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A Juicy Bit Of Software

Pineapple Software has produced an update for its popular PCB drafting program for BBC micro (reviewed in ETI March 1987). The update gives the program full auto track routing with up to 190 connections specified by the user then linked by the program. If the result is not quite what you'd hoped for components can be shifted about without redefining all your connection points, and to help the program in the right direction you can indicate a preferred track direction or even draw a few tracks yourself before the program does the rest.

The update is available to PCB owners for £55 from Pineapple Software, 39 Brownlea Gardens, Seven Kings, Ilford, Essex IG3 9NL. Tel: 01-599 1476.

DAT's The Way It Should Be

At last DAT players are 'officially' on sale in Britain about a year after they first appeared in Japan. The Sony DTC1000ES is retailing at £1130+VAT and includes the controversial built-in copyguard that stops it recording at the 4.1KHz sampling rate of compact disc. Since you can't record CD and commercial recordings aren't exactly widely available yet, the only use for this machine will be in the audio industry—and even then only among the less affluent. Richer studios will be more interested in the PCM2500 which can switch sampling rates (including 44.1KHz) and can match various formats (AES/EBU, SDIF-2 and Sony/Philips). It cannot however convert signals from one format to another. Ironically, considering Sony's own stand on copy prohibiting, the PCM2500 can add a copy prohibit code while recording. HBB is supplying both machines to the UK and should soon have stocks of the even more expensive PCM2000. For full details contact: HBB, 73-75 Scrubs Lane, London NW10 6QU. Tel: 01-960 1160.

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Young Electronic Designer Awards

The 1988 Young Electronic Designer Awards are up for grabs if you can come up with a commercially viable device with an everyday application. Last year's winners included a shrew for the blind, a digital pressure gauge, a speech synthesizer and an animal stress meter (?). There are three age groups (up to 15, 16-18 and 19-25) and applicants must be in attendance at a school, college or university in Great Britain. Apart from the cash on offer, Texas Instruments are offering course sponsorship and a job to the winner of the 19-25 group. (Even more exciting is the possibility of getting on Tomorrow's World.) Entry forms from: The YEDA Trust, 24 London Road, Horsham, West Sussex RH12 1AY. Tel: (0403) 211048.

Stereo TV broadcasts have not been shelved indefinitely as recent press reports have been claiming. Although the BBC Nicam broadcasts will not be off the ground for at least three or four years, the IBA will be introducing a service for London in 1989.

Other IBA transmitters will be updated to broadcast stereo as they become due for routine maintenance, which would put Yorkshire next on the list.

The majority of programmes going out in stereo will be pre-recorded films and music shows because of the lack of facilities for producing stereo in the TV studios.

The IBA are also considering multi-channel broadcasting, using the two channels as completely separate soundtracks (double channel sound).

Meanwhile BBC tests continue and some live programmes are already going out in London in Nicam stereo (Wogan for example) since none of the post-recording problems of stereo sound are encountered.

As for the general belief that the BBC originally scheduled Nicam for introduction in early 1988, the BBC says that this is complete rumour probably started by manufacturers such as Ferguson when they introduced stereo TVs and VCRs.

Meanwhile you may have noticed some peculiar transmissions going out on BBC2 in the early morning closedown period. These flickering distorted images of volcanos and skydivers are part of a BBC scrambling experiment with two applications in mind.

The first is the possibility of leasing unused air time to business and medical users to transmit high quality video pictures nationwide on a similar basis to BBC Enterprises' Datacast service on Ceefax.

The second more disturbing application stems from the Peacock Report's recommendations for Pay-As-You-View television where a scrambling system would be put on BBC broadcasts. The licence-fee would be abolished and consumers pay only for the programmes they watch (it is still unclear if ITV programmes could be watched free under such a system).

Although the BBC opposes this concept as removing any incentive to produce minority programmes, the Government may pass legislation that would give them no choice.
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In this third part of his series on satellite television, Keith Brindley considers the mathematical and theoretical side of signal reception and looks at ways of calculating antenna size.

Satellite TV systems (or TVRO for the pedantic) are of course examples of communications systems and as such can be looked at mathematically in terms of signal strengths to enable us to define certain minimum standards which are governed by the user's operating requirements.

However these minimum standards are related to television picture quality and as picture quality is largely subjective, some way of relating subjective picture quality to objective signal strengths needs to be devised.

What's more the relation must be made in such a way as to make easy comparisons between similar systems.

The process of calculating a satellite system's quality in this way is known as a link budget analysis — so called because it's an analogy with a financial budget, comparing gains and losses in the link between satellite and receiver to give a final result in terms of picture quality.

No Accounting For Taste

The main problem in any communications system (but particularly in TVRO systems) is the level of noise which the receiver output contains. A communications system's noise level is typically quoted in relation to received signal level as a signal-to-noise ratio (S/N), where the S/N ratio is given in decibels (dB) by:

\[ S/N = 10 \log \left( \frac{\text{signal power}}{\text{noise power}} \right) \]

Most readers will know this and no further time needs to be spent here. However, the S/N ratio depends on a number of other factors within the overall system, particularly regarding satellite television reception. So we need to consider further these related factors.

To do this we must consider some basic transmission calculations. A transmitting satellite rotating in the geo-stationary orbit (GSO) is sufficiently far away from the Earth to be considered an isotropic radiating source since it appears from an Earth viewpoint to radiate in all directions (Fig. 1a).

In reality of course this is not the case as all satellites use directional transponders (Fig. 1b) — they are only effectively isotropic (remember this term) as far as the earth-bound observer is concerned.

We can calculate an antenna's received signal from an isotropic source if we first calculate the flux density of the source at say a distance D, because all the radiated signal must pass through an imaginary sphere of radius D, around the source. The resultant flux density is given by:

\[ F = \frac{P_t}{4 \pi D^2} \text{ Wm}^{-2} \]

where \( P_t \) is the transponder power.

If a receiving antenna is presumed to be positioned on the circumference of the imaginary sphere then two main points are illustrated by this relationship:

- flux density decreases with distances from the source (fairly obvious) as an inverse square
- flux density (in watts) received at the antenna can be simply calculated if the area of the antenna (in m²) is known.

With a satellite's directional transponder a gain \( G(\theta) \) occurs in a direction \( \theta \), where gain is defined as the ratio of power per unit solid angle radiated in a given direction to the average power radiated per unit solid angle. That is:

\[ G(\theta) = \frac{P(\theta)}{P_s/4\pi} \]

where \( P(\theta) \) is the power radiated per unit solid angle,
P is the total power radiated and G is the gain of the transponder at an angle \(\theta\). The angle corresponds to the direction in which the transponder is pointing.

So for a transponder of power \(P\), watts and gain \(G\), the flux density is:

\[
F = \frac{P \cdot G}{4 \pi D^2} \text{ W/m}^2
\]

The product \(P \cdot G\) is usually called the effective isotropically radiated power or the equivalent isotropically radiated power (both abbreviated to EIRP). The terms merely signify the fact that the unidirectional transponder is looked upon by the observer as equivalent to a real isotropic source radiating in all directions.

As we’ve already seen, the power received by an antenna can be simply calculated knowing the flux density and the area of the receiving antenna. For an ideal antenna, the received power is given by:

\[
P_r = F \cdot A
\]

where \(P_r\) is received power, \(F\) is flux density and \(A\) is antenna area.

In practice however, no antenna is perfect and an efficiency, known as the aperture efficiency \(\eta\), of between 60-75% is normal. This aperture efficiency is said to reduce the area of the antenna’s receiving aperture to give an effective antenna area \(A_e\), where:

\[
A_e = \eta A
\]

For a practical antenna, the received power is therefore:

\[
P_r = F \cdot A_e
\]

So in terms of EIRP, received power of a practical antenna is given by:

\[
P_r = \frac{P \cdot G \cdot A_e}{4 \pi D^2}
\]

In other words, received power depends on the satellite’s EIRP, the antenna’s effective area, and the distance between the satellite and the antenna.

We can consider this relationship another way, where the term:

\[
L_s = \frac{1}{4 \pi D^2}
\]

is known as the spreading loss, or the path loss (in fact it is not a loss in the strict sense at all — it merely accounts for the way energy spreads out as the transmitted wave travels away from the source). The received power can now be calculated as:

\[
P_r = \text{EIRP} \times \text{effective antenna area} - \text{spreading loss}
\]

and, because all these terms usually are quoted in decibels (that is, logarithmically) then:

\[
P_r = \text{EIRP} + A - L \text{ dB}
\]

Again, this represents the ideal situation and in practice other losses such as attenuation due to weather (particularly rainfall) have to be taken into account.

**Practical Example**

To put things into perspective at this point, we can consider a real antenna site (such as in my back garden) and perform a few calculations to show the idea.

Figure 2 shows the footprint in terms of EIRP of the Eutelsat F1 satellite, positioned at 13° east in the GSO. Each ring of the footprint represents an EIRP ‘contour’ within which area the EIRP is numerically indicated in units of dBW. Over most of Britain, the EIRP is 46dBW.

The units dBW simply means that the power level is referred to 1W. In other words, 1W is signified by 0dBW. A positive dBW level means the power is greater than 1W, a negative dBW level means the power is less than 1W. Another common reference power is in units of dBm where the level quoted with those units is referred to 1mW — so that 1mW is signified by 0dBm. Both units are common in communications engineering.

Now in my back garden (in the wilds of the East Midlands) the distance from Eutelsat F1 and the antenna happens to be about 38747 km. So the spreading loss in my back garden is:

\[
L_s = \frac{1}{4 \pi D^2} = \frac{1}{4 \pi (38747 \times 10^3)^2}
\]

\[
= 5.3 \times 10^{-17}
\]

Expressed in decibels, this is: 10 log (5.3 \times 10^{-17}) or -162.76dB.

Using a 1.5 metre diameter antenna, the actual area is \(\pi r^2\) or 1.767m². With an aperture efficiency of, say, 60% (about right for most antennae), the effective antenna area is 1.06m², expressed logarithmically this is 0.25dBm².

It’s difficult to define exactly the loss which may occur due to weather conditions but generally 2dB is about right.

Calculating the received power is now just a simple matter:

\[
P_r = 46 + 0.25 - 162.76 - 2 = -118.51\text{dBW}
\]

which means that only 1.4 picowatts of signal power (work it out for yourselves) is received by the antenna — not a lot!

The received power we’ve worked out here is normally referred to as the carrier power and given the symbol \(C\). In frequency modulation (such as existing satellite television services) and phase modulation systems the two are equal because the carrier level is constant, whether or not a signal is modulated onto the carrier.

Most signal-to-noise calculations for satellite systems use estimates of carrier power rather than signal power and subsequent signal-to-noise ratios normally are quoted as a carrier-to-noise ratio (C/N).
Noise
Having calculated the received carrier power, we now have to consider the noise present so that we can compare the two.

In satellite television reception systems follows a classical receiver convention (shown in Fig. 3a) where the receiving antenna is connected to a demodulator through a high-gain amplifier. In a satellite reception system the high-gain amplifier is only one part of the low noise blockdownconverter (LNB). All important parts of the LNB are shown in Fig. 3b.

The noise we want to measure is that noise present at the demodulator input and this is made up of noise which is contributed by various parts of the system including the antenna itself, the waveguide and the high-gain low-noise amplifier (LNA). Although parts following the LNA also contribute noise, their effects are negligible in comparison and so can be ignored.

At the microwave frequencies involved in satellite communications systems, all objects with a temperature above OK generate noise and so contribute to the total system noise. This thermal noise is usually calculated in terms of noise temperature by relating temperature to the thermal energy produced by the random movement of electrons, with Boltzmann's constant. Noise is generated by an object at each frequency, so the bandwidth of the receiver also bears directly on the amount of noise present at the demodulator input.

The noise power present at the demodulator input is given by:

\[ P_n = kT_B \]

where \( k \) is Boltzmann's constant \((1.38 \times 10^{-23})\), \( T \) is the total system noise temperature in Kelvin and \( B \) is the receiver bandwidth in Hz.

Total system noise temperature is a combination of the noise temperatures of antenna and LNB. LNB noise temperature may be given by the manufacturer or can be calculated from the LNB’s noise figure (NF) which is usually quoted in decibels. LNB noise temperature is:

\[ T = \text{antilog} \left( \frac{\text{NF}}{10} \right) - 1 \]

where \( T \) is the temperature in degrees Kelvin (usually taken as 17°C → 290K).

Antenna noise temperature is a bit more difficult to estimate as it varies with antenna, elevation angle and received frequency but manufacturers should supply figures and results are typically in the range 40K to 90K.

In practical terms (my back garden again!) for an antenna with a noise temperature of 60K, an LNB noise figure of 2.5dB and a receiver bandwidth of 30MHz, we can now calculate the noise power present at the demodulator input. The LNB noise temperature is:

\[ 290 \left( \text{antilog} \left( \frac{2.5}{10} \right) \right) -1 = 225.7K \]

so the total system noise temperature is 225.7K+60K= 285.7K.

The noise power at the demodulator input can now be calculated as:

\[ P_n = 1.38 \times 10^{-23} \times 285.7 \times 30 \times 10^6 \]

or, logarithmically in decibels:

\[ = -228.6 + 24.6 + 74.8 \]

\[ = -129.2 \text{dBW} \]

Knowing this and using our previous calculation for carrier power the available C/N ratio can be calculated as:

\[ C/N = (-118.51) - (-129.2) \]

\[ -10.69 \text{dB} \]

Putting Us In The Picture
A simple C/N ratio like this doesn't mean a lot to the uninitiated so we need to look further into what it represents. Measurement of quality of a television picture is a highly subjective area (an adequate picture for one person is no good for another) but the International Radio Consultative Committee (CCIR) has defined a picture quality assessment plan which uses a five-point grading system. This, related to a system's CN ratio, is listed in Table 1 and shown in graphical form in Fig. 4.

