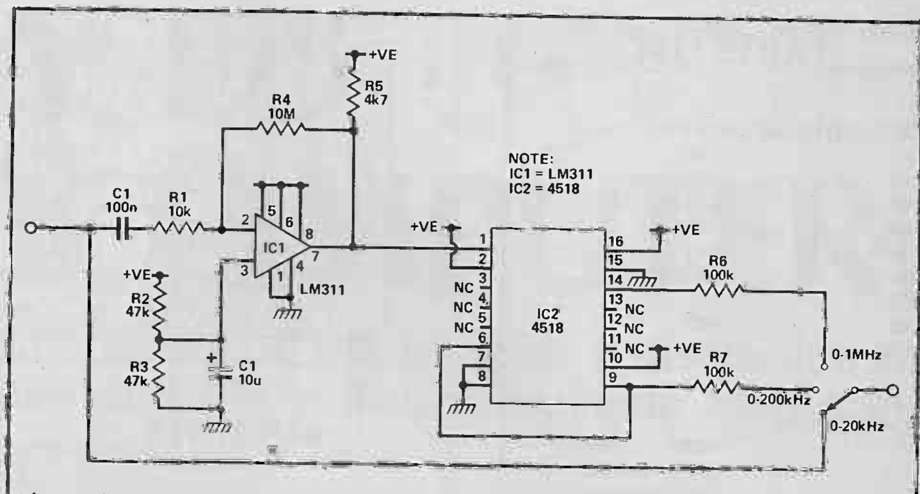


Fig. 2 Circuit diagram of the pre-scaler.

Pre-scaler

To read frequencies above 20kHz, the pre-scaling circuit of Fig. 2 can be used. The input is squared by IC1 and applied to a dual decade counter IC2, which divides the input frequency by 10 or 100. When the input frequency is divided by 10, 1V at the output of the frequency meter will represent 100kHz, and if divided by 100 the output will be scaled at 1V per MHz. You will certainly be able to read above 1MHz, possibly up to 2MHz, depending on the particular 311 IC used.

To make a multi-range instrument using the pre-scaler, you won't need a separate setting of RV1 for each range (unlike last month's capacitance meter). Just set it up for 0 to 20kHz and it will be just as accurate on the higher ranges.

Calibration

If you haven't got a signal generator or any other source of a known frequency, there are two possible ways of calibrating the meter. The easiest, but not entirely satisfactory, way is to use the mains frequency as shown in Fig. 3. This would be quite acceptable for low frequency ranges, but for a 0 to 20kHz range, the output would only be 10mV at 100Hz and you will probably have difficulty reading this accurately on your meter. Also, the linearity of the module is better in the centre of its range than at either extreme, so it's best to use a frequency around 10kHz. A 4060 IC and a crystal can be used to produce an assortment of frequencies (Fig. 4). To find what frequency comes out of any output, just divide the crystal frequency by 2 that many times; that is to say, from Q9 you will get the crystal frequency divided by 2 nine times, from Q14 you will get the crystal frequency divided by 2^{14}, and so on. Choose the output giving a frequency somewhere between 10kHz and 20kHz, connect that output to the frequency meter, and adjust RV1 until the correct reading is shown.

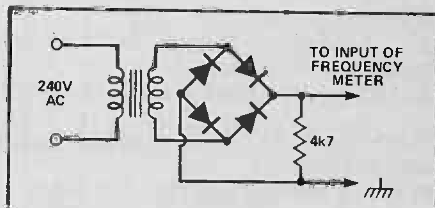


Fig. 3 A simple calibration arrangement using the mains frequency.

Construction

Construction should not present any problems as long as you check carefully to make sure all the components are in the right place. Since there are a lot of

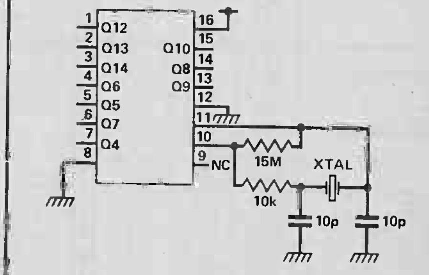


Fig. 4 A crystal and a divider arranged to give a number of calibration frequencies.

HOW IT WORKS

The input waveform is applied via C1 and R1 to a pair of diodes D1 and D2 which allow high input voltages, well in excess of the supply voltage, to be used without upsetting the LM358. R2 and R3 reduce the input voltage further to prevent the op-amp being taken outside its operating limits on negative excursions of the waveform. Strictly speaking, it is already operating outside its limits if pin 3 drops below 0V, but in this application it doesn't matter. If this worries you, or you'd like to know why — ask Auntie Static!

IC1a is used as a comparator and R4 supplies a smidgin of hysteresis to ensure it switches reliably. R5 pulls the output right down to 0V on low outputs and R6 and ZD1 set a fairly well defined upper limit to the voltage applied to C2.

When the output of IC1a is high, current from the 'right hand' terminal of C2 flows to ground via D3. When the output goes low, current flows into the right hand plate of C2 via D4, from the virtual earth at IC1b pin 6. On their own, RV1 and R8 would produce a voltage proportional to the instantaneous current from pin 6, by the action of IC1b, but C3 smoothes this out so the output from IC1b is proportional to the average current. R7 pulls the output down to 0V when no input is present. This is not a good idea in general, but works with the LM358 for reasons which I will also leave Auntie Static to explain.

A quick rule of thumb you can use to calculate component values for other ranges is:

$$v_{out} = fVC_2(RV1 + R8)$$

where f is the input frequency and V is the zener voltage.

C3 is chosen as a compromise between output ripple (reduced if C3 is increased) and settling time (increased if C3 is made larger). To measure very low frequencies you will find that a larger value of C3 will be needed to smooth out the ripple more effectively.

PARTS LIST

RESISTORS (all 1/4W 5%)

R1	10k
R2	10k
R3	22k
R4	10M
R5	2k2
R6	220R
R7	4k7
R8	68k
RV1	47k

CAPACITORS

C1	100n ceramic
C2	220p ceramic
C3	500n tantalum

SEMICONDUCTORS

IC1	LM358
D1–4	1N4148
ZD1	5V1 Zener

MISCELLANEOUS
PCB; PP3 battery connector; PP3 battery; BC3 case with battery compartment; On/off switch, rotary switch and knob and other parts for pre-scaler, if required (see text).

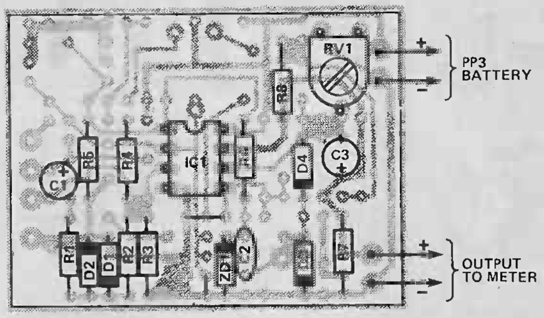


Fig. Component overlay for the frequency meter on the free PCB.

BUYLINES

Obtaining the parts for this project should not present any problems. You may be interested to know, however, that the LM358 ICs are available to ETI readers at the special price of 5 for £2 from: Specialist Semiconductors, Founders House, Redbrook, Monmouth, Gwent. The BC3 case with battery compartment is avialable from the same source at £2.80. Spare printed circuit boards can be obtained from us but stocks are limited. See the note in News Digest.

unused holes, some very close to the correct ones, it's easy to make mistakes. The best idea is to check the connections against the circuit diagram.

We built the final version into a box with a compartment for a PP3 battery — the same type as we used for the capacitance meter last month. The PCB can be held in place with sticky pads, since there are no mounting holes. With a bit of care, there should be plenty of room for the prescaler too although it will not, of course, fit on the PCB.

ETI

6809-BASED MICROCOMPUTER

Gary Mills concludes this series of articles with a discussion of the two main chips used and the software available to run on the machine.

Although the visual star of the show is the NEC7220A graphics controller, the 6809 is itself a very noteworthy chip. It was designed as a sophisticated answer to the 8080 and 8085 Intel chips, and incorporates a vastly improved set of addressing modes.

One of the most important features of the chip is the capacity to use program counter realtive addressing. This allows the programmer to generate Position Independent Code (PIC). PIC can be loaded and run anywhere in the available address space, and is easily ROMable. Both of these features are used to great advantage by operating systems which run on the 6809.

In terms of sheer speed the 6809 compares favourably with the Z80. A straight clock speed comparison is misleading because the Z80 takes an average of four clock cycles to perform instructions while the 6809 takes one. From benchmarks published in the 68XX Micro Journal, a two megahertz 6809 performs at almost exactly the same level as a 6 megahertz Z80. The 6809 in this design switches between 1 and 2 megahertz under software control.

