**EASY-PC, SCHEMATIC and PCB CAD**

**NEW VERSION:-**
**NOW DRAWS EVEN FASTER!**

![Image of EASY-PC, SCHEMATIC and PCB CAD]

**ONLY** £98.00

**Also:** over 400 Symbols!

Fast Professional Quality Output - Affordable Price

**SMITH CHART PROGRAM**

**Z-MATCH II**

For IBM, PC/XT/AT and clones.

Z-MATCH - Takes the drudgery out of R.F. matching problems. Includes many more features than the standard Smith Chart.

Provides solutions to problems such as TRANSMISSION LINE MATCHING for AERIALS and RF AMPLIFIERS with TRANSMISSION LINE TRANSFORMER and STUB MATCHING methods using COAXIAL LINES, MICROSTRIP, STRIPLINE and WAVES. The program takes account of TRANSMISSION LINE LOSS, DIELECTRIC CONSTANT, VELOCITY FACTOR and FREQUENCY. Z-MATCH is supplied with a COMPREHENSIVE USER MANUAL which contains a range of WORKED EXAMPLES.

£195 for PC/XT/AT etc

**CIRCUIT ANALYSIS BY COMPUTER**

**ANALYSER II**

For IBM, PC/XT/AT and clones.

"ANALYSER II" - Analyses complex circuits for GAIN, PHASE, INPUT IMPEDANCE, OUTPUT IMPEDANCE and GROUP DELAY over a very wide frequency range. Ideal for the analysis of ACTIVE and PASSIVE FILTER CIRCUITS, AUDIO AMPLIFIERS, LOUDSPEAKER CROSS-OVER NETWORKS, WIDE-BAND AMPLIFIERS, TUNED R.F. AMPLIFIERS, AERIAL MATCHING NETWORKS, TV.I.F. and CHROMA FILTER CIRCUITS, LINEAR INTEGRATED CIRCUITS etc.

STABILITY CRITERIA AND OSCILLATOR CIRCUITS can be evaluated by "breaking the loop". Can save days breadboarding and thousands of pounds worth of equipment.

£195 for PC/XT/AT etc

Write or 'Phone for full details:

**Number One Systems Ltd**

REF: EK, HARDING WAY, SOMERSHAM ROAD, ST. IVES, HUNTINGDON, CAMBS, PE17 4WR, ENGLAND.

Telephone: 0480 61778 (6 lines) International: +44 480 61778

The CAD Specialists

**The CAD Specialists**

REF: EK, HARDING WAY, SOMERSHAM ROAD, ST. IVES, HUNTINGDON, CAMBS, PE17 4WR, ENGLAND.

ACCESS, VISA, AMEX Welcome.
One of the greatest problems with speech recognition systems - or computers that listen - is that there is so much variation in the way people speak. A Scotsman may say a word one way, a New Yorker another, and even the same person may pronounce a word differently on separate occasions.

Traditional systems do not break sounds up into their component features, but instead use template methods whereby whole words are matched against stored patterns - and if they do not match perfectly, the system may not give the right answer. Seen in the photograph are scientists of the university's Centre for Speech Technology Research working on two stages of the process. In the background is Adlab, a commercial product developed at the Centre, and its display is showing voice patterns ready for processing. The equipment splits the patterns into different components or features, which are shown on the display in the foreground. It is these components that are used to identify the sounds that were spoken.

So far, scientists at the Centre have demonstrated the feasibility of their approach on a 4000-word system with 8500 pronunciations.

Centre for Speech Technology Research, University of Edinburgh, 80 South Bridge, EDINBURGH EH1 1HN Telephone 031 225 8883.
The electronic doorman enables a door to be opened automatically after a predefined delay from the moment a bell has been rung. It is intended for use in, for instance, waiting rooms and offices. The idea behind it is that the person who normally operates the door-opening control on his/her desk need not interrupt his/her work to open the door to visitors, patients or clients.

The voltage that serves to trigger the circuit is obtained by connecting inputs 'A' and 'B' in parallel with the bell, electric chime or buzzer. When a signal is detected, pin 2 of timer IC1 is pulled low by the phototransistor in optocoupler IC3. The time delay introduced by the timer may be set roughly between 3 s and 6 s with P1. After this delay, pin 3 of IC1 reverts to low. The trailing edge of this signal is converted to a short trigger pulse by C3-R4.

A second timer, IC2, is triggered and introduces a second delay of between 2 s and 6 s, set with P2. During this delay, the high output level at pin 3 causes the door opener to be actuated via driver TI and relay Rel. The relay contacts, C and D, are connected in parallel with the existing door-opening switch.

Resistor R7 eliminates inductive voltage peaks when the doorman is switched off to prevent erroneous triggering of the timers. Its value must be determined empirically. Clearly, it should not be so low as to cause the door to be opened at the moment the relay contacts are connected to the switch on your desk.

The presets on the board allow the wait time and the 'door open' time to be set to individual requirements. In practice, a 'door open' time of about 4 s gives the best effect.

(R. Dischler)

### PARTS LIST

**Resistors:**
- R1 = 1k0
- R2, R4 = 33 k
- R3, R5 = 22 k
- R6 = 4k7
- P1, P2 = 50 k preset

**Capacitors:**
- C1, C4 = 10 n
- C2, C5 = 100 µF, 25 V, radial
- C3 = 1 n

**Semiconductors:**
- IC1, IC2 = 555
- IC3 = TIL11
- D1, D2 = 1N4148
- T1 = BD139

**Miscellaneous:**
- Re1 = 12 V relay for PCB mounting (e.g. Siemens V23127-B00-A101)
- PCB Type 904002

---

**NUMERICAL PULSE-WIDTH CONTROL**

The control enables the pulse-width of a clock signal to be set with thumb-wheel switches. The pulse widths are set in five ranges: 1-999 µs; 0.01-9.99 ms; 0.1-99.9 ms; 1-999 ms; and 0.01-9.99 s. These ranges overlap to some extent and this is done on purpose to make available settings like 5.46 ms or 45.8 ms. The circuit has an error detector that gives a visual indication if the set pulse-width exceeds the period of the input clock signal. The permissible timing error is ±0.1 µs in all ranges.

HCMOS inverter IC6a and quartz crystal X1 form a 10 MHz oscillator whose output signal is divided by IC1, IC2 and IC3a to 1 MHz, 0.1 MHz, … 100 Hz. The required range is selected by S1. The signal is used to clock IC4 which, together with IC3b, forms the pulse-width counter.

The outputs of the pulse-width counter are applied to diodes and thumb-wheel switches. The AND function so created causes the output of IC6c to go low only when the count state in IC4 and IC3b is equal to the number set with the thumb-wheel switches.

The circuit operates at the leading edge of the input clock signal applied to IC5a. When a leading edge occurs, the output of IC5a goes low, thereby enabling counters IC1-IC4. The Q output, which forms the output of the circuit, goes high.

When the set time is reached, the output of IC6c goes low. Bistable IC5a is immediately reset whereupon its Q output goes high and the counters are reset. Consequently, the output of IC6c reverses to high so that IC5a is enabled again and ready to be clocked by the next leading edge at its CLK input. Meanwhile, the Q output of IC5a goes low and this marks the end of the output pulse.
If a leading edge occurs while the Q output of IC5a is high, a logic 1 is clocked into IC5b whereupon D13, the ERROR indicator, lights. It goes out again as soon as the error condition is ended by a change of range or set value.

IC6 is an unbuffered type that must not be replaced by an HC or HCT equivalent, otherwise the reliable operation of the oscillator is not guaranteed.

It should be noted that pulse-width settings between 0.1 μs and 99.9 μs with an input signal of 10 MHz may not work in all cases because the AND function formed by diodes D1-D12 and the associated pull-up resistor, R3, may not be fast enough.

The output pulses from IC5b may have to be cleaned or reshaped to eliminate overshoot.

The circuit draws a current of not more than 10 mA from a 5-V supply.

(C. Sanjay)

3 A, 5 V POWER SUPPLY

A standard voltage regulator Type 7805 is inexpensive and easily available, but its maximum current of 1 A can at times prove a handicap. However, this current may be increased by adding a power transistor (T3 in the circuit diagram below) on a heat sink. When the current drain is small, the 7805 continues to function as before. When the current rises above 15 mA, however, the potential drop across R4 is large enough to switch on T3. This transistor is protected against short circuits by T2. When the current through the MJ2955 rises above 3 A, the voltage drop across R3 is large enough to switch on T2. This limits the base-emitter voltage of T3, so that the output current can not increase much more.

In parallel with T2 is a transistor, T1.

Fig. 1. Circuit diagram of the power supply.
that switches on an LED as soon as current limiting occurs. Resistor R5 has been added to limit the current through the regulator as soon as the current limiting circuit operates, since R4 is then short-circuited by T2: in the absence of R5, the full current would flow through the 7805.

Of course, there is a price to be paid for the higher output current: the input voltage must be 10 V for an output current of 3 A, instead of 8.5 V for currents up to 1 A.

The current limiting comes in relatively gradually: when the output is short-circuited, a current of up to 6 A may flow for a short period. Obviously, that situation should not be allowed to last for long.

In the construction, take care that T2 and T3 are insulated from the heat sink. The 7805 does not really need a heat sink, but it does no harm to fit it also to the heat sink. If you follow the component layout in Figure 2 above, you should not experience any difficulties with the remainder of the construction.