From these we can see that a practical C/N ratio of 10.7dB (the system in my back garden) gives a picture which is classed as only just fair. Even more worrying than this, however, is the fact that such a low C/N ratio is probably close to the receiver's FM threshold margin (typically around 8-9dB) and so impulse noise interference showing up as black and white flashes or dots on the screen — the infamous 'sparklies' — could occur.

If this was a real system which I was about to buy, I'd want to know how I could improve the picture quality a bit. Let's go back to the calculations for C/N ratio to find out how. The overall ratio is given (in decibels by):

\[ \frac{C}{N} = \frac{P_r}{P_s} = \frac{\text{EIRP} + A_r - L_r - \text{weather loss}}{kT_B + B} \]
Some of the factors in the calculation are constant — EIRP, spreading loss Ls, Boltzmann's constant k — and cannot be altered. Weather loss (estimated at about 2dB, is of course not constant but cannot be ignored as worst-case weather situations must be accounted for. The others depend on physical parameters such as antenna size and efficiency, LNB noise temperature and receiver bandwidth, so a number of measures can be taken to improve C/N ratio and the picture quality.

A larger or a more efficient antenna will give a greater carrier power so improving the C/N ratio. Conversely, an LNB with a lower noise temperature will decrease the noise power, so improving the C/N ratio. Also, reducing the receiver bandwidth will decrease the noise power — but there is a limit to how small the receiver bandwidth can be! — 30MHz is about the minimum. Some satellite television channels (Sky Channel) have a bandwidth of as little as 27MHz but others (Superchannel) have a bandwidth of as much as 36MHz.

If receiver bandwidth is too low, not all the channel will be demodulated and results such as a tearing effect on video scenes with a sharp vertical edge can be caused. This leaves me and my back garden with a simple choice: get a bigger antenna, a better antenna or a higher quality LNB. Now, the wife's already up in arms about having a monstrous brilliant white 1.5m dish out there and she's not going to let it grow just to get a better picture so a better antenna and better LNB is the answer.

Using an antenna with an efficiency of, say, 65% and an LNB with a noise figure of, say, only 1.5dB (some of the most recently produced LNB's can meet this requirement) we can now re-calculate the C/N ratio.

Effective antenna area is now 65% of 1.767m² (1.15m²) or in decibels: 0.6dB. LNB noise temperature is now: 290 \( \left( \text{antilog} \left( \frac{1.5}{10} \right) - 1 \right) = 119.6K \)

and total noise temperature is 119.6K+60K=179.6K or in decibels: 22.5dB. Final C/N ratio now becomes: 

\[ C/N = 4+0.6-162.76-2-(-228.6+22.5+74.8) = 13.14\text{dB} \]

This result corresponds to a picture quality which is close to being good on the CCIR five point scale and I'll certainly accept it. Generally, a C/N ratio of about this 13dB figure should be aimed for when doing a link budget analysis. This should allow for extremes of weather conditions, minor antenna pointing inaccuracies and the like.

**Figure Of Merit**

Overall receiver performance is sometimes quoted as a figure of merit which is its gain over temperature (G/T). All receiver parameters such as antenna gain, total system noise temperature and so on are catered for in this G/T figure and so the figure of merit is related to the C/N ratio and one can be calculated from the other (approximately) according to the relationship:

\[ C/N = EIRP + G/T - 53\text{dB} \]

or inversely:

\[ G/T = C/N - EIRP + 53\text{dB} \]

So, the practical example of the system in my back garden (with improved antenna and LNB) has a figure of merit of:

\[ G/T = 13.14 - 46.53 = 20.14\text{dB} \]

Next month, to round off this series, I shall look at three specific STV systems in detail.

<table>
<thead>
<tr>
<th>Quality</th>
<th>Grade</th>
<th>Impairment</th>
<th>Approx C/N (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Excellent</td>
<td>5</td>
<td>Imperceptible</td>
<td>20</td>
</tr>
<tr>
<td>Good</td>
<td>4</td>
<td>Perceptible</td>
<td>15</td>
</tr>
<tr>
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<td>10</td>
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<td>Annoying</td>
<td>7</td>
</tr>
<tr>
<td>Bad</td>
<td>1</td>
<td>Very annoying</td>
<td>5</td>
</tr>
</tbody>
</table>

Table 1 CCIR picture assessment compared to satellite receiver carrier-to-noise ratio

---

**Figure 4 CCIR picture quality assessment compared to carrier-to-noise ratio at the demodulator input of a satellite receiver**
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ETI MARCH 1988
MICRO CIRCUITS

Malcolm Brown takes a look at two programs for the BBC micro and Amstrad CPC machines to help the amateur circuit designer.

In the March 1987 issue of ETI we took a look at several programs for the BBC micro to help you design and draw circuits and PCB foils. One of the winners of that selection was Diagram from Pineapple Software. Now Diagram has been completely rewritten with many additions and improvements as Diagram II. First, however, we'll look at a new contender - a PCB design package for Amstrad CPC micro users.

PCB Drafting Utility (cassette) £19.99
£21.99 (disk)
CADsoft, 18 Ley Crescent, Astley, Tyldesley, Manchester M29 7BD.

The humble Amstrad CPC range of micros provide a lot of computer for not very much money so it is about time they were used to help the struggling electronics enthusiast.

This package is a PCB design tool to produce paper artwork of foils for subsequent photographic transfer to celluloid for UV exposure and etching.

The program is available on cassette for CPC464 users and on 3in disk for those with a CPC664 or CPC6128.

The program is in two parts - the editor and the plotter. Both sections are based on Basic programs with machine code routines.

The editor is used first to create your PCB design. PCBs up to 25x25in can be designed. The screen display shows a greatly enlarged section - about 4x2 1/2in - at any time. However, the whole PCB can be drawn on at a time if you can manage to do this 'blind'.

This is more useful than it seems at first because it means a track can run for more than the width of the screen and still be drawn on as a drawing package which stores the start and end co-ordinates of the lines (tracks) and points (pads) and suffers from the problems associated with this method. Although the whole PCB is much larger than the screen window, you cannot scroll across it. Instead the screen must be entirely redrawn from a new viewpoint each time.

However, that aside, drawing a PCB with this package is easy enough. There are three structures which can be drawn - pads, tracks and DIL packages (which are really just collections of pads - so there's only two structures after all).

Tracks are drawn by rubber banding. The cursor is moved around in discrete 0.05in steps and the Tab key starts and stops line drawing. The lines drawn on the screen in this way do not directly correspond to the tracks on the final PCB. The final width of the tracks is set separately to 0.0125, 0.025, 0.0325 or 0.05in.

Pads are similarly positioned at the current cursor position and they too can be preset in size to 0.05, 0.065, 0.085 or 0.1 in.

Both pads and lengths of track can be deleted from the foil by moving the cursor to any part of them and pressing the appropriate key.

Sets of DIL pads are placed on the foil with the function keys and they can be set either horizontally or vertically. The pitch between the rows of pins can be altered to just about any conceivable value but sensible defaults are provided. Setting a row separation of zero provides SIL package pads.

Once on the foil design, the DIL pads are treated as individual pads and must be deleted one at a time.

The program can cope with double-sided boards and both sides are displayed at the same time in different colours. Which side is 'active' - being drawn on at the time - is selected with the U and L keys (upper and lower). All pads are always placed on both sides of the board which can occasionally be a problem.

The completed board can be saved to disk for further editing later or for printout.

The plotter routine uses a standard Epson-compatible dot-matrix printer. Either or both sides of the board can be printed and the printing can be either way around on the paper.

Printouts are made at twice size in four-inch strips which must be joined later for photography and are produced in either a quick draft mode or in the higher quality final mode.

The sections of the foil are drawn out on the screen before dumping to the printer and it is here that the real quality is shown. In this mode the tracks and pads
are expanded to their intended size and the pads take on the normal elongated oval shapes.

CADsoft's PCB Drafting Utility is adequate for most amateur use, but it is not to the quality of Pineapple's PCB program (see that March 1987 review) for the BBC micro.

The Amstrad CPC micros have a far greater potential for this kind of work because of their vastly increased memory. However, some silly scrappy Basic programming mistakes (such as overwritten menu prompts and no disabling of the potentially lethal Escape key) means this program cannot be considered a truly professional tool.

Diagram II
£55+VAT (exchange deals available for Diagram I users).
Pineapple Software, Brownlea Road, Seven Kings, Ilford, Essex IG3 9NL. Tel: 01-599 1476.

There are many drawing packages around for the BBC micro but this one uses an entirely new method of storing the picture to make it particularly suited to drawing circuit diagrams.

Many other drawing packages store the start and end co-ordinates of all the lines in the picture along with a code to determine whether a line is drawn or a rectangle or whatever. This can be wasteful of disk space and makes editing the picture slow and difficult.

Other programs (usually the more 'artistic' ones) store the memory mapped screen RAM. This is usually quite economical on storage space but useful editing is nearly impossible.

Diagram II uses neither of these methods. Instead, the picture is stored as an array of character codes (an extended ASCII coding system is used) with the bit map definitions corresponding to each code stored separately.

Of course with a picture (such as an artistic work) with a great deal of variation across it, this is an uneconomical method, to say the least.

However, with a diagram consisting of a great many repeated symbols and sections (as is a circuit diagram) the method is economical and has a few other great advantages.

Most important is the speed in which the diagram can be drawn. This not only means that it can be produced on the screen and edited quickly but also that the section shown on the screen can be just a part of a much greater whole. Diagram II allows smooth scrolling across a complete diagram of up to 30 Beeb mode 0 screens.

This method also allows the complete multi-screen diagram to be easily displayed on one screen with each pixel representing each character of the full sized screen. A complete diagram of up to 8x8 screens can be displayed in this reduced form for checking the 'whole look' of the diagram.

Despite this unique system, when using the program you would not know that Diagram II is any different from other drawing packages. The mechanics of it all are very cleverly hidden. The cursor is used to move around the diagram and a line is left trailing behind. The program automatically lays down the correct line character needed at that point in the diagram. Turn a corner and a right angle turn character is printed, cross another line and a cross character is printed at the intersection.

A 'blob' can be left at line intersections just by pressing Ctrl as you pass over the join. Now that just isn't possible with other types of drawing package. They require a precise positioning of a blob symbol after the lines are drawn.

This character system also means diagrams are naturally and easily kept aligned and 'square'. It is genuinely difficult to not draw neat and impressive diagrams using this software.

That's all very well for vertical and horizontal lines but Diagram II can also make use of 'rubber banding' to draw all the diagonal lines in between or for tracing an irregular shape. Amazingly, this still uses predefined characters, automatically selected and positioned to make up all the line sections.

As well as straight lines, Diagram II running on a Master or model B with Acorn's GXR ROM can produce circles, rectangles, ellipses, arcs and triangles, either filled or in outline, to build up large shapes. Without the GXR ROM only circles and filled triangles can be created.

Again, although drawing these shapes is performed with the cursor and is as easy as with any drawing package, the end result is made up of individual predefined characters which go to make up the complex shapes.

Text is positioned on the diagram simply by typing it in when the cursor is at the right position. Text can also be taken from a file created with Wordwise Plus and automatically fitted in a defined section of the diagram.

Diagram II can produce discrete symbols such as transistors, resistors and so forth. These are also made up of characters - up to 12 normal Beeb 8x8-pixel characters.

With a basic model B BBC micro, 335 characters can be defined, shared between the symbols. With a machine using shadow memory (such as the B+ or Master) 880 Beeb characters can be used.

A definite improvement over the old program is the ability to delete unwanted symbols and so free the characters for use with new ones. The complete set of symbols or just a limited subsection from one diagram can be saved to disk to use with another diagram - again a useful addition.

To design these symbols a very effective character definer is provided. There have been many such definers produced for the BBC micro but this is one of the best I've seen. It is fast and simple to use. As well as allowing the symbols to be defined one pixel at a time on an enlarged grid, complete symbols can be emptied.

\[\text{NOTE: When changing channels only XL and the original tone circuit need adjustment.}\]
or flipped horizontally or vertically.

The character definer can of course be used to define new styles of text characters and these can then be inserted in the diagram just like 'normal' text, from the keyboard.

Each character has a 'direction' defined along with it. This controls the cursor movement after the character is positioned on the diagram. The direction can be defined in any of the four compass directions so that, say, a complete set of upside down characters could be typed right to left straight from the keyboard without fiddling with the cursor between characters.

Any area of the diagram can be defined with the cursor and then deleted, moved, copied to another part of the screen or saved to disk to be inserted into another part of the program or into another diagram altogether.

Borders can also be added in one single movement and existing diagrams changed in size for additions or shifted across an enlarged diagram.

The completed diagram is stored on disk or printed on a standard Epson compatible dot-matrix printer at just about any size from three screens to just one character across a sheet.

All the parameters entered to control the size, position and codes used for the printout can then be saved to disk to save time for later repeat runs.

The quality of printing is the last factor which makes Diagram II so outstanding. This package really is capable of producing top quality diagrams. It is difficult to see how this package could now really be improved. For anyone looking for ways to edit, store and print out professional circuit diagrams with a bare minimum of equipment, Diagram II is the only answer.