The NEC7220A

The NEC7220A is a state-of-the-art graphics chip. While not a games chip, it has immense power for CAD applications and driving high resolution displays. It is currently used in the Epson QX10 and new QX16, and in the NEC personal computer series.

A full description of its capabilities would require several complete articles, but a brief list may help to convey the general idea:-

- line and arc drawing
- area fill
- zoom and pan
- up to four independently scrollable areas
- software definable character set
- resolution of 768h by 576v (in this computer)

It does all of this at the incredible drawing rate of 80 nanoseconds per pixel. To understand how fast this is, look at the four figures over the page. They can be drawn and erased on the screen in succession in under 7 seconds.

The publications listed below may be of interest to those who wish to explore in more detail the facilities offered by this chip.

uPD7220/GDC, uPD77220-1/ uPD7220-2 Graphics Display Controller

This is a fundamental document describing the chip's capacities and instruction set. However, it is terse and sparse on examples. It is published by NEC and can usually be requested when you purchase the chip, or ordered directly from them.

Product Description Graphics Display Controller uPD7220

Another NEC publication but better furnished with examples of the chip in use. It is available free-of-charge.

Application Note APN — 02 uPD 7220 Graphics Display Controller

Again from NEC, this document contains techniques and hints for programming the chip. It is available free-of-charge.

"Super Graphics Hardware from NEC"

An article by Steve Levin in BYTE magazine, April 1983. It gives a clear and concise summary of the chip's capabilities. A good introduction.

Monitor Routines Supporting Graphics

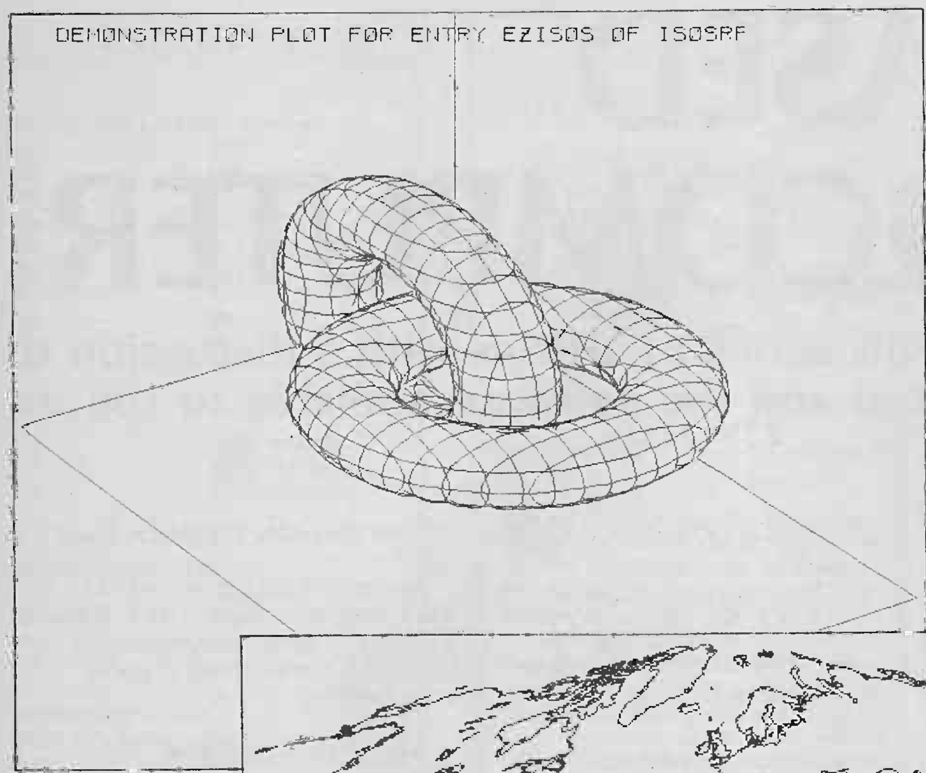
While contemplating the list of literature above, it may help to know that a great deal of work has been done for you. The kit for the project as supplied by Micro Concepts has a full set of assembly level graphics routines delivered in its firmware. They include:

- Set graphics cursor position
- Plot a point
- Plot a line
- Plot a rectangle
- Plot a circle
- Set a text cursor location
- Get a text cursor location
- Put a character on the screen
- Get or set a zoom factor
- Set an area fill pattern
- Fill an area

These routines can be called directly from your assembly language programs, and are fully described in the Microbox II literature.

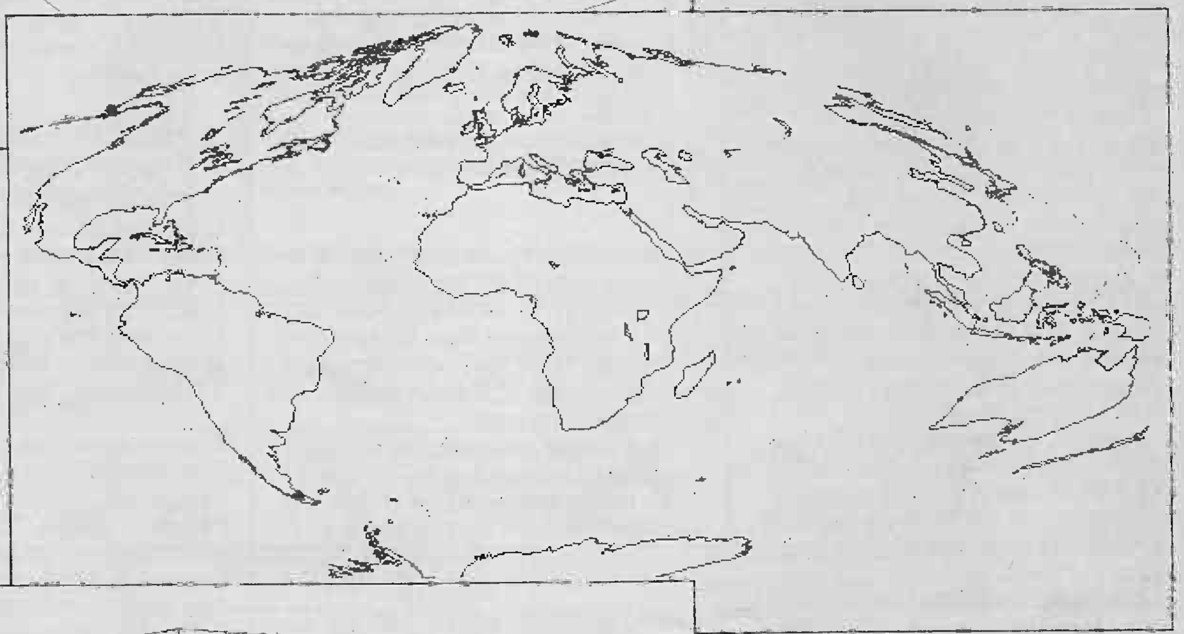
The Flex Operating System

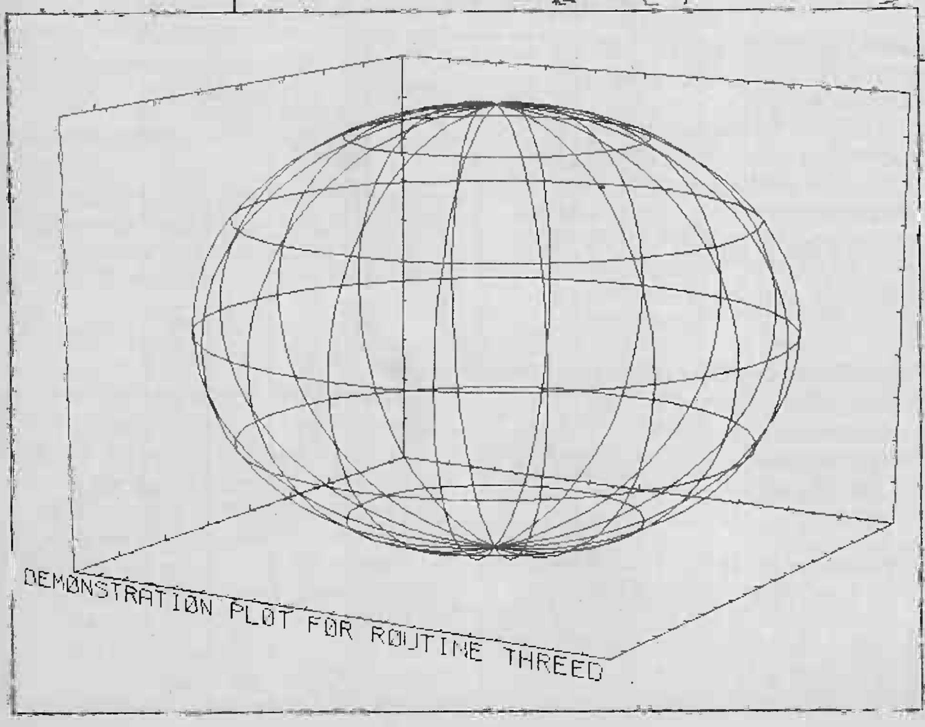
One of the strengths of this design is that it will run FLEX. Flex is a popular, standardised, and very useful operating system. It is perhaps not as well publicised as, for example CP/M because it has had different, less conspicuous areas of application rather than being used as a background for word processors and spreadsheets.



