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Supercapacitors

Design small loop antennas

Word generator for digital test

Refresh old valves

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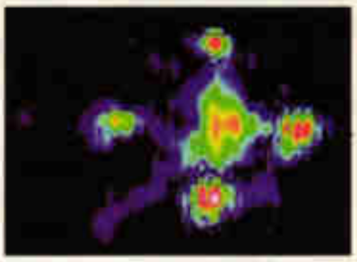

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REED BUSINESS INFORMATION

Enabling technology?

I first heard about microprocessors on a BBC radio broadcast in 1976 when I was an undergraduate. They were new, they were revolutionary and they were going to advance electronics to unprecedented levels. Or so I, along with many others, believed at the time. If there were wiser, cautionary voices around I wasn't aware of them.

I wanted to develop a microprocessor-based communication aid for severely disabled cerebral palsied patients. I had been shown round a school for spastic children, and had been horrified by the impoverished technology that was being used to help the children's communication problems.

In the world of medical research that I knew, we had access to all the latest, best, first-class equipment. By contrast, the unfortunate children were forced to use the Possum – or Patient Operated Selector Mechanism. Despite the euphonious acronym, this was an ugly device, based on electromechanical relays. It slowly, and laboriously, spelled out words one letter at a time onto an electric typewriter.

Possum looked like something out of Noah's Ark. Surely a microprocessor could do much better – very much better – and banish the old Possum forever. So later that year I took time out of my medical studies and set to work to put this idea into practice.

At the time, I was not aware of engineers' actual experiences with microprocessors. In fact, it was said that Intel made more money out of selling development systems in the early days, than out of shifting silicon! For most engineers, there was a prolonged learning curve merely to get anything out of their systems at all. I get the impression that many early microprocessor projects quietly failed.

True to form, my project also ran into trouble. And the steep learning curve wasn't the only problem. Try as I might, my efforts at technological revolution were just coming up with another Possum! Improved, it must be said, but not a radically different principle or solution.

The same is still happening. Today's 'advanced' communication aids for disabled people are still laborious to use, unsightly, and fail miserably to replace normal speech. Disabled people resent their normal faculties being bypassed, having to do everything in a different way from normal. To compound this, the more elaborate aids serve to reinforce the idea that the people using them are 'different'.

My first efforts were foxed and I diverted my attention from the problems of seriously and multiply handicapped people. I did build an experimental speech recognition system around my six-bit microprocessor system to demonstrate what the technology was good for, but I knew that this was not a serious answer to the original problem.

Now, twenty years on, I have much more experience with microelectronics. I have lived with disability myself and I am still pondering the same problems. But I am, I hope, somewhat wiser.

First of all, I had not correctly diagnosed the problem. There's a saying in medicine, 'diagnosis before treatment'. It was not just the obviously antiquated technology – or indeed the pedestrian performance – of the Possum that was the problem.

Basically, Possum was an aspirational product. It was about dreams. People hoped to find in it the brave new world of futuristic bliss, of technological miracle, of a science fiction future. One that our contemporary Western culture leads us all to expect.

Possum was a 'technology fix'. Problem was, the dream wasn't coming true. And even today's more advanced microprocessor technology doesn't deliver the dream. This is what was fooling me in 1976, and I think still fools people today.

There is something relevant here that is not specific to my design problem, but applies widely in modern life. It is easy enough to ask, "What do we really want?" But no technology answers that question for us.

The microprocessor is not the panacea I once naively imagined it to be. Originally, it was intended as a cost-aggressive approach to low-performance, high-volume production – essentially for making things cheap and cheerful.

Disabled people have very real problems that rarely lend themselves to the cheap and cheerful category. Does that mean that their problems should not be addressed, or that one should cheat?

So it is that a higher integrity, more honest, more thoughtful, more workmanlike, more whole-hearted approach than has been current is what I think disabled people – and indeed all of us – deserve from electronics in the future.

In my own work I want to hold to these ideals, and I dream that these reflections inspire someone to provide a real answer to the old Possum. ■

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UPDATE

Telecoms operators press chip makers on VDSL standardisation

Telecoms operators are pressing network equipment suppliers and chip makers to speed up moves to standard products for next generation broadband-to-the-home technology, known as VDSL.

BT, France Telecom and two US regional Bell operating companies – SBC and US West – have teamed up to accelerate the creation of an industry standard for VDSL systems, which will support video to the home services.

The move coincides with finalising the drafting of standards for VDSL, which are being sent for approval by European standard body ETSI, US standards body ANSI and world standards body ITU.

"The main motivation for the network operators is that they are starting to feel, more and more, the pressure from cable operators, as they are being forced to go into the same market as the cable operators – namely video," according to Thierry Pollet, Alcatel's VDSL standards boss.

The standards bodies are being asked to approve two approaches to VDSL – or very high speed digital subscriber line – technology. One is the multi-carrier approach of Alcatel, Ericsson, Nortel, STMicroelectronics and Texas Instruments, known as the VDSL Alliance.

The rival approach is the single carrier approach of Infineon Technologies and Broadcom, known as the VDSL Coalition. Lucent Technologies may support both standards.

"Operators would like to see both standards because they would like to explore both technologies," said Pollet. However it is thought the ITU is looking to adopt a single standard.

Meanwhile, BT's partner in the new grouping, France Telecom, is said to have ordered 100 000 lines of multi-carrier VDSL exchange equipment.

Alcatel's Paul Spruyt said: "We are in negotiation with France Telecom for DMT [discrete multitone modulation, that is multi-carrier] VDSL equipment."

Internet-ready fridge... A fridge fitted with a screen, aptly called the Screenfridge, is about to be tested in 50 Danish homes. As well as being a food storage facility the fridge will offer interactive broadband communication technology via a touch screen and also act as a TV and radio receiver. The five month trial is being run by TeleDanmark and e2 Home, a joint venture between Electrolux and Ericsson.

Flash memory stores two bits per cell using four voltage levels

Flash memory storing two bits per cell has been introduced by STMicroelectronics.

Storing two bits of data per memory cell, or four voltage levels, requires more complicated sensing circuitry. ST believes this is more than compensated for by the increased storage density.

Made using a 0.18μm process, the 64Mbit device is a NOR-type flash device, hence reasonable read and write speeds, making it suitable for storing both data and program code.

Data is erased in 1Mbit blocks, taking about a second, each of the 64 blocks being independent.

The memory array operates from a 2.7 to 3.6V supply, while a separate lower voltage powers the i/o.

Operating current is 30mA, in standby it is 50μA and in the automatic deep power down mode it is as low as 1μA, ST claims.

As a result, the device is aimed at hand-held applications, such as interactive PDAs, and also set-top boxes, Internet appliances, digital cameras and solid-state flash cards.

Higher density chips with 128 and 256Mbit capacities will also be available in the future, the firm said.

Mobile phone signals via aircraft comms

A system to let normal mobile phones be used on a plane anywhere, at any altitude, has been launched by BAE Systems. The Cabincall system uses the aircraft's existing communication system to carry calls. Signals are sent by the aircraft's transmitter, which connects via a satellite network to the ground. According to BAE, the cabin staff retain control of the system and can inhibit mobile phone calls during critical flight phases. The system is expected to enter service in autumn 2001 after in-flight demonstrations this year.

28GHz auction to go ahead

Government ignores calls for review of spectrum auction process

The Government has ignored industry calls to change its policy of auctioning chunks of the radio spectrum to the highest bidders.

The Federation of the Electronics Industry (FEI) has led calls for a review of the spectrum auction process, which it says is driving up costs for new operators and may not be the most efficient means of allocating radio spectrum.

Last week, the Department of Trade & Industry effectively snubbed the FEI and started its next auction by inviting applications from companies wishing to take part in the auction of 28GHz frequencies.

Could soon be a common sight... 28GHz wireless access antenna from Radiant Networks.

The auction of the 28GHz band, to be used for broadband fixed access wireless services, is due to go under the hammer in October.

The Government's decision to press ahead with an auction of the wireless fixed access licences, despite criticism following the highly priced 3G auction in May, demonstrates its commitment to a policy of raising revenues from frequency auctions.

"We have decided on an auction process as it provides a fast, transparent, fair and economically efficient way of allocating the scarce resource for radio spectrum," said Kim Howells, minister for competition and consumer affairs.

"The levels bid will reflect the bidders' own valuations of the licences," he said – a clear response to recent criticism of the eventual prices paid the the UK's five 3G mobile communications licences, which netted the government £24bn.

However, not all European government's share the DTI's preference for frequency auctions. Last week the French government announced it would award its 3G licences by fixed sealed bids – the so-called 'beauty-contest' approach.

- There will be three 28GHz fixed wireless access licences available in each of 11 English licence areas and in Scotland, Wales and Northern Ireland. Each licence will have forward and return channels of 112MHz. Licences will last 15 years. Bidders will only be allowed to hold one licence per area, but can hold a licence in any number of licence areas.

With a fixed price tag of $4bn per licence, this is seen as a way of keeping prices down.

The worry is that if operators are forced to pay too large sums for the licence they will need to pass on some of that expense to customers when the services go live or risk financial problems.

Simon Wilson, director of radio at the FEI, said: "We don't believe auctions are the most effective way of making use of spectrum in terms of innovation, impeding new entrants and distorting competition, if it takes money out of the supply chain through high costs."

Lithium batteries deliver power pulses ten years on

Need a 1A pulse in ten years time? Sonnenschein may be able to help. The company is producing long-life batteries developed particularly for delivering high current pulses at short notice.

Power comes from a lithium-thionyl-chloride bobbin-type battery. These have a very low self-discharge rate – typically one per cent a year, said Sonnenschein – and proven shelf and operating life of more than ten years.

Unfortunately for power hungry pulsed loads, the passivation process that gives them long life also makes them slow to wake up and awful at delivering high currents.

PULSE BATTERIES AT A GLANCE

Nominal capacity to 2.0V	1.2 Ah @ 1mA	1.65 Ah @ 2mA	2.5 Ah @ 2mA	8.5 Ah @ 4mA	19 Ah @ 5mA
Capacity to 3.0V 250mA, 1% pulses	1.2Ah	1.65Ah	2.5Ah	8.5Ah	19 Ah
Operating life 10yr @ average of	190µA	10µA	14µA	22µA	80µA
Height [mm]	46.6	55	71.5	69.2	80
Diameter [mm]	16.6	16.6	16.6	27.7	34.5
Volume [cm³]	10.1	11.9	15.5	41.7	74.8
Weight [g]	23	25	33	75	126
Models TLPxxxxx/A	96311	97311	91311	92311	93311

To get over this, the company has combined a lithium cell, made by Tadiran, which also developed this idea, with a low-leakage hybrid layer capacitor – all in a standard battery profile.

The capacitor charges slowly from the high impedance lithium cell. It can deliver, said Sonnenschein, "pulses up to several amps in the range of 50 to 3200A per pulse with a voltage window between 3.9 and 3.0V".

Proposed applications for these batteries, called Pulses Plus by Sonnenschein, include emergency call systems, portable defibrillators, GPS tracking devices and oceanographic transponders.

To meet varying requirements, there is a range of Pulses Plus's from AA-size (46.6mm long by 16.6mm round) with a nominal 1.2Ah capacity, to a 80 by 34.5mm cylindrical cell of 19Ah. In addition to its pulse capability, this last cell is rated to supply 190µA continuously for ten years.

Key to digital TV is choice of channels

The key incentive for taking up digital TV is the greater choice of channels offered according to an Oftel survey. Almost one in five households now subscribe to digital TV but fewer than one in five of these use the interactive services such as home shopping and e-mail. One third of digital TV subscribers signed up without having previously received analogue cable or satellite services.

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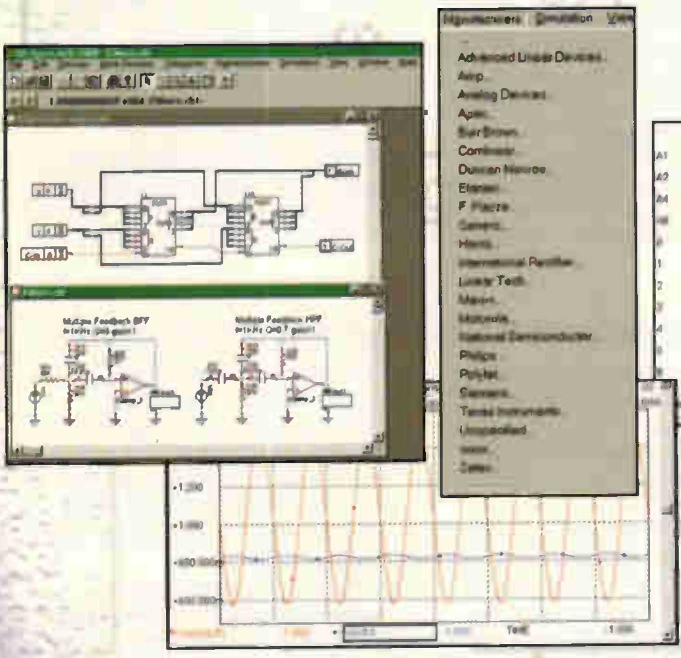

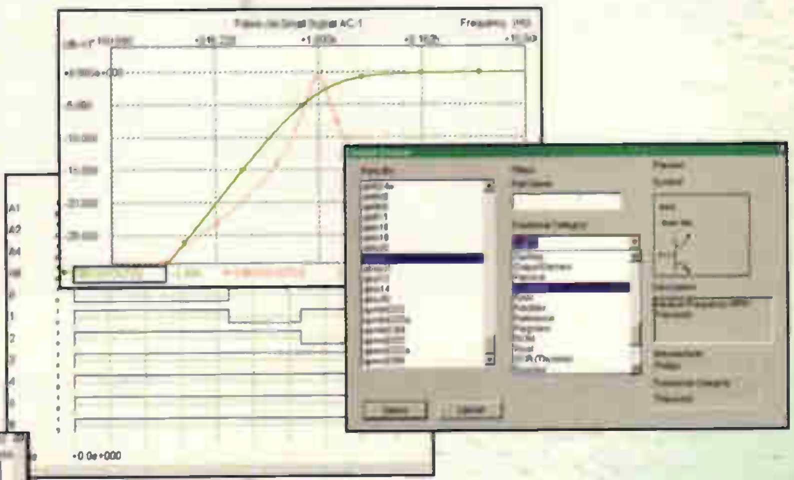

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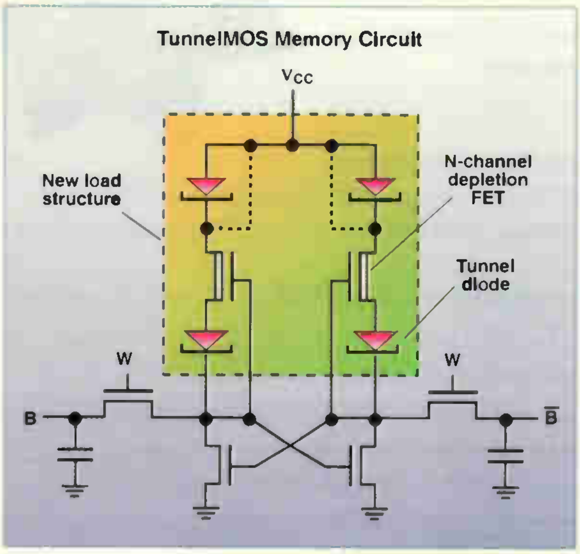
CIRCLE NO. 109 ON REPLY CARD

New static RAM cell is smaller *and* faster

In the new static RAM cell, n-channel FETs and tunnel diodes replace inferior p-channel devices

An alternative architecture for SRAM has been patented by National Scientific. Using purely n-channel transistors results in a memory cell 30 per cent smaller than conventional six-transistor SRAM.

Read and write speeds are improved, the latter by 50 per cent, while active power drops 45 per cent, the firm claims.



The company has devised a scheme for replacing the p-channel transistors used in standard SRAM with better performing n-channel devices and tunnel diodes.

Called TunnelMOS, the load portion of the cell is made from n-channel depletion FETs and tunnel diodes in series. The FET's gate connects to the diode's cathode.

Tunnel diodes were once touted as the saviour of electronics, and were used as storage elements. But their unstable nature meant they were unreliable for memory – especially after multiple operations and at changing temperatures.

TunnelMOS exploits the nature of the diodes – in particular their ability to switch extremely fast from one operating region to another.

When one of the flip-flop transistors in the cell is asked to switch on, V_{cc} is placed across the load. As current starts to rise, the voltage increases across the lower, larger, tunnel diode, reaching its peak voltage. Through tunnelling, it snaps to a higher voltage, making V_{gs} of the load transistor –0.7V and turning it off.

At the same time, the opposite happens to the other half of the circuit.

The smaller tunnel diode at the drain of the load transistor is included to reduce the gain of the transistor, lowering the chance of oscillation setting in.