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**REVIEW: Circuit CAD**

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**Audiokits Precision Components**

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**Sony Speakers**

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**ETI**

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**Sony Speakers**

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**ETI March 1988**
IT'S NEW, IT'S HERE AND ONLY AVAILABLE THROUGH ETI MICRO MINIATURE SURFACE MOUNT PROJECTS

You've heard about them, probably seen and marvelled at them inside your calculator or watch and now you can begin using them with the latest range of surface mounted component kits from SUMA DESIGNS. All the kits are of well proven, in-house design and come complete with all components individually packed for easy identification, high quality fibreglass, surface mount, solder resist P.C.B., full assembly and circuit instructions. All you need is a pair of tweezers, a fine-tipped iron and a steady hand. Using assembly techniques developed to cope with our own prototype and batch manufacture you can be assured of success. To make things even easier we have developed a custom made component jig which although not essential greatly simplifies assembly and is well worth the investment. A full range of SM components are also stocked for your own projects.

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Paul Chappell pulls his phasors together to find a use for mathematics in circuit design

Last month's article was concerned with the mechanics of manipulating complex numbers with hardly a mention of how this might apply to electronics. This month I'll redress the balance and concentrate on applying complex number arithmetic to circuit calculations.

We haven't reached the end of the mathematics by a long way but it's just as well to keep the final objective in sight — to make AC circuit theory as easy as π.

The Story So Far

In an earlier article (November 1987 Circuit Theory) I demonstrated that in a capacitor with a sinusoidal voltage applied, the current would also be sinusoidal. The ratio of the voltage and current amplitudes at any given frequency is constant so the capacitor can be thought of as having a 'resistance' to sine waves.

Since the apparent resistance varies with frequency and the voltage and current waves are offset by 90°, it's not quite the same as ordinary resistance. It is given the name reactance, the symbol X, and for a capacitor C at angular frequency ω it has the value -1/ωC.

For a sinusoidal current of amplitude I, the voltage amplitude will be given by V=IX, just as for an ordinary Ohm's law calculation V=IR. An inductor has similar properties and a reactance XL of ωL.

The problem we found with using the straight voltage-to-current amplitude ratio as the 'resistance' of reactive components was that the 'resistances' didn't add for components in series and the usual parallel resistor calculations didn't apply to components in parallel. So the concept of an AC 'resistance' was rather limited in application!

We concluded that this problem arose because the 90° phase difference between voltage and current was not being taken into account. With the aid of complex numbers we are now in a position to correct this flaw and bring out the full power of the ideas.

A Practical Example

Let's take a look at the circuit of Fig. 1(a). To keep our feet firmly on the ground I haven't used normalised values — the frequency and components are those you might find in a real circuit. I haven't specified the input for the moment. All we know is that it's a sine wave of angular frequency 5x10^6 radians per second (roughly 800kHz).

The reactances of the inductor and capacitor at this frequency will be XL=500 ohms and XC=400 ohms. This suggests that the entire circuit has an impedance of 1000Ω and that a 1mA amplitude current should produce a voltage across the circuit of 1V (both at 800kHz).

We already know that this doesn't happen and Fig. 1 shows why. Taking the cosine form of the input current and freezing the phasors at t=0 gives the situation depicted in Fig. 1(b) for the individual voltages across each component and Fig. 1(c) for the combined voltages. The voltage across the entire circuit (Vt) has an amplitude of 141mV against a predicted value of 1V. Our estimate was out by a factor of seven!

The diagrams show clearly what has gone wrong. To give the predicted 1V, the voltages would all have to be in phase so their phasors would add in a straight line. They aren't and they don't!!

Ordinary numbers can't take into account the direction of the phasors. All they can do is tell you how long they are. Complex numbers can do both!

Let's begin again. First of all we'll relabel the circuit as shown in Fig. 2. We have now entered the frequency domain! This means every time I mention a current I or a voltage V, they are automatically assumed to be sinusoidal in form (which saves having to state the obvious every two minutes) with amplitudes I and V respectively.

This also means the components are labelled with their complex impedances rather than with their normal component values. We'll see how it works right now.

First of all, we'll set the input frequency to ω=5x10^6 radians per second. Now we can work out the impedances of each of the components. ZL=j500, ZC=-j400 (note the minus sign — remember that 1/j= -j) and ZR=100.

To find the impedance of the entire circuit, just add up the individual impedances. 100+j500-j400=100+100. If we agree to work in mV and mA, then a 1mA input current (remember: frequency domain so this means a sinusoidal input current of amplitude 1mA and not a steady 1mA current) in cosine form will be I=1+j0 (or just 1 for short!). The voltage will be V=1XZ=1x(100+j100)=100+j100 and we're through. Well, almost. My voltmeter doesn't actually register voltages of 100+j100!
To express the answer in more familiar terms, we have to translate back to the time domain. The voltage 100+j00 has a modulus (length) of 141 and an argument (angle) of tan⁻¹(100/100)=45° = π/4. In full, we’ve discovered that an input current of cos(5x10⁻⁶t) at the input will give rise to a voltage 141 cos(5x10⁻⁶t+π/4) across the input terminals.

Now let’s look at the pictures to see what’s going on. First of all, we agree to draw our phasors in the complex or z-plane. The phasor representing a current of 1mA in cosine form points along the positive real axis and is described in complex number form as j00 (we’re working in units of mA and mV, remember).

This gives rise to a voltage of (1+j0) x (j500) across the inductor which comes to j500 or a phasor of length 500 pointing up the positive imaginary axis. There’s no need to draw it again — it’s there in Fig. 1(b)!

The voltage across the capacitor comes to -j400 — represented by a phasor of length 400 pointing down the negative imaginary axis (as in Fig. 1(b)). The voltage across the resistor you can work out for yourself!

Now we add up the three individual voltages (remember from last month that complex number addition corresponds to addition of phasors in the z-plane) giving the result shown in Fig. 1(c). The voltages due to a unit current add correctly so the impedances must also add correctly, since impedance is exactly that — the voltage per unit current.

**Time And Frequency Domains**

Just about every circuit you’ve ever seen in the pages of ETI is in the time domain. All this means is that if you know how the input behaves as a function of time, you can deduce the output — either by calculation, common sense or by reading the ‘How It Works’ section!

You might ask yourself what happens when the switch is pressed. What happens just after the pulse reaches 1C? When will the output of 1C go high? All questions to do with time. First one thing happens, then another, then another … OK, I won’t labour the point any more!

Figure 1(a) is a time domain circuit. You could ask what its response is to a voltage step then draw the output as a graph of voltage against time and so on. We know how to do that if we work from the time domain circuit.

Figure 1(a) was a sine wave, every single voltage and current would also be a sine wave. So asking how the circuit behaves in time is silly. We know already — it gives sine waves.

If we know the voltages and currents in the circuit at any instant in time, we can deduce what they will be for evermore. So to inspect the circuit we suspend the phasors spinning along gliding along its path. Since we haven’t found a way to do it on the pages of ETI, you’ll have to use your imagination!

If you think this is good, you ain’t seen nothing yet! If you still haven’t caught on, perhaps the more familiar f=1/(2πf0) will give you a clue!

Let’s see if we can work out the frequency at which this happens. Looking at Fig. 1(c), it’s clear that the condition for the voltage phasor for the whole circuit to point along the positive real axis is that Vc=-VL. If this holds, the voltages across the inductor and capacitor will cancel and the voltage across the entire circuit will be equal to the voltage across R. In mathematical terms we require 1/jωC=-jωL. Solving for ω we get ω^2=1/2πf0LC.

If you still haven’t caught on, perhaps the more familiar f=1/(2πf0) will give you a clue!

We’ve just derived the condition for the circuit to be in resonance. The fact it was obtained so easily is a demonstration of the power of this method of analysis — you’ll appreciate if you’ve ever tried to derive the same result in the time domain with the aid of a page of differential equations!

If you think this is good, you ain’t seen nothing yet! Watch this space next month.

**Frequency**

Sticking with the circuit of Fig. 1(a), let’s see what we can find out about its behaviour at different frequencies. First of all, you’ll notice the voltage across L is proportional to frequency. As ω increases so will the voltage across L. As the frequency decreases, so will the voltage.

For the capacitor, exactly the opposite happens. Up the frequency and the voltage goes down. Reduce the frequency and the voltage rises.

If you think about the effect of this on Fig. 1(c) (assuming an input current of 1mA at various frequencies) as the frequency increases, Vc gets longer and the VL phasor gets shorter. The resultant voltage across the entire circuit gets larger in magnitude and leads the current by a greater and greater angle.

Reducing the frequency increases Vc and reduces VL so the overall voltage swings downwards and lags the current by a greater and greater angle. The effect on the voltage across the circuit as a whole is shown in Fig. 2(b) for various frequencies. (It’s at times like this I wish I could produce animated graphics with all the phasors altering in length and the voltage phasor gliding along its path. Since we haven’t found a way to do it on the pages of ETI, you’ll have to use your imagination!)

For some value of ω, the voltage across the circuit will point along the positive real axis. It will be in phase with the input. It will also have its maximum amplitude at this point. (If you can see where this is leading, take the Mrs Beestly prize for supreme cleverness!)

Let’s see if we can work out the frequency at which this happens. Looking at Fig. 1(c), it’s clear that the condition for the voltage phasor for the whole circuit to point along the positive real axis is that Vc=-VL. If this holds, the voltages across the inductor and capacitor will cancel and the voltage across the entire circuit will be equal to the voltage across R. In mathematical terms, we require 1/jωC=-jωL. Solving for ω we get ω^2=1/(2πf0LC).

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**ETI MARCH 1988**
Arthur Bossard inspects this versatile IC and finds it's good for more than switch mode power supplies

Now that Les Sage has introduced you all to the mysteries of switch mode power supplies (ETI October and November 1987) it seems an appropriate time to look at one of the many control ICs available for the purpose — the LM3524.

This IC is not the ultimate in high-tech, just the opposite, in fact. It is chosen because it is cheap (so you can experiment without heart palpitations at the thought of ten quid going up in smoke!). It is also readily available, doesn't need many external components to make it do its stuff and because it can be used for other control functions (or even as an LED flasher if your imagination deserts you!)

For those not familiar with switching regulators, the LM3524 is a good introduction since the basic functional blocks are not obscured by a mass of control circuitry. There's an over-current sensor, a logic input to shut down the IC in case of faults in other parts of the circuit, and that's your lot.

The functional blocks inside the IC are shown in Fig. 1. The heart of the IC is the oscillator which runs at a frequency set by the resistor and capacitor from pins 6 and 7 respectively. The capacitor is repeatedly charged at a constant current to 3.5V, then quickly discharged back to 1V, resulting in a sawtooth waveform at pin 7.

Forgetting the input amplifier, current limit and all the rest of the paraphernalia for a moment, imagine applying a DC voltage to pin 9 of the IC. (You can actually do it if you want to, as the internal amplifiers are transconductance types and won't be upset by it).

If the voltage at pin 9 is below 1V, the comparator output will always be high since its positive terminal is connected to the sawtooth waveform which is always above 1V. If the voltage at pin 9 is a little above 1V, the comparator output will go low briefly after each discharge of the capacitor. A higher voltage on pin 9 will keep the comparator output low for longer and an input above 3.5V will keep it low all the time.

So the comparator output is a rectangular wave with a duty cycle controlled from zero to 100% by the voltage on pin 9.

If this wave were fed directly to a single output transistor, we would already have a very basic voltage controlled pulse-width modulator suitable for use in a switching regulator!

The LM3524 has two output transistors to allow for push-pull operation. Where this is not required, the two transistors are connected in parallel. The flip-flop switches between the two transistors on alternate cycles of the oscillator.

The gate input from the oscillator's output is a 'blanking pulse' to give a short period when both transistors are turned off, preventing cross-conduction at maximum duty cycle.

You can get a 'feel' for the operation of the IC so far by connecting up the circuit of Fig. 2. The LEDs flash alternately as the flip-flop switches from one output transistor to the other. RV2 sets the oscillator frequency and therefore the flashing frequency. RV1 controls the duty cycle of the out pulses.

At the ground end of its travel, as soon as one LED turns off, the other turns on (apart from the brief blanking pulse, which won't be visible). As the preset is rotated, the flashing becomes briefer, giving a more 'urgent' look to the display.

This could actually be quite useful for a warning display, or for use with model cars or trains to simulate danger signals.

Strictly speaking, the values of RT and CT (RV2 and C1 in Fig. 2) should not be above 100k and 100n respectively for reliable operation. What you'll find is that if you try to reduce the frequency too much the
flip-flop won't switch over, so you'll be left with one flashing LED and another that doesn't light at all.

Switching regulator circuits
The basic step-down regulator configuration is shown in Fig. 3. This is broadly equivalent to a standard series regulator. The input voltage can be anywhere between 8V and an absolute maximum of 40V. The output (for the component values shown) will be 5V. The maximum input current, without an external drive transistor, will be 80mA.

For the small currents involved, L1 will have to be quite high in value to sustain conduction when the transistors are turned off. A value of about 10mH will be needed. D1 can be a 1N4148.

The value of C3 will depend on the amount of ripple that can be tolerated on the output. A value of 1μF is a reasonable starting point. Increasing the value will improve the smoothing but will make the circuit less able to cope with rapid changes of input or load.