FLEX *can* run wordprocessors and spreadsheets (very good ones, too), but is primarily known as an engineer's operating system. It is consistently found on software development machines. Because of this there is a welter of languages, compilers, interpreters, assemblers and cross assemblers that run on FLEX, and because this machine can look like a standard FLEX machine to them, they all run on it.

The list of languages and applications that Micro Concepts sells is representative. The high-level languages include BASIC, both interpreted and compiled, COBOL, C, Small-C, PL9, Pascal, a Pascal cross-compiler for the 68000, Forth and BCPL. Assembly language packages include cross-





assemblers for just about every commercially available chip, relocating assemblers, debuggers, simulators and translators. Applications include word processors, mail merge packages, database managers, and spreadsheets.

Not only does FLEX support disk, EPROM disk and EPROM programmer in this design, but a graphics macro library compatible with FLEX's native macro-assembler is provided with the kit. Further, two of the FLEX compilers, PL9 and Small-C, have graphics libraries expressly written for this machine. Already there is a character set editor, graphics screen dump, save, and restore routines, several games and an interactive graphics package for The four figures with this article

SATELLITE TV ROUND-UP

Slow But Sure?

Britain's Independent Broadcasting Authority, the IBA, has adopted the slogan 'Evolution Not Revolution' in its attempt to win the broadcast industry over to its C-MAC television transmission system.

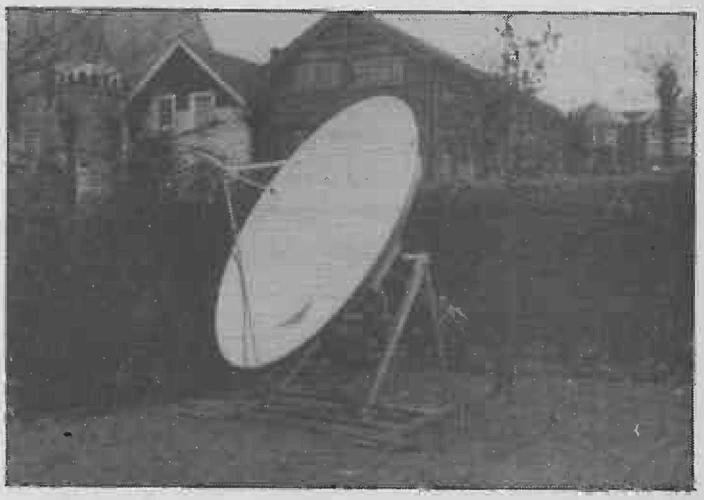
Designed for satellite-based systems in advance of the arrival of genuine DBS (direct broadcast by satellite), C-MAC is a step-up from the 625-line, 50Hz PAL system which dominates world TV broadcasting.

As a first step towards higher definition TV, C-MAC offers a 5:3 aspect ratio (like CinemaScope) and double the normal frame scanning rate. The result is a notably superior picture without too great a departure from current practice but capable of further improvement within a compatible system.

C-MAC has been accepted as the DBS standard by the European Broadcasting Union, is supported by Philips and other European manufacturers and is actually in use in Norway, but it faces strong competition from the 1125-line high definition TV system (HDTV) developed by Sony and NHK, the Japanese broadcasting authority.

Sony have recently demonstrated their system to plaudits from even their greatest opponents. The trouble is not that HDTV is technologically inferior. Far from it. The problem is that the system is not compatible with any existing broadcasting standard. Apart from that, it is expensive, requires tremendous bandwith and utilises a 60Hz field rate (alleged to prevent flicker). Even if the world adopts a 1125-line standard, 75% of all broadcast systems will still use a 50Hz field rate. Converters are inordinately expensive and cannot cope with the interference produced when 60Hz video equipment is used with 50Hz lights.

The Sony-NHK system may yet prove victorious — especially since the USA is one of the minority of countries operating on a 60Hz standard. Typically, we may expect Sony (with the aid of American friends) to try to batter us into submission, using a combination of seductive PR and crafty price manipulation. There is a real possiblity that the Japanese HDTV system may become a de facto international standard against the trend of the majority of TV users simply because those who might support the IBA and the EBU in Britain and elsewhere are too busy looking westwards for scraps from Ronald Regan's table to care what happens to European industry and innovation.



Good Connexions

A North London company claim to be the first to supply a full-band low-cost satellite TV receiving system aimed at the UK consumer. Connexions Satellite Systems, part of the SMC Supplies group, have three systems available comprising a 1.2 or 1.6 metre dish antenna, Low Noise Block converter and satellite receiver. The 1.2m system, designed to pick up 'Music Box' from the Eutelsat F1 satellite, costs £995 including VAT, but exclusive of delivery and installation. £1,045 will buy you a similar system with a fixed 1.6m antenna and for £1,295 you can get a system including a motor-driven 1.6m dish which can be readily adjusted to point at either of the two satellites currently broadcasting to Europe and the UK on the 12GHz band, Eutelsat F1 and Intelsat V. The component parts of the system can be bought separately and Connexions say they have 14 distributors throughout the UK. The company are mounting an advertising campaign in March, and they can be contacted at 125 East Barnet Road, New Barnet, Herts. EN4 8RF (tel: 01-441 1282).

X and Y

NEC Busines Systems (Europe) are claiming a first for their NESAT Satellite TV Receiver System. The system features a 1.8m antenna and a small-size Low Noise Converter mounted at the focus. apart from being particularly unobstuctive, the LNCs can be stacked at right angles and their IF outputs combined. In this way, the system is readily utilised for the reception of X,Y and mixed polarisation signals. NEC say that a complete system including antenna, one LNC and tuner with PAL UHF output will cost around £1,500.

Contact: NEC Business Systems (Europe) Ltd, 35 Oval Road, London NW1 7EA (tel: 01-267 7000).

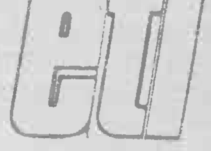
TV Times

● *A total of 18 channels are now available for satellite TV viewers in Europe, with two more channels from the Telecom 1 satellite planned for early this year. The channels are: Premiere (movies), Mirrorvision (entertainment), ScreenSport, The Children's Channel and Cable News Network from Intelsat; and Music Box, Sky Channel (entertainment, requires a decoder), TV5 (French language, general), RA1 (Italian language, general), Teleclub (German language movies), Filmnet (Dutch movie channel), Europa TV (general European programming), World Public News and Worldnet (US government news), SAT1 (publishers' channel), New World (religious broadcasts), 3SAT (German language, general) and RTL-Plus (entertainment) from Eutelsat. Only 14 of these, at most, are available at the same time, since Intelsat has only four and Eutelsat ten spare transponders.*

DB Prepared

A French subsidiary of Philips, Portenseigne SA, has announced production of a complete range of receiving equipment satellite TV aimed at every potential user from the large-scale cable and SMATV operator down to individual households. Portenseigne stress that they will be marketing an indoor adaptor for the reception of C-MAC broadcasts when higher-powered DB satellites come into use early in 1987.