Tunnel diodes allow the use of small n-channel depletion mode FETs, but improve their power consumption and switching speeds.

So far the only perceived disadvantage of TunnelMOS is it requires two extra mask steps in manufacturing. Also the load transistors seem to be on, if only slightly, in both states of the memory cell, which could lead to higher standby current.

National Scientific claims to be in the process of fabricating the memory cell and discussing licensing with larger memory manufacturers.

Richard Ball

Existing SRAM cells

Static RAM based on the cross-coupled transistor layout requires some form of load to maintain its memory. This load can be resistors, p-channel FETs or depletion mode n-channel devices.

Resistors need to be large in value to reduce standby currents and keep the speed high. Large in value equates to large in area, which makes the cell too big for most applications.

A better alternative is to use p-channel FETs in the classic six-transistor layout. While they overcome most of the resistor's problems, p-channel devices have a large die area, because of the need to diffuse an n-well into the substrate.

Using an n-channel depletion mode FET reduces area, but they have similar properties to resistors, causing high standby power and large current spikes when switching.

Strong pound blamed for stagnant pay levels

The continuing high level of sterling and "difficult trading conditions," are the reasons for stagnant pay levels in the UK's engineering industry, according to the Engineering Employers Federation (EEF).

"Pay settlements in the engineering industry are still at an historically low level," said a federation statement.

The EEF compiles a monthly report on wages, each covering the previous three months.

In its August issue it concluded that, in the three months to the end of July 2000, the average settlement level was 2.5 per cent – a slight decrease on the reported figure for the three months to the end of June 2000. This breaks down into averages of 2.5 per cent for May, 2.8 in June and 2.4 in July

Of the 212 settlements studied over the three months, 77 per cent were for three per cent or less, while 95 per cent were for four per cent or less. There were 29 pay freezes.

The August report covers settlements affecting 26 713 employers.

Tories promise to tackle high-tech brain drain

A future Tory government would tackle what it calls the high-tech "brain drain" and review BT's market power as part of a drive to make Britain the new technology capital of the world.

Opposition leader William Hague made the promises in the Conservative Party's draft election manifesto. He maintained that the country was well placed to capitalise on its lead in new economy technologies and promised to simplify the regulations for IT industries so that they could be left alone to grow by government.

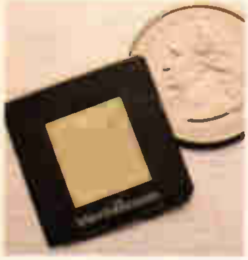
Measures would include reforming the so-called "brain drain" tax on contract engineers and IT consultants known as IR35 which the Tories claim has deprived Britain of some of its brightest, most productive workers.

They also seem set to target telecoms operator BT and are proposing to ask the Competition Commission to review BT's market power in local Internet connections as part of an examination of reducing industry costs and boosting consumer access to the Internet.

"We set out a comprehensive package of specific policies to help IT industries flourish in our country - including reforming damaging brain drain taxes, radically deregulating the telecoms industry and encouraging more competition in local Internet connections," said Hague.

In another topical vote catching move he also promised that the proceeds from future windfall sales of radio and mobile phone spectrums would be used provide endowment funds for Britain's top universities.

Fingerprint authentication is just around the corner



Silicon fingerprint sensing will be fairly common in mobile phones and PDAs in 18 months time, according to Veridicom.

"In 18 months, fingerprint authentication in mobile devices will be fairly common," said Naeem Zafar, Veridicom's v-p of marketing, following the launch of his company's latest fingerprint sensing chip.

With a profile of 1.4mm, the stamp-sized, 500dpi FPS200 silicon sensor is nearly half the thickness of Veridicom's previous chips, making the technology suitable for authenticating e-commerce transactions in ultra-thin laptop computers, PDAs, and mobile phones.

Veridicom's sensor technology is based on an array of capacitive plates, which sense ridges when a finger is placed on the surface. Sample-and-hold circuits capture the fingerprint images.

"The different capacitances are translated into a grey-scale image. We then look for unique minutia – bifurcations or ridge endings – and store the relationships between the significant points in a template that occupies about 300bytes," Zafar explained.

An important new addition on the FPS200 is auto-finger detect. This feature cuts standby power to below 20μA by interrupting the CPU only when a finger is placed on the sensor rather than, as in previous designs, constantly polling.

To ensure good images from all types of skin in a wide range of climactic conditions, the chip also introduces ImageSeek. This new function has a programmable gain control and feedback mechanism that takes several images of the finger in a fraction of a second and selects the best while changing the capacitive array bias levels.

According to Zafar, the FPS200 reduces the cost of implementing fingerprint authentication by nearly 40 per cent by integrating more peripherals on-chip. These include automatic gain control circuitry and three interface communications modes.

The on-chip USB core interface provides fingerprint image output up to 13 frames per second and the microprocessor control unit raises this to 30 frames per second. There is a peripheral interface mode for non-PC applications.

"Because the Internet enables stranger-to-stranger transactions, you need a way to know who's on the other end of the line. The element of personal trust, which is missing on the Internet, is the problem we're addressing," concluded Zafar.

US researchers demonstrate molecular level RAM storage mechanism

Molecule man... Fraser Stoddart, left, the person behind olympadene – the molecule that mimics the shape of the Olympic symbol, with James Heath.

A California-based team claims to have demonstrated the first resettable bistable molecular switch that works in solid state at room temperature.

"From here, molecular RAM may well be just a matter of time," said James Heath, professor of chemistry at the University of California, Los Angeles. "While there are many pitfalls between the demonstration of a technology and the actual invention, at the moment we can't see how any of those pitfalls could prove fatal."

The work is based on molecules developed by Scottish researcher Fraser Stoddart, now professor of organic chemistry at UCLA.

"These can be repeatedly switched on and off over reasonably long periods of time in a solid-state device under normal laboratory conditions," said Heath. "For the first time, we are able to turn the molecular switches on and off repeatedly."

Stoddart has developed bistable molecules before, particularly the dumbbell-and-ring-shaped rotaxane. The new ones are called catenanes.

"Where molecules were previously swimming around incoherently in solution, now we see coherent molecular motion in a solid-state device," he said. "The molecules have the ability to recognise one another and line up in an efficient manner."

Catenanes self-assemble to form a pair of interlocked rings. In catenanes containing the right two rings, one ring can be stimulated to move between two different states with respect to the other reference ring, giving it bistability.

Switching can be induced by taking away and giving back an electron.

The team is working on several other types of molecule, at least one of which looks like improving on the catenane's switching performance.

Within a few years, Heath thinks the research team will develop circuits that have molecular logic, molecular memory and nanometre-sized wires.

A hybrid computer that interfaces molecular memory with silicon logic is only a few years away, he said. "The demonstration of a nano-scale computer that is largely molecular – with molecular logic and molecular memory – will likely happen within the decade," concludes Heath. ■

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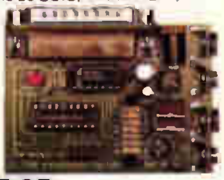

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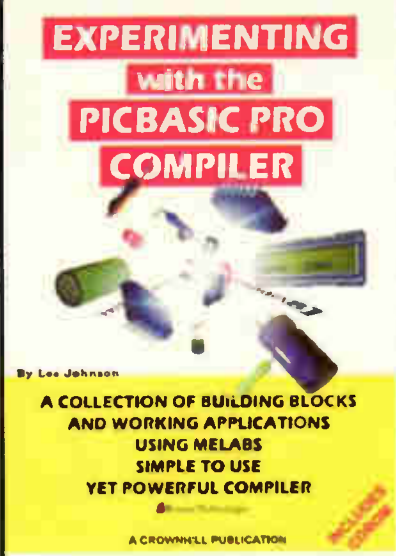

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Super capacitors

In recent years, the word 'supercapacitor' has taken on a new meaning.

Originally, it was used to describe a frequency-dependent negative resistance, of the type used in high-performance active filters. I described such circuits in an article that I wrote some years ago[1] but the term was first used well before that, in the late sixties[2].

More recently, the term has been given a new meaning. 'Supercapacitor', in this more recent sense, refers to a component that behaves like an ordinary capacitor, but exhibits a quite extraordinarily large capacitance for its size.

The traditional capacitor

An 'ordinary' capacitor uses a non-conducting medium, called a dielectric, to separate two conducting plates. The larger the plates and the smaller the gap between them, the higher the capacitance.

The dielectric serves two purposes. Firstly, it supports and separates the plates. Secondly, it increases the capacitance from the lower value it would have if the two plates could somehow be supported and separated by air, in the same configuration.

In a capacitor with a gap instead of a dielectric, the capacitance C in farads would be $\varepsilon_0 A/d$. Here, A is the area of the plates in square metres, d the separation in metres and ε_0 is called the permittivity of free space – i.e. a vacuum.

Permittivity ε_0 has a value of 8.85×10^{-12} farads/metre. In fact, the capacity would be 0.0536% higher than this, as ε_r, the permittivity of (dry) air relative to free space is 1.000536.

Dielectrics used in high-quality capacitors tend to have an ε_r larger than unity by a modest but useful amount. In the case of polystyrene for example, it is 2.6.

By using a dielectric with an even higher permittivity, a larger capacitance can be obtained in a smaller volume. Various types of rutile ceramic, for example, have an ε_r running into hundreds or thousands, permitting disc ceramic capacitors to be made in values up to 100nF or more. Ceramic capacitors in the larger sizes typically are somewhat 'lossy' – that is, they have a poor Q – but they are widely used for decoupling purposes at both audio and radio-frequency applications.

'Electrolytics'

Where even higher values of capacitance are needed, electrolytic capacitors can supply the answer. Unfortunately, the name is a bit of a misnomer, as, in use, electrolysis is the very last thing you need.

In the process of manufacture, a 'forming voltage' is applied between the aluminium foils, which are separated by a porous membrane impregnated with electrolyte. A current causing electrolysis flows. This results in the coating of the positive or anode foil or plate with a thin non-conducting layer of aluminium oxide.

After this process of 'anodisation', no more current can flow, or at least, only a very small residual leakage current. Thus electrolytic capacitors are 'polarised'. When the are used, they must be connected the right way round.

Electrolytic devices are widely used as 'reservoir capacitors' in power supplies, and for coupling and decoupling purposes at audio frequencies. Even though the relative permittivity of aluminium oxide is only around 10, the extreme thinness of the oxide dielectric results in a component with a very high capacitance – the conducting electrolyte itself can be considered part of the other plate.

Non-polarised or 'reversible' electrolytics consist of two capacitors in series in the same case. As they are connected negative plates together, whatever polarity is applied, one is always the right way round, preventing reverse polarisation of the other.

More on how exactly the basic capacitor works can be found in the panel entitled 'How a capacitor works'.

On capacitors and batteries

A 'supercapacitor' is basically a half-

Recently, the term 'super capacitor' has reappeared in an entirely different context. The new 'supercapacitor' is a discrete component that behaves like a capacitor, but has vastly greater capacity – so much so that it can be used as a back-up battery. Ian Hickman examines how they work, and what they can be used for.

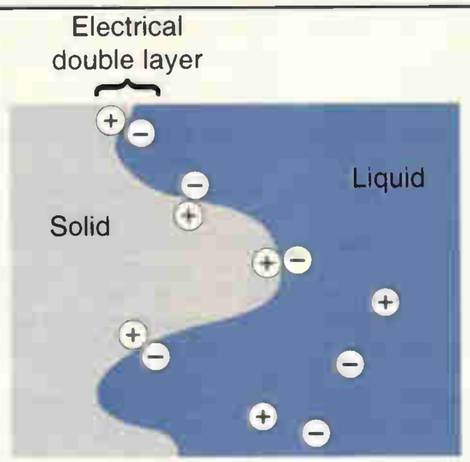

Fig. 1. Showing how opposite charges appear at the boundary interface between a conducting solid and an electrolyte. Reproduced from the Tokin Supercapacitors User Manual.

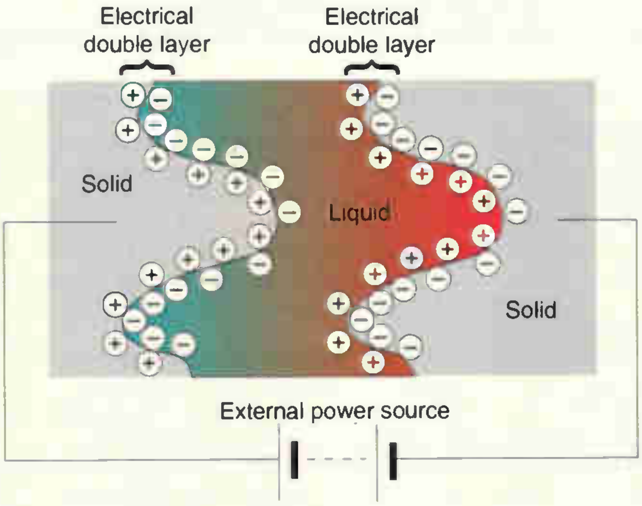

Fig. 2. Diagrammatic representation of the workings of a double layer supercapacitor. Reproduced from the Tokin Supercapacitors User Manual.

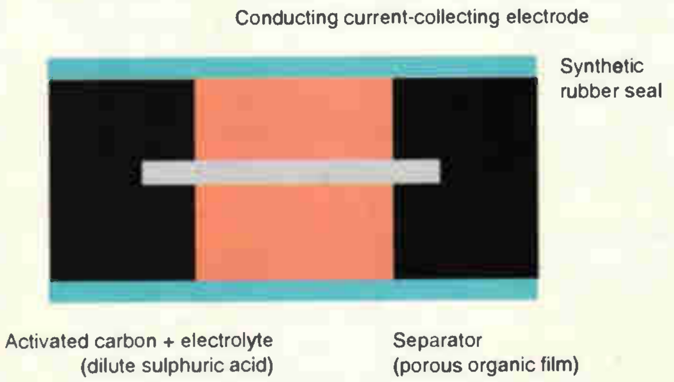

Fig. 3. Showing basic construction of a supercapacitor, using two identical activated carbon electrodes. Reproduced from the Tokin Supercapacitors User Manual.

way house between a conventional capacitor and a secondary – i.e. rechargeable – cell.

As illustrated in the panel, the electrical charges locked up in the atoms of a conventional dielectric can 'give' a little under the action of the electric field between the plates of the capacitor. But they cannot actually move through the dielectric; they are anchored in place by their parent atoms in the solid dielectric.

In a battery, such as a lead acid accumulator, the dielectric is replaced by an electrolyte. The active constituents of this are ionised, and the positively and negatively-charged ions will, during charging, be attracted to the negative and positive plates respectively. There, they react with the material of the plate to form different compounds, storing electrical energy in the process.

These chemical reactions are reversible, so that when a conducting path, a load, is connected between the plates, current flows and some of the stored energy is supplied to the load.

Different secondary cells use different electrolytes, plate materials and chemical reactions, resulting in different cell voltages. Thus the common nickel-cadmium, or NiCd, rechargeable cell has a fully charged voltage of 1.2V, compared with just over 2V for a lead acid cell.

A battery of cells is used where a higher supply voltage is required, e.g. the 12V battery in a motor car, or the 48V central office battery of a telephone exchange.

To charge a secondary cell, the applied charging voltage must marginally exceed the basic cell voltage, in order to make the chemical reactions happen. If the applied voltage is less than this, although the ions may migrate through the electrolyte under its action, the chemical reactions will not occur.

Thus there is a voltage barrier, in some ways similar to the band-gap in a semiconductor junction. But note that the ions *can* move, unlike the charges associated with the molecules of the dielectric in a conventional capacitor. This mass mobility of charge acts as a dielectric with an enormous relative permittivity: an admittedly analogical view of things: a more specific explanation follows.

The double-layer capacitor's principles

The supercapacitor is also known as a 'double layer capacitor', for reasons that will become apparent.

Various makes are available, in various styles, designed for various applications. The ones to be described below are designed for use as memory back-up 'batteries', though being capacitors, they are rated not by their capacity measured in mAH, but by their capacitance measured in farads.

Various values are available, from 0.01F (10000μF) up to several farads, or even 100F, depending upon the manufacturer.

When a conducting solid and a liquid electrolyte come into contact, positive and negative charges distribute themselves in opposition to each other, in a boundary layer at the interface, i.e. at the surface of the solid, **Figure 1**. This is called an 'electrical double layer'.

In one make of supercapacitor, the solid phase is highly porous activated carbon, while the liquid phase is dilute sulphuric acid. **Figure 2** shows, again purely diagrammatically, an external voltage applied to a capacitor with two identical such carbon electrodes sharing the same liquid.

Experiments using a test electrode of mercury indicate a resultant capacity of around 20 to 40μF per square centimetre. One gram of activated carbon may well have an effective surface area of 1 000 square metres, and hence should be able to provide a capacitance of 200 to 400F/gram.

In practice, figures not far short of this are obtained in production. **Figure 3** shows the basic construction of a supercapacitor, using two identical activated carbon electrodes.