A Step Up
The basic step-up regulator configuration is shown in Fig. 4. Once again, L1 can be about 10mH, C3 1μF and for the low frequency and small currents involved D1 can be a 1N4148. The output voltage will be 27.5V for the component values shown. The input can be anywhere from 8V to about 20V.

The output voltage can be varied by changing the value of R5. If a current limit is needed, it can be connected in the same way as for Fig. 3.

The component values given are intended as a starting point for experiment — this isn't a project. You might like to try finding the limits of each circuit for the values given — particular of the inductor. What is the smallest different in output and input voltage that the circuits will cope with? What is the largest? (Bear in mind that the IC will be damaged by input voltages of over 40V.)

How does it cope with variations of the load? What is the disadvantage of using too large an inductor? Or too small? What happens if you use a 1N4001 in for the rectifiers? Or a Schottky diode? What happens if you make the value of the smoothing capacitor 10μF or 1000μF?

My theory about electronics is that an hour spent in intelligent experimentation (where if something goes wrong you ask 'why?' and make damn sure you come up with an answer!) is worth a week reading a text book. That's not to devalue theory but a touch of practical experience gives meaning to what would otherwise be dry and uninteresting facts.

If your amplifiers oscillate and your regulators don't regulate, it gives you the incentive to get to the bottom of what's going on!

Other applications
The LM3524 is quite a versatile little IC and can be used for other purposes besides switching regulators. One thing that occurs to me is that it could be used for temperature regulation of sensitive components — a crystal oven for example.

In its simplest form, the heating element could be a resistor connected to the output transistor collectors (Fig. 5). The temperature sensing element could be a common or garden diode, arranged so that increasing temperature causes a rise in voltage at pin 1 of the IC. The set point would be controlled by varying the DC offset at pin 1 or the voltage at pin 2.

Another possibility is to make optically isolated analogue outputs, such as the ones in the recent ETI EEG Monitor. A different approach would be to vary the mark-to-space ratio of the outputs in sympathy with the analogue signal. The circuit (Fig. 6) is similar...
to Fig. 2 with the LEDs replaced by the opto-isolator and the two output transistors connected in parallel. Some level shifting is needed on the input to match the available analogue voltage range with the requirements of the LM3524, and IC2 achieves this.

Infra-red remote control? Once you start thinking about it, the possibilities for applying this IC seem endless. Have fun experimenting, and don’t forget our Tech Tips page if you come up with something good.

ETI MARCH 1988
GAS ALERT

Greg Thompson promises not to set the world alight with this top-spec gas detector design

Gas explosions in the home, caravans and the like are becoming virtually a daily occurrence. The ETI modular gas detection system will help to reduce these horrific accidents.

The system will detect and provide early warning of the presence of domestic gas, bottled gas based on methane and explosive vapours such as petrol fumes and other hydrocarbons.

Good Sense

Most inflammable vapour sensors are based on what is known as the 'hot-wire' principle. This involves a small heater element inside the sensor.

The sensor chosen incorporates the usual heater element which is designed to operate at 5V (±0.2V). The sensor output is quite simply a varying resistance.

This resistance remains stable whilst the sensor is in clean air but the resistance across the sensor drops when gas or explosive vapour is detected.

The heater element and resistive detector are two ideally independent circuits within the sensor.

HOW IT WORKS

Power enters the circuit via a protective fuse. The specified transformer supplies about 10.5V on the +V rail.

C1, R1 and C2 form a smoothing filter which acts as a safeguard should the unit be powered from an unsmoothed supply. IC1 is a 5V regulator from which the sensor and IC2 are powered. C3 is an additional smoothing capacitor.

LED1 and its series resistor R2 are positioned across the supply rails to provide power-on indication.

The heater element of the sensor is fed directly from the 5V rails. IC2a acts as a voltage comparator biased by resistor R8 and the sensor itself. R3 and C4 raise the input impedance and also stabilise the op-amp inputs.

When the sensor detects gas its resistance drops below that of \textit{R} causing the voltage at the inverting input (pin 2) to swing negative which in turn causes the output (pin 1) to go high. The non-inverting input (pin 3) is biased to half the supply potential by R6 and R7 enabling the op-amp to make its comparative decision against the voltage at pin 2.

R10, C5 and R11, 01 form the time delay stage. When gas is detected, pin 1 goes high which will after approximately 13 seconds take pin 5 high (the non-inverting input of IC2b).

If the gas dissipates within 13 seconds, pin 1 returns low (its normal state). C5 discharges through R11 and D1. Pin 1 must remain high for the full duration of the 13 seconds before the final alarm is triggered.

Any brief encounter with gas will be shown by the illumination of amber LED2. This is turned on whenever pin 1 goes high via R9 and Q1.

IC2b again serves as a voltage comparator also biased by R6 and R7. If gas is detected for a longer period than 13 seconds the non-inverting input (pin 5) is held high causing the output (pin 7) to also go high. When this happens Q2 and Q3 are also turned on thus switching on the buzzer, red LED3 and reed relay RLA1 via Q3.

The coil of the reed relay also serves as a series resistance for LED3. If gas is detected for a longer period than 13 seconds the non-inverting input (pin 5) is held high causing the output (pin 7) to also go high. When this happens Q2 and Q3 are also turned on thus switching on the buzzer, red LED3 and reed relay RLA1 via Q3.

As with all hot-wire sensors, respect must be given to current consumption. The heater element in this sensor will draw approximately 170mA when supplied by a 5V regulator. This current drain compares favourably with other available devices.

Talking of other devices, use of the 'two dome' sensor/compensator type detector should be avoided. These detectors consume more current and are rendered useless if allowed to absorb silicone. Gas alarms that are utilised in domestic applications can easily become contaminated with the silicone used in many household spray polishes.

It may seem strange or even
The Sensor
Whilst most of us are more interested in the electrical and environmental characteristics of the sensor, the physical philosophers and philosophically fit will be eager to learn how it works. You will see from Fig. 2 that the heater element passes through a small ceramic tube. This is coated with a layer of tin oxide, SnO₂.

If tin oxide is heated in a 'clean air' atmosphere, oxygen is absorbed into the surface layer. The rate of absorption remains constant at a given temperature. If, however, a contaminant combustible gas is introduced, this will also be absorbed.

The reaction of a combustible gas with the oxygen causes electrons to be released from the oxygen giving the tin oxide greater electrical conductivity. This all takes place between the two electrodes at either end of the ceramic tube. The resultant factor is increased conductivity which is equal to the lowering of its electrical resistance.

Fig. 2 The hot wire gas sensor

absurd to incorporate a heater element in a gas sensor as such a system appears to be a source of ignition itself. The sensor type used here has been vigorously tested in an atmosphere of 2:1 hydrogen/oxygen — a very explosive mixture. These tests were carried out under normal conditions and with an internal spark, both without causing ignition. This is due to the extremely fine stainless steel gauze used in the construction of the sensor. The sensor itself is an internationally proven and accepted device and we stress it cannot itself be a source of ignition. However, careful consideration must be given to the circuit and system in which it is used.

It's OK for your electric toaster to blow you off the face of the earth but if your gas alarm does the same, it's not on, is it?

It is worth pointing out at this stage that due to relatively high current power consumption it is neither practical nor recommended to operate such a gas detector from dry cell or NiCd batteries.

In our design we have incorporated an independent relay output in the form of a simple single pole make switch. The relay used is of the sealed reed type. This is important as relay contacts create a spark when thrown.

The relay output is included so the detector can be used to trigger existing alarm systems or any external device for that matter. Some of the more ingenious readers may decide to develop subtle luxuries such as switching in your extractor fan to help to 'clear the air.'

You may well decide to do this by using the internal reed relay to throw an external mains relay. Mains relays are usually of the open type and will without doubt cause a spark sufficient to ignite a gas filled room. It's recommended that a triac switching circuit be used for switching in any mains powered add-on.

We could also add that the electric motor inside extractor fans will also produce an arc from its commutator brushes but for the moment we'll assume most extractor fans are sufficiently sealed and in any event are normally exposed to clean air on one side.

Having scared the living daylights out of you we will now...
tell you that the alarm will function well before an explosive mixture is allowed to accumulate.

The point at which a mixture of air and explosive vapour or gas will ignite is known as its lower explosive limit (LEL). Such mixtures are defined in parts per million (PPM). This is the molecular count of explosive mixture per one million molecules of air. The LEL is dependent on the PPM — got it!

Our gas alarm system will trigger when an atmosphere of between 10% and 40% LEL has been reached. This may seem a rather large tolerance window but it includes all explosive atmospheres — methane (natural gas), butane, propane and so on, and other vapours (petrol, methanol, ethanol, propanol Navy Rum, etc).

Alarming Falsified

The time honoured expression 'false alarm' is a day-to-day saga in the majority of warning and detection devices of all types. Sophisticated electronic devices incorporating failsafe and fail failsafe are normally so safe that they fail to register anything or they are continually being triggered by just about everything except the gas they were designed for.

It's not a bit of good having a gas detection system triggered off by a quick squirt of hair spray or cooking deposits in the air. It's just switched off when everybody is so fed up of grabbing their wallets and purses and running out into the street and going four or five doors down the road to use someone's phone (next door is too near, they'll go up as well).

Now Hear This

The British Standards Institute is in the process of receiving (this is confidential you understand) a draft proposal from British Gas in respect of domestic gas alarms. Although the document will delve into meticulous detail of test procedures involving gas concentrations, temperature and air pressure, notwithstanding wind velocity, it would appear that some of the more important factors have been overlooked.

The whole purpose of an alarm is to let someone know something is happening. Quite sensibly, the Standard has opted for an audible alarm which it stressed must have a sound level output of 85dB at 3m. Having delved into the depths of Einstein's archives to create test procedures it omits to specify the frequency of the sound the audible alarm should generate!

Well, British Gas, have you ever heard 85dB at 25kHz? No, that's not surprising because neither has anyone else on this planet. 25kHz is all very well for dogs, they'll be saved!

It may well be that if you buy a gas alarm that has been manufactured to British Standards, your first indication of an imminent explosion is the dog going berserk or your mynah bird flat on its back in the cage pushing up the daisies.

About 400-700Hz would be more suited. However even at a desired frequency, 85dB is certainly not over loud. In an average house you may have an alarm installed in your kitchen. If the kitchen door is closed, your bedroom door is closed, it's the middle of the night and you're asleep, it is open to interpretation as to whether you would hear an alarm operating at this sound pressure level.

The ETI system has an internal buzzer rated at 75dB at 1m but serious provision has been made for external louder audible sounders positioned wherever required. This is of particular importance in larger houses.

Sensors

In the near future British Gas may invite you (at your own cost, of course) to have a gas detection device fitting in your kitchen. That's fine if you get a gas leak in your kitchen but what about the gas meter under the stairs, the gas fire in the lounge, the central heating boiler in the garage or what have you.

In the light of that, consideration should be given to the installation of more than one sensor in order to provide comprehensive cover.

The ETI gas alert is designed to operate from an AC or DC supply producing 12V. Up to now we have dealt with practical considerations for domestic installations and for this purpose a
240V mains transformer is used. Boats and (more so) caravans are also prone to gas leaks from faulty low pressure bottled gas (LPG) systems. This system can be powered from standard 12V lead-acid car batteries — the normal source of power for boats and caravans. The current consumption of around 200mA under normal working conditions, would not be considered a serious drain on such battery systems. The consumption of the unit rises above 200mA when the alarm is triggered but the last thing you are going to worry about is a flat battery if you’re about to spontaneously combust!

The Circuit
As previously mentioned, false alarms are not acceptable in any gas alarm system. The sensor supplied for this project requires no setting-up or calibration. Each individual sensor has been calibrated prior to leaving the manufacturer's factory. During the manufacture of a complex device such as this component, material tolerances are unavoidable. The inclusion of a resistor (R7) which is supplied with the sensor enables the device to operate within the prescribed parameters.

If you are assembling more than one detector, care must be taken to ensure each sensor and its companion resistor do not get mixed-up. As a precaution it is worthwhile measuring the resistance of the resistor supplied and writing its value on the side of the sensor with a fine permanent marker. You will need to use an accurate digital meter if you choose to do this. IC1 acts as two voltage comparators. As with the rest of the components, a good quality op-amp is chosen — the Toshiba TA75358. In fact this is a dual op-amp in an 8-pin DIL package — an LM358. It features distinct advantages in a circuit of this type.

It is specifically designed to operate on single supply rails, has very low current consumption and its output will swing fully low. Whilst on the subject of Toshiba, the TBC546B transistors were chosen, again being a quality device and are equivalent to BC546. All right Tosh!

Construction
Assembling the Gas Alert should provide no particular problems, particularly if the recommended PCB (Fig. 3) is used.

In the recommended case, the component leads should be trimmed close to the board underside to enable the backpanel to be fitted. The LEDs should be soldered in first with 8mm sleeves over the leads so that they stand proud of the case top. The test button should be soldered in next. It must be positioned exactly vertical and with careful comparison to the height of the LEDs, again for correct positioning. Although it is a little awkward to have this test button protruding throughout construction, it must be fitted at this stage since the amount of heat required might damage more delicate components. The nut on the test button is unused and the pins need trimming off. The voltage regulator IC1 is bolted flat to the board with its pins at right angles. A small heatsink may be placed on top (held in place by the bolt) but this is not imperative. The fuseholders should be installed with the fuses in place. Solder in the tags and trim. Fit the bridge rectifier and check its polarity.
The transistors and ICs can now be fitted, again with attention to their orientation. Solder the relay into place, treating it like an IC. If the relay is not required, insert a 1K0 resistor between pinholes 2 and 6 to load LED3.