(K. Walters)

**AF ISOLATING AMPLIFIER**

An isolating amplifier, also called buffer, is used to match two dissimilar impedance points and isolate one stage from a succeeding one in a cascaded system, and thus prevent undesirable interaction between them.

The present isolating amplifier has a bandwidth of 40 Hz to 40 kHz and a distortion of not greater than 1% at a 1 kHz signal of 70 mV r.m.s. The current drawn by each section is not greater than 10 mA.

The amplifier is based on an opto-isolator that provides the separation between the two sections. The LED in the isolator, a Type CNY21 or IL10, is driven by opamp IC1, a Type LF356.

Because the feedback resistor, R2, follows the LED, a large portion of the distortion produced by the LED is suppressed by the opamp. The bias current for the LED is adjusted with P1. In the present circuit, the level of this current is chosen at 1 mA, since that gives a reasonable compromise between the overall power consumption and the non-linear distortion.

The bias-current setting is not the only factor that determines the total distortion: the alternating current through the LED also plays a part. This is the reason that the primary section of the amplifier has been designed to cause an a.c. through the LED whose level is about 10% of that of the bias current at a signal level of 70 mV r.m.s. (100 mV p-p). When this input level is exceeded, the distortion increases significantly, and it is, therefore, necessary, to limit the input level to the value stated.

The direct and alternating currents through the LED, IL and IL respectively, are calculated from:

\[ IL = \frac{U_{PI} (R2+R3)}{R1R3} \]

\[ IL = \frac{U_I (R1+R2+R3)}{R1R3} \]

where \( U_{PI} \) is the voltage at the wiper of P1 and \( I_I \) is the input voltage.

---

**PARTS LIST**

- **Resistors:**
  - R1 = 330 Ω
  - R2 = 470 Ω
  - R3 = 0.1218
  - R4 = 47 Ω
  - R5 = 18 Ω

- **Capacitors:**
  - C1 = 4700 μF; 16 V
  - C2 = 10 μF; 16 V

- **Semiconductors:**
  - D1 = LED, red
  - T1 = BC5578
  - T2 = BD140
  - T3 = MJ2955
  - IC1 = 7805

- **Miscellaneous:**
  - K1, K2 = 2-way PCB connector

---

**Fig. 2. The printed circuit board for the power supply.**
Adjustment of the primary section is effected by setting P1 to obtain a reading of 1 mA through a milliammeter connected in place of JP1.

The signal received by the photocell in the opto-isolator is amplified by a second LF356 whose gain is controlled by P2. After the LED current has been set, this preset may be adjusted to ensure unity gain of the entire amplifier.

The circuit needs two completely separate power supplies and this means two transformers or one transformer with two isolated secondary windings. The primary section needs a symmetrical supply, whereas a single 8–15 V supply will suffice for the secondary section.

---

**DEBOUNCING CIRCUIT WITH TWO OUTPUTS**

Any switch or key in a digital circuit may cause problems because mechanical contacts bounce up and down a few times before they close. Normally, this weakness is negated by an RS bistable, but this article shows that it also may be achieved by a monostable.

The two gates in the circuit diagram form a monostable with a mono time of 100 ms (the bounce time of a key is typically 20 ms).

In quiescent operation, the input of inverter IC1b is at the level of the supply voltage, so that its output is low. This low level is connected to the input of IC1a via R3. The output of IC1a is thus high and C1 is not being charged.

When the switch, S1, is closed, the input of IC1a goes high because R1 has a smaller value than R3. The output of IC1a then becomes zero, which is immediately connected to the input of IC1b via C1. This low level remains at the input during a time determined by R2 - C2. Any bounce of the switch during this time has no effect whatsoever, because the output of IC1b, and thus the input of IC1a, is high.

When a switch or key is released, it will be noticeable at output B but not at output A, because C1 takes time to discharge. Only after it has discharged, can the monostable be triggered again.

The gates should be CMOS types, preferably of the HC/HCT series. The circuit works best with Schmitt trigger inverters, although most run-of-the-mill inverters work perfectly well.

The current drawn from the supply is negligible.

(From an idea by H. Smits)

---

**CLOCKWISE AND ANTI-CLOCKWISE DC MOTOR CONTROL**

This straightforward circuit, based on four darlings, enables a d.c. motor to rotate clockwise or anti-clockwise under the control of two digital signals provided by, say, a computer.

As may be seen from the diagram, the circuit consists of two identical sections. Concentrating on the left-hand section, when a high logic level (+5 V) is applied to input I1, T2 is switched on and a current can flow to earth via D1. T1 is cut off, because its base is negative with respect to its emitter owing to the voltage drop across the diode (−0.6 V). When a low logic level (0 V) is applied to I1, T2 is cut off and T1 obtains base current via R1. The motor can then draw current via T1.

The right-hand section operates in an identical manner.

By applying different logic levels to the inputs, that is, logic high to I1 and logic low to I2, or vice versa, the motor may be made to rotate clockwise or anti-clockwise, as the case may be. When the levels at the inputs are identical, the motor is at a standstill.

With component values as shown, motors needing up to 45 V at 2 A may be controlled. However, when the current exceeds 0.5 A, the transistors need heat sinks.

The circuit may be used to control the motor speed by pulse-width modulation. This requires a constant level at one input (depending on the direction of rotation), while the pulses are applied to the other input.

(R. Mensis)
As more and more electronics enthusiasts appear to have overcome their initial doubts, misgivings and fears of working with surface-mount technology (SMT) components, there is a growing demand for a universal board that allows prototype SMT circuits to be assembled quickly and reliably.

Since SMT component have no wire terminals, they can be fitted only by being soldered direct to the copper pads. The board presented here provides pads that are arranged in a pattern that enables virtually all types of SMT component to be accommodated.

(M. Fabisch)

Foldback Voltage Regulator

The usual series 7805 and 7812 three-pin voltage regulators are excellent for normal applications. If currents up to 3 A are required, an additional transistor, such as T2 in the diagram, is used. That solution works well, but the overall dissipation in case of a short-circuit can get fairly high. That creates a difficulty, particularly when a Type 7812, 7815 or 7824 is used. This difficulty may be overcome by so-called foldback regulation. This ensures electronically that the maximum current is reduced when the output voltage decreases. In the prototype, the maximum current with the output short-circuited was only 0.5 V so that over heating did not occur.

Only a few additional components are needed for foldback regulation. In the diagram, T1 provides current limiting. As soon as the voltage drop across R2-R3 becomes greater than 0.6-0.7 V, the transistor is switched on, which reduces the base current of T2 to virtually zero. The voltage regulator is then more or less on its own, but it has very good thermal protection and limits its output current well before any harm is done. The voltage at which the protection circuits come into operation is the sum of the potentials across R2 and R3. Resistors R3 and R4 form a voltage divider for the potential across T2. The dissipation in T2 is directly proportional to the collector-emitter voltage, which is thus used here to control the current. In this way, the regulation characteristic is a function of the level of the input voltage.

It is educational to experiment with the values of R2 and R3. When a short-circuit occurs, the drop across R3 should be large enough to drive T1 into virtual saturation. There is then practically no output current.

During testing, it will be noticed that 78xx regulators can withstand currents that are considerably larger than specified by the manufacturer (1-1.5 A), that is, until they get hot, when the maximum current level decreases.

(M. Fabisch)

Low-Frequency Sawtooth Generator

The most noteworthy element in this circuit is T1, which is known in data books by no fewer than three different names: thyristor tetrode, programmable uni-junction transistor (PUT) and silicon-controlled switch. In fact, the BRY39 is a four-layer (p-n-p-n) component. One of the characteristics of this type of component is that the junction of the two outer layers, that is, anode and cathode, begins to conduct when the potential across it exceeds a certain value. It ceases to conduct when
The current flowing from anode to cathode drops below a given level.

The sawtooth generator, which makes use of this property, is little more than an integrator of which the input is connected permanently to the negative supply rail. If it were not for Th1, the output of IC1 would rise slowly from 0 V to +15 V after switch-on and stay there. This voltage would also remain across C1 and Th1. This does not happen, however, because before the output of IC1 reaches +15 V, Th1 is switched on, which causes C1 to discharge rapidly so that the output of IC1 drops to 0 V. It does not reach that level, however, because before then the current through Th1 is too low to keep the device conducting. As soon as Th1 is switched off, the output of IC1 will rise slowly to +15 V, and so the action continues.

The voltage at which Th1 switches on may be preset within certain limits with P1. If it is set at 8.3 V (the positive peak of the sawtooth), the period would be of the order of 0.5R1C1, but this could, of course, be set to the required value with P1.

Note, however, that because of other properties of Th1 the value of R3 must remain between 300 kΩ and 2.2 MΩ. The value of C1 may lie between 1 nF and 200 μF. If large values are used, it is advisable to connect a 15 Ω resistor in series with the capacitor to limit the peak current during the discharge.

(R. Sanjay)

The LM3915 from National Semiconductor contains virtually everything for making a simple, yet reliable, audio power meter with a bar-type read-out. The circuit has one drawback in that it requires a separate power supply. This is compensated, however, by the fact that it is pretty sensitive (0.2 W min.) and does not degrade the sound quality in any way, since it does not present an additional load to the amplifier (in contrast, many inexpensive AF power indicators derive their display current from the amplifier).

The value of resistor R1 depends on the loudspeaker impedance as shown in the table inset in the circuit diagram. The resistor may be replaced by a wire link in the relevant position on the PCB if it can be fitted inside the plug that connects the indicator to the loudspeaker. This makes it convenient to use the indicator with loudspeakers of different impedance: use a dedicated cable for each impedance.