Note that while a very high capacitance can be achieved, it is not permissible in principle to apply a voltage higher than the decomposition voltage of the electrolyte. If you do, electrolysis will occur, with the evolution of gases.

Double-layer capacitor construction

The limited permissible voltage requires practical supercapacitors to be a structure of several or many individual cells, like that of Fig. 3, in series.

Figure 4 is the cross section of a practical device, showing the arrangement diagrammatically. The separator is a porous inert organic film. It prevents a short circuit between the activated carbon powder positive and negative electrodes, while allowing ions of the sulphuric acid electrolyte to pass.

The breakdown voltage of each of the basic cells forming the complete capacitor depends upon the electrolysis volt-

How a capacitor works

Figure a) shows an uncharged capacitor with a solid dielectric. The atoms of the dielectric are electrically neutral, having a positively charged nucleus surrounded by a cloud of circulating negative electrons.

In **Fig. b)**, a voltage has been applied between the plates of the capacitor, with the result that the electrons are attracted slightly, in their orbits, towards the positive plate. This movement of charge results in the capacitance being larger than if the space between the plates were a vacuum.

If the *permittivity* – in my younger days called 'dielectric constant', and by my then elders 'specific inductive capacity' – of the dielectric is 10, then nine parts of the current 'through' the capacitor are due to the presence of the dielectric. The rest would have been there anyway – even with a vacuum between the plates.

Current through a vacuum? But how can a current flow through a vacuum? No one has satisfactorily answered the question, but Kirchoff's first law assures us that it must do.

If current flows through a wire to the plate of a capacitor to charge it up, an equal current must flow away from the plate, i.e. through a perfect vacuum if need be. We now know that electrons pile up on one plate, while the other plate accumulates an equal dearth of electrons, but that was after Kirchoff's time.

Furthermore, Maxwell's equations tell us just how much the current flowing through the vacuum will be. The current through the empty space between the plates equals the plate area times the current density per unit area. This is the 'displacement current' – an item that most electronic engineers learn about on their degree course, and then promptly forget about for the rest of their lives.

We tend to remember that magnetic flux density B in webers/m^2, is related to magnetic field strength H in amperes/m by $B=\mu_0\times\mu_r\times H$, but don't forget that electric flux density D in coulombs/m^2 is related to electric field strength E volts/m by $D=\varepsilon_0\times\varepsilon_r\times E$.

These equations apply not only in the static DC state, but also for a time-varying AC field strengths – electric or magnetic. In the AC case, since D varies with time, then the coulombs/m^2 value varies with time and a time-varying charge is a current.

Maxwell's equations express this in the general three-dimensional case, but $dD/dt=\varepsilon_0\times\varepsilon_r\times dE/dt$ covers the bill in the simple one dimensional case in the narrow confines of the space between the plates of a capacitor.

The rate of change of electric flux density, i.e. the current density in the dielectric – be it a vacuum or otherwise – is represented by dD/dt. As it is proportional to the rate of change of the electric field strength E with time, it is zero in the DC case, when the capacitor is sitting there quietly, fully charged, or instantaneously in the AC case, at the peak of a sine wave.

As the displacement current is at a maximum when E is changing most rapidly, at zero volts, the current 'through' a capacitor is inherently 90° out of phase with the voltage.

Whether you believe in the reality of displacement current is up to you. A year or two ago I was working on an L-band project, namely an EPIRB. I mulled over the problems of a two-band antenna with the manager in charge of the development – a knowledgeable and experienced engineer. He vowed he'd never heard of displacement current, and doubted if any such thing existed.

Certainly there must be displacement currents in the air surrounding a half wave dipole, to complete the path of the current flowing into the antenna from the feeder, if Kirchoff's first law is to hold. But in calculating the resultant field at a point remote from the antenna, one takes into account only the current flowing on the elements of the antenna.

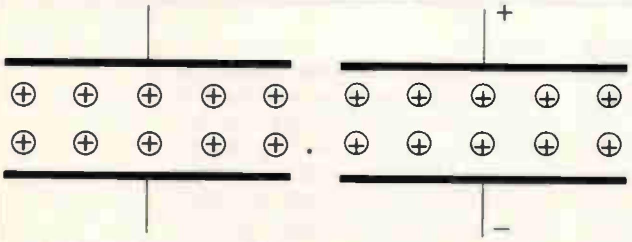

Fig. a) An uncharged capacitor with a solid dielectric.

Fig. b) When a voltage is applied between the plates of the capacitor, the electrons in the atoms of the dielectric are attracted slightly towards the positive plate.

age of the electrolyte. In turn, the electrolysis voltage depends upon the electrolyte chosen and also upon its concentration.

In the case of dilute sulphuric acid, the electrolysis voltage is about 1.2V. Thus for a memory back-up capacitor rated at 5.5V, a stack of five or more basic cells is required.

Due to the tortuous nature of the current paths, in among the passage ways of the activated carbon, this type of supercapacitor exhibits a comparatively high internal resistance or 'ESR', which is short for equivalent series resistance. This is in the range from several hundred milliohms to 100Ω, depending upon the capacitance, type and construction.

Because of their high internal resistance, such supercapacitors cannot be used as reservoir capacitors for ripple smoothing in power supplies. However they can be used in parallel, where a higher capacitance is required than is available in a single component. They can also be used in series where a higher voltage rating is required, provided care is taken to ensure equal distribution of the applied voltage between the supercapacitors.

Characteristics

Figure 5 shows typical change of capacitance with temperature, with the temperature cycle regime shown. Clearly there is a positive temperature coefficient of capacitance, but of quite manageable proportions.

You will also notice small changes in ESR, both with temperature variations, and as a function of frequency. Like both electrolytic capacitors and secondary batteries, supercapacitors exhibit a self-discharge characteristic.

Figure 6 shows the self-discharge characteristics for 5.5V memory back-up capacitors with various values ranging from 0.047F to 1F. It is of course a decreasing exponential curve. It looks rather different here, because the *x* axis time scale is logarithmic.

Note that like electrolytic capacitors, when left indefinitely to charge up to an applied dc voltage, the charging current drops to a very low but finite value, defined as the leakage current.

Supercapacitors are supplied with the polarity marked, just like electrolytic capacitors. This is because in the manufacturing process, the device is subjected to an applied voltage in the indicated polarity. Bear in mind that, as supplied, a small amount of charge may remain.

Supercapacitors also exhibit the 'soakage' effect, whereby even though short circuited, when the short is remove, a small residual voltage may reappear.

However, as is clear from the diagrams, the construction is basically symmetrical. No long term harm will come to the device if used in the wrong polarity. In the short term, if used back to front, the self discharge characteristic will be poor, but if subjected to inverse charging in excess of 100hours, the device is effectively reprogrammed to the new polarity. It will then show the same self discharge characteristic as originally.

Many makers, many types

My ancient copy of Technical Indexes' Electronic Engineering Index does not cover 'supercapacitors', so I don't have a list of all the various makes available. But I do know that besides Tokin, they are available from Panasonic – called 'electric double layer capacitors' – and Elna, which calls them 'Dynacap electric double layer capacitors'. I believe that NEC and Korea Chemicon also make supercapacitors.

Another type of double layer capacitor, from Epcos, has been designed specifically with very low ESR in mind. Its ESR is so low in fact that one suggested use is for energy recovery by regenerative braking in electric vehicles – an application for which lead-acid batteries are unsuited, due to their inability to accept large amounts of charge in seconds.

Finally, my thanks go to Semicom UK Ltd, UK agents for Tokin Supercapacitors, for permission to use certain material supplied by them. ■

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1. Hickman, Ian, 'Filters using negative resistance.' *Electronics World*, March 1997, pp. 217-221.
2. Bruton, L.T., 'Network transfer functions using the concept of frequency dependent negative resistance.' (1969) *IEEE Transactions on Circuit Theory*, Vol. CT-16, August, pp. 405 - 408.

Fig. 4 Cross-section of a Tokin supercapacitor.

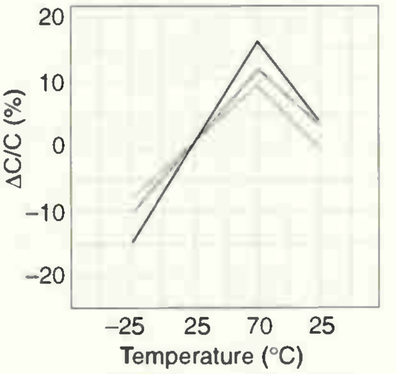
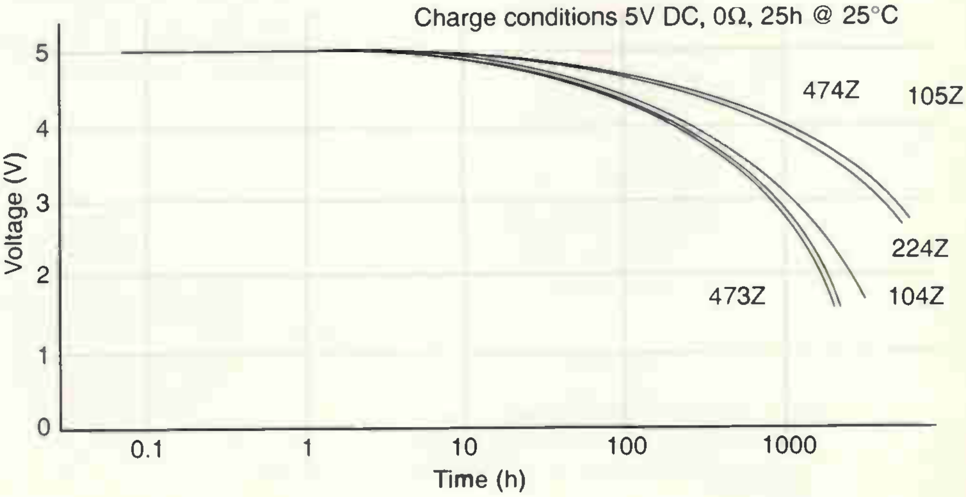


Fig. 5. Showing the typical change of capacitance with temperature, with the temperature cycle regime shown. Reproduced from the Tokin Supercapacitors User Manual.

Tokin's half-farad 5.5V supercapacitors, left.

Fig. 6. Below, Showing the self-discharge characteristics for various value 5.5V memory back-up capacitors ranging from 0.047F to 1F. Reproduced from the Tokin Supercapacitors User Manual.



Small-loop receiving antennas

Small-loop receiving antennas are invaluable for receiving difficult signals anywhere up to VHF. Their strong directivity also makes them ideal for locating transmitters. Joe Carr explains how easy it is to develop your own.

Radio direction finders, and people who listen to the AM broadcasting bands, VLF, medium-wave or the so-called low-frequency 'tropical bands' are all candidates for a small-loop antenna. These antennas are fundamentally different from the large loops and other sorts of antennas used in these bands.

Large loop antennas have a length of at least 0.5λ and most are quite a bit larger than 0.5λ. Small-loop antennas, on the other hand, have an overall length that is less than 0.22λ, with most being less than 0.10λ.

The small-loop antenna responds to the magnetic field component of the electromagnetic wave, instead of the electrical field component. One principal difference between the large loop and the small loop is found when examining the radio frequency currents induced in a loop when a signal intercepts it.

In a large loop, the current varies from one point in the conductor to another with voltage varying out of phase with the current. In the small-loop antenna, the current is the same throughout the entire loop.

Big loop, little loop

The differences between small loops and large loops show up in some interesting ways. Perhaps the most striking is the directions of maximum response – the main lobes – and the directions of the nulls.

Both types of loop produce figure-of eight patterns, but in directions at right angles with respect to each other. The large loop antenna produces main lobes orthogonal, at right angles or 'broadside' to, the plane of the loop. Nulls are off the sides of the loop.

The small loop, however, is exactly the opposite: the main lobes are off the sides of the loop (in the direction of the loop plane), and the nulls are broadside to the loop plane, Fig. **1a**).

Don't confuse small-loop behaviour with the behaviour of the loop-stick antenna. Loop-stick antennas are made of coils of wire wound on a ferrite or powdered iron rod. The direction of maximum response for the loop-stick antenna is broadside to the rod with deep nulls off the ends, Fig. **1b**).

Direction finding

Both loop-sticks and small-wire

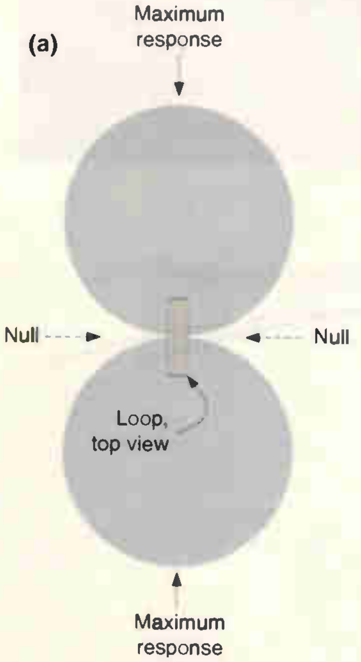


Fig. 1. In a), the small-loop antenna, the lobes are off the sides. In contrast, the lobes of the loop-stick antenna, b), are broadside to the stick.

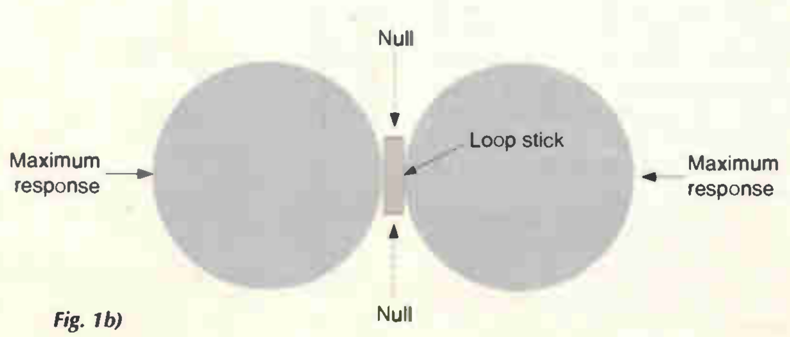


Fig. 1b)

loops are used for radio direction-finding and for shortwave, low-frequency medium wave, AM broadcast band, and VLF listening.

The nulls of a loop antenna are very sharp and very deep. Small changes of pointing direction can make a profound difference in the response of the antenna.

If you point a loop antenna so that its null is aimed at a strong station, the signal strength of the station appears to drop dramatically at the centre of the notch. But turn the antenna only a few degrees one way or the other, and the signal strength increases sharply.

The depth of the null can reach 10 to 15dB on sloppy loops and 30 to 40dB on well-built loops – 30dB is a very common value. I've seen claims of 60dB nulls for some commercially available loop antennas. The construction and uniformity of the loop are primary factors in the sharpness and depth of the null.

At one time, the principal use of the small-loop antenna was radio direction-finding – especially in the lower frequency bands. The RDF loop is mounted with a compass rose to allow the operator to determine the direction of minimum response.

The null was used, rather than the peak response point, because it is far narrower than the peak. As a result, precise determination of direction is possible.

Because the null is bi-directional, ambiguity exists as to which of the two directions is the correct direction. What the direction-finder 'finds' is a line along which the station exists.

If the line is found from two reasonably separated locations, and the lines of direction are plotted on a map, then the two lines will cross in the area of the station. Three or more lines of direction – a process called triangulation – yields a pretty precise knowledge of the station's actual location.

Loops for general reception

Today, these small loops are still used for radio direction-finding, but their use has been extended into the general receiving arena – especially on the low frequencies.

One of the characteristics of those bands is the possibility of strong local interference smothering weaker ground wave and sky wave stations. As a result, you can't hear co-channel signals when one of them is very strong and the other is weak.

Similarly, if a co-channel station has a signal strength that is an appreciable fraction of the desired signal, and is slightly different in frequency, then the two signals will heterodyne together and form a whistling sound in the receiver output.

The frequency of the whistle is an audio tone equal to the difference in frequency between the two signals. This is often the case when trying to hear foreign BCB signals on frequencies – called split frequencies – between the standard spacing. The directional characteristics of the loop can help if the loop null is placed in the direction of the undesired signal.

Loops are used mainly in the low-frequency bands even though such loops are either physically larger than high-frequency loops or require more turns of wire. Loops have been used as high as VHF and are commonly used in the 10-metre ham band for such activities as hidden transmitter hunts.

The reason why low frequencies are the general preserve of loops is that those frequencies are more likely to have substantial ground wave signals. Sky-wave signals lose some of their apparent directivity because of multiple reflections.

Similarly, VHF and UHF waves are likely to reflect from buildings and hillsides, so will arrive at angles other than the direction of the transmitter. As a result, the loop is less useful for the purpose of radio direction-finding. If your goal is not RDF but listening to the station, that is hardly a problem. A small loop can be used in the upper shortwave bands to null a strong local ground wave station in order to hear a weaker sky wave station.