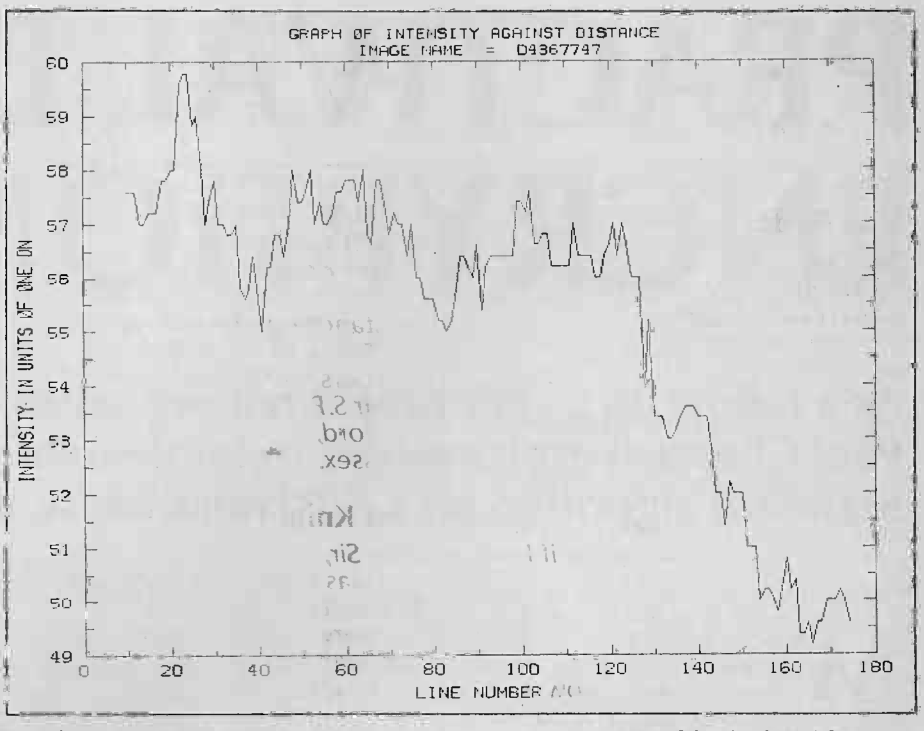
PS



A Mr. Gale informs us that he has bound volumes of ETI for disposal. The volumes in question cover the years 1972 to 1977. Mr. Gale can be contacted on 91-205 0221.

were dumped to printer using the routines above.

Another aspect of a machine or system's useability is the number of people using it, and their availability. For FLEX there are several sources of information and news. One is the "68XX Micro Journal", an American publication that can be purchased in London at Stirling Microsystems on Baker Street or subscribed to directly. There are also various Dragon and Tandy Colour Computer organs.



6809 User's Group

The 68 Micro Group is a very active user's group which includes a large number of user's who have already built this kit. They hold meetings every four weeks in London, and maintain a disk library of around 30 volumes for FLEX and the 6809. You can have access to this library as a member either at their meetings or through the mail. At the moment they have 3 full disks devoted especially to this machine including the character set editor, screen save and dump routines, and the PL9 graphics interface mentioned above. You can contact the membership secretary about joining. Jim Turner, 63 Millais Road, London E11, tel 01-558 3681.

A basic kit for this design (the Microbox II) is available from Micro Concepts, 2 St. Stephens Road, Cheltenham, Gloucestershire GL51 5AA, tel, 0242-510525.

ETI

PRACTICAL DATA ENCRYPTION

As a follow up to our recent feature article on data encryption, Paul Chappell explains how to implement the Data Encryption Standard algorithm on a BBC microcomputer.

Message scrambling, or encryption, is an area of growing interest, set against a background of industrial espionage and ever more sophisticated computer fraud. An introductory article in the September 1985 issue of ETI described some current encryption techniques, and we now present a simple add-on for the BBC micro-computer which will put one particular encryption method into practice. The circuit consists of a single IC, the Western Digital WD2002, which implements the Data Encryption Standard (DES) algorithm. For those who missed the original ETI article, the following is a brief description of the main features of DES.

The easiest way to use DES is known as the 'electronic code book'. The data input is presented to the DES IC in the form of a 64-bit block. This block could be eight successive bytes from an 8-bit computer, or perhaps eight successive samples from our sound sampler circuit if you want to scramble voice transmissions. The DES IC takes this block and performs a complicated mathematical process to produce an output block, also 64 bits long, of scrambled (enciphered or encrypted) data. At the receiving end, the encrypted data block is fed into another DES IC which will reverse the process and give the original (plaintext) data back again.

The term 'code book' is quite appropriate because you could achieve the same result manually by having a list of all possible input data blocks with their encryptions as a huge look-up table. As there are 2^{64} possible input blocks —somewhere around 20,000,000,000,000,000,000 —the table would be rather long. The process, as described so far, could be implemented as a pair of huge ROMs — one of 2^{64} locations of 64 bits each for encryption and another the same size for decryption. To make the system secure you would personalise it by choosing your own permutation of the input block for the encryption process. Unfortunately, ROMs of this size are not easy to come by, and would take about 4,000,000 years each to program (at 1 bit per second)!

The DES algorithm is personalised by means of a 56-bit long 'key' which determines how the encryption process is carried out. The same input block fed into any number of DES ICs, each with a different key, will always produce a different output block. In the electronic code book system, the sender would use a certain fixed key and the receiver would use the same key to unscramble the message. Anyone intercepting the message would find it virtually impossible to 'crack' — even with a DES IC at their disposal — unless they too knew the key. They could certainly have a bash at putting in keys by trial and error, but 2^{56} is an awful lot of possibilities to search through, and it is currently way beyond the scope of even the most powerful computers to decode a message by trying keys at random. The pundits suggest that computers capable of performing a complete search of all keys within a day will be available within the next ten years. Whether or not this need concern you depends on just who you expect to be intercepting your messages.

The security of the DES algorithm is a controversial topic and the subject of a great deal of research. At the time of its adoption by the USA National Bureau of Standards, rumours were rife that the algorithm had been deliberately weakened to allow interception of messages by government agencies, and that a trapdoor in the algorithm would allow rapid deciphering without knowledge of the key and so on. As far as I know, nothing has ever been published to substantiate the rumours.

Anyone intercepting the message ... could certainly have a bash at putting in keys by trial and error, but 2^{56} is an awful lot of possibilities to search through ....

Leaving this aside, it is certainly true that DES encrypted messages will eventually succumb to an attack by statistical methods — given sufficient material encrypted from the same key, a skilled cryptanalyst would be able to deduce the key and decipher the messages. It is also well established that some keys result in weakly encrypted outputs.

A number of ways to improve the security of DES-based systems have been proposed. One is to encrypt the same input block several times under different keys. The plaintext block would be fed into the first DES IC and encrypted; the output would be transferred to the next DES IC where it would be encrypted again with a different key, the output of this IC would feed the next, and so on as many times as you like. Although it seems intuitively obvious that the more times you encrypt the input the more scrambled it will become, this method must be used with caution. There is no guarantee that a sequence of keys chosen at random will, in fact, strengthen the encryption.

The very worst case would be a sequence which took the input block through several encryption processes whereby it eventually emerged unchanged. To see how this might happen, just imagine a very simple encryption process which changed all 1s of the input to 0s and 0s to 1s. If you tried to apply this twice in an effort to scramble it further, all the 0s would end up as 0s and all 1s as 1s. The data would not be encrypted at all! The chances of this happening with the DES algorithm are negligible, but you can see that there is no guarantee of improving security. Satisfactory multiple encryption systems have been devised, however, and you may like to follow the matter up in the paper mentioned at the end of this article.

A second method is to change the key block regularly — the more often, the better. Changing the key manually becomes impractical if it is done too often (remember that the person receiving the message must also know which key you are using at any time). Various ways have been devised to provide automatic key changes, in particular to use the DES IC to change its own key. The chain block system described in the September 1985 issue of ETI is one example.

The method used in this project — output feedback — is yet another way to confuse the issue. In this system the DES IC does not perform the encryption itself, but provides pseudo-random numbers for a Vigenere substitution — a variation of the shift cipher. In its simplest form, a shift cipher involves replacing each letter of the message with another letter a fixed number of places away in the alphabet. For a shift of two, every A in the original message is replaced by C, B becomes D, C become E, and so on. A cipher of this form could be cracked within a few minutes by any competent boy scout, yet it forms the basis of some very secure systems, including the unconditionally secure 'one time pad' (again see ETI September 1985).