For use with stereo systems, the circuit is either built in duplicate or the signals across the loudspeakers are applied to two series-connected resistors R1, whose common junction is connected to pin 5 of IC1. The latter method may raise some eyebrows, but it works fine in practice.

The power supply for the indicator is

**PARTS LIST**

<table>
<thead>
<tr>
<th>Resistor</th>
<th>Value</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1</td>
<td>10 kΩ</td>
<td>(see text)</td>
</tr>
<tr>
<td>R2</td>
<td>10 kΩ</td>
<td></td>
</tr>
<tr>
<td>R3</td>
<td>390 kΩ</td>
<td></td>
</tr>
<tr>
<td>R4</td>
<td>2 kΩ</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Capacitor</th>
<th>Value</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1</td>
<td>22 μF, 25 V, axial</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Semiconductor</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>D1–D10</td>
<td>LED, rectangular</td>
</tr>
<tr>
<td>IC1</td>
<td>LM3915</td>
</tr>
</tbody>
</table>

**Miscellaneous**

- Enclosure: Pan-Tec Type PIN1064-01
derived from a simple a.c. mains adaptor that provides a d.c. output of 12-20 V.

Finally, it should be noted that the actual power measurement is an approximation only since the LM3915 reacts to the positive half-cycles of the signal only. This causes the top LED in the bar to light at a slightly reduced intensity. (K. Walters)

**011**

LINEARLY VARIABLE POWER SUPPLY

In most variable power supplies, the correlation between the setting of the wiper of the relevant potentiometer and the output voltage is non-linear. This results in the output voltage either having to be measured or indicated on an integral meter. If the relation were linear, a simple linear scale on the potentiometer would suffice.

The only difference between the present and a traditional power supply is that the wiper and the 'earthy' terminal of the relevant potentiometer are inter-linked. This simple fact allows a number of equivalent circuits to be improved. It is, of course, necessary to use a linear potentiometer.

Some parameters of the circuit: input-voltage: 28-37 V; output voltage: adjustable from 2 V to 25 V; maximum output current: 2 A.

It should be noted that the dissipation of the MJ3000 can rise to 50 W: it is, therefore, necessary that this component be fitted on a 1.5 K/W heat sink. (N. Karber)

**012**

POWER SUPPLY DOWN TO ZERO VOLTS

Most power supplies, particularly switch-mode types, are not designed to provide output voltages smaller than a volt or two. It is often desirable, however, especially in experimental work, to raise the output voltage slowly from 0 V.

The circuit shown here may be used with almost any power supply. The use of an auxiliary voltage, provided by T1-R3-D6, makes the supply act as if its output is equal to the internal reference voltage of 5.1 V, whereas in fact it is lower.

When the wiper of P1 is at earth potential, the circuit is an ordinary supply whose output may be varied with P2 between 5.1 V and 30 V. For the present purposes, P2 is permanently adjusted to give an output of 30 V: from now on the output will be varied by P1.

When the wiper of P1 is at earth potential, the output voltage is 30 V. Opening the potentiometer causes a larger and larger voltage from wiper to ground, which is derived from the auxiliary source based on T1. From the regulator, it therefore seems as if the output voltage increases, which, of course, is not so. The output voltage thus drops further and further the more P1 is opened.

In the prototype, an auxiliary voltage of 6 V was rather too low to reduce the output voltage to zero. A zener of 8.2 V was found more suitable.

The Type L296 regulator can provide a current of up to 2 A. A Type L4960 may also be used, but the maximum current is then somewhat lower. (SGS application)
A pulser is a special type of pulse generator that produces short-duration (fast) pulses. Such pulses are very useful for test and measurement purposes. For instance, in conjunction with an oscilloscope, they enable short-circuits and breaks in PCB tracks to be located quickly.

The pulser also makes it possible for the logic level to be ascertained at any point in a circuit where the probe is held. When spring-loaded switch S1 is pressed, a 1 μs pulse is generated at the opposite logic level from that at the relevant point in the circuit under test. Because of C5, a current of up to 500 mA may be drawn for the duration of the pulse.

The logic level at the test point is ascertained by IC2a, a D-type bistable to whose D input the measured signal is applied. At the instant S1 is pressed, the bistable receives a clock pulse upon which the signal at the D input is transferred instantly to the Q output and—inverted—to the Q output.

Two monostables, IC1a and IC1b, drive the two output transistors that generate the pulse.

The signal from S1 is applied to the B inputs of the monostables via delay line R2-C2. The monostables can be triggered by a leading edge at their B input only when their CLR input is logic high. Therefore, depending on the logic level at the D input of IC2a, one of the monostables will be disabled. The other is triggered and switches on the relevant transistor at its output for 1 μs. If, when S1 was pressed, the logic level at the probe was '1', transistor T2 is switched on by IC1b. If the level had been '0', T1 would have been switched on by IC1a.

Capacitor C5 ensures that sufficient current is provided at the test point. When no pulses are generated (pressing S1 generates only one pulse), C5 recharges via R7.

During operation, the pulser draws on average not more than 5 mA from a 9 V battery.

With circuit values as shown, this charger allows up to seven series-connected NiCd batteries to be charged. This number may be increased by raising the input voltage by about 1.65 V per additional battery. Provided T2 is fitted on an appropriate heatsink, the input voltage may be increased up to 25 V.

In contrast to many conventional NiCd battery chargers on the market, the present one provides protection against polarity reversal. That is, when the battery terminals are connected the wrong way to the charger, no charging current is produced.

Another useful property of the charger is that it does not load the battery when it is switched off.

NiCd batteries are normally charged over a period of 14 hours at a current equal to about 1/10th of the battery capacity. In other words, a 500 mAh battery is usually charged at 50 mA for 14 hours. Normally, a slightly longer period of charge will not harm the battery. If, however, the current is higher than indicated, the charging time must be reduced proportionally to prevent damage to the cells.

The level of charging current is controlled between 0 mA and 1 A with preset P1. The current may be measured with a voltmeter across R3, which, to avoid calculations, has a convenient value of 1 Ω.

Operation of the circuit is straightforward. Transistor T1 is on when a battery is connected with correct polarity and also if the output terminals are open-circuit. The collector current of this transistor results in a reference voltage of about 2.1 V across...
diodes D1-D3. Part of this voltage is applied to darlington transistor T2 via P1. The emitter resistor of T2, R3, provides the constant-current function. Note that T2 must be fitted on a heat sink.

In case the BD679 is difficult to obtain, it may be replaced by any n-p-n medium power darlington with a collector voltage/current specification of 30 V at 2 A.

The maximum output current may be increased above 1 A by lowering the value of R3.

The charger draws a quiescent current of 15 mA at 12 V.

(Ever Ready application)

**015 LOW-LOSS VOLTAGE REGULATOR**

The voltage regulator in portable, battery-operated equipment must have minimum dissipation, good temperature stability, and be able to deliver an adequate current. These requirements are met by the IRF9530 MOSFET series regulator from Precision Monolithics Inc. (PMI). One of its main attributes is the low drive current required: the two Type OP90 opamps that drive it and at the same time generate the reference voltage draw only 20 μA.

Circuit IC2 compares a portion of the output voltage existing across divider R6-R7 with the reference voltage provided by IC1. Its output adjusts the gate potential of the IRF9530 in a manner that ensures a constant drain voltage of 5 V. Resistor R7 is a pull-up resistor. At output currents between 0 A and 1 A, the gate voltage lies between 3.75 V and 1.9 V respectively.

Transistor T1 and circuit IC1 form a band-gap voltage reference. Because of its high stability and low current drain (about 5 μA), this type of reference is eminently suitable for this application.

Owing to the presence of R3 in the emitter circuit of T1b, this transistor draws less current than T1a. Because of this, the base-emitter voltage of T1b will be somewhat smaller than that of T1a. None the less, the collector voltages of the two transistors are practically equal since the value of R2 is much higher than that of R1. These collector voltages are applied to the two inputs of IC1 and their (magnified) difference is fed back to the bases.

The ratio R1:R2:R3 has been chosen to ensure good compensation of the temperature coefficient of the transistors.

The optimum operating point of the circuit coincides with a reference voltage of 1.23 V.

The circuit was originally designed with a Type MAT01 in the T1 position, but it will work satisfactorily with most types of dual transistor. It should be noted, however, that some other types were found very sensitive to tolerances. In some cases, this prevented the circuit from stabilizing on a fixed operating point, which resulted in the output remaining high or low. To en-
sure a well-defined output voltage at power-up, the offset adjustment (pin 5) of IC1 is connected to earth. This results in the output of IC1 always being logic high on power-up.

When a dual transistor other than the MAT01 is used, it is advisable to adapt the values of R3 and R4 by disconnecting pin 6 of IC1 from the bases of the transistors. Next, apply a voltage of 1.23 V (via a voltage divider) to the bases and vary the values of R3 and R4 (if necessary) until IC1 just does not toggle. Finally, test the temperature stability with a small blow lamp.

The IRF9530 only needs a heat sink if the input voltage exceeds 6.25 V.

The circuit is intended for use with input voltages from NiCd or low-capacity lead-acid batteries, which provide an inherently stable voltage. Therefore, the circuit does not contain any ripple suppression components. Furthermore, large variations in current are compensated only slowly. These small drawbacks may be avoided by the use of a standard reference and a faster opamp in the IC2 position. That will, however, result in an increase in current, which in the circuit as shown is only 45 μA.

(PAM application)

The fluid level indicator may be used, for instance, in the fresh water tank on board yachts, or in mobile homes or caravans.