Finally, loops can be useful in rejecting noise from local sources, such as a 'leaky' electric power line or a neighbour's outdoor light dimmer.

Next, I'll examine the basic theory of small-loop antennas, and then take a look at some practical implementation methods.

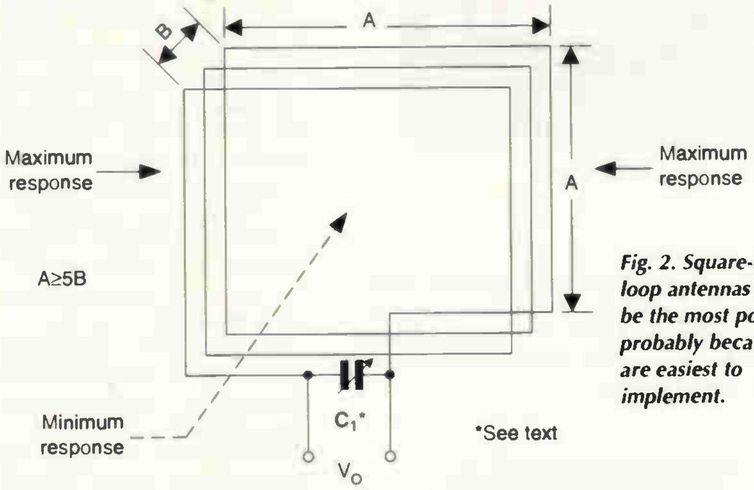


Fig. 2. Square-shaped loop antennas seem to be the most popular, probably because they are easiest to implement.

Box loops

A wire-loop antenna is made by winding a large coil of wire, consisting of one or more turns, on some sort of frame. The shape of the loop can be circular, square, triangular, hexagonal, or octagonal.

For practical reasons, the square loop seems to be most popular. With one exception, the loops considered in this section will be square so you can easily duplicate them.

The basic form of the simplest loop is shown in **Fig. 2.** This loop is square, with sides the same length 'A' all around. The width of the loop, 'B', is the distance from the first turn to the last turn in the loop, or the diameter of the wire if only one turn is used.

The turns of the loop in Fig. 2 are depth wound, meaning each turn of the loop is spaced in a slightly different parallel plane. The turns are spaced evenly across distance 'B'. Alternatively, the loop can be wound such that the turns are in the same plane. This is called planar winding.

In either case, the sides of the loop, 'A' should be not less than five times the width 'B'. There seems to be little

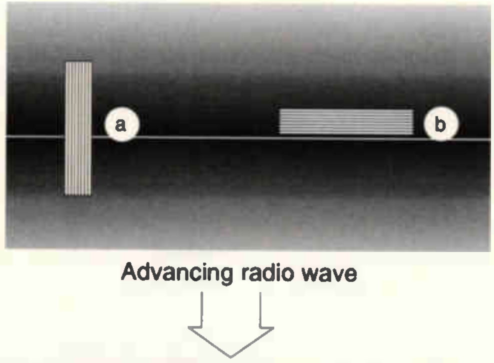


Fig. 3. When the loop antenna is perpendicular to the signal passing it, as in a), a potential difference develops in the wires. When the antenna is broadside, the signal's strength is similar throughout the antenna, no significant potential difference is developed hence the antenna 'nulls'.

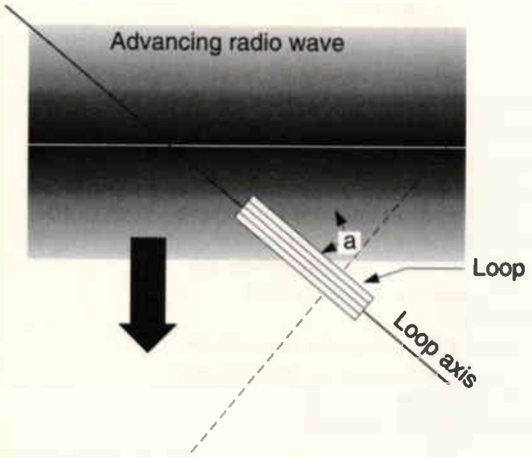


Fig. 4. Voltage across the output terminals of an untuned loop is a function of the angle of arrival of the signal a.

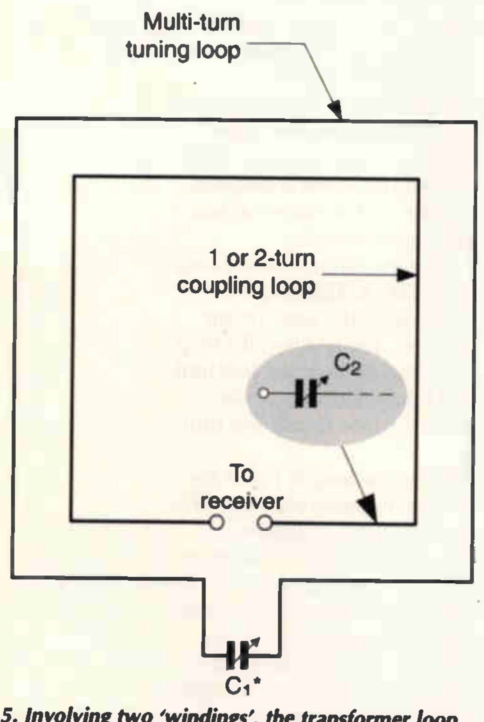


Fig. 5. Involving two 'windings', the transformer loop antenna is capable of providing much improved matching.

difference between depth and planar wound loops. The far-field patterns of the different shape loops are nearly the same if the respective cross sectional areas (πr^2 for circular loops and A^2 for square loops) are $<\lambda^2 \div 100$.

The reason why a small loop has a null when its broadest aspect is facing the signal is simple, even though it seems counterintuitive at first blush.

Take a look at **Fig. 3.** Here you have two identical small-loop antennas at right angles to each other. Antenna 'A' is in line with the advancing radio wave, while antenna 'B' is broadside to the wave.

Any given level of grey on the wave front, i.e. any given horizontal cross-section. represents a given signal strength – a so-called 'isopotential line.' When the loop is in line with the signal – antenna 'A' – there is a difference of potential from one end of the loop to the other, so current can be induced in the wires. But when the loop is turned broadside, all points on the loop are on the same potential line, so there is no difference of potential between segments of the conductor. Thus little signal is picked up – and the antenna therefore sees a null.

The actual voltage across the output terminals of an untuned loop is a function of the angle of arrival of the signal *a* of **Fig. 4**, as well as the strength of the signal and the design of the loop.

Voltage V_o is given by,

$$V_o = \frac{2\pi ANE_f \cos(a)}{\lambda}$$

Here, V_o is the output voltage of the loop, A is the area of the loop in square metres m^2, N is the number of turns of wire in the loop, E_f is the strength of the signal in volts per metre (V/m), a is the angle of arrival of the signal and λ is the wavelength of the arriving signal.

Loops are sometimes specified in terms of the effective height of the antenna. This number is a theoretical construct that compares the output voltage of a small loop with a vertical piece of the same kind of wire that has a height of,

$$H_{eff} = \frac{2\pi NA}{\lambda}$$

If a capacitor – such as C_1 in Fig. 2 – is used to tune the loop, then the output voltage V_o will rise substantially. The output voltage found using the first equation is multiplied by the loaded Q of the tuned circuit, which can be from 50 to 100:

$$V_o = \frac{2\pi ANE_f Q\cos(a)}{\lambda}$$

Even though the output signal voltage of tuned loops is higher than that of untuned loops, it is nonetheless low compared with other forms of antenna. As a result, a loop preamplifier is usually needed for best performance.

Transformer loops

It is common practice to make a small-loop antenna with two loops rather than just one. **Figure 5** shows such a transformer loop antenna.

The main loop is built exactly as discussed above: several turns of wire on a large frame, with a tuning capacitor to resonate it to the frequency of choice.

The other loop is a one or two turn coupling loop. This loop is installed in very close proximity to the main loop. It is usually – but not necessarily – on the inside edge not more than a couple of centimetres away. The purpose of this loop is to couple signal induced from the main loop to the receiver at a more reasonable impedance match.

The coupling loop is usually un-tuned, but in some designs a tuning capacitor, C_2, is placed in series with the coupling loop. Because there are many fewer turns on the coupling loop than the main loop, its inductance is considerably smaller. As a result, the capacitance to resonate is usually much larger.

In several loop antennas constructed for purposes of researching this article, I found that a 15-turn main loop resonated in the AM broadcast band with a standard 365pF capacitor, but the two turn coupling loop required three sections of a ganged three 365pF capacitors in parallel to resonate at the same frequencies.

Use computer ribbon cable? In several experiments, I used computer ribbon cable to make the loop turns. That type of cable consists of anywhere from eight to 64 parallel insulated conductors arranged in a flat ribbon shape.

Properly interconnected, the conductors of the ribbon cable form a continuous loop. It is no problem to take the outermost one or two conductors on one side of the wire array and use it for a coupling loop.

Benefits of tuning the loop

Loop performance is greatly enhanced by tuning the inductance of

The sports fan's loop

OK, sports fans. What do you do when the best game of the week is broadcast only on a low-powered AM station and you live at the outer edge of their service area where the signal strength leaves much to be desired? You use the sports fan's loop antenna, that's what!

I first learned of this antenna from a friend of mine, a professional broadcast engineer, who worked at a religious radio station that had a pipsqueak signal but lots of fans. It really works... one might say it's a miracle.

The basic idea is to build a 16-turn, 60cm square tuned loop and then place the AM portable radio at the centre so that its loop-stick is aimed so that its null end is broadside of the loop. When you do so, the nulls of both the loop and the loop-stick are in the same direction.

The signal will be picked up by the loop and then coupled to the radio's loop-stick antenna. Sixteen-conductor ribbon cable can be used for making the loop.

For an extra touch of class, place the antenna and radio assembly on a dining room table 'lazy Susan' to make rotation easier. A 365pF tuning capacitor is used to resonate the loop. If you listen to only one station, then this capacitor can be a trimmer type.

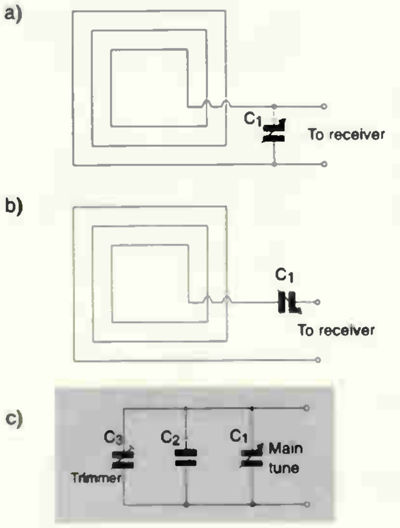


Fig. 6. Tuning the loop significantly improves performance and parallel tuning, as in a), is by far the most popular.

the loop to the desired frequency. The bandwidth of the loop is reduced, which reduces front-end overload.

Tuning also increases the signal level available to the receiver by a factor of 20 to 100 times. Although tuning can be a bother if the loop is installed remotely from the receiver, the benefits are well worth it in most cases.

There are several different schemes available for tuning, and these are detailed in **Fig. 6**. The parallel tuning scheme, which is by far the most popular, is shown in **Fig. 6a)**.

In this type of circuit, capacitor C_1 is connected in parallel with the inductor, which in this case is the loop. Parallel-resonant circuits have a very high impedance to signals on their resonant frequency, and a very low impedance to other frequencies. As a result, the voltage level of resonant signals is very much larger than the voltage level of off-frequency signals.

The series-resonant scheme is shown in **Fig. 6b)**. In this circuit, the loop is connected in series with the capacitor.

A property of series resonant circuits is that they offer a high impedance to all frequencies except the resonant frequency – exactly the opposite of the case of parallel resonant circuits. As a result, current from the signal will pass through the series resonant circuit at the resonant frequency, but off-frequency signals are blocked by the high impedance.

There is a wide margin for error in the inductance of loop antennas. Even the precise-looking equations to determine the required values of capacitance and inductance for proper tuning are actually only estimations.

The exact geometry of the loop 'as built' determines the actual inductance in each particular case. As a result, it is often the case that the tuning provided by the capacitor is not as exact as desired, so some form of compensation is needed.

In some cases, the capacitance required for resonance is not easily available in a standard variable capacitor. Some means must be provided for changing the capacitance. **Figure 6c)** shows how this is done.

The main tuning capacitor can be connected in either series or parallel with other capacitors to change the value. If the capacitors are connected in parallel, then the total capacitance is increased – all capacitances are added together. But if the extra capacitor is connected in series then the total capacitance is reduced. The extra capacitors can be switched in and out of a circuit to change frequency bands.

Tuning of a remote loop can be a bother if done by hand, so some means must be found to do it from the receiver location – unless you enjoy climbing into the attic or onto the roof.

Traditional tuning methods called for using a low-speed DC motor, or stepper motor, to turn the tuning capacitor. A very popular combination was the little 1 to 12rev/min motors used to drive rotating displays in retail store show windows. But that approach is not really needed today. You can use varactor voltage variable capacitance diodes to tune the circuit.

A varactor works because the junction capacitance of the diode is a function of the applied reverse bias voltage. A high voltage – such as

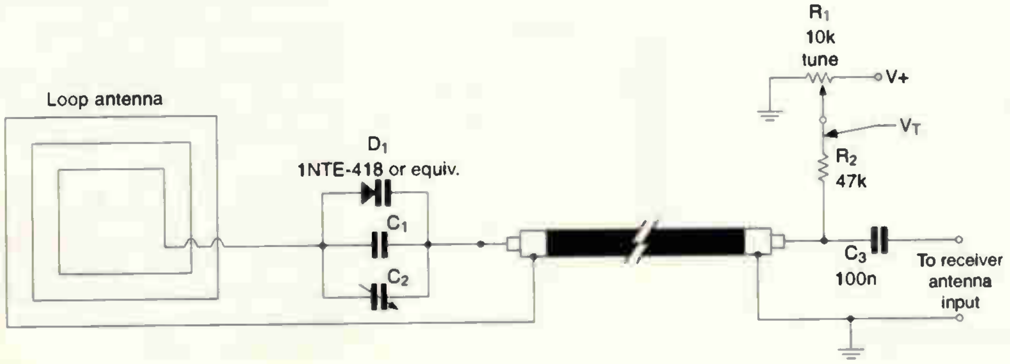


Fig. 7. Remote tuning of the loop used to be done with a motorised variable capacitor, but nowadays it is much more practicable to remotely tune using a variable-capacitance diode.

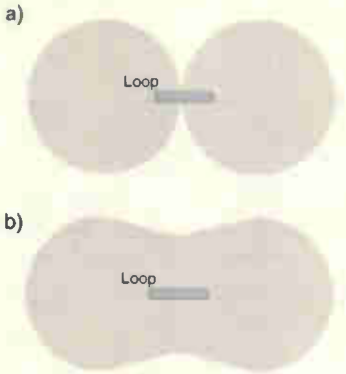


Fig. 8. Normal 'free-space' loop, a) is ideal, but in practice, the loop interacts with its surroundings, resulting in a sensitivity pattern more like that in b).

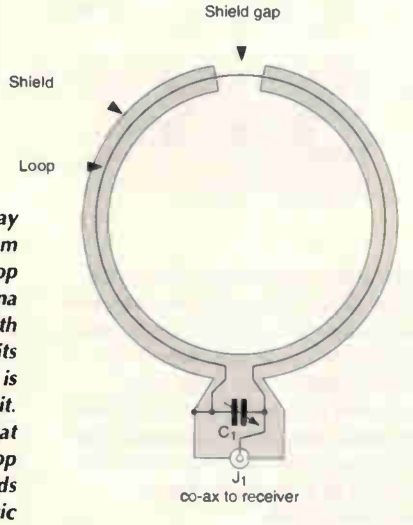


Fig. 9. One way round the problem of the loop antenna interacting with structures in its surroundings is simply to shield it. Remember that the small-loop antenna responds to the magnetic field component of the electromagnetic wave – not the electrical component.

30V – reduces the capacitance; lowering the voltage increases it.

Varactors are available with maximum capacitances of 22, 33, 60, 100, and 400pF. The latter are of most interest to us because they have the same range as the tuning capacitors normally used with loops.

Figure 7 shows how a remote tuning scheme can work with loop antennas. The tuning capacitor is a combination of a varactor diode and two optional capacitors: a fixed capacitor (C_1) and a trimmer (C_2). The DC tuning voltage (V_t) is provided from the receiver end from a fixed DC power supply (V+). A potentiometer (R_1) is used to set the voltage to the varactor, hence also to tune the loop. A DC blocking capacitor (C_3) keeps the DC tuning voltage from being shorted out by the receiver input circuitry.

Shielded-loop antennas

The loop antennas discussed thus far in this article have all been unshielded types. Unshielded loops work well under most circumstances, but in some cases their pattern is distorted by interaction with the ground and nearby structures such as trees, buildings.