The advantage ... is that the algorithm has been thoroughly investigated and you can be confident that your communications will be secure ...

```
 10 REM     ***DES PROGRAM***
 20 DIM Key(8),Int(8)
 30 REM       ***KEY BLOCK***
 40 M=&3000
 50 FOR I=1 TO 8
 60 READ Key(I)
 70 M?I=Key(I)
 80 NEXT I
 90 REM ***INITIAL I/P BLOCK***
100 M=&3008
110 FOR I=1 TO 8
120 READ Int(I)
130 M?I=Int(I)
140 NEXT I
150 P%=&3100:REM sets aside memory
              for assembler code
160 [
170 OPT 2
180 .entry LDA #&FF
190        STA &FE62
200        LDA #2
210        STA &FE60
220        LDX #0
230 .loop1 LDA &3001,X
240        STA &FD00
250        INX
260        CPX #8
270        BNE loop1
280        LDX #0
290 .loop2 LDA &3009,X
300        STA &FD00
310        INX
320        CPX #8
330        BNE loop2
340        LDX #0
350 .loop3 NOP
360        INX
370        CPX #&FF
380        BNE loop3
390        LDX #0
400 .loop4 LDA &FC00
410        STA &3009,X
420        INX
430        CPX #8
440        BNE loop4
450        LDA #1
460        STA &FE60
470        RTS
480 ]
490 REM ***MAIN BASIC PROGRAM***
500 INPUT "Input E to encrypt or
              D to decrypt";R$
510 IF R$="E" THEN 540
520 IF R$="D" THEN 650
530 CLS:PRINT "Mistake":GOTO 500
540 REM ***ENCRYPTION***
550 PRINT"Plaintext","Ciphertext"
560 INPUT P$
570 P=ASC(P$)-65
575 IF P<0 OR P>25 THEN 560
580 CALL entry
590 K=?&3010
600 X=K MOD 26
610 C=(P+X) MOD 26
620 C$=CHR$(C+65)
630 PRINT P$,C$
640 GOTO 560
650 REM ***DECRYPTION***
660 PRINT "Ciphertext","Plaintext"
670 INPUT P$
680 P=ASC(P$)-65
685 IF P<0 OR P>25 THEN 560
690 CALL entry
700 K=?&3010
710 X=K MOD 26
720 C=P-X
730 IF C<0 THEN C=C+26
750 C$=CHR$(C+65)
760 PRINT P$,C$
770 GOTO 670
775 REM ***KEY BLOCK DATA***
780 DATA 161,176,193,208,224,241,161,176
785 REM ***INIT I/P BLOCK DATA***
790 DATA 11,22,33,44,55,66,77,88
```

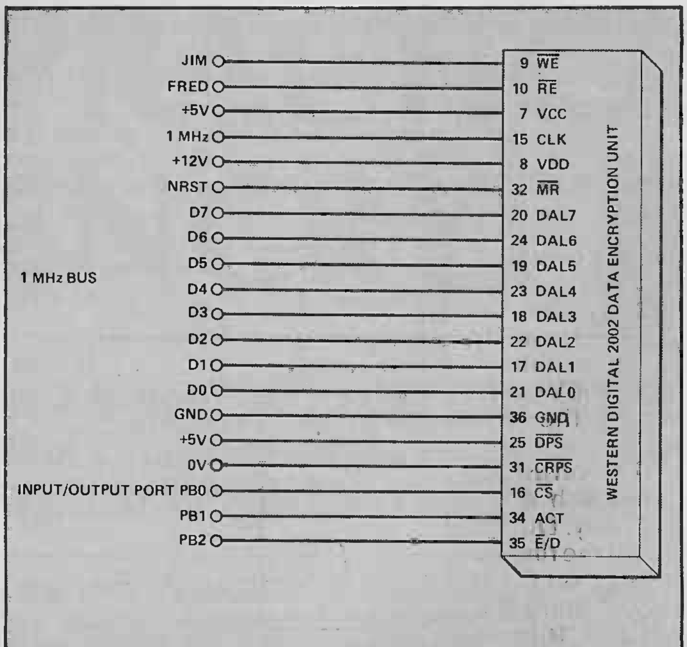


Fig. 1 Connecting a WD2002 to a BBC microcomputer.

The shift cipher is made secure by varying the shift for each letter; the more unfathomable the system for varying the shifts, the more secure the resulting ciphertext. The program for our project alters the shift length from the pseudo-random numbers generated by the DES IC. In the form presented here it will actually substitute one letter of the alphabet for another (instead of substituting one data block for another as required for more general use) and so can be used for ciphering written communications.

When you run the program, the computer will first ask whether you want to encrypt or decrypt. Type 'E' to 'scramble' your message, or 'D' to unscramble one you have received. The computer will then print two column headings: 'plaintext' and 'ciphertext'. Now type in your message, one letter at a time, followed by ENTER. If you are encrypting, the letters you type in will appear under the plaintext heading and the corresponding encrypted characters will appear in the ciphertext column. You will find that if you continuously input the same letter, the ciphertext will be an apparently random series of different letters, which should confuse any boy scouts who may be lurking around. The program, as it stands, will not encrypt spaces and punctuation in any sensible way, but modification of the program to do this should be a trivial matter for anyone moderately fluent in BBC BASIC. If you have received a message, enter 'D' to decrypt and type in the scrambled version, one letter at a time as before. The original message will appear under the plaintext heading.

It may seem that you could produce similar results by writing your own routine to generate pseudo-random numbers, without using the WD2002 at all. So you can, as long as you bear in mind that the receiver of the message must be able to generate the same string of numbers to be able to decrypt your message, which precludes using the computer's own random number generation. The advantage of using the DES IC is that the algorithm has been thoroughly investigated and you can be confident that your communications will be secure, in as much as the number of organisations which have the vast resources of skill, time and hardware needed to make any sense of them — if it can be done at all — is very small indeed. This will not be the case for most home-brew systems.

The DES algorithm is not secret — it is available in FIPS Publication 46, which you can obtain through your local library, by special request. The security of the system depends solely on the secrecy of the key you are using (in the context of this project, the secrecy of the key and the initial input block). The algorithm can be implemented entirely in software if your aim is to produce secret written communications as described above. For serious use, you must bear in mind that steps must be taken to protect your key — it's no good writing it into a BASIC data statement! You should also change the initial input block after each message, otherwise the same sequence of random numbers will be used each time you encrypt, allowing a relatively simple statistical analysis of your messages.

In the program for this project, the DES key is in the DATA statement in line 780. To change the key, you can either select eight bytes at random, as long as each has odd parity (an odd number of 1s in its binary representation) and substitute these for the ones given in the program. Otherwise, you can write down your chosen 56-bit key, split it into eight groups of even bits, add a 1 or a 0 to each group to give it odd parity, then use the eight bytes you end up with in the DATA statement. The initial input block is in line 790, and you can alter this in any way you choose — no need to worry about parity.

At the time of its adoption ... rumours were rife that the algorithm had been deliberately weakened to allow interception of messages by government agencies ...

For data transmission and real-time speech communication, software implementation is much too slow to be of any practical value, even on computers a good deal faster than the BBC. The various DES ICs, since they are pieces of hardware dedicated to that one function, are very much faster, making speech scrambling and high speed encrypted data transmission a practical proposition. I would suggest the analogue board of the sound sampler with a DES IC and some control logic as a suitable starting point for experiments in speech scrambling.

Finally, I have had a metaphorical slap on the wrist from John Pritchard, data encryption expert at the National Computing Centre, for using the term 'scrambling' in connection with the DES algorithm. He tells me that this suggests the output is merely a permutation of the input block, which it certainly isn't. My apologies if anyone was misled.

The Data encryption standard is published in full in: 'Data Encryption Standard', FIPS Publication 46, National Bureau of Standards, Washington DC, USA. (1977).

Further details of multiple encryption can be found in: Merkle, Ralph C., 'On the Security of Multiple Encryption', Communications of the ACM, vol. 24, no. 7 (1981).

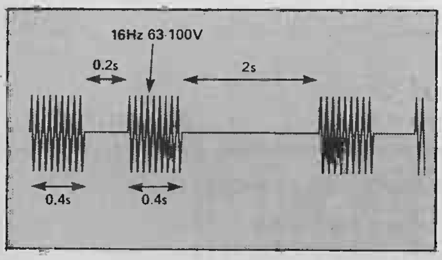
ETI

TECH TIPS

Telephone Answering Device

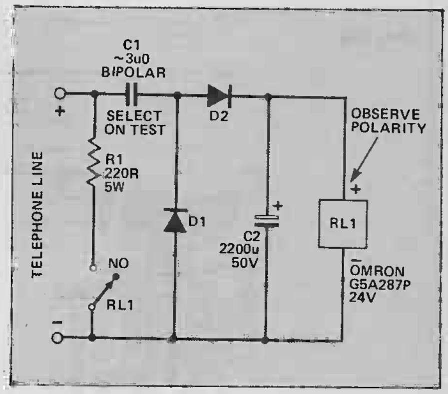
Geoff Phillips
Durham

The phone rings, you answer, and the caller hangs up. It's an uneasy feeling: is somebody checking that you are at home? when you are out you can't help wondering if they are checking again, and whether you'll come home to a burgled house.



This device answers your telephone after a few seconds delay and then hangs up on the caller. The circuit requires no external power but operates on the AC ringing signal which is transmitted by the exchange to your phone. This is usually a 16Hz voltage with an amplitude in the range 63 to 100V with cadences as shown in the first figure.