The drivers in IC1 and IC2 are darlington transistors (of which the ULN2803 has eight and the ULN2003, seven). The base of each of these is connected to a sensor, formed by a carbon rod, or a strip of aluminium or copper. These sensors are fitted at the appropriate level in the fluid vessel.

The 15 LEDs ate the outputs of the drivers are arranged to display a bar that provides an easy-to-read fluid level indicator when S1 is pressed. Because of the relatively high current drain (about 300 mA) and the limited capacity of the battery on board the yacht or vehicle, it is strongly advisable not to use the indicator continuously (this would also cause rapid erosion of the sensors).

Since the ULN drivers are capable of supplying a peak current of up to 500 mA, one or more of the LEDs may be replaced by a relay, small buzzer, or other means of providing an audible warning for too high or too low a fluid level.

The ULN 2003 and ULN2803 may be replaced by similar devices from the same family of power drivers, for instance, the ULN2005 and ULN 2805, or the ULN2001 and ULN 2801. A word of warning if the 2001 and 2801 are used: DO NOT CONNECT any of the inputs of these devices direct to +12 V, since this is likely to destroy the relevant IC.

(D. Lorenz)

FLUID LEVEL INDICATOR
Described here is a maximum-level indicator that does not, or hardly, react to fast peak powers. It is intended to be driven by the output stages of an AF amplifier.

The section based on T1 is a traditional voltage regulator for the incoming supply. The values of R1 and R2 depend on the level of the supply voltage as shown in the table. The transistor does not need a heat sink.

The remainder of the circuit consists of two identical sections for the left- and right-hand channels respectively. Only the left-hand channel section is discussed.

The incoming signal is rectified by D4 and D5 and then charges C4. This capacitor discharges via R5 and R6.

Circuit IC1a acts as a comparator and is connected as a Schmitt trigger. It compares the rectified input signal with a reference voltage that is set with P1. Diode D3 gives this voltage an offset of 0.7 V, which is necessary in view of the hysteresis of IC1a. The hysteresis is determined by R6 and R7, but is reduced by the d.c. feedback network R8-R9.

The LED is driven by an emitter follower, T2.

The opamp is a Type LM358, which has the advantage of performing well even if operating with an asymmetrical supply when the inputs are at earth potential.

The signal input may lie between 2 Vpp and 10 Vpp. The toggle level of the circuit is arranged with preset potentiometers. The hysteresis is 400 mV over the whole range. The circuit is triggered at the same level over the frequency range 100 Hz to 15 kHz.

The circuit reacts to short bursts (such as those from the kettle drum) only if their amplitude is at least 50 per cent higher than the trigger voltage.

The single supply voltage may have a value of between +20 V and +60 V. Current drain varies between 10 mA (quiescent) and about 50 mA when both LEDs light.

(A. Ferndizen)

Supply voltage R1 R2
20 V 330 Ω, 1 W 3k3
30 V 470 Ω, 1 W 3k3
40 V 560 Ω, 2 W 3k3
50 V 680 Ω, 5 W 5k6

Almost all households have at least one infra-red remote control unit, be it for the video or audio equipment. Unfortunately, it sometimes happens that the control does not function properly and it is then difficult to ascertain whether the receiver or sender is at fault. The detector described here can help by telling us whether the sender works or not.

The IR light from the sender is detected by T1, an IR photo transistor. When IR light falls on to T1, it switches on T2 and this results in the LED lighting at the rhythm of the IR signals.

The brightness of the LED depends on the strength of the IR light falling on to T1 so that the remaining capacity of the batteries may be estimated.

Although a Type TIL81 is used here in the T1 position, virtually any IR photo transistor is suitable.

Since the current through the LED is fairly small, it is advisable to use a high-efficiency type.

(R. Systermans)
A 4-to-16 decoder is eminently suitable for constructing a simple four-position selector with integral debouncing of the switches.

The circuit is a cunning design that uses few components. The switches are connected to the inputs of the decoder via a resistor, an LED and a transistor. The transistor is in parallel with the switch, so that pressing the button has effect only when the transistor is off.

The transistors are driven by four outputs of the decoder via 1 k resistors.

When the supply is switched on, none of the transistors conducts, so that the inputs of IC1 are '1111' and pin 17 of the decoder is logic high. If in this situation one of the switches is pressed, the associated input bit to IC1 becomes logic low. This changes the input code and the output associated with the switch goes high, that is, pin 16 for S1, pin 15 for S2, pin 13 for S3 and pin 8 for S4.

The signal at the relevant output pin is used to switch on the transistor associated with the closed switch. From then on, the transistor assumes the task of the switch, so that this may be released. This condition is indicated by the relevant LED.

If one of the other switches is then pressed, the input code to IC1 contains two zeros. This actuates a decoder output that is not fed back to the transistors. The conducting transistor is then switched off, which results in the decoder input code changing once again, but this time to the correct code associated with the currently pressed switch. The decoder output associated with that switch then goes high and switches on the relevant transistor, which assumes the task of the switch.

The outputs of the circuit may be used, perhaps via an additional amplifier, to actuate four relays.

The current drawn by the circuit is determined primarily by the LEDs: as shown, the circuit draws about 10 mA.

(H. Smith)

---

It happens frequently that a circuit powered by a single supply voltage is extended and then needs a second supply voltage of opposite polarity to that already available. That requirement can be met by the circuit described here.

Normally, an auxiliary negative voltage would be provided by half-wave rectification by C1, D1, D3, and C3, but it is better to use full-wave rectification, because of the higher current and smaller ripple, and that is effected by the addition of C2, D2 and D4.

In the diagram, bridge rectifier D3–D8 and smoothing capacitor C4 provide the existing positive supply voltage. The added components provide a negative voltage at practically the same level as the positive supply, but that depends on the value of C1 and C2 and the current required. With values as shown, the negative supply can provide up to about 200 mA.

It should be noted that in this type of
circuit the current drawn from the positive supply must always be larger than that from the negative supply. If the positive supply is not loaded, the negative supply can not provide any current! Note also that in the diagram R1 and R2 are the respective loads, not actual resistors.

If it is required that the current from the negative supply is greater than that from the positive rail, the circuit should be inverted. The bridge rectifier should then provide the negative voltage and the additional components, the positive voltage. All diodes and capacitors should also be inverted.

(K. Walters)

021 
INEXPENSIVE TRANSISTOR TESTER

Resistance checks on suspect transistors normally fail to provide a conclusive answer as to whether the transistor is all right or defect. Moreover, such checks mean that the transistor must be removed from its circuit and connected in six different ways, that is, b-e test, b-c test and c-e test, each two times with reversed meter polarity.

The tester presented here allows the transistor under test (TUT) to remain in circuit, provided the circuit has relatively high resistance values.

The tester is suitable for p-n-p as well as n-p-n transistors and also accepts darlington. The selection between p-n-p and n-p-n is made by reversing the supply voltage with S2.

When the necessary connections are made, the transistor under test forms part of a collector-coupled astable. The transistor is almost certainly all right if it enables the astable to oscillate at about 2 kHz, which is indicated by buzzer Bz1. It should, however, be noted that transistors with relatively low current gain may pass this test and still be defect.

The small size of the printed-circuit board, and the use of two series-connected 1.5 V batteries allow the tester to be housed in a compact case. Connections to the TUT are by means of flexible test leads terminated in small crocodile clips.

In operation, the tester draws a current of about 20 mA.

(J. Ruffell)

022 
POWER SUPPLY INPUT ADAPTOR

This circuit enables the input voltage of a power supply to be adjusted in accordance with the required output voltage to eliminate undue power dissipation in series regulators. This is achieved by monitoring the output voltage and altering the unregulated input voltage as necessary with the aid of two relays.

The sensing input, derived from the positive supply terminal, is applied to potential divider P1-R1-R2-R3-R4. Comparators IC1a, IC1b and IC1c detect respectively when the voltage at the relevant junction of the divider is 1/4, 1/2 or 3/4 of the maximum sensing input set with P1.

ELEKTOR ELECTRONICS JULY/AUGUST 1990
Fig. 2a. The voltage drop across diode D1 serves as a 0.7 V reference.

Double-pole relay Rel is energized (a) when the output of IC1a is high and that of IC1b is low, and (b) when the output of IC1c is high.

Condition (a) pertains when the output voltage of the supply is between 1/4 and 1/2 of its maximum value, while (b) occurs when the output voltage is greater than 3/4 of its maximum value.

When IC1a is the only opamp whose output is high, the +ve input of IC1d is held at half the supply voltage. Since in that condition the −ve input is at 1/3 of the supply voltage, IC1d toggles and this causes Rel to be energized.

If the output of IC1b also goes high, the −ve input of IC1d becomes 2/3 of the supply voltage, the comparator returns to its original state and Rel is de-energized.

When the output of IC1c also goes high, IC1d toggles again and the relay is re-energized.

To prevent unwanted voltage dips when the circuit switches between the taps on the secondary winding of the supply transformer, it is better to use two relays as shown in Fig. 2b. This arrangement has advantages over that in Fig. 2a in which one relay controls four contacts. An even better arrangement is shown in Fig. 2c, since that obviates the use of a DPDT relay. Note, however, that a mains transformer with two separate secondary windings is required.

Preset P1 is adjusted to give the required supply output at which the relay is actuated. Switching levels of 10 V, 20 V and 30 V, for instance, are obtained with P1 set to a value of 125 kΩ.