In my own tests, trips to a nearby field proved necessary to measure the depth of the null because of interaction with the aluminum siding on my house.

Figure 8 shows two situations. In **Fig. 8a)**, you will see the pattern of the normal 'free-space' loop, i.e., a perfect figure-of-eight pattern. But when the loop interacts with the nearby environment, the pattern distorts.

In Figure **8b)** you will see some filling of the notch for a moderately distorted pattern. Some interactions are so severe that the pattern is distorted beyond all recognition.

The solution to the problem is to reduce interaction by shielding the loop, as in **Fig. 9**. Loop antennas operate on the magnetic component of the electromagnetic wave, so the loop can be shielded against voltage signals and electrostatic interactions. In order to prevent harming the ability to pick up the magnetic field, a gap is left in the shield at one point.

There are several ways to shield a loop. You can, for example, wrap the loop in adhesive-backed copper foil tape. Alternatively, you can wrap the loop in aluminum foil and hold it together with tape. Another method is to insert the loop inside a copper or aluminum tubing frame. Or... the list seems endless.

Using a loop antenna

Most of you will use a loop for DXing rather than hidden transmitter hunting, navigation, or other RDF purposes.

For the DXer, there are actually two uses for the loop. One is when you are a renter or live in a community that has routine covenants against outdoor antennas. In this situation, the loop will serve as an active antenna for receiving AM broadcast and other low-frequency signals without the neighbours or landlord becoming PFJs (purple-faced jerks).

The other use is illustrated by the case of a friend of mine. He regularly tunes in to clear channel WSM (650kHz, Nashville) in the wee hours between Saturday evening – 'Grand Ole Opry' time – and dawn.

However, that 'clear' channel of WSM isn't really so clear – especially without a narrow filter in the receiver. He uses a loop antenna to null out a nearby 630kHz signal that made listening a bit dicey, and can now tape his 1940s/1950s vintage country music.

It isn't necessary to place the desired station directly in the main lobes off the ends of the antenna, but rather place the nulls (broadside) in the direction of the offending station that you want to eliminate.

So what happens if the offending station and the desired station are in a direct line with each other with your receiving location in the middle between them? Both nulls and lobes on a loop antenna are bi-directional, so a null on the offending station will also null the desired station in the opposite direction.

One method is to use a sense antenna to spoil the pattern of the loop to a cardioid shape. Another method is to use a spoiler loop to null the undesired signal.

The spoiler loop is a large box loop placed one to three feet – found experimentally – behind the reception loop in the direction of the offending signal. This method was first described by Levintow and is detailed in **Fig. 10**.

The small loop-stick may be the antenna inside the receiver, while the large loop is a box loop such as the sports fan's loop. The large box loop is placed about 33 to 100cm behind the loop-stick and in the direction of the offending station.

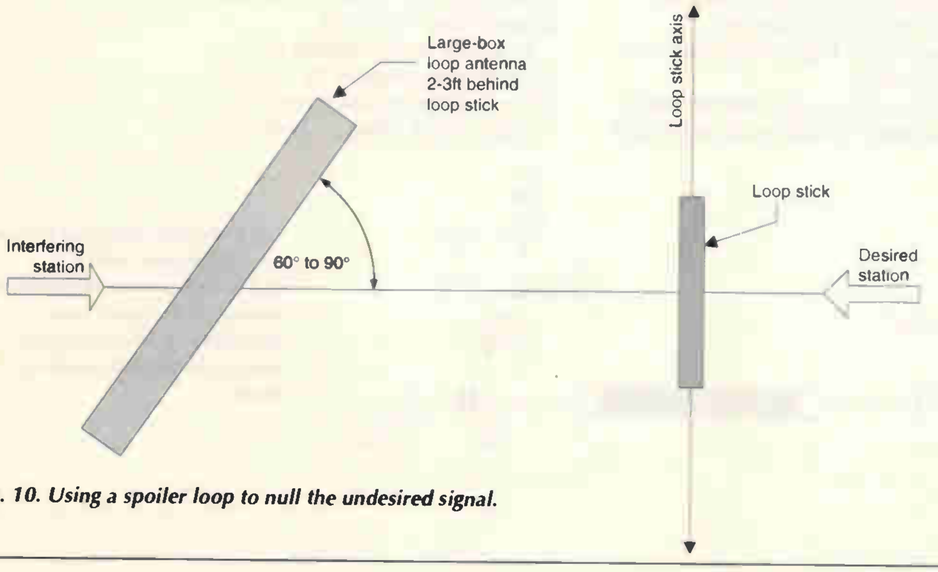


Fig. 10. Using a spoiler loop to null the undesired signal.

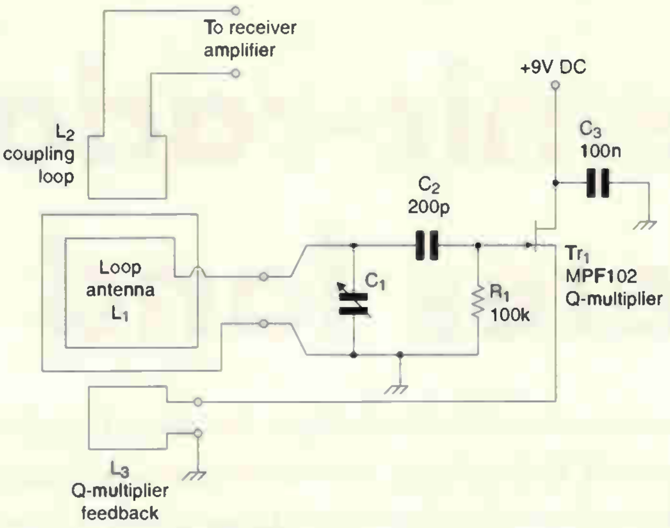


Fig. 11. Sharpening the loop using a Q multiplier.

The angle with respect to the line of centres should be 60° to 90°, which is also found experimentally. It's also possible to use two air-core loops to produce an asymmetrical receiving pattern.

Sharpening the loop

Many years ago the Q-multiplier was a popular add-on accessory for a communications receiver. These devices were sold as Heathkits and many construction projects were seen in magazines and amateur radio books.

The Q-multiplier has the effect of seeming to greatly increase the sensitivity of a receiver, as well as greatly reducing the bandwidth of the front-end. Thus, it allows better reception of some stations because of increased sensitivity and narrowed bandwidth.

A Q-multiplier is an active electronic circuit placed at the antenna input of a receiver. It is essentially an Armstrong oscillator, as shown in **Fig. 11**, that doesn't quite oscillate.

These circuits have a tuned circuit, L_1/C_1, at the input of an amplifier stage, and a feedback coupling loop, L_3. The degree of feedback is controlled by the coupling between L_1 and L_3. The coupling is varied both by varying how close the two coils are, and their relative orientation with respect to each other. Certain other circuits use a series potentiometer in the L_3 side that controls the amount of feedback.

The Q-multiplier is adjusted to the point that the circuit is just on the verge of oscillating, but not quite. As the feedback is backed away from the threshold of oscillation, but not too far, the narrowing of bandwidth occurs as does the increase in sensitivity.

It takes some skill to operate a Q-multiplier, but it is easy to use once you get the hang of it and is a terrific accessory for any loop antenna. ■

Seismic-radar locates land mines

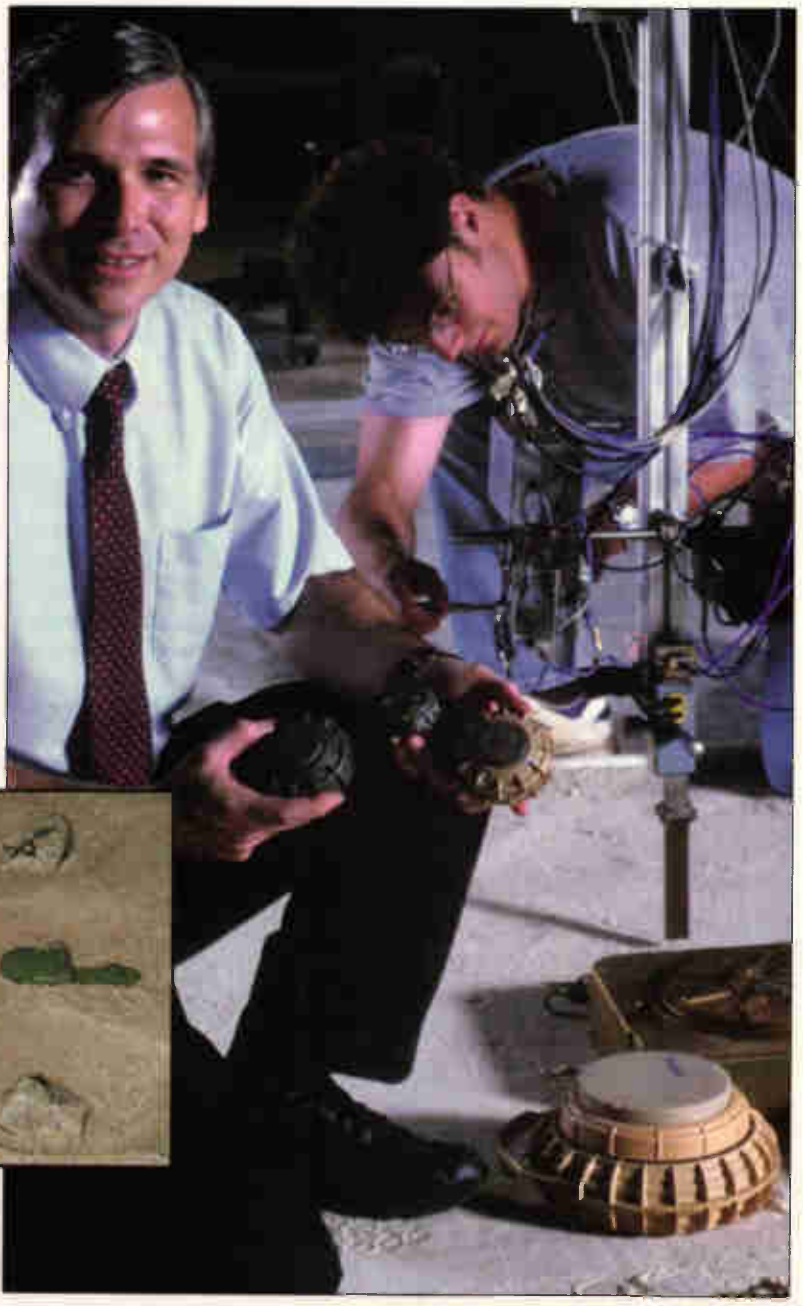
Steve Bush reports on a new technique for detecting land mines. Developed at Georgia Institute of Technology, the technique involves sending out a seismic wave then looking for mine refraction signatures using a specially-developed radar.

By using seismic waves and precision radar, researchers at the Georgia Institute of Technology have developed a new method for finding buried land mines.

"Detecting land mines is a very difficult thing to do," said Dr Waymond Scott, associate professor in Georgia Tech's School of Electrical and Computer Engineering. "Every existing method for mine detection has conditions under which it will work very well and conditions under which it will fail."

Scott is not claiming that his method is foolproof, but tests show that is is good at rejecting the sticks, stones and other 'clutter' that defeat some methods.

Originally proposed before the technology to implement it existed,



"You could put this mine 1 000 miles out in space and be able to detect it more easily than if you put it one centimetre under the soil," says Dr Waymond Scott, here seen on the right kneeling in the foreground. Co-worker Christoph Schroeder adjusts the radar system used to measure soil displacement in their detection technique.

Before... Mines, from anti-tank size to the smallest anti-personnel, are laid out in sand with potentially confusing rocks before being buried in more sand and covered with pine needles.

the detection technique involves creating seismic waves that travel through the soil containing land mines. These elastic waves causes the soil and everything buried in it to be displaced slightly.

Solid objects move with the wave, whereas land mines, due to their shape and hollow construction, act like a drum and re-radiate vertical waves. These new waves rise and disturb the surface of the ground.

This is where the radar comes in. "You could use a geophone to detect the surface movement although this would involve touching the soil and there could be a mine buried underneath," said Scott.

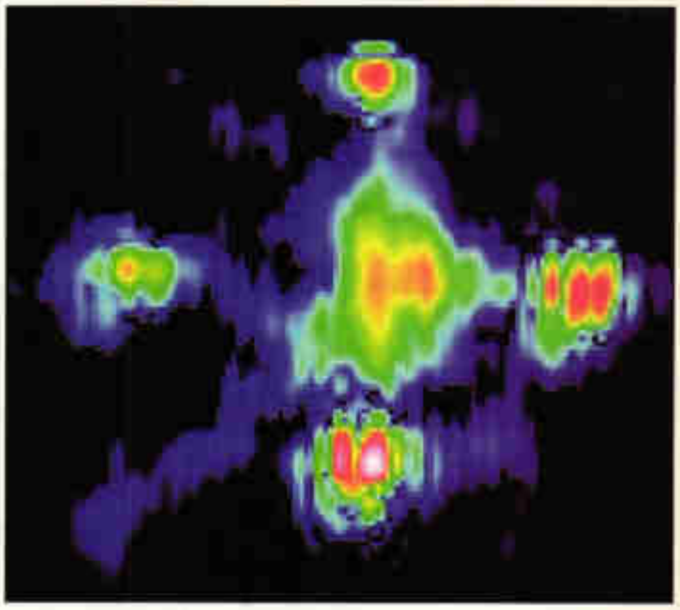
The movement in the surface of the soil is a damped sinewave of around 1μm displacement according to Scott and the team's in-house developed radar can pick this up.

At the moment, the group is experimenting in 50 tonnes of damp sand. "This is a pretty good model for typical soil, except sand is more homogeneous. The technique does not work with dry sand as this has no shear strength," he said.

A synthetic minefield map is made up using a step-and-repeat process with a radar head 2.5cm from the ground. This gives a resolution of 2cm and will operate through a covering of pine needles simulating light ground cover.

Operation from 40cm above the soil using beam-forming algorithms has also been demonstrated. "To get this far has taken three years," said Scott. "We have to get more than a few centimetres through heavier cover to be practical."

As well as increasing range and using an area-scanning radar, Scott intends to use different seismic waveforms to better select mines. Step-sinewaves and chirps are under investigation. M-sequences will be examined in the future. Non-contact wave sources such as an electric arc, loudspeaker, microwave, laser and water jet will completely remove the need to touch the ground. ■



After... The seismic-radar method spots the mines and rejects the rocks. You can see the trace of a rock in the bottom left-hand corner. Mines over 10cm down can be found in this way. Scientist Dr Waymond Scott points out that only a lot more research will show whether a practical, reliable, mine detector will come from this work.

Make your own WAP SITE

WAP sites – the equivalent of WEB sites but available on a WAP enabled mobile phone – have only a limited text and graphic capability, but to outweigh that, they can be accessed on a pocket-sized battery-powered appliance from almost anywhere in the World. Peter Marlow explains how easy it is to set up your own site.

It may be slow, over-expensive – and certainly over-hyped – but wireless Internet is here. Predictions about the Internet are often conservative, but by 2003 it is reckoned that 95 percent of all mobile handsets shipped will be Internet enabled and third-generation phones will be appearing with multimedia capabilities. Some 30% of all Internet data will be accessed from wireless terminals.

What does this mean? Can you surf the Internet with a mobile phone now and order your groceries? Well, in theory yes. In practice though, not at the moment, because of the size of the display and keyboard, and the narrow bandwidth.

However, you can do some very useful things with it now and this set of three articles will show you how.

WAP backgrounder

Wireless data has been around for while. The GSM standard allows a Short Message Service, or SMS, that can be used to deliver text messages, e-mail and news services to subscribers. But it is limited to 160 characters per message and there is latency in the system.

GSM phones also have data ports that allow lap-top computers to be connected to them, albeit at 9.6kb/s unless you have GPRS (see later).

In the early 1990s, when the Internet was young, it became apparent to the major cell-phone manufacturers that wireless voice and data and the Internet would converge. People wanted to access data on the

Motorola Timeport showing a typical menu at the Phone.com WAP developer site.

The main menu screen at Club Nokia

Exclusive free WAP site evaluation software

On this month's cover-mounted CD, you will find a copy of the unique shareware program 'WAP Wizard Lite'. You can use this program to compile and upload your own WAP site on a PC. It can produce either of the two popular WAP phone formats. This is a fully-enabled version of the program, yours to try out for a thirty days.

move without having to lug a laptop computer around.

But could full Internet access be put on to a mobile phone? There were major bandwidth and ergonomical obstacles to be overcome.

In June 1997, the WAP Forum was set up by Ericsson, Motorola and Nokia as a non-profit making industry association to, "create the standards for delivering Internet access to consumer-class wireless devices – mobile phones, pagers, two-way radios, smart-phones and communicators." This statement is taken from the association's web site at www.wapforum.com.