The transformer may be mounted with an additional fuse (500mA) in the mains live input if desired. Rectification and smoothing are performed on the PCB. The transformer may be mounted with an additional fuse (500mA) in the mains live input if desired. Rectification and smoothing are performed on the PCB. The transformer can either be mounted in a plug-topped box or in the complete unit with the alarm sounder.

Testing
The unit can be tested by pressing a cigarette lighter over the sensor for about 15 seconds (don't light it unless you want to melt your PCB). The 'Alert' LED (LED 2) should light almost immediately and the alarm should sound after about 20 seconds.

If this doesn't happen (and the supply voltages are known to be operational) then check polarity of components, terminal numbers on the transformer and look for soldering errors.

Installation
Some thought should be given to siting the Gas Alert module or module. Try to place them near likely sources of leaks — the kitchen, gas meter, gas fires and so on.

If used for detecting LPG bottle gas, (butane, propane and so on) the sensor should be 6-12in from the floor (LPGs are heavier than air). For domestic or other natural gas, the sensor must be about 12in from the ceiling as these gases are lighter than air.

The placing of the extension siren (if used) will depend on your habits! Don't put it under the stairs with the gas meter — unless you're in the habit of sleeping there.

For use in caravans, I recommend the module with sensor and 75dB sounder in a single box which simply connects to the caravan's 12V DC supply. Similarly for boats, where the unit could be installed in bilges or engine compartments to detect excess fuel vapours.

It is possible to mount the sensor away from the PCB (a case is available to do this — see Buylines). In this case resistor R1 should be wired across the sensor (to maintain the resistive calibration) not into the PCB. This separate unit can then be connected to the PCB using lightweight 3-core cable (do not use screening as one of the cores).

A louder extension siren can also be fitted, powered via the reed relay contacts. A suitable 85dB siren module is available (see Buylines).

Although many of the parts used for the Gas Alert are available from usual suppliers, some components are specific to this project. These can be ordered from Live Audio Systems, Unit 52, Tafarnaubach Industrial Estate, Tredegar, Gwent NP2 3AA.

The gas sensor and matching resistor cost £10.95, the mains transformer £2.25 and the PCB £2.25.

A kit of the PCB, sensor and other PCB-mounting components costs £19.95.

A complete kit of the PCB, components, buzzer and screen printed case suitable for 12V caravan or boat use costs £28.75. The small extension sensor case costs £2.95. A box for the transformer incorporating a built-in mains plug is available for £2.75. The extension siren module on its own costs £4.50.

A larger kit including the PCB, components, transformer, siren and a box suitable for mains operation is available for £32.85.

All prices include VAT. Please add 50p per order for postage.

Please address all orders to Live Audio Systems. Enquiries can be answered on (0495) 717462 from 3.00 to 5.00pm.
Richard Penney finds there's scope for a scope on his BBC micro

Oscilloscopes are an expensive piece of test gear for the electronics hobbyist. Flashy glowing knob-filled units can have pricetags stretching into the thousands and even the cheapest are well out of birthday present range.

This project is a low-frequency four-channel oscilloscope which is basically a fast analogue to digital converter that adds onto your BBC micro using the 1MHz bus.

A model B BBC is specified because the software needs to poke directly to the screen memory and to the FRED page of the 1MHz bus to optimise running speed. Models with shadow screen RAM can be used provided this is disabled first. The software can be adapted relatively easily to suit different screen addresses.

The circuit diagram for the Beeb-Scope is shown in Figs. 1 and 2. The Beeb-Scope has four inputs (only one is shown) with input resistance switchable up to 1MΩ on DC coupling. On AC coupling a 1μF capacitor C1 is placed in series with this (such a large capacitor ensures there is minimal additional attenuation and phase-shift even at low frequencies).

The four inputs have range switches to select one of nine full-scale input voltage ranges — 0.5V, 1V, 2V, 5V, 10V, 20V, 50V, 100V and 'infinite.' On AC coupling these refer to the peak levels not

Fig. 1 The analogue and power circuits of the Beeb-Scope
The circuit diagrams are shown in Figs. 2 and 3. For an explanation of the circuit we shall concentrate on channel one of the unit.

Inputs come (directly or via a capacitor) from BNC socket (SK1) or terminal post (SK2, 3) to an input attenuator consisting of resistor chain R1-R13 tapped at various points by a rotary switch SW2. This switch determines the input sensitivity and couples the input to analogue switch IC1 which provides channel switching under software control. The selected channel signal is amplified by a factor of 6.375 by IC2, thus boosting the input FSD (from the attenuator) of 400mV to 2.55V (suitable for the ADC chip on the digital board). The gain is slightly variable (by RV1) to allow maximum accuracy to be obtained. Even uncalibrated. However, the unit is adequately accurate for most uses.

Overload indicators LED1, LED2 are driven from the output via a potential divider and two transistors. The first starts glowing for inputs above 80% FSD and one each is provided for positive and negative levels in excess of FSD.

The amplified signal now goes to the digital board where it feeds the analogue to digital converter (IC4). An operational differentiator is used to derive a rising-edge pulse. Most of the ICs on the digital board are concerned with decoding the signals available on the 1MHz bus to ensure access to the device only on receipt of the correct addresses from the computer. The data-bus is connected directly to IC4, as this has tri-state outputs. A simple potential divider is used to ensure that inputs in the range ±2.55V are fed to IC4 as 0 to 2.55V. The data bus is also connected (via various buffers) to the rising-edge and external trigger signals and to the channel-select pins of IC1 on the analogue board.

The external trigger input is simply a TTL input configured in a 'monostable' arrangement. An 'in-use' indicator is driven by this board, which lights when the board is being accessed by the computer.

The device is memory mapped to the BBC micro's 1MHz bus as shown in Table 1. The power-supply is a simple unit using a PCB-mounting mains transformer and standard 78M05 and 79M05 voltage regulators to fulfil the ±5V power requirements of all the electronics.

**How It Works**

<table>
<thead>
<tr>
<th>FCC0</th>
<th>Write: Channel selection (0-7) start conversion (9µs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Read: Bit 6-External trigger pulse</td>
<td></td>
</tr>
<tr>
<td>Bit 7-Rising edge pulse (high=rising edge)</td>
<td></td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>FCC1</th>
<th>Write: start conversion (9µs)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Read: Read digital voltage level</td>
<td></td>
</tr>
</tbody>
</table>

**Table 1 Addresses used by the Beeb-Scope**
the RMS levels. The infinite range grounds the input to the ADC. The levels can be easily changed by altering the attenuation resistors. If you decide to do so, bear in mind that at FSD the attenuators should yield 400mV and that keeping a suitably large resistor between the most sensitive input and the analogue board is sensible. This will avoid blowing the protection diodes should the maximum input be exceeded.

Calculating the resistor values is simply a matter of applying potential divider theory to the situation. Start with the least sensitive input and work up the resistor ladder.

Each input is protected by two 1N4148 diodes in inverse parallel. Both the analogue switch (IC1) and ADC (IC4) are fairly tolerant of excessive inputs but the diodes provide a simple and effective way of avoiding damage, keeping inputs to the switch between ±0.65V. These diodes also protect the analogue switch, which is a CMOS device, against static charges when a particular channel is unused.

The primary purpose of the analogue switch is to alleviate the need for swapping over input leads when monitoring various signals and not to provide a full-blown dual (or rather quadruple) trace display.

A multi-trace display is only practical for low frequencies (under 500Hz) since if the switch is "flicked" too rapidly, a meaningless...
The circuit as it stands does not have a sample-and-hold module and the prototype functions perfectly normally without it. The trace is smooth, even when approaching the highest displayable frequency.

**Construction**

The prototype was built with four input attenuators although the analogue switch and software can cope with up to eight. To keep costs to a minimum, just one or two could be used.

The components used only in additional inputs (beyond the one illustrated in Fig. 1) have prefixes 100, 200 and 300, so omit some or all of these if you are using less than four channels.

More inputs can be added by replacing each of the 1M0 pull-down resistors R15-R18 (which represent channels 4, 6, 7 and 5 respectively) with further attenuators and protection diodes.

**The Tedious Bit**

Construction should be started with the input attenuators, made by soldering resistors directly to the tags of the rotary switches as shown in the wiring diagram (Fig. 3).

This is a tedious task (increasingly so with larger numbers of channels) but care should be taken to avoid errors which can be difficult to correct. Remember to set the stop on the rotary switch to allow nine positions.

**Fun Bit**

Assembly of the PCBs is quite straightforward. The power supply listing:

Listing 1 The BBC Basic/assembler program to operate the Beeb-Scope
Assembly order here is not critical. However, it is wise not to insert the transformer until last. Bend the regulators so they lie flat against the board, to make them less vulnerable. This unit should be tested, to ensure the +5V and -5V supply rails are at the correct voltage.

Connect up a temporary (though not slipshod) mains supply. Care should be exercised, as 240V mains will be present on the underside of the transformer.

If all is well, the analogue board should be assembled next (Fig. 5). Again, assembly order is not critical as there is plenty of space on the board. Care should be taken when handling IC1 which is a CMOS device. This should be the last device to be inserted and an IC socket is essential.

The board can be partially tested at this stage by applying power and attaching a suitable signal (a signal-generator is by no means essential — a tape recorder is quite sufficient) to R23 and checking that this appears amplified at the output of the board and that the overload indicators light when they should do.

Lastly the digital board may be tackled (Fig. 6). Start with resistors, capacitors and diode, then insert the IC sockets, presets, links and finally the ICs themselves.

There are a number of links, including some on the underside of the board. There are five pairs of points which must be linked by insulated lengths of wire on the underside of the board. These are shown on the overlay as bracketed numbers. Link (i) to (i), (ii) to (ii) and so forth.

An extensive system of flying links is used above the board to carry power to the ICs. Wires are simply soldered at one end in the holes in the PCB and at the other to the bus-bars formed from tinned copper wire at the top of the board. Use suitably coloured wires to help avoid (possibly expensive) errors.

Testing
It is prudent to make a thorough check of this board before soldering in the IDC cable and to do all possible checking now, as errors are far easier to correct before the whole device is put in a case. After preliminary checking, link up the power-supply lines to
## Parts List

<table>
<thead>
<tr>
<th>Resistors (all 1/4W 5%)</th>
<th>Capacitors</th>
<th>Semiconductors</th>
</tr>
</thead>
<tbody>
<tr>
<td>R1-4, 101-104, 201-204, 301-304, 200k</td>
<td>C1, 101, 201, 301, 1μF polyester</td>
<td>IC1, IC2, 3, 5, 6</td>
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<tr>
<td>R5, 105, 205, 305, 120k</td>
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<td>IC4, ZN427E-8</td>
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<td>R9, 109, 209, 309, 12k</td>
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<td>IC8, 74LS04</td>
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<td>R19-22, 35, 36, 4k7</td>
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<td>IC13, 78M05</td>
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<td>IC15, BC184L</td>
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<td>Q1, BC214L</td>
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<td>Q2, 1N4148</td>
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<td>Q3, WO1</td>
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<td>R38, 150k</td>
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<td>Q4, Red LED</td>
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<td>R39, 82k</td>
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<td>Q5, Green LED</td>
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<tr>
<td>R40, 330R</td>
<td></td>
<td>Q6, Yellow LED</td>
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<tr>
<td>RV1, 47k</td>
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<td>PCB; case (300 x 150 x 90mm); wire; fuseholder; nuts and bolts</td>
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<td>6-0-6V 500mA</td>
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**Fig. 5** The component overlay for the analogue board
Fig. 6 The component overlay for the digital board

Software
A program in BBC Basic and assembler is given to allow basic operation of the unit (Listing 1). Those parameters which can be altered are displayed at the top right of the screen. The parameter for alteration is selected with the ↑ and ↓ keys and the value incremented or decremented with the ← and → keys.

The time base is given in μs/point, there being 256 points across the trace. The ‘Fast’ option allows the fastest possible sampling (11μs/point) to be achieved. The ‘Expand’ option allows the time-base to be halved, quartered and so on, with intermediate points being added between the genuine readings.

Remember to take into account the settings of these three options when calculating frequencies, periodic times and the like.

BUYLINES
The 74HC4351 (IC1) is stocked by Maplin (part no. UF140). Note that both 18 and 20-pin devices exist. The PCB is designed for the 20-pin version only.

The case used for the prototype is also as supplied by Maplin as part no. YMS5F.

The IDC cable used to connect the unit to the computer is most easily brought ready assembled. Maplin offer a service for making up such leads, for a small fee.

The PCB is as ever available from the PCB service.
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Clap hands here comes Chappell (Paul) with a sound-activated trigger for light shows on the stage or curtain-drawing in the home.

Back in the golden days when the world was young and electronics was fun, before designers became obsessed with creating ever more sophisticated multi IC circuits, a great deal of enjoyment was to be had from the simple pleasure of trying to squeeze the most spectacular performance from the smallest number of components. If your sensibilities have been jaded by auto-ranging multi functional megachips and rubbed raw from exposure to high-tech gadgetry, just call on J J Flash to bring a breath of fresh air into your life.

Rumour has it that Jack Flash was born when some friends known collectively as Filament started to achieve notoriety around the local pub and club circuit with their special brand of rock music. For a simple way to give impact to their stage presence, Flash had all the answers. A microphone in the bass drum would pick up the driving beat of the song and relay it to powerful lights which could be trained on any member of the band or even on the audience. For an outrageously OTT effect, a microphone and 100W lamp inside each drum would cause the drum itself to flash each time it was struck!