The adaptor is particularly suitable for use with fairly large power supplies, for example, 40 V, 5 A types. The fourfold reduction in the unregulated input voltage results in a decrease in dissipation from 200 W to 50 W.

The adaptor draws a current of 5 mA, excluding that required by the relay(s).

(K. Walters)

**REPEATING “FIRE” BUTTON**

A joystick is an indispensable aid with most computer games. Apart from the stick that enables movement to the left, the right, upward and downward, the unit also contains one or two “fire” buttons. In some games, there is a lot of firing to be done, and this means continuous pressing and releasing of the fire button.

To prevent your getting a “tennis thumb” or “tennis finger”, it is advisable to use the circuit presented here, which provides automatic actuation of the “fire” button.

Two gates of a Type 4011 chip, IC1a and IC1b, form an astable multivibrator, whose frequency is varied with P1. The generated square wave pulses are applied to the “fire” button via gate IC1c, which is connected as a buffer.

Power for the circuit is derived from the computer via DI and CI.

The circuit is small enough to be housed in the base of the joystick.

(K. Nischalka)
The "MIDI signal redistribution" unit we published in May 1987 (p. 22) has even more possibilities than we thought at the time.

If the redistribution unit is used to the full, there is a veritable mass of interconnecting cables. Any changes, and a number of cables have to be rerouted. There is, however, a way of avoiding a lot of this work, as will be described here.

Requirements are: one MIDI redistribution unit and four two-pole, four-position switches. Connected as shown in the diagram, these make it possible to interconnect four MIDI instruments in a variety of ways without a vast number of cables. As a bonus, far fewer DIN connectors are needed than for the original set-up.

In the preset circuit, the redistribution unit is used as a four-way throughput. The original switches, S1 and S2, are not required: they are replaced by a simple wire link between contacts M and I. A wire link is also required between '2' and '3'. The remainder of the wiring is as shown in the present diagram.

From each quarter of the unit, cable connections run to the switches in the other quarters. Since the outputs of the quarters carry the same signals, it does not matter which switch is connected to a particular output. This is the reason that the connections are shown as a "bus", in spite of their being called '4' or '5' (that is, the pin numbers of the DIN connectors). The result is a much clearer circuit diagram.

A summary of the properties of the circuit is therefore:

- each MIDI output may be connected to the inputs of the other instruments;
- each input may be switched off;
- on each instrument, a MIDI output remains externally available.

(J. Blankaert)

The design of this meter is in direct response to the intention of most electricity generating boards in Europe of standardising on a mains voltage of 230 V, 50 Hz. The change from the current 220 V or 240 V (mainly UK and Eire) will be made gradually during the present decade.

The instrument provides an accurate indication of the mains voltage on an analogue meter, the scale of which ranges from 210 V to 230 V or from 230 V to 250 V, depending on the current mains voltage. This means that the current mains voltage is read at the centre of the scale. Once the mains voltage in your locality has been changed to 230 V, it will be a simple matter of adjusting the scale appropriately.

A 50 µA moving-coil meter is connected between two voltage sources. One, the reference, is divider R1-C1-R2-D1. The other, divider R5-R6-P1-P2, is variable and

ELEKTOR ELECTRONICS JULY/AUGUST 1990
PARTS LIST

Resistors:
R1 = 470 kΩ
R2 = 4 MΩ
R3 = 220 kΩ
R4 = 27 kΩ
R5 = 100 kΩ, 1.5 W
R6 = 8 kΩ
P1 = 10 kΩ preset
P2 = 2 kΩ preset

Capacitors:
C1 = 100 nF, 400 V
C2 = 100 μF, 10 V
C3 = 22 μF, 350 V

Semiconductors:
D1 = zener diode, 47 V, 1 W
D2 = 1N4007
D3, D4 = 1N4148

Miscellaneous:
M1 = 50 μA moving-coil meter
K1 = 3-way PCB terminal block
Enclosure, ABS, e.g., Bopla SE432DE

is used for calibration.

Resistor R1 protects zener diode D1 by limiting the charging current of C1 to a safe value when the circuit is first connected to the mains.
Both voltage sources provide half-wave rectification of the mains voltage.
The value of capacitor C2 determines the meter response to relatively fast variations of the mains voltage. Its value may lie between 20 μF and 220 μF. Since the reference voltage is not a direct voltage, a change in the value of C2 may require the circuit to be realigned.
The circuit may also have to be readjusted after a time to compensate for the drift caused by the heat generated by D1 and R5. Since R5 is located quite close to R4 on the PCB, it may in some cases be necessary to use a 27 kΩ metal film resistor in the R4 position.

Connect a variable transformer to the input of the circuit and set its output to the minimum expected mains voltage, that is, 210 V or 230 V (UK and Eire). Adjust P2 until the meter reads 0 μA. Next, increase the transformer output to the maximum expected mains voltage, that is, 230 V or 250 V (UK and Eire). Adjust P1 until the meter reads 50 μA.

The circuit may be aligned without the use of a variable mains transformer by connecting the primary of a mains transformer with an open-circuit secondary voltage of about 10 V to the mains, and the secondary winding in series with the mains. When the voltages across the secondary and primary windings are in phase, a total voltage of 230 V or 250 V (UK and Eire) is obtained. When the connections of the secondary winding are reversed, 210 V or 230 V is obtained. In both cases, either P1 or P2 should be adjusted as described when a variable mains transformer is used.

It is recommended to provide the meter with indication marks at, say, 10 V increments.

(T. Gifford)

VARIABLE DIVIDER

The divider is based on a dual hex counter Type 74HCT393 and two 4-bit comparators Type 74HCT85. The division factor is set between 2 and 256 with two hex thumb-wheel switches.
The signal to be divided is applied to the clock input of the first counter, IC1a via IC1a. The outputs of the counter are compared by IC4 with the setting of the first thumb-wheel switch. When the data at the Q inputs and P inputs are identical, the output of IC4 becomes logic high, provided
that the = input (pin 3) is high, which here is always the case.

The = output (pin 6) of IC4 is linked to the = input of IC5, so that the comparator output of this IC can be high only when the data at the inputs of IC4 are identical.

The input signal is combined via the OR gates in IC3 with the outputs of IC2a in a way that arranges for the second counter, IC2b, to receive a clock pulse when IC2a has counted to 16 (and has thus returned to zero).

Circuit IC5 compares the content of IC2b with the position of the second thumb-wheel switch. Only when both comparators have identical inputs, that is, when the position of the counters corresponds to the setting of the two thumb-wheel switches, does a '1' appear at pin 6 of IC5.

The counters are then reset via IC7b and IC1c, after which the counting and comparing can start afresh.

When pin 6 of IC5 is high, the output of the divider is a positive pulse whose width is equal to half the period of the input signal, irrespective of the set divide factor. Therefore, the output signal is not symmetrical. If symmetry is required, a D-type bistable should be added at the output. The divider output then serves as the clock for the bistable (Q output linked to the D input). Remember, however, that the bistable divides the signal by 2.

When, after the counters have been reset, the input signal becomes low, it would be expected that the second counter is immediately set to '1'. This is prevented, however, by the still active reset section.

The highest counter position is 256 and that is reached when both thumb-wheel switches are set to 0. Both counters then go their entire range.

It is possible to extend the divider by adding one or more stages consisting of a counter, comparator and thumb-wheel switch, and connecting this (them) to the preceding counter via the four OR gates. The reset signal must, of course, always emanate from the last stage.

The thumb-wheel switches may be replaced by standard DIL switches.

(H. Snits)

**BEDSIDE LIGHT WITH AUTO OFF**

This simple bedside light is suitable for telling the time at night, finding one's way to the door, and so on.

It is based on the well-known CMOS Type 7555 timer, connected as a monostable, whose time constant, T, with values as shown gives a delay of about 5 minutes before the light goes out. Different delays may be calculated from

\[ T = 0.69 \times R1 \times C1 \] seconds

where R1 is in ohms and C1 in farads.

Resistors R2 and R3 serve as pull-ups on the trigger and reset pins respectively.

Touch pads are used to trigger the monostable or to reset it before the time delay has lapsed.

The output of the timer drives T1 via R4. The transistor can switch up to 250 mA without needing a heat sink.

Depending on the application, higher or lower wattage bulbs may be used, but bear in mind the cost of the batteries!

Standby current of the prototype was 35 µA at a supply voltage of 4.5 V.

(R.G. Evans)
The adjustable flashing rate and low off-state current drain make this circuit eminently suitable for use as an on-off indicator where battery power is at a premium, or as a fake car alarm.

Transistors T1 and T2 form a relaxation oscillator whose frequency may be set between 1 Hz and 10 Hz by P1.

The LED has an on-time of about 5 ms, so that it may draw a relatively high current to produce intense flashes.

Because of the low duty factor, 0.005 at 1 Hz and 0.05 at 10 Hz, the average current drawn by the circuit is between 0.1 mA and 1 mA at a supply voltage of 12 V. This compares favourably with special high efficiency flashing LEDs that typically require an average current of 2 mA, depending on their series resistor.

The circuit may be operated from supply voltages between 6 V and 25 V.

Resistor R6 and transistor T3 should be omitted, and a wire link fitted between 'A' and 'B', if the circuit is used as an 'on' indicator.

In the fake car alarm application, T3 is switched off when the contact key, Sk, is closed. The oscillator then draws a current of not more than 2 µA and the LED does not flash. When Sk is opened, T3 draws base current via R6, starts to conduct and actuates the oscillator.