Now, the WAP forum has over 500 members, representing 90% of the global handset market. It aims to create and develop a global protocol specification – the Wireless Application Protocol, "to work on all wireless networks in all parts of the world".

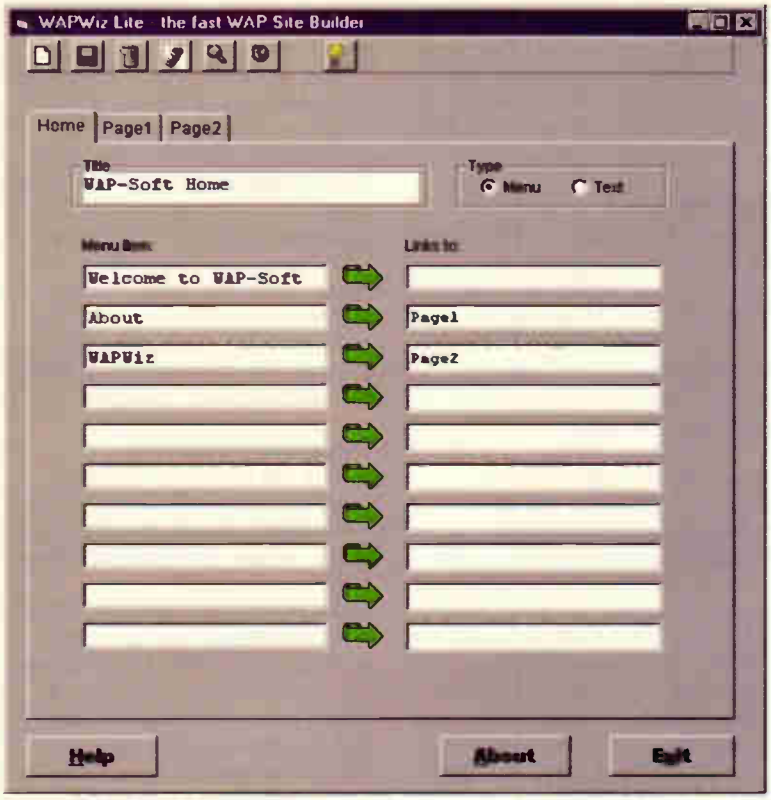
WAP is an 'end-to-end' solution, which uses existing standards as far as possible, promotes new open standards, gives air-interface independence and provides device independence.

It was obvious from the outset that the point-and-click HTML interface as used on the World Wide Web would not suit the current design of mobile phone, with its small display and keyboard, limited processing power and low bandwidth. So a company called Unwired Planet, now called Phone.com, developed the Handheld Device Markup Language (HDML), derived from Extensible Markup Language (XML).

The language XML was itself developed from Standard Generalised Markup Language (SGML) – the mother of HTML – which dates from 1974. Using HDML, the ATT Pocket Net service went live in the US in the autumn of 1997. Phone.com provided the mobile phone gateway and the microbrowser.

Through the WAP Forum, HDML was developed into Wireless Markup Language (WML), although Phone.com still have some proprietary extensions. Currently at version 1.1, revision 1.2 is due out by the end of 2000.

Jumping the gun, however, was the Japanese operator NTT Docomo. It started a wireless Internet-access service called 'i-mode' in February 1999. Although a member of the WAP Forum, Docomo decided that it couldn't wait for the standard to be settled



Producing a WAP menu page is as easy as filling in the boxes. To produce text for a page, simply select 'Text' in the 'Type' panel.

and decided to go with HTML text.

I-mode has been a great success. It was probably responsible for the hype in the UK about WAP. It is rumoured that one Japanese company has made a fortune selling mobile phone screen-savers and wallpaper for i-mode.

A late arrival on the mobile phone scene is Microsoft whose Mobile Explorer microbrowser can handle HTML, HDML and WML. There is wisdom in waiting!

How WAP works

To access the Internet from your mobile phone, you dial the number of your WAP Gateway. This is just like connecting to your ISP.

The phone number and password are stored in the phone – you just select them from a menu. It is possible for companies to have their own WAP Gateways for employees on the move. Gateway software is currently supplied by Phone.com, Nokia and Infinite Technologies.

The Phone.com offering allows the user to request pages of information that can be sent to a fax machine of their choice. It can also send alerts – like an SMS message.

The Gateway connects directly to the wider Internet, but it translates and compresses the data before it sends it back to your phone over the 9.6kb/s link. Some Gateways can translate HTML pages. But beware. Even without graphics, these can be bulky and impossible to navigate – particularly with frames.

It's better to go to a WAP site with its WML pages designed for a phone. Most, but not all, phones allow you to type in the full URL of the WAP site. You can then bookmark it for easy access later.

Some directory sites have large menus with many links already in. There are search sites to help you find what you want. However, mobile browsing is just not practical.

Once you have established (quickly) what you want, you need to bookmark it for future reference. If you don't, the bill can quickly mount up. When I tried to surf, I clocked up 40 minutes in the first session, which cost me £8.

BT is showing the way with GPRS (General Packet Radio System) launched in June. With this you are always on-line. You only pay for the data received or transmitted. Incidentally, the speed is 40kb/s, with plans to reach 115kb/s.

If you want your own WAP site simply add a WAP folder to your present site and put a home page written in WML into it (see later). For example,

http://www.mycompany.co.uk/wap/index.wml

When accessing it from your phone

the microbrowser needs the full URL – the 'index.wml' for the home page is not assumed. If you want to look at your page on the web with Internet Explorer 5 it will only show you the source code. Things will change.

Uses for WAP

So what do you want a WAP site for? John Dvorak was particularly damming about WAP is the October 2000 issue of *PC Magazine*. But WAP has already found its niche. It is particularly useful for supplying time-sensitive data such as news, stocks and share prices and sports scores.

Other uses include transport timetables, take-away menus, 'what's-on' notice board, and company phone books. You can dial a number shown on the screen by just pressing a button.

The killer-application is 'where am I'. Mobile phone operators know where their customers are to within 20 metres and can supply information to them on request about restaurants, garages, cash-point locations, hotels, etc.

BT is already supplying such a service – albeit verbally – with its 1500 facility. However, due to privacy rules this information is closely guarded and not available to the average website. It is not possible to insert a cookie into the mobile phone to find out which cell it is in.

If applications don't require much text input, they can provide practical interactivity for services such as mobile banking. The link between the microbrowser and the WAP gateway is encrypted. But the link from there to your web site may not be. So if you want to set up an M-commerce site – beware.

Other applications include locating and booking a video from your local library or making a doctor's appointment without hanging on the phone. There are many more. The big players like Amazon.com and LastMinute.com are already there.

The bottom line with mobile phones is that they possess a big advantage over PCs – you turn it on and it's working within seconds, not minutes. The trick is to design WAP sites that fulfil their niche role.

Designing a WAP site

So how can I design a WAP site? There is a shareware package called WAP Wizard Lite on the cover CD-ROM – or you can download it from www.wap-soft.com.

This package enables you to build a WAP site without needing to know anything about WML. It runs on any PC with Windows 95 or higher. It allows you to write simple menus with links to text pages and a selection of graphics are included.

After installation, start up the program, Start\Programs\WapWizLite\Wap Wiz. It opens with a blank Home Page configured as a menu. First type in a page title and type in a menu item on the left and the link, either a page or internet URL which you want the phone to go to if selected.

To create a new page, click on the page icon on the far left of the toolbar. This will be configured as a text page, but you can make it into a menu by clicking the 'Menu' radio button at the top right. The text page has a title and one graphic, selectable from a library – courtesy of Phone.com. Remember that the phone screen is quite small.

You can view your page by clicking on the magnifying class icon on the toolbar. Note that mobile phones can display pages differently.

To move between pages click on the tab strip. To compile your WAP site into WML click on the disk icon.

This builds a file called index.wml in your 'Program Files\WAPWizLite\Web' folder. Take a look at the resulting WML code. It's only a text file so you can use Windows Notepad. I will expand on this in the next article.

To publish your site, upload the contents of the folder 'Program Files\WAPWizLite\Web' to your web site. This will consist of Index.wml and any graphics files. Use your ISPs FTP standard software or use Microsoft Web Publisher which you can launch by clicking the Mobile Phone icon on the toolbar.

Next, try your site out on a mobile phone. The best phones at present are the Nokia 7110 or the Mitsibushi Truim, which allow you to directly enter a URL. Motorola's Timeport inexplicably does not allow you to do this, but there are other ways of viewing URLs with it (talk to Genie).

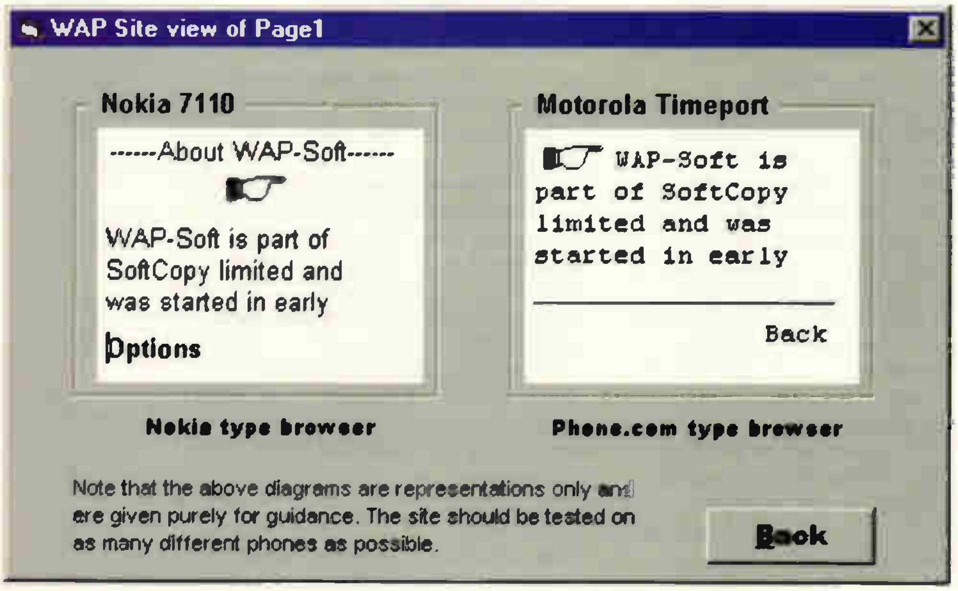
For more details, print out the help file on the cover CD-ROM with Internet Explorer: d:\wap\WapWizLite Help.htm.

What's coming up next?

In the next article, I take a close look at WML, best practice for WAP page design, and the pitfalls. I will also give you an overview of WML script, which allows you to write games for a mobile phone.

In a third article, I will discuss the resources available to build and test WAP sites, including the Phone.com and Nokia toolkits. ■

Once you've produced your WAP page, you can see what it looks like in either one of two popular formats.



The author

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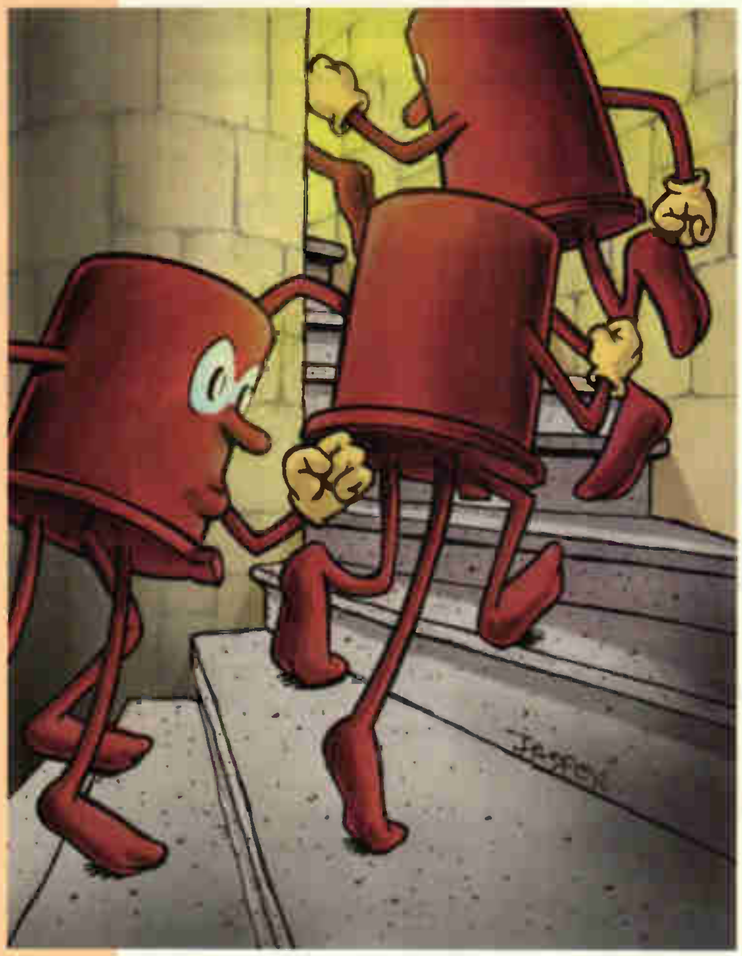

Better buffers III

Dave Kimber explains in simple terms how to design a Class-B output stage using complementary compound emitter followers. He also explains how to minimise crossover distortion – and why it's impossible to eliminate it.

In the first of these three articles[1], I looked at the complementary compound emitter follower, or CCEF, as a Class-A buffer, **Fig. 1**. My second article[2] looked at a variant with voltage gain, the complementary feedback pair, **Fig. 2**.

In both cases, the critical parameter was transconductance, g_m. In this article, I look at using a complementary pair of compound emitter followers as a Class-B output stage of the type found in audio power amplifiers.

I explained earlier that the operation of the CCEF can be considered to take place in three current regions. First a summary of how.

Looking at Fig. 2, in the low-current region there is insufficient voltage across R to turn on Tr_2, so Tr_1 operates alone. At slightly higher currents, there is a sharp transition to the medium-current region, in which Tr_1 can be considered to be a voltage amplifier driving the transconductance of Tr_2.

There is a smooth transition to the high current region, in which Tr_2 acts as a current amplifier on the transconductance of Tr_1. This is illustrated by **Fig. 3**. Resistor R's function is to boost the g_m of Tr_1 by increasing the current flow through it, but too small a value for R reduces g_m by loading the collector circuit of Tr_1.

Having chosen a suitable quiescent current I_{c2}, then an optimum value for R is given by,

$$R = \frac{25 \times \beta^2}{I_c^2} \qquad (1)$$

Class-B in outline

A complementary pair of CCEF is a popular Class-B output stage, **Fig. 4**. This topology has been explored by Self[3] via simulation and measurements on real circuits.

A few amplifiers have used a complementary pair of feedback pairs to give an output stage with voltage gain; probably most famously in the *PW* Texan[4]. Apart from one blown output transistor, my Texan is still going strong after more than 25 years service.

An output stage may require slightly different approximations from those which apply to a low-signal Class-A CCEF. Currents will generally be higher, although perhaps not under quiescent conditions. The transistors will be beefier, which generally means lower current gain.

So far it has been assumed that a transistor turns on at about 600mV, but then V_{be} remains more or less constant. An output transistor might begin to have some significant conduction at a lower V_{be}, of, say, 500mV, but the larger current range becomes significant.

The ratio between quiescent and full signal current could be 500, which will require a change of V_{be} of around 160mV. This will require some change in driver current – particularly at low signal levels.

The lower β of the output transistor will cause an increased demand for base current. In turn, this increases the driver current at higher signal levels. The net result is that the sort of algebra used for the Class A case is likely to be less predictive for Class B, although it still provides some useful insights.

Self found that for low crossover distortion, a CCEF output stage needed a low quiescent current of 7 to 15mA. You may find this surprising, until you learn that the figure is related to the low-medium transition in the CCEF. This is the point where Tr_2 begins to turn on and contribute to the transconductance.

For R, Self used a value of 100Ω, so you would expect the

David P Kimber B.Sc.

transition to occur somewhere near 6mA. The transconductance rises rapidly just above the transition from a low value to a high-ish value, before levelling off due to the combined effect of the medium-high transition and the degeneration provided by r. The correct quiescent current will be somewhere near the point where the g_m for each half of the output stage is half the value for higher currents.

At high currents the output stage g_m is dominated by r and is a little under $1/r$, so the quiescent setting should have gm for each half of about $1/2r$, or a CCEF g_m of $1/r$ or slightly less.

Crossover distortion

Below the quiescent current setting, the transconductance drops rapidly down to the transition knee. Below this point, it reverts to being proportional to current.

Above the quiescent current setting, the transconductance rises – but more slowly – until it is eventually limited by the small resistor r. The result is that it is impossible to eliminate crossover distortion, because the 'below' and 'above' curves do not match Fig. 5.

If the quiescent current is too low then the total transconductance has a dip in the middle. As the quiescent current increases this splits into two dips symmetric about the middle with a central peak.