Of course there are those who would dispute this version of JJ's provenance. Many say that he was called into being as an aid to photography. Capturing a balloon about to burst and a bullet in flight are his finest achievements, they claim. Yet others insist that his main talent is in his versatility as a capable and powerful sound to action controller.

But what of the man himself? What did Flash claim to be? Perhaps we will never know the truth behind the legend but the story of his life may shed some light on one of the most intriguing mysteries of our time.

Jumping Jack Flash Is A GaAs

One matter which is undisputed is JFF's circuit. It is shown in Fig. 1.

To the left, a microphone picks up sound vibrations and converts them to an electrical signal. This is amplified by an amount set by RV1 and fed to a circuit which responds only when the peaks are above a pre-set level. A timer IC (the good old 555) is triggered by the peaks and gives a pulse of controllable length at its output. The pulse drives an opto isolated triac, which drives a power triac, which in its turn drives a lamp. All 500W of it!

The experience of operating the circuit in a small room is not to be missed — it gives a ridiculous sense of power to have every snap of your fingers accompanied by a blinding flash of light. Five hundred Watts in a living room is bright!

To pick up a hand clap from the opposite side of the room, it will have to be set somewhere near the maximum.

Construction

The component overlay for the PCB is shown in Fig. 2. There is nothing to cause any difficulties here.

The only point to watch is the mounting of the triac which must make good contact with the heatsink if the circuit is controlling a respectable amount of power. First lay it on the board with the hole in its tab over the mounting hole in the PCB. You can then judge exactly where to bend the leads. Bend them at right angles to the body of the triac (not too sharply, or you'll weaken them) then insert the triac leads temporarily into their holes to check that the tab hole is aligned with the PCB hole.

Take the triac away from the board, smear the tab with a little heatsink compound then bolt the triac to the heatsink and PCB before soldering the leads. A flat washer between the bolt and the triac tab will help to spread the pressure and to prevent the tab from becoming distorted.

Figure 3 shows a suitable power supply for the circuit. Fig. 4 is a suggested layout for the circuit in a small plastic hobby box. There is no need to use a PCB for the power supply unless you are really keen to make a neat job of it. The smoothing capacitors and transformer will support the rectifiers.

The microphone insert can be fixed to the case or can be remote at the end of a length of screened cable. If you splash out on a 3.5mm jack plug and socket, you can leave all your options open.

You'll notice the power supply has no regulator ICs. In the days
HOW IT WORKS

The signal from the microphone is amplified by IC1a.

Immunity to mains hum is achieved by the crude but effective expedient of giving C1 a low enough value to reduce the gain almost to unity at 50Hz while amplifying the fast spikes of percussive sounds anything up to about 2,000 times.

Components D1, R4, R5 and R3 hold pin 6 of IC1 slightly lower than pin 5 in the absence of a signal, keeping the output high. C2 tames the output of IC1b a little but no attempt is made to rectify or smooth the output since all IC2 requires is a signal which drops briefly below \( \frac{1}{3} V^+ \).

IC2 is the familiar 555 timer connected in monostable mode with time period adjusted by RV2. R6 limits the current to the opto isolator.

When connected as shown in Fig. 3, Q1 diverts the timer's output current from the isolator whenever any appreciable voltage is present across BR2, forming a simple zero crossing detector. The opto triac triggers the main triac via the surge limiting resistor R9. The main triac switches on the load.

when regulators had to be made from discrete components, they would only be used when absolutely necessary. These days they are included in every circuit almost as a reflex action. Just to be different (and since they are not necessary) I've left them out. Saves a quid, doesn't it?

When you come to test the circuit remember that everything beyond the opto coupler (IC3) is live, including the heatsink which is connected via the triac tab directly to the mains input. If you think anything may be wrong with the low voltage side of the circuit, disconnect the mains from the PCB (by removing the connections to points L and N) before checking it out.
Testing
The circuit really is too simple to need much explanation. The fault finding procedure is simply to check that the signal is reaching pin 1 of IC1 and pin 7 of IC1, triggering the timer IC2. The outputs from IC1 can be checked with a crystal earpiece if you haven’t got a scope and the output from pin 3 of IC2 can be checked with a meter on 25V range. It should normally be high, going low when a sharp sound is present. Tapping the microphone with the gain set to maximum should be enough.

If you find that the lamp remains on at all times when RV1 is advanced to the maximum sensitivity (even when the room is perfectly quiet) the value of R4 should be reduced slightly. Use the largest value that will allow the lamp to turn off reliably.

Other Applications
If your pulse doesn’t race at the thought of JJF’s electric stage performances, perhaps your mind is on photography. Sound triggers for cameras were all the rage a while back, allowing spectacular photographs of such things as

Fig. 3 Jack Flash in stage costume

Fig. 4 Suggested layout

Transistor Q1 and the connection to the J terminal form a zero crossing detector to cut down on the interference generated by SCR1. If this feature is not required, D2 and the link to terminal J can both be omitted.

Fig. 5(a) Circuit modifications for the sound trigger. SW1 is optional (see text). Q1 and R8 are not required, but there is no need to remove them. All parts of the circuit not shown are the same as for Fig. 1. (b) Connections to the PCB for photographic sound trigger
bursting balloons, popping balloons, exploding balloons, balloons going bang and so on.

**BUYLINES**

Most component suppliers will be able to provide suitable parts for this project. The only awkward device is the opto triac (IC3). If you use a type other than the IS608, it would be wise to increase the value of R9 just in case the surge current rating is lower. Most samples of the more readily available MOC3021 will work without modification to the circuit but if it does not turn on reliably this can be cured by reducing the value of R6 (not less than 1kΩ).

To get you started a parts set for the main circuit including PCB, triac, heatsink, pots and microphone is available from Specialist Semiconductors (see their ad in this issue). Power supply components are available separately from the same source, although your spares box will probably have suitable components in it.

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The photographs are made either by using the sound of the exploding balloon to trigger flash lights in a darkened room (the camera shutter would be left open throughout the procedure) or by triggering the camera itself if it was fitted with a solenoid operated shutter.

A touch of the soldering iron will transform JJF into a camera sound trigger. The way to do it is shown in Fig. 5. The power triac and heatsink are removed from the PCB and the opto triac is replaced by an ordinary opto-isolated transistor.

The only other difference is that the circuit is run from a pair of PP3 batteries instead of the mains power supply. You will probably want to put an on-off switch in the power supply as well.

The circuit can be armed by simply turning up the sensitivity. Otherwise, a switch in the 'reset' line of IC2 will give an arm/disarm function. You'll need to cut the PCB track for this.

For general sound switching purposes, the circuit of Fig. 1 can be used for mains applications and the modifications shown in Fig. 5 will make the circuit suitable for low voltage applications. The other feature you may want to add — latching — is shown in Fig. 6. With this configuration the circuit will latch when a sound triggers the switch and will remain on until the reset button is pressed. So — open your curtains with a whistle or your garage doors with a toot! Or just have fun experimenting.

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Having covered the design and construction of the Z80 CPU card last month, it is now time to move on and look at the main dynamic RAM cards. These provide the co-processor with a large area of storage which may be used for program or data. Each DRAM circuit is built on a standard 160 × 100mm eurocard sized double-sided PCB and consists of 256K of store.

Since the Z80 has an address range restricted to 64K, the DRAM card uses a paging mechanism to enable the Z80 to accommodate this increased address range. Connections to the board are once again made via a DIN41612 connector, the pinout being the same as for the CPU card described last month.

This DRAM card was also designed to be strictly general purpose and it can be used in any Z80 or Z80A based application you like — for example upgrading the available memory in a Z80 based home computer.

**DRAM**

Dynamic RAM has traditionally been tricky stuff to use. Although DRAM is cheap and can offer large storage capacities on small chips, it has the awkward requirements of periodic refresh and a multitude of carefully timed control signals. These have often been enough to turn the average hobbyist to the alternative expensive and bulky static RAM devices.

Early DRAM devices needed everything doing for them. All refresh strobes and addresses had to be externally generated. As generations of dynamic memory have come and gone, the external hardware overhead has been reduced. Now DRAM has in effect "grown-up." It is much more self sufficient and needs a lot less in the way of external prodding to facilitate refresh and data access. The "state of the art" DRAM as far as performance goes must be the AAA2800 series of CMOS 256Kbit devices, designed by Inmos and produced by NMB in the newest, most advanced CMOS VLSI manufacturing plant in Japan.

These memories combine low-power operation, fast access times and a number of versatile operating modes, making them the world's fastest monolithic 256K DRAM available.

One of the most interesting features included as far as the amateur is concerned is the so-called 'CAS before RAS' refresh mode which makes the process of refresh simplicity itself. So the timing signals required to drive the AAA2800 series are very straightforward. It is for these reasons that the AAA2800 family was chosen for use in this design.

One special feature of this circuit is that all the timing on the board is derived from the main CPU clock using synchronous logic techniques. So it is usable in any Z80 or Z80A based system regardless of the clock frequency. Also, expensive LSI DRAM controller chips and unreliable RC generated relays are completely avoided. The design is based on ordinary TTL logic and operates reliably under all conditions with no setting up required at all.
Refresh Schemes
The usual method used to achieve the obligatory DRAM refresh process is termed RAS-only refresh. This is shown in Fig. 1.

Since the refresh operation only requires that each row within the memory array is accessed, CAS is held inactive high and RAS is strobed whilst a suitable row refresh address is supplied to the memory chips. To ensure correct operation of the AAA2800, all 256 rows must be refreshed every 4.4ms.

You may be wondering why we cannot just use the built-in refresh mechanism provided by the Z80 microprocessor itself — the processor provides an incrementing row address for the DRAM after every instruction fetch (while it is internally decoding the new instruction and the system buses are otherwise unused).

This would be great but for the fact the Z80 dates back to the days when sub-64K DRAM devices were the best available, and so only a 7-bit refresh address is generated, whereas we need an 8-bit address. This problem can be overcome by software but that introduces an unacceptable time overhead.

The usual solution is an external refresh address generator but using the AAA2800 we have another more elegant option.

The AAA2800 provides a CAS before RAS refresh mode. During normal DRAM accesses, RAS is strobed before CAS to latch-in the required address. If this order is reversed, the AAA2800 ignores the address present on its pins and uses an internal 8-bit counter as a source of row addresses, performing a refresh operation all on its own.

The trick is to strobe CAS before RAS (Fig. 2) every time the Z80 authorises a refresh cycle and the result is a completely transparent refresh scheme with no wasted time overhead and no external refresh logic required.

Paging
In common with most 8-bit microprocessors, the Z80 is capable of directly addressing only 64K of memory. However, this limited address range can be extended by paging. The large block of storage to be addressed is partitioned into areas called pages, each 64K locations or less in size.

The processor addresses the contents of a page directly, because its address range can handle the entire page in one go. However, the actual page accessed is selected, one at a time, by some other means.

The usual method of page selection is via a bank select register. This is a multi-bit latch which holds the currently selected page number and can be updated by the processor to change pages. The memory hardware responds to the contents of the bank register by enabling only the memory page required for access on the Z80 buses.

In a Z80 based system, this register can be addressed as an I/O port so it is conveniently separate from the memory map — thus avoiding the possibility of being able to page itself out!

The disadvantages of simple paging schemes like this revolve around the fact that the memory space is no longer randomly accessible.

The addressing of a location is now a two-stage process, requiring an update of the bank register as well as the memory access itself. This can prove awkward if the program being executed is stored in the paged memory when jumps and subroutine calls may entail page changes.

Unless the machine's operating system has some kind of
monitoring capability to handle such situations, the simplest way round this problem is to provide a small area of permanently enabled memory which is accessible regardless of the current page selected.

When code execution requires a change of page, a routine residing in this unpaged memory is called to change page and then to jump to the required location in the new page. This is the scheme used in this design. A memory map PROM (IC19) is used to permanently disable a section of each page in thepaged memory, to allow the unpaged memory to sit undisturbed.

This unpaged memory consists of the operating system EPROM and the scratchpad RAM on the CPU card, described last month. The PROM allows unpaged program areas to be selected in steps of 256 locations so the least possible DRAM space is wasted in any given application.

For this co-processor design, the bottom 5K locations of each 64K page are disabled by the PROM but this can be easily changed for other applications by blowing different map PROM data.

Construction

The construction of the RAM card should not create too many problems, particularly if the suggested PCB layout is used. Like the Z80 CPU card described last month, this board is double sided but for reasons of expense, through-plating is not used to interconnect the two layers.

The through connections are shown in the overlay diagram (Fig. 4) and made by either wire links through the board, by the passive component leads, or by means of the IC pins. The various techniques for employing these methods were described last month. Once again, 'turned pin' DIL sockets should be used to allow through connections via the IC pins.

As high currents are likely to flow in the memory power and ground lines, multiple through connections are provided to reduce the impedance of these links. These take the form of two or three closely situated through links. All must be made to ensure reliable operation.

This board also includes a large number of supply decoupling capacitors and it is important not to omit any of these. Each RAM chip has its own associated capacitor, connected close to the power and ground pins. These capacitors must be long leaded axial types to fit the suggested PCB foil and these ensure low supply impedances at high frequencies where noise problems might otherwise exist.

Short M2.5 nuts and bolts should be used to fix the edge connector to the PCB for added strength. Note that two of the connector pins must also be soldered on the topside of the PCB to link through the board.

At this stage you must decide whether to program the card for operation as pages 0-3 or pages 4-7. This determines what you do with the link pads on the PCB.