(J. Ruffell)

Sine wave oscillators normally generate near-perfect sine waves, but the stability of their output signal is often not very good. The circuit presented here is aimed at improving the stability.

The improvement is brought about by limiting the fed-back output signal by two series-connected zener diodes, D1 and D2.

The oscillator proper consists of two sections: IC1, R1, R5, C1 and C2 form a second-order low-pass filter, while IC2 is connected as an integrator.

The sinusoidal output signal of IC2 becomes trapezoidal after the limiting action and is then applied to the low-pass section. This means that the amplitude of the fed-back signal is constant, so that the peak value of the sine wave output of the oscillator is also constant.

The circuit provides two sine wave outputs at A and B respectively that are mutually 90° out of phase. The waveform at output B is slightly purer than that at A.

The third harmonic is about 40 dB down on the fundamental.

With values as shown, the circuit generates a signal at a frequency of 3.3 kHz with a peak-to-peak value of 11 V. The frequency may be altered by changing the values of C1, C2 and C3 proportionally.

The circuit draws a current of 3 mA at a supply voltage of 15 V.

(National Semiconductor application)
This circuit behaves like a real mosquito: as soon as it gets dark, it begins to buzz irritatingly. As soon as it gets light, it becomes mute once more, so that, again like a real mosquito, it is very difficult to pinpoint.

The potential at input 2 of IC1a determines whether the mosquito hums or not. The level of this voltage depends on the resistance of photovaristor RI and the setting of P1. If the level is higher than the trigger threshold, pin 3 goes low, which causes C3 to charge via R3.

It takes about 90 seconds for C3 to become fully charged and during this time pin 6 of IC1b is high. As soon as the voltage at this pin drops below the lower threshold of IC1b, the output (pin 4) of the gate goes high. This results in IC1c commencing to oscillate. A square-wave signal is then produced at the output of the oscillator, which is buffered by IC1d and then applied to the buzzer. The buzzer operates as long as T2 is switched on.

The output of the oscillator is used also as a clock for 14-bit binary counter IC2. The consequent signals at outputs Q7, Q9 and Q11 of this circuit are added together and used to drive T2. The whining tone is therefore not continuous but totally random. This random signal is used also by T1 to influence the output of IC1c to a small extent.

As soon as it gets light, the potential at pin 2 of IC1a drops, which results in C2 discharging rapidly via D1 and R4, so that the oscillator ceases at once.

The circuit draws a current of only 2-5 mA, so that a 9-V battery will last for quite some time.

(J. Beckers)

Siemens' Type PID20 is a passive infra-red detector that transforms heat radiation into electrical pulses. Adding two opamps and some miscellaneous components to the device is sufficient for the construction of an efficient infra-red (IR) detector.

The magnitude of the output signal of the PID20 is determined by the load at its pins 3 and 4 as illustrated in Fig. 2 that shows the output at loads from 1 kΩ to ∞.

From the curves, it is clear that a reasonably high load is needed for a good output signal. In the circuit, both output pins are loaded by 100 kΩ (R1 at pin 3 and R3 shunted by R4 at pin 4).

The output signal at pin 3 is compared with a reference voltage equal to half the supply voltage. This reference voltage is derived from potential divider R2-R3-R4-R5. The output increases at the approach of an object that is warmer than the surroundings or at the removal of an object colder than the surroundings. Note the ringing in Fig. 2 after a change in surroundings has taken place. The sudden appearance of a warm object causes the output voltage first to rise and, after a few seconds, to drop, sometimes even below the reference voltage level. This behaviour needs to be taken careful account of in practice.

The change in voltage at the output of
the sensor is compared by IC2a and IC2b with voltages respectively 0.5 V below and 0.5 V above the reference voltage. Depending on the output level, one of the comparators toggles and switches on T1. This transistor may be used, for instance, to drive a relay (up to 100 mA). It is also possible to connect a transistor (and diode) to each opamp, so that it is immediately seen whether a source has caused an increase or decrease in heat.

As shown, the circuit draws a current of about 1 mA (of which roughly 0.2 mA is on account of the sensor). The current drain may be reduced by replacing the TLC272 by a TLC27L2.

The sensor should be fitted by means of a special connector (ask your dealer).

(Siemens application)

There are many ways of designing an analogue-to-digital converter: the present design is aimed particularly at a high speed. The speed is determined primarily by the reaction time of comparators (here about 1 μs). The design has a drawback: the number of components is directly proportional to the number of levels that the converter can cope with. In the present design, this is limited to ten (if none of the comparators detects the exceeding of a level, the input is below the lowest level, and this is recognized as zero [10th level]).

The open-collector outputs of the comparators are connected to a priority detector, IC4, which is a kind of decimal-to-BCD decoder. If, however, several inputs are low at the same time, priority is given to that BCD code that corresponds to the input with the highest number. Four NOT gates invert the BCD code generated by IC4.

Because the comparators have open-collector outputs, they can work from a higher supply voltage than IC4 and IC5. Pull-up resistors R11-R19 must, therefore, be connected to the +5 V rail and NOT to the comparator supply rail.

The voltage levels with which IC1-IC3 compare the input level are provided by potential divider R2-R10.

The input range of the converter may be modified by changing the value of R1. It should be noted, however, that the upper threshold, that is, the level at pin 5 of IC1, must always be at least 2 V lower than the supply voltage. With values as shown, and a supply of 12 V, the upper threshold is 5.68 V and each step is 632 mV.

(P. Coster)
Although the multimeter whose circuit is shown here is fairly simple to make, you may experience difficulties in obtaining some of the components, since most retailers do not stock 99 MΩ or 100 MΩ resistors. It then becomes matter of making up these resistors or omit the associated (current) ranges.

The use of these high-resistance components is made possible by the use of a Type TL061 opamp, which has very high-resistance inputs.

For the voltage ranges, the opamp is connected as a non-inverting amplifier with an amplification of \( 1 + \frac{R12}{R13} = 10 \). If the meter has a full-scale deflection (f.s.d.) of 6 V, the input sensitivity is 600 mV. Input attenuator R1–R4 provides additional ranges up to 600 V.

When the meter is used to measure current, the unknown current will flow through one of the resistors R7–R10, depending on the selected range. Since the value of these resistors is much higher than that of R13, the output voltage is equal to 10x the potential drop caused by the unknown current across R7–R10. With the values as shown, this gives current ranges of 6 nA, 60 nA, 600 nA and 6 μA. These ranges are fairly small, but that is because the opamp can not deliver more current and also, in order to keep the circuit simple, R7–R10 must be much larger than R13. Note that these resistors have nothing to do with the input resistance of the instrument. The + terminal is at earth potential and the – terminal is a virtual-earth point: that results in a very low-resistance input (as should be the case).

Apart from the unknown current, the input offset current of the opamp also flows through R7–R10. To eliminate the effect of this current (some nA) in the most sensitive range (R7), resistor R5 has been added.

Since a centre-zero meter and symmetrical supply were used, the polarity of the input signal does not matter. The meter is set to zero with P1.

Resistor R6 and diodes D1 and D2 serve to protect the opamp against too high input voltages.

The supply is formed by two 9-V batteries that, since the current drain is less than 1 mA, will last a long time.

(Texas Instruments application)

In much equipment, such as computers, amplifiers, and the like, where high tolerance components are used, it is important to know at all times whether the supply voltage is within specification. The circuit presented here was designed to show just that. It does so by comparing two d.c. voltages. As soon as there is a difference of more than 10 mV between these, the indicator LED goes out to signal that something is amiss. It is, therefore, necessary to use a very accurate reference voltage, corresponding to the required voltage level.

Circuits IC1a and IC1b form a difference amplifier: the voltage to be monitored is applied to input 1 and the reference voltage to input 2. Differences between these two potentials as small as 10 mV are sufficient to cause one of the two comparators, IC1c or IC1d, to toggle.

The output of IC1d goes high when the voltage level at input 1 is 10 mV or more greater than that at input 2. Comparator IC1c toggles when the voltage level at input 1 is 10 mV or more lower than that at input 2. The indicator LED is switched off via diode D1 or D2, as the case may be.
The tiny a.f. power amplifier presented here provides an output of up to 250 mW and may be used in a number of applications, for instance, in a stereo version as a booster for personal radios.

The design is straightforward: a BC3417 transistor drives a balanced power amplifier consisting of a BC337 and a BC327.

The quiescent current is arranged by diodes D1 and D2. Owing to the simplicity of the circuit, the quiescent current varies with temperature. This drawback is particularly noticeable when the output transistors get much hotter than the diodes. In that case, either the output power has to be reduced or the output transistors must be fitted on heat sinks. Another solution is inserting 0.47 kΩ resistors in the emitter circuits of the output transistors.

The amplification is determined by the values of R1 and R3 and, of course, P1. With values as shown, and dependent on the setting of P1, the amplification is about 15. This may be varied by changing the value of R1. It is not recommended to alter the values of R2 and R3, since these resistors determine the d.c. operating point of the amplifier.

The input sensitivity for a power output of 250 mW into 8 Ω and an amplification of 15 is about 95 mV. The circuit then draws a current of around 180 mA.

The 50/75 Ω driver is based on a Type OP64, which was designed specifically for use in pulse and video applications. Because of internal compensation, the IC is stable at amplification factors of 5 or greater. Since the output can deliver a current of up to 80 mA, the IC can be loaded by 150 Ω (i.e., a 75 Ω system) with a supply voltage of ±15 V without limiting taking place.