As quiescent current is increased further, the central peak will eventually exceed the high current value. This is known as 'g_m-doubling' and its effects may sound worse than normal crossover distortion.

The smoothest curve that can be achieved has a slightly broader dip in the middle, but this is still well below the level of the transconductance for higher signal currents. Some degree of 'dip split + peak' gives lower crossover distortion than the smoothest curve because the total deviation of g_m from the ideal flat line is less. These features are illustrated by Self, although he plots output voltage 'gain' – with an 8Ω load – rather than g_m which makes the curves look slightly less alarming!

Some correctly applied negative feedback will greatly reduce the remaining crossover distortion, but is there anything else that can be done? There's a fairly sharp knee below the quiescent point, and a fairly smooth one above it. If you could make the shape of these two transitions more similar then this should help reduce crossover distortion.

Given the CCEF circuit, the shape of the lower knee is fixed by the relationship between g_m and I_c in Tr_2. However we can influence the shape of the upper knee.

Sharpening the upper knee

Two quite separate effects produce the upper knee. One is the medium-high transition in the CCEF i.e. the point where Tr_2 base impedance begins to significantly load R. Above this, the g_m continues to slowly rise with increasing current.

The other is the degeneration provided by r, which causes the g_m to level off. If these two effects coincide then the upper knee will be softened. Given that we now want to separate them, which way round should they be? If we want low levels of large signal distortion then it would be best to ensure that for large currents the g_m is flat, i.e. it is dominated by r.

However, as you will see, this may require unreasonably high values of beta. On the other hand, the rising g_m above the medium-high transition may be limited by beta-droop at high currents in the output transistors so it might not be as bad as it seems.

In either case you need to ensure that the medium-to-high transition does not coincide with the point where r begins to control things. In other words, you want $g_{m(m-h)} \neq 1/r$, and probably at least a factor of two away from equality.

In the first article, I showed that the transconductance of a CCEF in this region is given by,

$$g_m = \frac{24000}{R} \times \left(1 + \frac{\beta_2}{1 + \frac{I_{2(m-h)}}{I_{c2}}}\right) \text{ (mA/V)}$$

So you want,

$$\frac{24}{R} \times \left(1 + \frac{\beta_2}{2}\right) \neq \frac{1}{r}$$

or,

$$\beta_2 \neq \frac{R}{12 \times r} - 2 \qquad (2)$$

or,

$$\frac{R}{r} \neq 12 \times (\beta_2 + 2) \qquad (3)$$

Inserting typical values of R=100 and r=0.22 into (2), and assuming a safety factor of 2, you will find that β_2 should be outside the range 18 to 72. Given the current gain of most output transistors, it seems likely that some amplifier designs will not meet this criterion.

Conversely, if you insert β_2=25 into equation (3), then you will find that R/r should be outside the range 162 to 648. This could be satisfied by R=68 and r=0.47 – with possibly reduced peak output because of the high r. Alternatively, R=100 and r=0.1 could be used, but with a possible penalty of increased large signal distortion.

Self found that with R=100, he obtained best crossover performance with r=0.1. His graph[5] appears to show r=0.33 as slightly worse than r=0.47.

This would confirm my analysis if his output β was somewhere near 25. Of course, satisfying the criterion does not guarantee good crossover performance, but it should at least provide scope for avoiding poor performance.

You can now estimate the quiescent current. It is set by the requirement that transconductance of the CCEF alone, i.e. without including r, is about – and typically just below – $1/r$. This should be well below the medium-high transition so we can use a simple form of g_m. The requirement is,

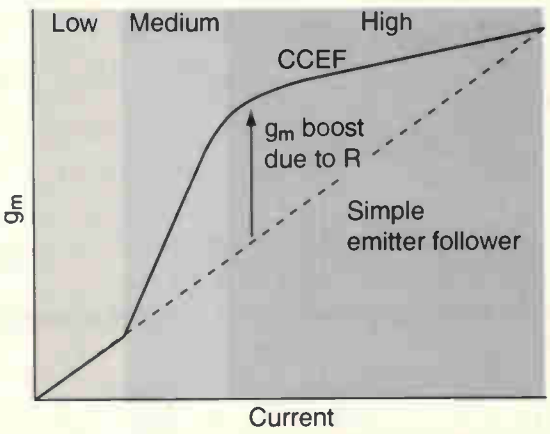

Fig. 3. In a complementary compound emitter follower, transconductance varies with current – an important factor when working out the follower's performance.

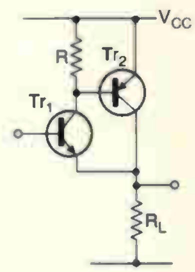

Fig. 1. Class-A complementary compound emitter follower, or CCEF, often used as a buffer.

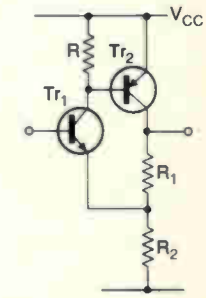

Fig. 2. In this variant of the complementary compound emitter follower, some voltage gain is obtained.

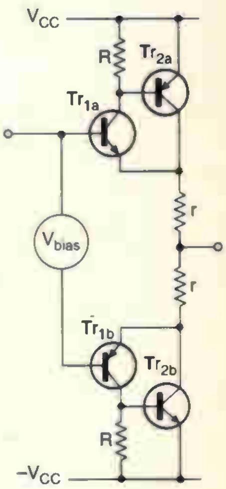

Fig. 4. Complementary Class-B output stage using complementary compound emitter followers.

Thermal stability

Before considering possible improvements to the CCEF output stage, it is necessary to look at maintaining the existing performance, despite changes in junction temperature.

Both driver and output transistors will get quite hot in normal operation. This could cause changes in quiescent current. As a transistor heats up it tends to draw more current. Conversely, to maintain the same current it is necessary to reduce the base-emitter voltage by about 2.5mV/°C.

Naively one might think that the critical thermal issue for the CCEF is the V_{be} change for the driver transistors Tr_1, and that thermal effects in the output transistors can be ignored because these are current driven. It turns out that this simple view is wrong.

Bias for the output stages is normally produced by a V_{be}-multiplier circuit, which shares the driver heat sink. The result is that V_{be} changes in Tr_1 are compensated, provided that the junction temperature of the bias generator accurately tracks the junction temperature of the drivers. Unfortunately, good temperature tracking turns out to be surprisingly difficult to arrange in the real world.

However, I have already shown that at the quiescent point, the output transistors are largely voltage driven. So their thermal V_{be} changes *do* matter. There is local feedback around the CCEF due to *r*. It turns out that at the quiescent point this halves the thermal change in I_{c2}.

If there is to be no thermal change, then Tr_2 V_{be} must fall by 2.5mV/°C. Because of the voltage gain in Tr_1 this requires an additional reduction of around 0.1mV/°C in the base-emitter voltage of Tr_1. So our bias supply must overcompensate Tr_1 by about 4% if I_{c2} is to remain stable – assuming Tr_1 and Tr_2 are at the same temperature.

There is another factor to be considered. Until now, transconductance has been calculated as 40×*I*, but the coefficient of 40 should really be *e/k*T, where *e* is the charge on an electron, k is Boltzmann's constant and *T* is absolute temperature. This is 40 at when *T* is 290K, i.e. 17°C.

As the transistor junctions heat up, the g_m reduces. To maintain good crossover performance it is important to maintain stable g_m, but not necessarily stable current.

In fact the current needs to increase linearly with absolute temperature. If your amplifier has a CCEF output and maintains rock-steady quiescent current as it heats up, then it is thermally over-compensated!

A junction temperature increase of 60°C requires a current increase of 20% to maintain transconductance.

The current in Tr_1 needs to fall very slightly as temperatures rise, so the current in Tr_2 needs to rise faster than absolute temperature to compensate for the thermal g_m drop in Tr_1.

I'll pull some figures out of the air here. Assume that Tr_1 junction rises by about 30°C, and Tr_2 by 60°C. Current I_{c2} needs to be increased by about 30%, which requires a V_{be} increase of about 7mV.

There is the plain thermal reduction of 150mV, so a total change of 143mV at the base Tr_2 is needed. This is about 7mV at Tr_1 base, in addition to its own thermal requirement of 75mV, so it is necessary to over-compensate by about 9%.

Unfortunately most simple bias circuits will tend to under-compensate. This is because it is impossible to properly track the junction temperature of Tr_1, given access only to its case temperature or – even worse – its heat-sink temperature.

It should be possible to achieve something like the correct compensation, although this will require careful design. Fortunately the temperature rises of driver and output transistors remain roughly proportional as the signal varies, with the ratio between them depending on thermal resistances and current gain.

The biggest problem is probably the differing thermal time constants. If the outcome is that Tr_2 quiescent current gently rises with temperature then the design is probably not too far wrong.

$$\frac{24000}{R}+960\times I_{c2}=\frac{1000}{r}$$

or,

$$I_{c2}=\left(\frac{1}{r}-\frac{24}{R}\right)+0.96\approx\frac{1}{r}-\frac{25}{R}$$

$$I_q=I_{c1}+I_{c2}=\frac{600}{R}+\frac{1}{r}-\frac{25}{R}=\frac{575}{R}+\frac{1}{r} \qquad (3)$$

So for *R*=100, the quiescent current for typical values of *r* are,

r	I_q (from (3))	I_q (Self)
0.1	15.8	15.3
0.22	10.3	11.5
0.33	8.8	8.54
0.47	7.9	7.64

Given the simplicity of my model, the agreement with Self's results is remarkable.

Softening the lower knee

Instead of – or in addition to – sharpening the upper knee, is there a way to soften the lower knee? This would bring improvements – especially if it could be moved to a slightly higher current too.

The transition is sharp because Tr_2 turns on over only a narrow range of base-emitter voltage, i.e. a narrow range in Tr_1 current. It immediately receives an amplified signal from Tr_1, which has a voltage gain of about 24. You get a kick when the turbocharger comes in!

One option might be to simply reduce the voltage gain of Tr_1 using emitter degeneration. This would soften the lower knee, but at the expense of reducing total transconductance.

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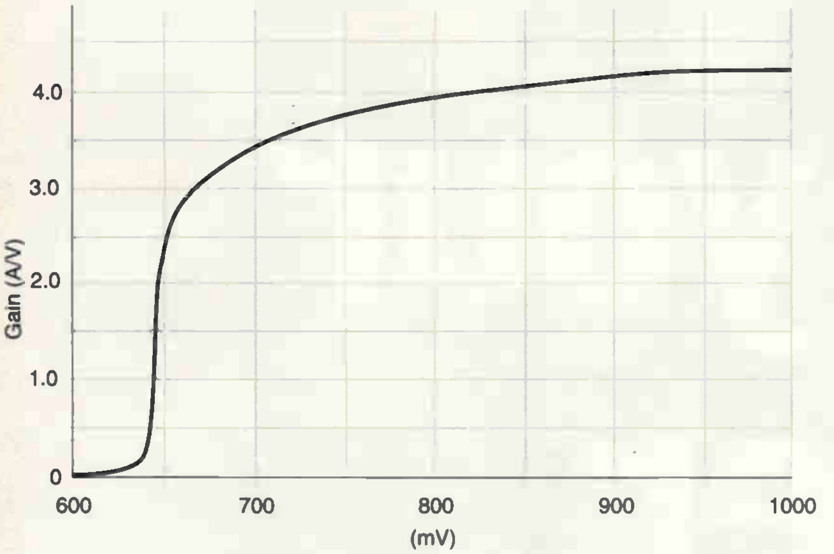

Fig. 5. Transconductance, i.e. g_m, of the Class-B stage, assuming that r is 0.22Ω, R is 100Ω and β_2 is 30. Up to the point where the quiescent current is reached, the curve is steep but gain increase starts to level off thereafter. Because of the difference in these two curves, it is impossible to eliminate crossover distortion

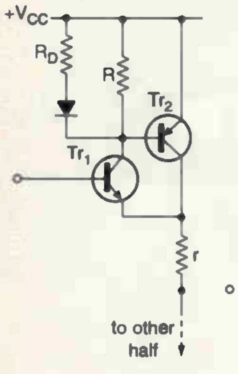

Fig. 6. Reducing the voltage gain of Tr_1 around the point where Tr_2 switches will improve the CCEF's linearity. One way of doing this is to add what is sometimes called a 'linearising diode'.

It would have its greatest effect at high currents, rather than the low current area we are interested in. The likely outcome would be to further soften the upper knee – which defeats the object.

A way needs to be found to reduce the voltage gain of Tr_1 around the point where Tr_2 is switching on, but without significantly reducing the overall transconductance for higher currents.

Putting a diode across R reduces the voltage gain of Tr_1, but then you need to add a resistor in series with the diode to avoid an overall reduction in g_m, **Fig. 6.** This is sometimes known as a 'linearising' diode. Unfortunately this combination is quite difficult to solve analytically.

If the diode is in fact a transistor with its base and collector connected together, then it will obey the Ebers-Moll model. For example, you know its dynamic impedance will be 25/I (mA). Assume that it is the same type of transistor as Tr_2 – although in reality it is more likely to be the complement of Tr_1. Also assume that R is still 100Ω. The problem is to choose a value for R_d. Why not pluck some figures out of the air?

By the time I_{c2} has reached 1mA, the total transconductance is rising quite sharply, so by this point the diode needs to be loading R significantly. Let's say, arbitrarily, that diode current I_d should be more than 0.5mA when I_{c2} is 1mA. This is a ratio of 2 in current, which needs a voltage reduction of ln(2)/40=17mV for the diode, i.e. R_d is less than 34. This will put an impedance of 84Ω or less in parallel with R.

At higher currents you don't want the diode network stealing too much base current from Tr_2. So choose a value of less than 2mA arbitrarily for I_d when I_{c2} is 20mA. This is satisfied if R_d is more than 29.

It looks like R_d will have to be 33Ω. Of course, if the two inequalities for R_d could not be simultaneously satisfied, you would need to pluck different figures out of the air!

It is now possible to make a rough estimate of the new quiescent current. Current I_d is going to be somewhere around 1mA, so the collector of Tr_1 will see an impedance of about 37Ω. Current Ic_1 will be about 7mA, so the voltage gain of Tr_1 will be about 10.

If r is 0.22 then you want a total g_m of 4545mA/V, of which 280mA/V will be contributed by Tr_1. So Tr_2 needs to provide 4265/10 mA/V, which will require Ic_2 to be 10.7mA. This is over twice the previous value, so at least the crossover region has been widened.

I have run a few simulations, with inconclusive results. However it appears that using a linearising diode softens the upper knee as much as it softens the lower knee. The overall effect is to reduce the g_m slope in the crossover region, which might reduce higher-order harmonics a little.

Local feedback

There is another way to soften the lower knee. This is to add a resistor in parallel with Tr_1, such that under quiescent conditions nearly half of I_{c1} is diverted to the resistor.

The aim is to keep Tr_2 nearer the 'switch-on' point, and reduce the g_m of Tr_1 by reducing its current. At the low current extreme – i.e. output voltage near the opposite supply rail – Tr_1 will then have very little current so the g_m 'tail' is lengthened. At high currents, the resistor will have little effect. This is vaguely reminiscent of thermionic valve techniques known as 'lengthening the grid base' or 'ultralinear'.

There should be no effect on HF stability or slew rate because there is already a feedback path here at high frequencies via Tr_2 base-collector capacitance.

Again both the lower and upper knee are softened, but the picture looks a little more promising than the linearising diode. The crossover region is widened and made slightly smoother. Higher harmonics should be reduced and the quiescent current setting may be slightly less critical.

Unfortunately, the power supply rejection ratio will be impaired because the feedback signal reference is not a clean earth but a potentially dirty supply rail, so this technique will not always be appropriate.

It should be easy to add this resistor as a retro-fit to an existing amplifier. The value should be a little over $4 \times R \times V_{cc}$. The quiescent current will need to be reset to about twice the previous value in Tr_2, which will require a reduction in V_{bias} of about 10-15mV.

Don't try this on an amplifier still in warranty though, or if you are unable or unwilling to carry out any resultant repairs!

In summary

I started this whole train of thought when designing a birdy filter buffer for my FM tuner. I have concluded by gaining some insight into the behaviour of a popular class of output stage, which may prove useful when designing an amplifier that may one day partner the tuner.

Along the way I have offered guidance on choosing component values for Class-A stages, and confirmed that even a well designed Class B stage has residual crossover distortion. I have deliberately exposed the maths, because this enables other people to do similar things in different situations.