If you want the card to appear as pages 4-7, point B on the circuit diagram must be connected to point C with two through connections at pads Y and Z. The page indication LEDs are connected to point C on the circuit diagram by soldering a through link at pad X.

If the card is to appear as pages 4-7, point B on the circuit diagram must be connected to point C with two through connections at pads Y and Z. The page indication LEDs are connected to point C on the circuit diagram by soldering a through link at pad X.

The address mapping PROM IC19 must be fitted to the board before power is applied for the first time. The How It Works section explains the PROM programming. See Buylines for details of pre-programmed PROMs with the correct specification.

Testing

Testing the completed memory card is not easy at this stage. If you are intending to build up the whole co-processor project, the procedure for testing all the hardware together will be given next month. If, however, the DRAM card is intended for your own particular application then testing must consist of connecting the card into your target system and powering it up, not forgetting to initialise the bank select registers.

If data can be reliably stored and retrieved, your DRAM card is fine. If there are problems, first check that you have soldered everything in, particularly on the topside of the board, then ensure all the on-board control signals are active and that addresses and data are reaching the DRAM chips.

Like all microprocessor based projects fault finding is extremely difficult without the right gear, and a decent logic analyser will set you back five grand or so. Having said that, a little perseverance with a scope should pay dividends.

Next month we shall look at the final bit of hardware to complete the Spectrum co-processor design — the interface card through which the co-processor can communicate with the Spectrum. We shall also start to examine the software side of things, so don’t miss the next exciting episode!
HOW IT WORKS

The heart of the circuit (Fig. 3) is of course the eight dynamic RAM devices, IC22-29. The rest of the circuit can be considered as two distinct areas around this DRAM — the paging/address logic and the circuits which manage the memory chips by generating the required control signals.

Looking first at the paging/address logic, this enables the memory card for access when both a valid page on the card is selected and a valid location within that page is addressed.

The bank select register, consisting of three D-type flip-flops (IC14 and IC33b) is mapped onto the Z80 output port at address 254 by IC20 and IC31a, b, d. The data value sent by the Z80 to this port will determine which page is chosen.

IC18 decodes the port data to see whether one of the four pages on this card is selected. IC33b stores the result, its Q output being low if the current page is on the card. At the same time, IC14 stores the states of the two least significant port data bits.

If the required page does reside on this card then these two bits signify which page from the four is selected and will form two extra address bits to the DRAM devices.

IC21 is used as a two-to-four line binary decoder to light up one of four LEDs to indicate which page is currently selected. This decoder is disabled if the current page is not present on the card so all the LEDs are blanked.

For circuit simplicity inverse logic is used for page selection. Sending 255 to the bank register switches in page zero, 254 switches in page one and so on.

The user has the option of making the card either pages 0-3 or pages 4-7. This is achieved by means of a link on the PCB. If point A is linked to point C, the output of IC18 will be active low when the most significant six bits of data to the bank register port are high. So, pages 0, 1, 2, 3 are decoded onto the card. If point B is linked to point C then port bit D2 is inverted before it gets to IC18 and pages 4, 5, 6, 7 are decoded onto the card.

In this way, up to eight pages of DRAM (two DRAM boards) may be used in the system. The 8-bit port decoding however, allows up to 256 memory pages to be controlled from the processor and if more pages are required then further data bits must be inverted on their way into IC18. This would of course entail modifications to the PCB design itself.

IC19 is an address decode PROM used to disable the memory card when certain addresses occur, allowing other unpaged memory (such as EPROM) to reside in the system.

Its address inputs are connected to the CPU's eight most significant address lines and so the PROM changes its addressed location every 256 Z80 locations.

The PROM is programmed to produce a logic low at its least significant bit output (pin 12) when the card is to be disabled. For example, if the first 1K locations are to be reserved for an unpaged EPROM, IC19 must be programmed with the first four locations containing zeros in the LS bit position leaving the rest of the device holding all ones.

For this design, where the first 5K locations must be disabled, IC19 must have its first 20 locations programmed to E(hex) and the following 236 locations programmed to F(hex).

The rest of the circuitry manages the memory devices themselves allowing...
A further signal is required to drive the address multiplexers themselves, selecting whether the row or column addresses are being supplied to the DRAM.

In this design the RAS signal is directly obtained by using the Z80 memory access strobe MREQ via buffer IC32c. The DRAM write control signal W is the Z80 write strobe WR buffered by IC32b.

The generation of the DRAM CAS and the multiplexer select signal, is a little more complicated. Their precise timing has to tie in with RAS and W.

D-type flip-flop IC33a is activated at high speed by the system clock and on each rising clock edge passes the state of MREQ to its output. The relative timing of MREQ with the clock is such that the Q output of IC33a follows MREQ with a delay of about 50ns minimum (assuming a 4MHz clock).

This delayed signal is used to drive the select line of the memory address multiplexer formed by IC15, 16 and 30. The 50ns switchover delay after RAS ensures the row address hold time is not invalidated.

The address multiplexed into the DRAM consists of 16 bits of address from the Z80 and two bits of address from the bank select register, IC14. This results in the required configuration of four 64K pages with each page being directly accessible to the CPU.

The Q output of IC33a goes on to form CAS via IC34b, c and IC35b, c. The DRAM chips used in this design (in common with most dynamic memories) require for normal access that CAS goes low after RAS and that the W strobe has settled before CAS goes low.

The delayed RAS signal from IC33a is guaranteed to occur after RAS itself. Similarly IC35a detects when the Z80 RD or WR strobes are active. When this happens it signifies that the DRAM W control is also present. The output of IC35c will go low only when both the delayed RAS and the output of IC35a are active, thus fulfilling the timing requirements for CAS in relation to RAS and W.

The output of IC35c is given a little extra delay by IC34b, c to ensure the recently multiplexed column address has had time to settle. This signal then passes via IC35b and IC34d to drive the DRAM CAS line.

IC33a can be disabled via its asynchronous preset input by IC32a, thus holding CAS inactive when a memory access is not required on this card.

This occurs when either the current page is not on this card (IC33b Q is low), when an invalid address has been received (detected by IC19) or when a refresh operation is taking place (Z80 RFSH line low).

When the Z80 authorises a refresh cycle to occur, it drives the RFSH line low before the MREQ (RAS) strobe. This RFSH strobe forces CAS low via IC33b to accomplish the automatic CAS before RAS refresh cycle. The result is that refresh occurs trans-
**PROJECT: Co-processor RAM**

**HOW IT WORKS**

CONT.

Parently to the user during the time the Z80 is internally decoding its next instruction.

Data from the Z80 data bus is buffered onto this card by IC17, an octal tranciever chip. The direction control for this device is simply derived from the Z80 RD strobe via IC35 and its enable control is the same signal as the multiplexer select — this being active low during the time that data is moving into or out of the card.

Since each of the AAA2800 devices used is configured as a 256K by one bit memory, each chip handles one bit of the Z80 data word and the address information is fed to all the DRAM chips in parallel.

Numerous supply decoupling capacitors are used throughout the circuit (especially around the RAM chips themselves) to avoid problems with high frequency supply noise. A high value electrolytic capacitor C15 acts as a bulk energy store to reduce problems associated with remotely connected power supplies.

**PARTS LIST**

**RESISTORS** (all 0.25W 5% unless otherwise stated)

- R13-18 1k0
- R19 220R 0.5W 5%

**CAPACITORS**

- C15 100µ 6V ultramin electrolytic radial
- C16-32 100n ceramic

**SEMICONDUCTORS**

- IC14, 33 74LS74
- IC15, 16, 30 74LS157
- IC17 74LS245
- IC18, 20 74LS330
- IC19 24S10 PROM
- IC21 74LS138
- IC22-29 AAA2800
- IC31 74LS111
- IC32 74LS11
- IC34 74LS14
- IC35 74LS100
- LED1-4 green LEDs

**MISCELLANEOUS**

- SK2 DIN41612 type C 64-pin right angle PCB mounting plug
- PCB; IC sockets; tinned copper wire for through connections; M2.5 nuts and bolts for DIN connector.

Fig. 4 The component overlay for the RAM board

ETI MARCH 1988
## IBM COMPATIBLES

- IBM 5151, 2762, 6710
- IBM 5152, 2763, 5171
- IBM PC/XT, 5170, 386, 680
- IBM AT, 386, 680
- IBM PS/2, 386, 680
- IBM PS/2 Model M, 386, 680
- IBM PS/2 Model 1, 386, 680
- IBM PS/2 Model 2, 386, 680
- IBM PS/2 Model 3, 386, 680
- IBM PS/2 Model 4, 386, 680

## CASED/UNCASED FLOPPY DISC DRIVES

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## DISK DRIVE ACCESSORIES

- Diskettes: £3.50<br>- Double Sided Doulble Density: £2.00
- 3.5" CFDD: £5.00
- 3.5" Drive: £10.00
- 3.5" Drive: £25.00

## M.COLOUR MONITORS

- Philips 8833 Colour Monitor: £225.00

## SPECIAL OFFER

- 512K DDS MITSUBISHI DISC DRIVE: £75.00
- 512K IBM MITSUISHI DISC DRIVE: £30.00

## DISK DISC DRIVES

- IBM 5151, 2762, 6710
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- IBM PC/XT, 5170, 386, 680
- IBM AT, 386, 680
- IBM PS/2, 386, 680
- IBM PS/2 Model M, 386, 680
- IBM PS/2 Model 1, 386, 680
- IBM PS/2 Model 2, 386, 680
- IBM PS/2 Model 3, 386, 680
- IBM PS/2 Model 4, 386, 680

## DIL SWITCHES

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## ELECTRO-MECHANICAL, PCB AND CABLE ASSEMBLIES

- IBM Mark II, IBM Mark III, IBM Mark IV, IBM Mark V, IBM Mark VI
- IBM 5151, 2762, 6710
- IBM 5152, 2763, 5171
- IBM PC/XT, 5170, 386, 680
- IBM AT, 386, 680
- IBM PS/2, 386, 680
- IBM PS/2 Model M, 386, 680
- IBM PS/2 Model 1, 386, 680
- IBM PS/2 Model 2, 386, 680
- IBM PS/2 Model 3, 386, 680
- IBM PS/2 Model 4, 386, 680

## ENVIROMENTAL

- IBM Mark II, IBM Mark III, IBM Mark IV, IBM Mark V, IBM Mark VI
- IBM 5151, 2762, 6710
- IBM 5152, 2763, 5171
- IBM PC/XT, 5170, 386, 680
- IBM AT, 386, 680
- IBM PS/2, 386, 680
- IBM PS/2 Model M, 386, 680
- IBM PS/2 Model 1, 386, 680
- IBM PS/2 Model 2, 386, 680
- IBM PS/2 Model 3, 386, 680
- IBM PS/2 Model 4, 386, 680

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Door Chimes

G Landry
South Africa

This circuit is not just another two tone doorbell but is envelope modulated to give a realistic 'ding-dong' chime sound. It also has the battery saving feature of consuming no power in standby mode.

Pressing SW1 turns on Q1 to power the rest of the circuit. IC1a is a low frequency oscillator providing clock pulses to the decade counter IC2. Q2 is switched on when the counter passes zero and maintains the power for a complete cycle.

The fast attack, slow decay envelope is formed by D4, C3 and R8 at the first count. Q4 modulates the output of audio oscillator IC1d which is fed to amplifier Q5 and the loudspeaker.

At count 5 from the counter, another envelope is started from D5, C3 and R8. The carry out pin goes low and C4 is introduced in parallel with C5 to produce the lower 'dong' sound.

At count 9, the counter is reset, Q1,2 are turned off via IC1b and the circuit is ready for another ring.

Motorcycle Regulator

D Noble
Kyle

This circuit was devised to provide regulation for the lighting circuit added to an off-road motorcycle without lights.

The principle is straightforward. Any extra energy from the bike's generator is dumped. The 741 acts as a comparator. The output is low when the supply voltage is less than the desired value. When the supply voltage is greater than the desired output, the output of the 741 goes high and the transistors are switched on, taking current from the generator and reducing the output voltage.

The switch on voltage is set by RV1. A bridge rectifier was chosen to provide greater power at low revs (tickover).

The transistors are mounted on a large heatsink and the whole circuit coated with a paint to keep out water.

Battery Regulator

G Landry
South Africa

This voltage regulator circuit provides a 6V output from a 9V battery input and can maintain the output with an input as low as 6.5V. Up to 180mA can be supplied.

On application of power, Q2 switches on via R2. Q3 drives Q2 to ensure the voltage across ZD1 and the base-emitter voltages of Q2 and Q3 are constant with varying supply voltages and output currents.

The output voltage can be trimmed between 5.7V and 6.7V with RV1 as the zener voltage is dependent on the current through it.

Q1 is switched on and lights LED1 to indicate that regulated output is available. If the supply voltage drops below 6.5V then Q1 and Q3 turn off, the LED goes out and Q2 is driven hard on connecting the output to the supply with minimum power dissipated in the idle regulator.
Printer Switch

M Pill
Nazeing

This design offers a convenient way of enabling two different printers or a printer and a plotter to be used with a computer which has only one parallel printer port. It uses tri-state buffers to switch the signals from the computer to the printer currently required.

Which printer is active at any one time is determined by the toggle switch connected to the 'enable' inputs of the buffers. With the switch in the position shown in the diagram Printer 1 will be active, as the buffer control pin is pulled low by the switch, activating the buffer outputs. The buffers feeding Printer 2 will conversely be open circuit due to the action of R2. The printer control lines are treated in a similar fashion to the data lines.