The circuit in Fig. 1 is intended as a simple output amplifier in a 75 Ω system or, if the value of R1 is changed, as an impedance adaptor. With the values of R2 and R3 as shown, the amplification is 5. The bandwidth depends to a large extent on the board layout. According to the manufacturers' data, the roll-off frequency, at this level of amplification, may be as high as 20–30 MHz.

The circuit in Fig. 2 is intended to couple a 75 Ω system to a 50 Ω system. The gain in this set-up is 0.5 dB. Potential divider R4–R5 ensures an output impedance of 50 Ω and attenuates the output of IC1 (=5× input signal) to make the overall amplification unity. At the same time, R4 ensures that the load on the OP64 does not become too large in a 50 Ω system.

The OP64 is very fast: with an amplifi-
Fig. 1. Simple output amplifier or impedance adaptor.

Fig. 2. Converter for coupling a 75Ω system to a 50Ω system.

cation of 5, the slew rate at a leading edge is about 135 V/µs and at a trailing edge, around 120 V/µs. There is then some overshoot, but that disappears when the amplification is increased to 10.

At higher frequencies, the output impedance of the OP64 increases to some extent: to about 20 Ω at an amplification of x10. The impedance may be increased to about 2 kΩ via the disable pin 6 (which is active low). This is a handy way of limiting the current drawn by the IC. The normal current drain is 6–6.5 mA, but in the disable condition this reduces to about 0.5 mA for the positive half and around 0.12 mA for the negative half.

According to the data sheet, the IC is protected against short circuits for about 10 seconds, so care must be taken that no lasting short circuits or overloads occur.

In a board layout of the circuit, the current through the load must not return via the input earth, but via a common earth, to which the decoupling capacitors should also be connected.

(T. Gifford)

037 FOUR-MONITOR DRIVER FOR PCs

The driver described here allows up to four monitors for IBM PCs or compatibles to be driven by a single display adaptor card. This card and associated monitors must be types that operate with TTL (digital) signal levels, such as:
- CGA—colour graphics adaptor;
- Hercules card—monochrome adaptor;
- EGA—enhanced graphics adaptor.

Applications of the circuit may be found in teaching computer science, PC slide shows, presentations and demonstrations.

The circuit consists of four octal three-state buffers Type 74HC541: IC1–IC4. The TTL levels from the computer are applied to the inputs of (permanently enabled) buffers via connector K5, which is a 9-way sub-D male type.

The signals carried by this connector depend on the display adaptor card used in the PC as shown in the accompanying parts list.

**PARTS LIST**

<table>
<thead>
<tr>
<th>Resistors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1 = 8-way SIL resistor array, 4kΩ</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Capacitors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>C1 = 47 µF, 16 V</td>
</tr>
<tr>
<td>C2–C5 = 100 n</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Semiconductors:</th>
</tr>
</thead>
<tbody>
<tr>
<td>D1 = 1N4001</td>
</tr>
<tr>
<td>IC1–IC4 = 74HC541</td>
</tr>
<tr>
<td>IC5 = 7805</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Miscellaneous:</th>
</tr>
</thead>
<tbody>
<tr>
<td>K1–K4 = 9-way female sub-D connector for PCB</td>
</tr>
<tr>
<td>K5 = 9-way male sub-D connector for PCB</td>
</tr>
<tr>
<td>Enclosure: Hoodie Type 222</td>
</tr>
</tbody>
</table>

ELEKTOR ELECTRONICS JULY/AUGUST 1990
The circuit is powered by an on-board 5 V supply based on the familiar Type 7805 voltage regulator, IC5. The unregulated input to the supply may lie between 9 V and 15 V. The total current drain is not greater than 10 mA.

(F. Tronchet)
Many people find that when they want to add a CD player to their audio system, there is no available input left on their power amplifier. It seems obvious to use the phono input, but that presents a couple of problems. This input is far too sensitive for a CD player and, moreover, it passes the signal via an RIAA frequency correction network to the amplifier. The present network can solve both problems without any internal changes to the audio system, since it is merely plugged into the phono input socket. 

The circuit is not much more than a filter and an attenuator that reduces the output of the CD player (here assumed to be 200 mV) to 2 mV and has a frequency response that is exactly the opposite of that of the RIAA network. The resulting overall frequency characteristic is straight within ±1.5 dB.

It is recommended to fit the circuit (left- and right-hand channels separately) in a suitable metal enclosure, since it is susceptible to hum and noise. This is caused by the high-impedance input of about 1 MΩ at 50 Hz. The impedance decreases to 100 kΩ at 1 kHz, 10 kΩ at 10 kHz and about 1 kΩ for frequencies between 100 Hz and 500 Hz. Because of the variable input impedance, it is important that the internal impedance of the signal source is not much higher than about 2 kΩ in order to keep the -1 dB point above 30 kHz (provided the main amplifier is of reasonable to good quality).

Figure 2 shows the calculated frequency response of the circuit, which is identical to that of the recording amplifier. Figure 3 shows the overall frequency response of an (idealized) audio system with the present network connected to the phono input.

For good results, all capacitors should be polystyrene types, and metal film resistors should be used in the R1 and R2 positions.

(T. Gifford)

With the availability of portable CD players, a need has arisen for a means of connecting such a machine to the radio-cassette player now found in most private cars. Unfortunately, few of these players have a suitable line input. However, a CD-to-cassette adaptor will solve the problem. Although such adaptors are freely available, it is fairly simple (and great fun) to make one yourself.

To build the adaptor, you need an old cassette case (screwed together, not glued), a stereo cassette recording head, a stereo 3.5 mm phono plug, some two-core screened cable, two 820 Ω resistors and two 15 nF capacitors (and some dexterity).

Remove all tape guides and fastening strips from the recording head. Cut two 20 mm long, 7 mm wide strips of thin tin plate and bend them at right angles about 90°.

(T. Gifford)
5 mm from each end. Drill a 3 mm hole near one end of the long part of the bracket so formed. Solder the brackets to the sides of the recording head as shown in Fig. 1; gently fasten the head in a small vice to prevent it getting too hot.

Unscrew the cassette case and remove everything from the inside. With a small saw or sharp knife, remove the rib behind where the screening shield and pressure pad were.

Drill 2 mm holes in the remaining ribs at the same distance as the holes drilled in the tin strips. Fasten M2x7 screws in these holes and slide appropriate insulating sleeve over the protruding screw-thread. Next, slide the recording head assembly (slot at the underside) over the insulated screw-threads, fit light springs (such as those from an old ballpen) over the screw-thread and hold these in place with a nut (see Fig. 1). The springs will then push the recording head forward.

The electronics are very simple as may be seen in Fig. 2. One resistor and one capacitor form a low-pass filter for one channel that interfaces between the CD player output and the frequency-dependent recording head. These components may be soldered direct to the head. Finally, the short cable is connected to the components, taken out of the cassette case via a small hole in one of the short sides of the case and then connected to the stereo phono plug.

It pays to experiment with the setting of the volume control on the CD player to find which position gives the best reproduction.

(J. Ruffell)

**10 MHz Reference from France-Inter**

The circuit published last year can easily be modified for reception of the French station. The shaded areas in the block diagram indicate what changes are needed. The two figures showing sections of the original PCB indicate which connections at IC5 and IC6 must be cut and what new connections must be made. Furthermore, the table shows the changes in value of a number of original components.

(A.N. Other)

**COMPONENT CHANGES**

- **Capacitors:**
  - C6, C9 = 2n7
  - C7, C10 = 15 n

- **Inductors:**
  - L2, L3 = 10 mH
An electric guitar with three single-coil elements, the well-known Fender model, can furnish more facilities than the manufacturers provided. A simple modification makes it possible to produce a real stereo signal or to connect one or more elements in anti-phase—see the circuit diagram.

If the guitar is fitted with a 5-position switch, this must be replaced by four in-line miniature toggle switches. Switch S4 determines the operating mode: upper position = stereo; lower position = normal or with elements in anti-phase.

Switches S1, S2 and S3 determine how the throat element, the central element and the bridge element are switched, although this is also dependent on the setting of S4.

With S4 set to 'normal/anti-phase', the other switches function as follows. In the upper position, the relevant element is switched into circuit, in the centre position, the relevant element is inactive, and in the lower position, the element is in anti-phase. The phase shift is varied with P2.

With S4 in position 'stereo', each element is connected to the left-hand channel if the relevant switch is in the upper position, and to the right-hand channel if that switch is in the lower position. If the associated switch is in the centre position, the relevant element is inactive. In this condition, P2 acts as the volume control for the right-hand channel.

Potentiometer P1 is the volume control already fitted to the guitar (500 kΩ or 1 MΩ log). Potentiometer P2 should have the same value as P1 and be fitted in the hole previously occupied by the tone control. It is, of course, possible to retain the original tone control, but an identical one must then be fitted for the second channel (in exactly the same way as the original tone control).

The original jack connectors must be replaced by stereo versions. When the guitar is switched to stereo, it is connected via a good-quality 2-core screened microphone cable, terminated at each end in a stereo jack. The left-hand channel is then available at the tip of the plug, and the right-hand channel at the ring. The channels may be separated by a stereo splitter and can then be transmitted via mono jacks and cables. When the guitar is switched to normal/anti-phase, a standard guitar cable may, of course, be used.

The modifications described here were carried out on a Yamaha SC1000 and gave excellent results. They are particularly recommended for good quality studio operation.

(A. Ferndouni)
This simple design, consisting of a TL071C driver and two MOSFET power amplifiers, can deliver up to 45 W into 8 Ω.