The answering device consists of a diode pump circuit which energises a high sensitivity relay. During the positive half cycles of the ringing tone, some of the charge on C1 is transferred to C2 via D2. During the negative half cycle, C1 is 'topped up' again via D1. Charge builds up gradually on C2 and when the pull in voltage of RL1 is reached, its normally open contact closes and places R1 across the telephone line. This has the same effect as lifting your handset and answering the call. The exhange then discontinues the ringing tone. The charge on C2 discharges into RL1 and when the drop-out voltage is reached the relay disconnects R1 from the line again. The impression a caller gets is that the line is answered after an acceptable delay, after which there



are a few seconds of silence and a subtle click as RL1 drops out.

The delay before answering is controlled by the value of C1 and this may have to be selected on test.

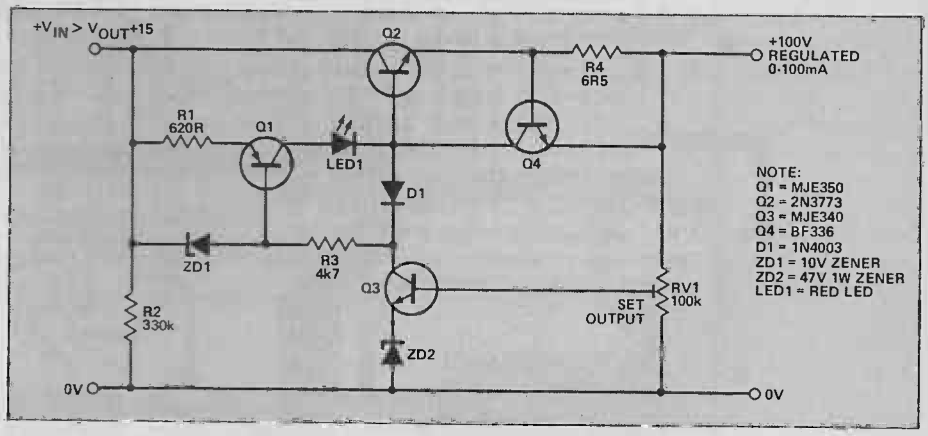
It should be pointed out that BT do not like indiscriminate connection of unapproved paraphernalia to their lines and this circuit should only form the basis of a design submission for approval with BT. the recommended RL1 type is the Omron G5A287P-24V which has BT approval.

Regulated PSU with Current Limiting And Short Circuit Protection

A. J. Holme
York

This design can provide from 50V to 115V and the output is current limited to 100mA maximum. Higher currents may be supplied with simple modifications. A somewhat novel feature is that the supply will switch off if a short circuit or excessively low load resistance causes the output value to drop below its preset value.

Q4 current-limits the output to the desired maximum by depriving the pass transistor Q2 of base current. Within the limits set by the mains transformer you are using, the limiting current can be increased by inserting diodes in the emitter lead of Q4 or by reducing the value of the current sampling resistor, R4.

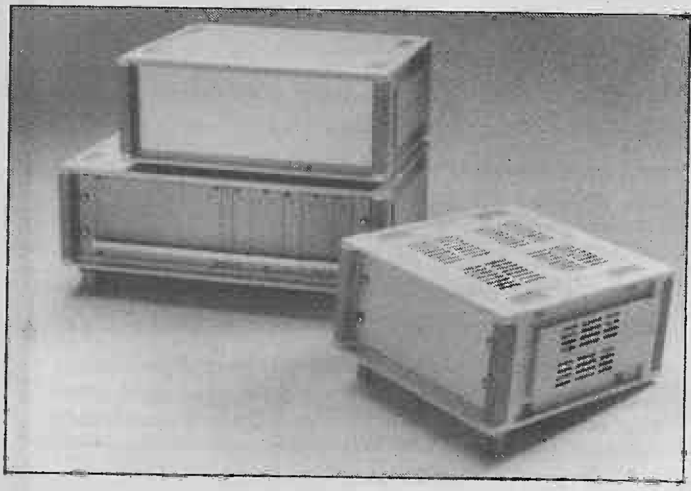


Regulation of the output voltage is controled by Q3 which shunts drive current from the base of the pass transistor, Q2. The resultant voltage is set by RV1 and the 47V zener.

Q1 is a constant current source which produces 15mA for Q2, Q3 and Q4 to share according to load conditions. If the output voltage is pulled low, Q3 switches off and Q1 base current will cease. This causes the whole circuit to turn off, thus providing overload and short circuit protection. Vin must be removed and re-applied to re-start the circuit. The LED extinguishes when the supply is tripped.

Q2 is capable of providing much higher currents if required (provided suitable heatsinking is used), but make sure you understand the consequences of any modification before attempting it. 100V or more can be very destructive!

ETI



Moulded 19″ Cases

New from West Hyde Developments is a lightweight, moulded enclosure designed to house 19″ racking units.

The Internorm range is available in half, three-quarter and full width versions in heights from 3U (5¼″) to 6U. They incorporate fold-away tilt legs which are released by buttons at the sides and an integral handle recess at either end, and West Hyde say they are strong enough to support a man's weight. Wall-mounting and panel mounting versions are availabel as well as free-standing types and other options include ventilation slots and dust/water protection to IP54.

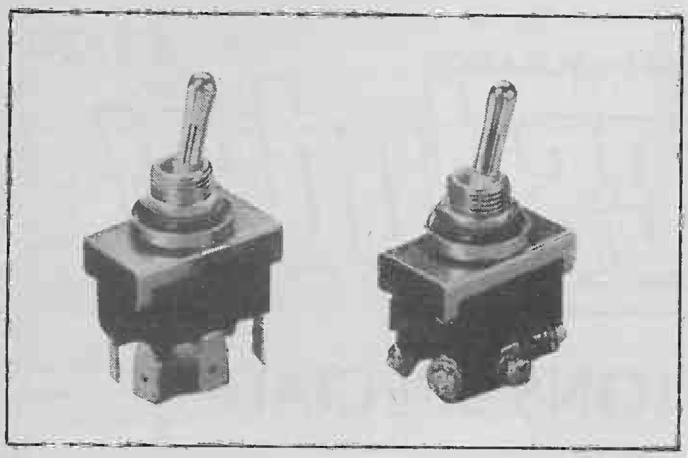
3U high cases are available now ex-stock and larger cases will be supplied to order as they become available. West Hyde Developments Ltd, 9-10 Park Street Industrial Estate, Aylesbury, Buckinghamshire HP20 1ET, tel 0296-20441.

● The latest publication from BICC-Vero is a sixteen page, full colour catalogue which describes their range of plastic instrument cases. The range includes small held-held types, flip-top cases in various sizes, free-standing instrument cases with moulded or aluminium front and back panels, sloping-front cases and tilt-up cases which have a carry-handle cum stand. Accessories such as battery compartments and battery connectors are also listed. For a free copy contact BICC-Vero Electronics Ltd, Unit 5, Industrial Estate, Flanders Road, Hedge End, Southampton SO3 3LG, tel 04892-5824.

● **The Institution of Electrical Engineers have issued further amendments to the 15th Edition of the Wiring Regulations which take effect as from the first January 1986. The principal change is the addition of regulations to cover the use of low voltage control circuits in shower units and other bathroom equipment, but there are also some changes to existing regultions and a number of points have been clarified. Copies of the amendments can be obtained from IEE, PO Box 26, Hitchin, Hertfordshire SG5 1RS for £1.00 including postage. Copies will also be on sale to callers at the IEE, Savoy Place, London WC2.**

● *Dage have released a catalogue of international standard connectors manufacturered by Positronic Industries. Thirteen connector families are described in its 108 pages with photographs, detailed drawings and full dimensions. Also included are cross-reference guides to equivalent products from other manufacturers and to military parts numbers. Contact Terry Reeve, Dage (GB) Ltd, Eurosem Division, Rabans Lane, Aylesbury, Buckinghamshire HP19 3RG, tel 09296-33200.*

● Two people who work for Maplin Electronic Supplies at Southend have raised £420.00 in a sponsored parachute jump for charity. Technical Artist Lesley Foster and Deputy Sales Manager Paul Ridler undertook the jump with four friends in aid of the Child Care Unit at Rochford Hospital. None of them had ever jumped from a plane before. They plan to repeat the jump next year with other Maplin staff and hope to double their sponsorship target.



Waterproof Toggle Switches

Mains-rated toggle switches for use in damp or humid atmospheres are among the new items listed in the latest edition of the B & R Electrical products catalgoue.

The Series AT switches are available with current ratings from 6a to 25a at 125 or 250V aC, come in single and double pole versions and offer eight different switch functions and five terminal styles. They are resistant to cold, vibration and mechanical shock and B & R claim that they will continue to operate correctly even if the terminals are loosened or the case deformed by heat.