An added advantage of using buffers in this way is the driving capability they add to the printer port. The printer ports on most computers tend to be unbuffered. This can lead to corruption of data when long cables are used.

The circuit will need a 5V supply which in most instances can be obtained from the computer and fed down an unused wire in the cable.

The circuit could be expanded to drive more than two printers or even reversed to allow two computers to share one printer.

Soft Turn-On Dimmer

F Choy
Singapore

High current surges through the low resistance of a lamp's cold filament when power is first applied. This can be fatal to the lamp, particularly if switched-on at or near the voltage peaks.

Such a disaster, however, can be avoided by incorporating soft turn-on, that is, to let a lamp be dully lit for a few initial moments, pre-heating its filament. This circuit provides a warm-up period (determined by R7 and C2) of about one second during which analogue switch IC1b opens leaving the wiper of RV1 unconnected. The timing resistance from Q3 is therefore the full resistance of RV1, corresponding to minimum lamp intensity while C2 charges via R7.

After one second, C2 is charged up to the threshold of IC1b closing it for as long as power remains on. Now the wiper of RV1 is effectively shorted to its upper end resulting in a conventional phase control lamp dimmer. IC1a is switched on by the rectified mains through D1 so the potentials on each side of switch IC1b are equal.

When power is removed, R6 discharges C2 via D6 in preparation for a new warm-up period. The rest of the circuit provides a 12V supply and synchronisation for Q3.

ETI
PLAYBACK

In many areas of electronics good design ideas are thought up and sometimes patented years before they are really practicable. Maybe an effort is made to apply the idea commercially for a while, and the unsuccessful attempt deters further development, so that it becomes feasible. The idea of class D audio power amplifiers may well fall into this category.

A class D audio amplifier works somewhat like a series switch, switched mode voltage regulator. Of course, the output of the class D amplifier is AC instead of DC at any point, so it is not possible to optimise the design for a specific output voltage but many of the ideas are the same. At this point consideration which bears a striking resemblance to a class D audio amplifier is a switched mode uninterruptible power supply. In this system, a very close mark/space ratio is maintained and the switching signal is generated and filtered to remove the switching frequency leaving a 50Hz sine wave on the output of the supply.

The same idea can be applied to audio signals. The problem is more difficult because a wide frequency range must be accommodated. Also, a distortion level acceptable on a power supply would be far too much for an audio signal.

If the problems were solved, the resulting design would have two great advantages over currently available amplifiers. The power dissipation would be very low and there would be no crossover distortion. Why? Because there is no crossover region.

At zero output the switching stage would provide a square wave at the switching frequency, and the mark/space ratio of this square wave would alter in time with signal excursions. There is no discontinuity at any point. I admit the same is true of class A amplifiers but for any substantial power rating the heat generated is a problem and the noise of the cooling fan can spoil the music.

Duplication

Recently, over a pint at the local, a friend who works for a consultancy asked my views on high frequency audio amplifier design. Apparently they had been asked to design an audio distribution amplifier for use with audio cassette duplication equipment.

The basic specification may interest or shock audiophiles, particularly those who advocate amplifier bandwidths exceeding that of a bat, or distortion figures of around 1E-6 per cent.

Apparently the duplication equipment has to work at 128 times normal running speed in order to make the process economic.

The bandwidth specified for the distribution amplifier was 400Hz to 1.6MHz, which corresponds to a single speed audio bandwidth of 3.125Hz to 12500Hz.

Low Response

I do not regard this as impressive at the high frequency end, but the low frequency performance is anomalous. This low frequency response is needed to record the low frequency cue tones used on some tapes.

The basic amplifier was required to drive ten 50R coaxial cables, terminated at both the sending end and the receiving end. This attenuates the signal to half its original amplitude, so a corresponding gain is required in the amplifier.

The normal operating signal level was specified as 4V peak to peak, with an overload margin of 26dB. On this level, this corresponds to a maximum signal of 4V peak to peak, which represents 80W into the maximum combined load of 10R.

Specification

This specification may well be partly the result of a specification writer's nightmare but assuming it is realistic I would imagine that an amplifier designed to meet it would use a quasi-complementary output stage with hunky RF power transistors.

How likely is such a device to exhibit distortion figures of 0.01% or less, I wonder?

Here is just another example of how the source material is mistreated before it reaches the punter.

If this type of signal handling is rife in the music publishing industry then it would seem to make that last fractional improvements in hi-fi equipment rather academic.

Andrew Armstrong

OPEN CHANNEL

A couple of months ago (January ET1) Open Channel reported on the second generation of cordless phones, code named CT2, currently in development. CT2 cordless phones are digital and will operate at similar frequencies to current cellular radio phones (around 900MHz).

Then I said the first CT2 phones will be similar to existing cordless phones.

I now prophesise that future CT2 cordless phones will be developed as a single base station with a number of portable handsets — rather like a private internal Private Branch Exchanges without the wires.

It appears, however, that another use for CT2 cordless phones is just around the corner. Telereport in which base stations are located in public places. Anyone having a portable handset within a short range of the base station can make their exchanges without the wires.

Typical public places would be train stations and shopping centres and the base station areas would be known as Phone Zones. Users won't need cash to make a call as they will be billed direct to their home addresses.

This new telephone service (if it gets off the ground) is to be operated by Ferranti and has been developed by a company called Libera, although Libera is in fact part-owned by Ferranti.

Libera has reported that operators in France have already been found and the company is currently recruiting operators in other European countries.

But When?

Operation is hoped for by the middle of the year but this depends on appropriate licensing by the Department of Trade and Industry.

It appears that licensing of systems installed in Phone Zones places will be more easily obtained than licensing required for system use, say, in the street.

One big problem in the DTI's system licensing procedure is the fact that little or no standards have yet been produced for the proposed system — basically because its such a new concept.

I would hope this should present few problems to the DTI as the system appears to have found market-places both at home and abroad, long before any foreign competition has even thought of a similar product.

This move by Ferranti into telephone operations is very interesting, particularly when viewed with two other happenings. First, the company's recent merger with American International Signal and Control. ISC is a large supplier of defence equipment and services, so any experience Ferranti gains from telephone based operations will be a big plus in the defence market.

The second notable happening is Plessey's buy-out of Ferranti's semiconductor division. Plessey is reported to have paid £300m for this and will now be by far the largest chip producer in the United Kingdom. As Plessey concerned this can only have advantages in terms of economy and pricing.

Six months ago Ferranti was basing a large chip manufacturer and finding life very difficult. Today it is a systems supplier. It will be interesting to watch Ferranti's fortunes in the near future.

No Competition

Plessey, of course, has recently merged its telecommunications division with that of GEC to prevent self-destructive competition.

Rumours abound that Ericsson too, wants to get in on the act to make a European telecommunications force which will further reduce competition.

If such a merger came off, there would be little to prevent the combined companies from upping prices considerably.

Although this would suit Plessey, GEC and Ericsson, it would not be agreeable to companies (like British Telecom) which as a matter of principle always prefer to foster competition in equipment procurement by inviting tenders from as many individual suppliers as possible.

Given the truth of such rumours Plessey/GEC should take care not to give too much thought to a merger with Ericsson. BT and other customers are not going to change procurement policies just for their benefit. Non-European suppliers will be asked to tender and competition may then be even fiercer.

Keith Brindley
Rubbish! I am referring to the deluge of inaccurate advertising and parrot-fashion media reporting that digital audio technology drags in its turbulent wake.

While not being in quite the same league as relative quanters. Almost all the subleties of analogue/digital conversion have certainly proved sufficiently complex to cause wholesale confusion. Many CD players sound better than bad CD players, despite proving sufficiently complex to achieve a respectable 86dB against conversion speed with- over a number of clock cycles.

So is the fact that Sony use the Burr-Brown convertors in their own professional recording products.

Most of the inaccuracies of convertors are attributable to thermal resistor lattices in the strays capacitor and ground loops. A follow-up to this suggests that error in precision converter circuits is actually the stability of the clock. A timing jitter of one part in 65,536 can correspond to the loss of as much as one bit in precision in a 16-bit system. A well designed crystal oscillator is more stable than this but what about glitches and variable loading effects in the intermediate stages of the digital supply rails? This source of error applies equally to DACs and ADCs.

Peek and 12-bit ADCs are definitely not yet firing on all cylinders. Almost all ADCs in current audio use are based on the successive approximation (SA) technique, in which a DAC and voltage comparator placed within a digital feedback loop home in on the input voltage over a number of clock cycles. A 16-bit SA ADC must perform 16 internal A/D conversions per sample and so cannot be expected to be as accurate as a simple DAC because higher speed generally results in lower precision.

Most 12 and 16-bit ADCs used by the audio industry are Burr-Brown SA types, notably the 16-bit PCM75. This £30 device is quoted as having a dynamic range of 90dB when used in conjunction with a perfect sample—hold (which Burr-Brown are unfortunately not able to provide). With a precision discrete component error-canceling sample—hold and with a lot of careful testing and loving attention, the PCM75 can reach down as far as 88dB — on a lab bench, when the wind is blowing in the right direction.

In the harsher environment of a commercial production line a figure of 80dB is more typical.

SA is not the only viable approach to A/D conversion but other approaches currently offer little or no advantage. For example, Sony’s 16-bit CX20018 dual ADC employs a dual-ramp integration conversion process to achieve a respectable 82dB accuracy according to the data sheet but what gives the game away is that Sony use the Burr-Brown convertors in their own professional recording products.

Given that precision audio DACs are at best rather suspect, it is worth considering next the thermal resistor lattices, the strays capacitor and ground loops. A follow-up to this suggests that error in precision converter circuits is actually the stability of the clock. A timing jitter of one part in 65,536 can correspond to the loss of as much as one bit in precision in a 16-bit system.

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SA is not the only viable approach to A/D conversion but other approaches currently offer little or no advantage. For example, Sony’s 16-bit CX20018 dual ADC employs a dual-ramp integration conversion process to achieve a respectable 82dB accuracy according to the data sheet but what gives the game away is that Sony use the Burr-Brown convertors in their own professional recording products.

Given that precision audio DACs are at best rather suspect, it is worth considering next the thermal resistor lattices, the strays capacitor and ground loops. A follow-up to this suggests that error in precision converter circuits is actually the stability of the clock. A timing jitter of one part in 65,536 can correspond to the loss of as much as one bit in precision in a 16-bit system.

A well designed crystal oscillator is more stable than this but what about glitches and variable loading effects in the intermediate stages of the digital supply rails? This source of error applies equally to DACs and ADCs.

Peek and 12-bit ADCs are definitely not yet firing on all cylinders. Almost all ADCs in current audio use are based on the successive approximation (SA) technique, in which a DAC and voltage comparator placed within a digital feedback loop home in on the input voltage over a number of clock cycles. A 16-bit SA ADC must perform 16 internal A/D conversions per sample and so cannot be expected to be as accurate as a simple DAC because higher speed generally results in lower precision.

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Telephone Alarm (July 1987)
In the component overlay (Fig.2) IC1 and IC2 should be swapped. The capacitor to the right of IC1.2 is C1 and the inductor between them is L1. The unmarked resistor to the left of L1 should be a wire link. In the circuit diagram (Fig.1b) IC4.a,b should be AND gates. IC5 should be NAND gates. The parts list is correct.

Kappellmeisters (July 1987)
The position of the speaker port in the front panel was omitted from Fig.2. This should be a 7%4\%in ellipse centred across the panel with its top edge 2\%in below the panel top.

Knight Raider (August 1987)
In Fig.1(a) pins 4 and 5 of IC1 are swapped. IC2-3 show the correct pin-out.

Car Alarm (August 1987)
In Fig. 1 Q7 is not numbered and its emitter is shown unconnected. This connects to earth. The transistors in the parts list went a little awry. Q2-6 are BC232 and Q7 is a TIP31.

Boiler Controller (September 1987)
In Fig. 2 (a) the primary of T2 is shown connected to Earth. This should be neutral. In Fig. 2(b) one of the bridge rectifier diodes, D6-9, is shown the wrong way around. This is correctly shown in Fig. 5.

EEG Monitor (September 1987)
In Fig.3a the pins of IC1 connected to the power rails are shown swapped around. In Fig.4a R7 is unlabelled and is between C3 and C6. In Fig.5 C20 should be £10 and R18 is unlabelled. It lies between R17 and R19.

ETI Concept (October 1987)
The Power Board parts list wrongly lists R6 as 270R. This should be 270k. Also, note that the power board's OV rail must not be connected to Earth or the OV rail of the CPU board.

Printer Buffer (November 1987)
The software for the EPROM had three errors listed. The byte at 039A should read 20, at 039B 14 and at 0492 30. All numbers are in Hex.

Dream Machine (December 1987)
The transistors used in this project are ST1702. BC108s can be substituted.

RGB Auto-Dissolve (January 1988)
In Fig. 5 there are marked two D6's. The right hand one should be D5 (they are both 1N4148's anyway). In the text the reference to zener diode D5 should read zener diode ZD1.

PASSIVE INFRA-RED ALARM (January 1988)
Fig. 2(a) shows the base of Q1 connected to ground and to R14. It should be connected only to R14.
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The April issue sees the start of the hi-fi Virtuoso power amp project to complement the Virtuoso pre-amp of last year. The Spectrum Co-processor is in its third installment and we have the start of a new series of easy projects for the beginner to electronics or the expert short of spare time.

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