The design is basically a Siliconix application. The principle of this application is that the output transistors are driven by the voltage changes across two resistors inserted in the supply lines of the opamp driver. In other words, the currents drawn by the opamp determine the drive to the power amplifiers. Use is made of current feedback that is effected by connecting the output of the opamp to a potential divider across the amplifier output.

In the diagram, FETs T5 and T6 are driven by the potential differences across R8 and R13. Since the supply voltage is considerably higher than normal for a standard opamp, transistors T1 and T2 have been inserted in the supply lines to the IC. These transistors are provided with a fixed base potential of ±15 V by zener diodes D1 and D2. The supply voltage to the opamp, whatever its drive, is therefore always 14.4 V.

Setting of the quiescent current is effected with the aid of T3 and T4. It is essential that T4 is coupled thermally to T5 so as to ensure stability of the quiescent current.

Transistor T3 draws its current via R8, and this ensures a correct operating point for T5. When the temperature rises, the base-emitter voltage of T4 drops; the current through T3 decreases; and the gate-source voltage of T5 drops. The opamp ensures equilibrium of the circuit, so that the current through T6 is also adapted as appropriate. The total current drawn by the amplifier is adjusted to about 75 mA with P1. The current drawn by the FETs is then around 70 mA.

The presence of T1 and T2 makes the amplifier slightly slower than it would be without these transistors. However, the fairly large capacitors associated with the MOSFETs can only discharge via resistors R8 and R13. This results in an increase of the quiescent current at frequencies above 40 kHz. Because of this, the bandwidth is limited to 20 kHz by C3. A resistor is connected in series with this capacitor to further improve the stability.

The MOSFETs must be fitted on a heat sink of not less than 1 K/W. In contrast to the usual emitter or source follower, the configuration used here may be driven virtually up to the supply voltage, so that an efficiency of up to 70% may be attained. In the prototypes, cross-over distortion was not greater than 0.2% at 20 Hz (10 W into 8 Ω). With a stable supply of ±30 V, the amplifier can provide 45 W into 8 Ω or 70 W into 4 Ω.

Note that the amplifier is not protected against short circuits. Check therefore at all times what load is connected to it before the supply is switched on.

(T. Giffani)

The triggerable sawtooth generator presented here is intended to convert an old-fashioned oscilloscope with only a synchronized time base to a modern triggered instrument.

The circuit is straightforward: transistor T1 is a current source that charges one of capacitors C1–C4, depending on the position of switch S1. The linearly increasing voltage across the capacitor is compared with two reference voltages by IC1a and IC1b. The reference voltages are derived from potential dividers R3-R4 and R5-R6 respectively.

As soon as the voltage across the capacitor reaches 5 V—the toggle level of IC1a—bistable IC2a-IC2b is reset. Transistor T2 is then switched on, which causes the capaci-
tor to discharge. Once the potential across the capacitor has dropped to the toggle level of IC1b, the output of IC1b goes high and the bistable is set via the trigger input. Transistor T2 is then switched off and the voltage across the capacitor increases.

The pulse that switches on T2 may also be used as the blanking pulse for the oscilloscope and it is, therefore, available at a dedicated blanking output. For the duration of this pulse, the electron beam in the oscilloscope is suppressed.

Circuit IC2 is an AND gate that ensures that the trigger pulse can set the bistable only if the capacitor is really discharged. This prevents spurious triggering of the sawtooth generator.

The generator output must be terminated in a high-impedance load to prevent distortion of the sawtooth.

The generator draws a current of only 5 mA from a single 10 V supply.

(P. Coster)

---

044

SLAVE MAINS ON-OFF CONTROL

This circuit enables the automatic on/off switching of mains-powered equipment after a master unit has been switched on or off. One particularly useful application is in audio racks where a number of signal sources, such as the tuner, CD player, and tape recorder are to be switched on and off in unison with the power amplifier.

The circuit works on the principle of current sensing in the mains input lines to the master apparatus, which should be connected to K2. When the master unit is switched on, a voltage drop appears across diodes D9 and D10. This potential triggers thyristor Th1. When it conducts, the thyristor connects junction D6-D7 to the neutral line. The resultant output voltage of the bridge rectifier—about 25 V—is smoothed by C1 and then applied to the coil of relay Rel.

When the relay is energized, indicated by D2 lighting, its contact links the live line of the mains to K3 to which the slave units are connected.

When the master unit is switched off, the thyristor stops conducting so that the relay is de-energized and the mains live line is removed from K3.

The reactance of capacitor C2 limits the
Nickel-cadmium batteries should normally be charged at 1/10 their capacity in Ah. The capacity of a 9 V NiCd battery is usually 110 mAh and this type of battery should therefore be charged in about 10 h. However, the efficiency of the charger is only around 70%, so that the real charging period should be of the order of 14 hours.

In the present charger, the charging time is measured by IC1, whose oscillator is set to a frequency of 1/6 Hz, ensuring that output Q13 goes high after 14 h.

When Q13 goes high, pin 1 of IC2 goes low, which causes bistable IC2a-IC2d to be reset. At the same time, pin 3 of IC2 goes high, resulting in the base voltage of current source T1 becoming equal to the supply voltage. This causes the transistor to be switched off, which interrupts the flow of charging current.

The charging cycle is started with S1 or
the detection circuit based on IC3. This circuit ensures that batteries left in the charger are topped up after the batteries have lost some charge through self-discharge. This ensures that these batteries are always fully charged.

The voltage at which the battery is assumed flat is set with P2 to, say, 8.4 V. To this must be added the drop across D1. This diode prevents the battery from discharging through the charger in case the supply fails.

The detection circuit also ensures that as soon as a flat or partly discharged battery is connected to the charger, the charging cycle is started immediately.

Resistor R9 ensures that even when the current source is off, pin 3 of IC3 retains a certain positive voltage. In that condition, the resistor also maintains a small charging current (5-6 µA) through the battery. The drop across D1 then decreases to around 100 mV and this lowers the voltage at which the battery is assumed to be flat.

If two batteries are alternately used and charged, the value of R9 may be reduced to 1 kΩ. The self-discharge of the battery is then countered by a trickle charge at about 5 to 6 mA.

The charging current is adjusted to 11 mA with P1 as indicated by a milliammeter in series with the battery. When the charging cycle is completed, buzzer B2 sounds. Square-wave generator IC2c may be set to a suitable buzzer tone with P3. The buzzer may also be muted by strapping pin 8 of IC2 to earth via a switch. The buzzer may, of course, also be replaced by a LED. The current through the buzzer or LED is limited by R14.

Diode D2 only lights when a charging current flows: when no battery is connected or if there is a bad contact, the low current through R7 will cause a drop of only about 1 V across the diode and this is not enough to switch it on.

When a connected battery is defective, D2 and D3 light and the buzzer is mute. With a good battery, there will come a time in the cycle when D3 goes out, while D2 remains on.

(T. Cifarei)

046  MESSAGE PAD

This little circuit provides both a visual and an audible indication that a message has been left.

CMOS inverters IC1a and IC1b form a touch-sensitive on/off bistable in which R1 provides positive feedback so that the circuit will latch in whatever state it is left in.

The output of the bistable controls astable IC1b-IC1c, which is designed to oscillate at a rate of about 11 Hz. This rate is determined by R2 and C1.

Although the astable could just about drive the LED and buzzer, a p-n-p driver transistor is used to make the circuit more versatile. When the 'on' pad is touched, the LED flashes and the buzzer bleeps intermittently.

Power is derived from a 9-V PP3 battery. Current drain in the 'off' condition is negligible.

(R.G. Evans)
BOARDMAKER 2 from TSIEN

BOARDMAKER 2 can help you turn a Netlist into a PCB that's right first time, quickly and easily.

Feature Summary

- Ultra fast Redraw
- Easy to use
- WYSIWYG display
- 10 Circuit Layers
- Net highlighting
- Design Rule Checking
- Surface Mount
- Mouse/menu or keys
- BGA,EGA,VGA & Matrix Printer O/P
- Laser Printer O/P
- NC Drill Output
- Photoplot Output
- 360 page manual
- Schematic Drawing
- Library Editor
- Auto via placement
- Fully Integrated
- Hotline Support
- And much more...

TSIEN (UK) Ltd's vigorous development policy has produced BOARDMAKER 2, with all the excellent facilities and features of BOARDMAKER 1 supplemented by full Netlist capability. This means that netlists generated by schematic capture (OrCAD,Sema II etc), by hand or generated within BOARDMAKER's ratsnest editor can be used to assist and check routing. The major benefits of this are much quicker routing and getting it right first time.

BOARDMAKER 2 maintains TSIEN's philosophy of making powerful facilities immediately available to the designer by keeping them logical, visual and easy to use. This unique collection of tools for just £295 outstrips those on many packages that have commanded a higher price.

Full upward compatibility from BOARDMAKER 1 to BOARDMAKER 2 allows current users to make a painless move to a more powerful system. Non supported BOARDMAKER 1 users will be able to upgrade.

BOARDMAKER 1 is £195 plus carr. & VAT

BOARDMAKER 2 is £295 plus carr. & VAT

Tsien (UK) Limited
Cambridge Research Labs.
181 Huntingdon Road
Cambridge
CB3 0DJ
Tel. 0223 277777
Fax 0223 277747

Send for your FREE demonstration disk now and discover how easy PCB CAD on an IBM PC can be.
SPECIAL ANNOUNCEMENT: Our Retail Shop now open at 116 New City Rd, Plaistow, London E13 8JX.