The Series At switches are just one of many new lines included in the fourth edition of B & R's mail order catalogue, which, at nearly 200 pages, is almost twice the size of their previous catalogues. Copies are available free-of-charge from B & R Electrical Products Ltd, Templefields, Harlow, Essex CM20 2BG, tel 0279 4561.

Cortex Users Group

KPH Computaware inform us that they have taken over responsibility for the Cortex Users Group from Powertran.

The group was originally set up by Powertran when they were marketing kits of the Cortex microcomputer based upon the ETI design. In its new guise, the group will offer a quarterly newsletter and a range of software and hopes to be able to supply hardware in the near future. The annual subscription is £5.00 and all enquiries should be addressed to KPH Computaware, 63 Highlands Road, Andover, Hampshire SP10 2PZ.

● RR Electronics can now supply the LB5410 blue light emitting diode from Siemens which was described in News Digest in January 1985. They can also supply the Siemens Light Bar displays which we described in the December 1985 issue. Contact RR Electronics Ltd, St Martins Way, Cambridge Road, Bedford MK42 0LF, tel 0234 - 47211.

● **The Royal Television Society are inviting nominations for two of its annual awards. The John Logie Baird Travelling Scholarship is worth £1,500 and will be awarded to a post-graduate student at a UK educational establishment carrying out research in television, electronic engineering or a related field. The money is to be used for a period of study overseas. The Geoffrey Parr Award is presented in recognition of an outstanding contribution to television engineering or an associated science. Contact Francesca Smith, the Royal Television Society, Tavistock House East, Tavistock Square, London WC1H 9HR, tel 01-387 1970.**

● *The London Electronics College celebrates its 80th anniversary this year and plans to mark the occasion by tracking down its oldest former student. It was founded in 1906 as the British School of Telegraphy and reckons to have trained some 5,000 students. If any elderly ex-student is so hip and up-to-the-minute as to be reading this magazine, the college would love to hear from you on 01-373 8721.*

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We have a small number of these 3½-digit LED panel meters supplied to us by a well-known manufacturer and as used in professional equipment. The bright, 0.5" display is easily readable even in low-light conditions. The units are powered from a single 5V supply to give, without any external circuitry, a full-scale reading of 199.9mV. They are supplied complete with instructions giving details of calibration and range-setting via on-board links and trimmers.

These units are accurate and sensitive and can be used as the basis of any kind of meter. They are unbeatable value at the offer price, which is inclusive of VAT and post and packing, being less than half the price of the cheapest comparable unit available from regular suppliers. In fact, they cost no more than a reasonable quality moving-coil panel meter.

As mentioned last month, we plan to bring you some applications circuits for these units. Unfortunately, illness has prevented us including such circuits this month.

Cut out and send to ASP Readers' Services (RO ET2), PO Box 35, Wolsey House, Wolsey Road, Hemel Hempstead, Herts. HP2 4SS. Or telephone your ACCESS or VISA number to 0442-41221.

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NEWS: NEWS: NEWS

DIARY

Digital Techniques In Television: Recording — March 4th
The IBA, Bromptom Road, London SW3, 6.30 pm. Ninth in a series of lectures organised by the Royal Television Society. Cost is £13.80. Contact Liz Rankin on 01-387 1970

Power UK '86 — March 4–6th
Kensington Exhibition Centre, London. See March '86 ETI or contact TCM Expositions Ltd, Exchange House, 33 Station Road, Liphook, Hampshire GU30 7DN, tel 0428-724 660.

Telecommunications R&D: The Door To Tomorrow — March 5th
The IEE, London, 5.30 pm. Lecture given by W.T.G. Jones of BT, in association with the IERE. Contact the IEE at the address below.

The Application of Surface Mounting Techniques in Broadcast Engineering — March 6th
The IEE London, 5.30 pm. Lecture by A.M. Stark of Mullard Ltd. Contact the IEE at the address below.

System-On-A-Chip Seminar — March 6th
The Meeting Rooms, London Zoo. Seminar by Hitachi on their 64180 microprocessor. Includes introduction of a single-board evaluation system for the 64180. Cost is £25.00 plus VAT. Contact Miss S. Walker at Hitachi, tel 01-861 1414.

Electronic Production Efficiency Exposition — March 11–13th
Olympia, London. See November '85 ETI or contact Network Events Ltd, Printers Mews, Market Hill, Buckingham MK18 1JX, tel 0280-815226.

Digital Techniques In Television: Advanced 625-line Broadcasting — March 11th
The IBA, London, 6.30 pm. See entry for March 4th above.

Digital Techniques In Television: High Definition Television — March 18th
The IBA, London, 6.30 pm. See entry for March 4th, above.

Electro-Optics/Laser International — March 18–20th
Metropole Convention Centre, Brighton. See March '86 ETI or contact Cahners at the address below.

Low Energy Ion Beams — April 7–10th
University of Sussex, Falmer, Brighton. See March '86 ETI or contact the Meetings Officer of the Institute of Physics, 47 Belgrave Square, London SW1X 8QZ, tel 01-255 6111.

Design of Printed Circuits — April 7–11th
Cranfield Institute of Technology, Bedford. Course aimed at draughtsmen/designers starting work and at those converting from mechanical to electrical design. Contact Brian Phelps on 0234-750113 extension 2737.

British Electronics Week — April 29–May 1st
Earls Court and Olympia, London. Single event combining the All Electronics/ECIF Show, Fibre Optics, the Circuit Technology Show and Electronic Product Design. Contact Evan Steadman Services Ltd, The Hub, Emson Close, Saffron Walden, Essex CB10 1HL, tel 0799-26699.

Electrical Insulation Conference — May 19–22nd
Brighton. See March '86 ETI or contact the British Electrical and Allied Manufacturers Association, Leicester House, 8 Leicester Street, London WC2H 7BN, tel 01-437 0678.

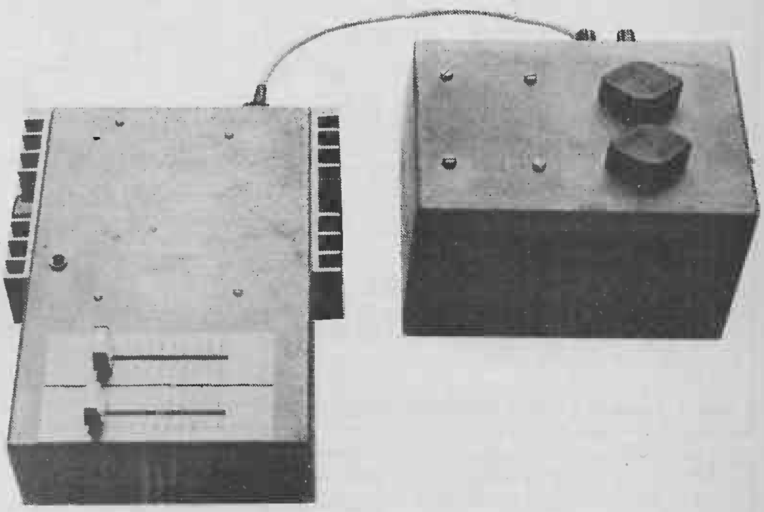
Advanced Infrared Detectors And Systems — June 3–5th
Institution of Electrical Engineers, London. See March '86 ETI or contact the IEE at the address below.

Addresses:
Cahners Exhibitions Ltd, Chatsworth House, 59 London Road, Twickenham, Middlesex TW1 3SZ, tel 01-891 5051.
Institution of Electrical Engineers, Savoy Place, London WC2 0BL, tel 01-240 1871.
Online Conferences Ltd, Pinner Green House, Ash Hill Drive, Pinner, Middlesex HA5 2AE, tel 01-868 4466.

ETI

CONVERTER FOR PA AMPLIFIERS

A portable public address system of reasonable quality would prove useful at many outdoor events, but the absence of a mains supply poses a problem. John Linsley Hood paves the way for his PA amplifier design with a DC-DC converter which produces 55V from a 12V car battery.



One usually sees electronic circuit designs only in their final form, as though the designer had put down on paper precisely what was required and then, perhaps, instructed some minion to assemble and test the design — just to make sure it did indeed perform exactly as specified.

If the design required is very simple or if the designer has had considerable experience in producing similar designs, the above procedure might be followed. More often, without the services of a careful and trustworthy minion and where the requirements are fairly straightforward, the circuit will probably go straight from brain to hardware and nothing will be put on paper until all is complete and satisfactory.

However, especially when working in unfamiliar territory, a prudent designer will look through the files and the back numbers of electronics periodicals to see if anyone has done something similar before. If the previous author has also described all the snags, much trouble can be spared.

In the case in point, the design of the audio ampifier would be quite straightforward for me but the provision of a DC supply would be much less familiar territory. Since the final design involved a couple of false starts, abandoned because of afterthoughts, it occurred to me that it might be interesting to describe the thought processes involved as the final design took shape.

The DC-DC Converter

The process began with the decision that the required audio power output was to be 50 watts into a 4 ohm load. Four ohms is a fairly common value for public address speaker units and 50 watts does make quite a loud noise, especialy when you bear in mind that PA driver units are usually a lot more efficient than their hi-fi equivalents.

This power/load combination would require an output AC drive voltage of 14.14V RMS ($V = \sqrt{P.R}$). This is 40V peak-to-peak. Since the power amplifier output stage will inevitably have a DC voltage drop at full power of some 12-15V, a DC power supply of 55V will be needed. A similar calculation shows that the RMS current into the load will be 3.54A, which implies a half-peak (average) DC supply current of some 1.77A. Allowing for other amplifier needs, some 2A should be allowed.

Since 2x55 = 110W, the overall efficiency of the system will be 45% and the heatsinks will need to be able to dissipate some 50 watts (1.77x55-50.) Also, because the DC-DC converter will be only 85% efficient at best, the demand from the battery will be 10.5 amperes at maximum output. If the converter system is reasonably well designed, the quiescent current demand at zero or low volume levels should be a good bit less than this, say 0.5A. However, the thought of a battery drain of 10.5 amperes makes one feel that there is a good case for not making it a 100W or 200W system.

The Circuit Design

Inverter circuits consist of some form of power oscillator driving a coupling transformer, usually at a frequency well above 50Hz in order to make the transformer a smaller and lighter unit. The several forms which this oscillator can take include single-ended or push-pull, self-oscillating or driven from some external source, forward or fly-back, and I have shown the layouts employed in Fig. 1.

The forward converter system is that in which the output transformer secondary voltage is generated during the period in

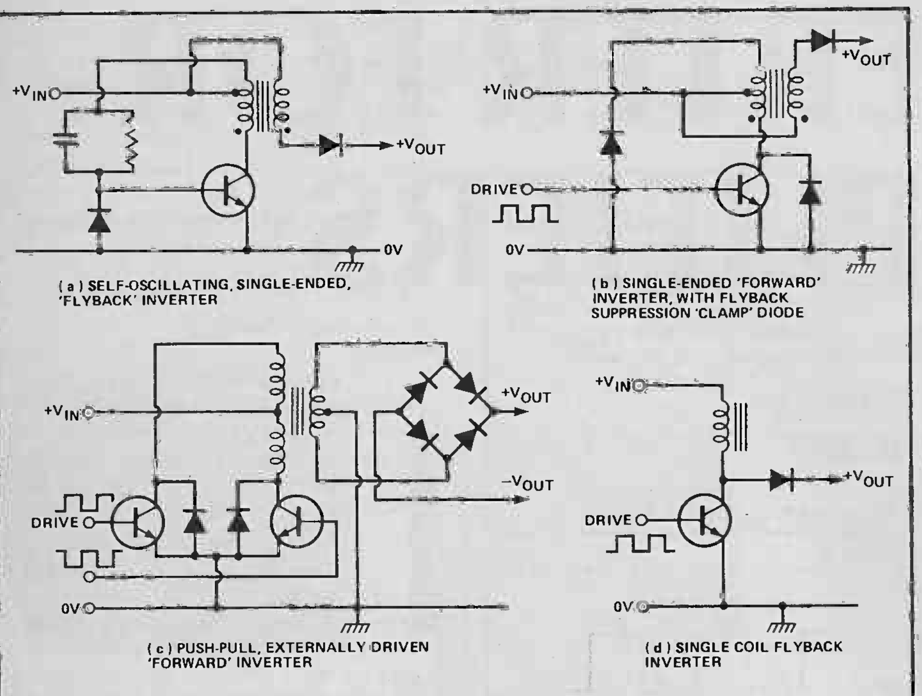


Fig. 1 Various circuit arrangements for an inverter.

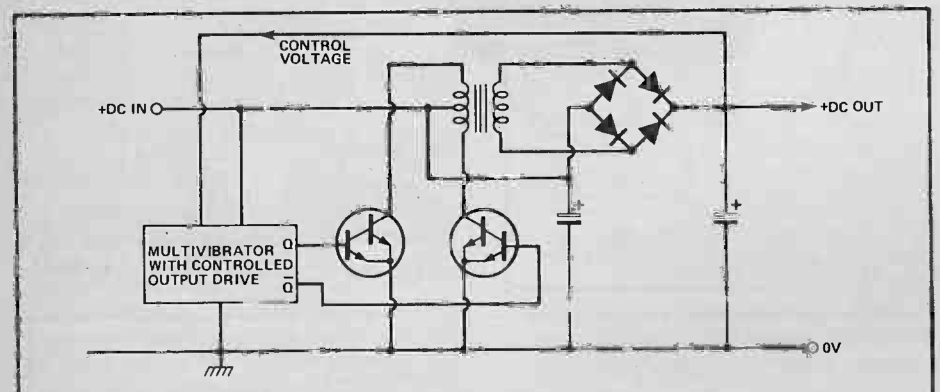


Fig. 2 A DC-DC converter with push-pull drive.

which current is flowing in the primary circuit — as in normal transformer operation. In general, this is a more efficient system than the flyback converter in which the secondary or even the primary, output voltage is that developed when the primary current is suddenly interrupted. The conventional motor car ignition coil is an example of a flyback converter.

Since we are hoping for the best practical efficiency in order to minimise battery drain, the forward system is to be preferred. Again, in the interests of efficiency, a push-pull system will be preferable, especially since there will then be two (or four) power switching transistors to cope with the 10 A primary current demand. This suggests the layout of Fig. 1c.

The next choice is between a self-oscillating circuit, in which the base excitation signal for the switching transistors is derived from a separate small winding on the step-up transformer, or one in which the switching waveform is derived from some external circuitry. This is also an easy choice to make. Self oscillating systems are notably dependent on the output load, and this means that their operating frequency can fluctuate, or that they can stop oscillating or even fail to start in the first place.

Also, if the excitation voltage is derived from some external small signal circuitry, it will be much easier to manipulate this to provide output power or voltage control. This leads to a circuit layout of the kind which I have shown in Fig. 2. The choice of power Darlington devices as the switching transistors, Q1 and Q2, is suggested because they are not a lot more expensive than ordinary power transistors but have drive requirements that are so small (10mA for full output) that the control circuitry need only use small signal transistors.

The Drive Circuit

Having decided upon the switching transistor configuration to be used at the output of the converter, the next thing was to decide what kind of mulitvibrator I should use to drive it.

The symmetrical emitter-coupled free-running multivibrator of Fig. 3, using a pair of PNP transistors, can directly drive the output (switching) transistors. Power control could be achieved by the circuit modification shown in Fig. 4, in which the current through the current source transistor, Q1, could be throttled back by way of the potential divider R6/R1 and D1 if the output voltage exceeded some predetermined value. This circuit works well, but requires that the supply voltage to Q1 (R7) should be held at a fixed level, to provide a reference voltage.

Unfortunately, for a power switching circuit it is essential that the switching devices should be

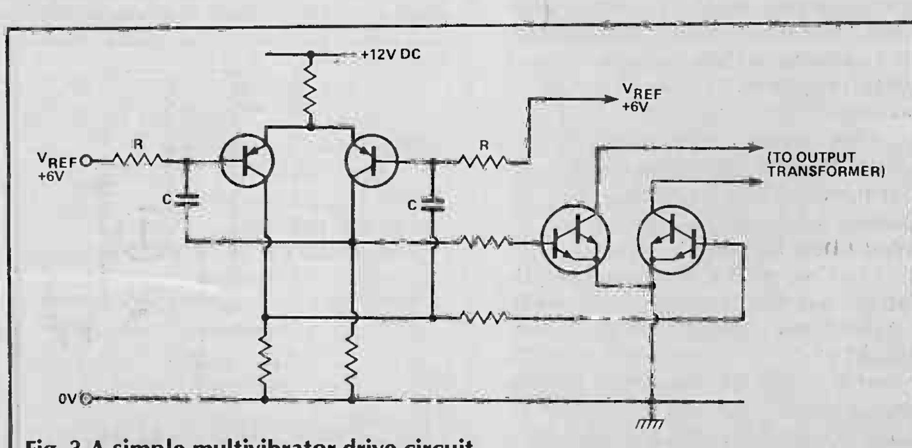


Fig. 3 A simple multivibrator drive circuit.