My thanks to Colin Homer by the way for commenting on an early draft of these articles. ■

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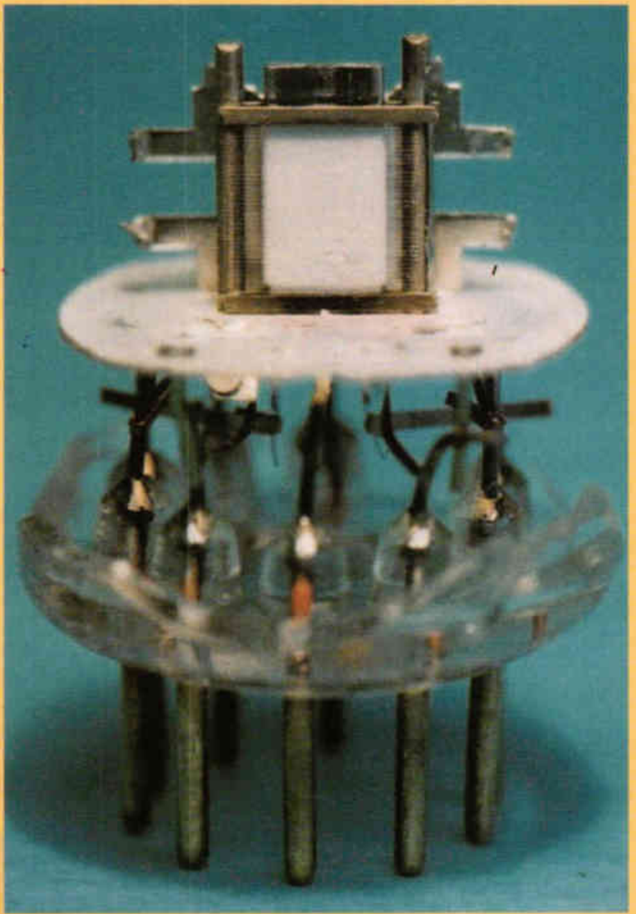

Morgan Jones had 29 unused but old valves drawing on average just over a milliamp of grid current. Such a high current is unacceptable for small-signal valves, so he devised a method of bringing them back to life – and it works remarkably well.

New life for OLD VALVES

Although the thermionic valve has now been obsolete for decades as a mainstream electronic device, there are still niche markets that use valves. As is well known, musicians use valve guitar amplifiers because of the distortion that can be produced under overload, whereas some hi-fi enthusiasts are willing to pay the high prices that valve equipment commands because of its legendary smoothness.

It is perhaps less well known that valves are extremely popular in recording studios, and that all the major microphone manufacturers feature at least one microphone with a valve head amplifier in their condenser microphone range.

Valve, or tube, microphones tend to be large capsule designs – around 18mm in diameter – and they are popular for vocals. Because a microphone level signal is, by definition, uncontrolled, it is not uncommon for recording engineers to want the entire vocal channel to have valve electronics until it reaches the channel fader. Consequently, other valve studio electronics includes outboard microphone channels, equalisers, and compressor/limiters.

Although the demand for audio valves has been demonstrated, in comparison with the electronics market as a whole, the audio valve market is minuscule. It is not worthy of significant investment.

Unfortunately, manufacturing valves is a high-technology enterprise, so production runs must be maximised, resulting in fewer than thirty different types currently being made. Worldwide, there is now only a handful of factories producing audio valves in significant quantities. Despite this, production runs are short, so quality control is difficult, and contemporary production engineers are having to rediscover the skills of their Cold War counterparts in the 1950s and 1960s.

The Cold War generated huge stockpiles of unused valves that are gradually released by governments. These valves are charmingly termed 'new old stock' – or NOS for short. There is a very great variety of NOS valves. Because these valves were made by manufacturers at the height of valve production, quality control is rarely an issue. As a result, many modern designers choose to specify NOS, rather than modern valves.

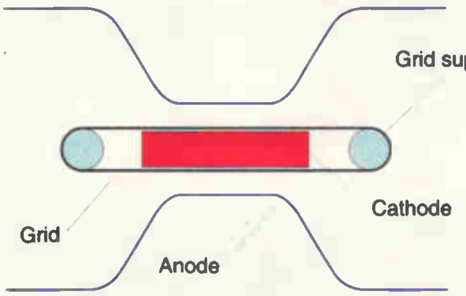

Fig. 1. Cross-sectional plan view of a 417A triode, drawn in proportion.[1]

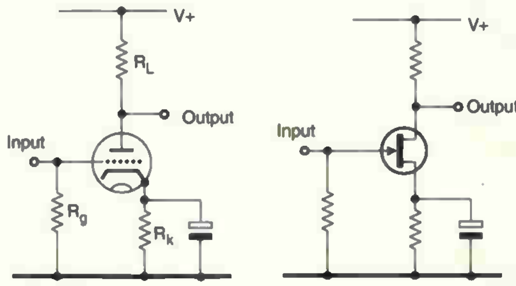

Fig. 2. Bias arrangements of thermionic triode, a), versus JFET, b). Both are voltage controlled and need a negative bias on their grid/gate to limit anode/drain current.

However, there can be problems with NOS valves. Scarcity can cause valves such as a NOS GEC KT66 to command a high price, so buyers will want to be sure that the product still meets its original specification – which is no mean feat after more than thirty years.

To understand one of the most common problems of NOS valves, it is worthwhile to briefly review the physics of the thermionic valve...

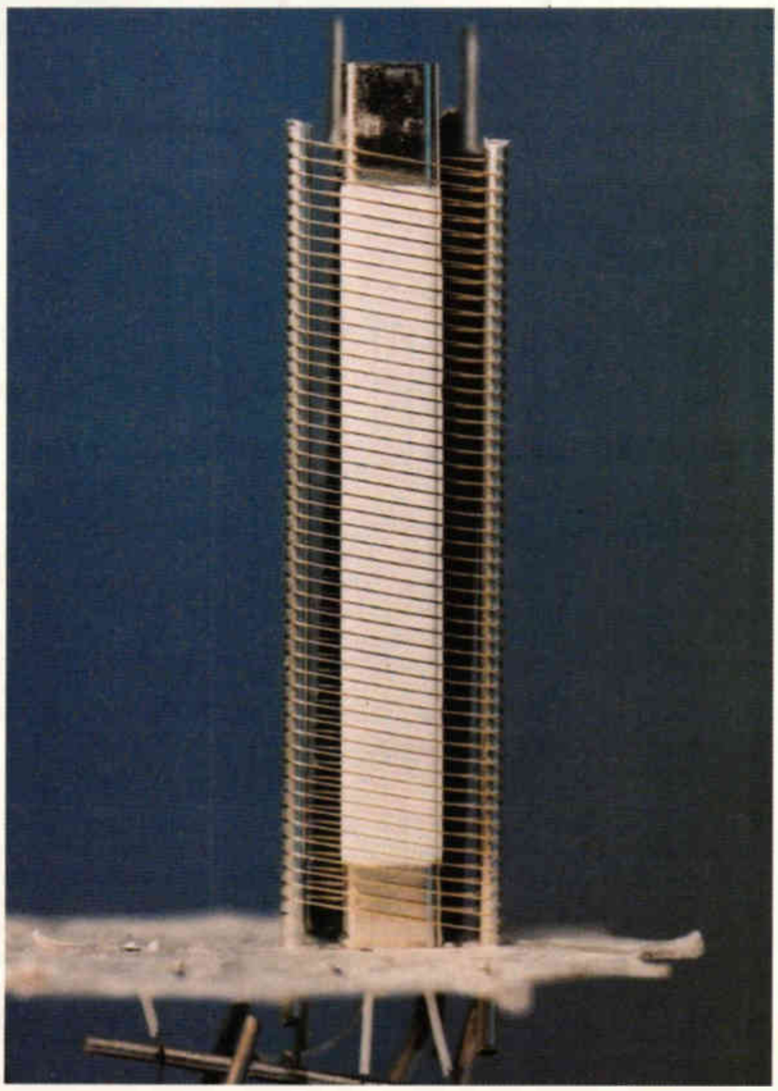

Fig. 3. 6080 power triode grid/cathode structure – note the gold plating on the grid to reduce grid emission.

Thermionic emission and diodes

All metals have free electrons within their crystal structure. Some of these electrons must be at the surface of the metal, but they are bound there by the nuclear forces between them and the adjacent atoms.

However, the atoms and electrons constantly vibrate due to thermal energy. If the metal is heated sufficiently, some free electrons may gain sufficient kinetic energy to overcome the attractive forces of the atoms and escape.

The heated metal in a valve is the cathode. When it is heated to a temperature determined by the work function of the metal, an electron cloud, or space charge, forms at its surface. Because electrons are negatively charged, and like charges repel, the cloud eventually attains a sufficient charge to prevent other electrons escaping from the surface, and an equilibrium is reached.

If a conductive plate, or anode, is placed some distance from the cathode, and charged to a positive voltage, electrons will be attracted from the cloud towards the anode. The electron cloud has now been depleted, and no longer repels electrons so strongly, so more electrons leave the surface of the cathode to replenish the electron cloud.

Current cannot flow in the opposite direction because only the cathode can emit electrons, and only the positive anode can attract electrons.

Electron velocity. At the instant that an electron leaves the cloud, it has almost zero velocity, but it is constantly accelerated by the electric field between the cathode and anode, and acquires energy proportional to the accelerating voltage,

$$E = q_e V = \frac{1}{2} m_e v^2$$

Rearranging, and solving for velocity,

$$velocity = \sqrt{2V\frac{q_e}{m_e}}$$

The ratio q_e/m_e is the electron charge/mass ratio. It has an approximate value of 1.76×10^{11}C/kg. If 100V is applied between the anode and cathode, the electrons will collide with the anode with a velocity of around 6×10^6m/s, or 13 million miles per hour. Note that the cathode-to-anode distance is immaterial because an infinite distance would allow an infinite time for acceleration, so the collision velocity would still be reached even if the rate of acceleration was very low.

When the speeding electrons are abruptly halted by the anode, their kinetic energy is converted into thermal energy. It is this heat that is responsible for the anode dissipation rating.

Many effects within valves can be understood by having an appreciation of the collision velocity of the electrons as they hit the anode.

Ionization current. The vacuum in a valve cannot be perfect, so there will always be stray gas molecules between the cathode and anode. As an electron nears the anode, it has considerable velocity. If it collides with a stray gas molecule, it may easily knock an electron from the molecule, which will promptly be captured by the anode.

Now, the gas molecule lacks an electron, so this positive ion is now accelerated towards the cathode in exactly the same way as the electrons were accelerated towards the anode.

Because a gas ion contains one, or more, neutrons, it is thousands of times heavier than an electron and does not attain the velocity of the electrons. However, if an ion manages to reach the cathode without being neutralised by a low-velocity electron, it collides with considerable momentum. It may be able to dislodge a molecule from the surface of the cathode. Unfortunately, the active surface of the oxide coated cathode in a small valve is only one molecule thick and therefore susceptible to damage by ion bombardment.

Table 1. Test results of triode-strapped D3a on VCM163 tester with an anode potential of 175V and –2V on th grid.

Valve No	I_a (mA)	g_m (mA/V)	I_g (µA) From the box	I_g (µA) After baking
1	15	28	1	0.13
2	21	35	1	0.20
3	6.5	-	soft	0.15
4	24	40	1.5	0.22
5	23	41	1.5	0.24
6	21	34	1	0.12
7	20	33	1	0.06
8	15	29	1	0.17
9	21.5	34	1	0.14
10	20	35	1	0.05
11	23	40	1	0.21
12	5	-	soft	1.20
13	15	27	1	0.22
14	25	40	1	0.19
15	21	32	1	0.22
16	16	28	1	0.30
17	15	29	1	0.17
18	4	-	soft	1.20
19	20	37	1	0.12
20	19	32.5	1	0.14
21	19.5	35	1.5	0.44
22	19.5	35	1	0.23
23	9	-	soft	1.3
24	22	33	1	0.24
25	22	36	1	0.27
26	22.5	38	1	0.19
27	3	-	soft	0.17
28	20	37	1	0.22
29	25	39	1	0.20
Average	20.25	34.5	1.06	0.195

The control grid

In order to control the flow of electrons from cathode to anode, and produce amplification, a grid of fine wires is placed between the cathode and anode, **Fig. 1**.

As you can see, to maximise its effect, the control grid is placed close to the cathode surface, where the velocity of the electrons is low, rather than near the anode, by which time the electrons have acquired considerable momentum, and are not easily repelled.

The control grid and ionization current. Valves are voltage-controlled devices and are operated in a similar manner to JFETs, in that the grid must be biased negatively with respect to the cathode in order to limit anode current,[2] **Fig. 2**.

A grid-leak resistor, whose value is typically around 1MΩ, holds the the grid at 0V. The value of the cathode bias resistor in conjunction with individual valve characteristics determines the anode current because it sets the grid-to-cathode voltage, V_{gk}.

From the point of view of a gas ion, the grid and cathode are at very nearly the same potential, so they are equally attractive. As a result, the probability of a gas ion striking the grid is largely determined by the relative dimensions of the grid wire diameter and its pitch, **Fig. 3**.

As you can see, the grid wire diameter is quite fine compared to the gaps between the wires, so most of the gas ions strike the cathode. Whether the ions strike the grid or the cathode, they are immediately discharged by a balancing number of electrons flowing up through the external paths to ground.

Even though little of the ionization current flows into the grid circuit, it is usually the grid ionization current that is significant, not the cathode ionization current. This is because the external grid circuit typically has a much larger resistance to ground.

Table 2. Test results of triode-strapped EF184 on VCM163 tester with 175V anode potential and –2V at the grid.

Quantity	I_g	R_{hk}	Comments
16	–	>25MΩ	No defects
1	0.5µA		
6	1µA		
3	2µA		
1	–	25MΩ	
1	–	12MΩ	low emission
1	–		low emission

Bias stability. If the ionization current is sufficiently large, it can develop a large enough voltage across the grid-leak resistor that V_{gk} is reduced, causing anode current to increase significantly.

Because power valves are usually operated at maximum anode dissipation, valve manufacturers specify a maximum value of grid-leak resistor to ensure that any change in V_{gk} due to

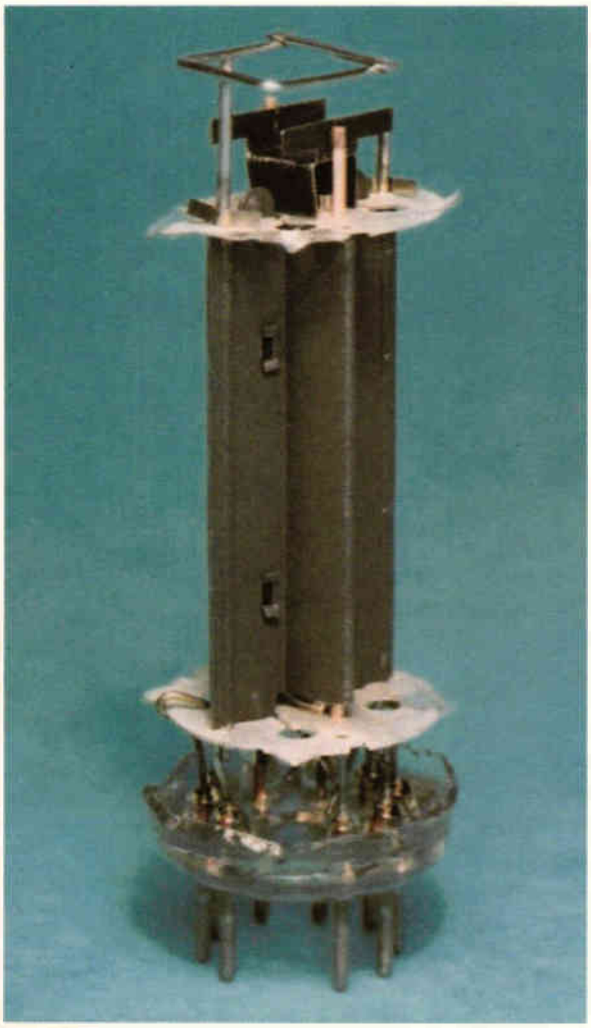

Fig. 4. A2293 power triode: the active getter material is in the trough completing the far side of the square loop aerial at the top of the structure.

grid ionization current is kept within safe limits.

Noise. Since the formation of ions and their subsequent discharge by the grid is random, the ionization current has a noise component. This current is converted by the grid-leak resistor into a noise voltage, and is amplified by the valve.

It is not unusual for high resistance circuits such as condenser microphone head amplifiers to use 500MΩ grid-leak resistors, so ionization current must be minimised to avoid degrading the signal-to-noise ratio.

Minimising ionization current

Clearly, the fewer gas molecules within the valve envelope, the lower the ionization current, so valve manufacturers strive to achieve as hard a vacuum as possible. During manufacture, the gas in the valve is pumped out, but some gas will remain that cannot be removed by pumps. This is removed by a structure called a 'getter'.

The getter is a metal structure, often near the top of the valve, coated with a highly volatile powder – usually a barium compound similar to the cathode emissive surface. Once the valve has been pumped and sealed, the getter is exploded, consuming the remaining gas, and oxidising some of the getter material. The force of the explosion throws molten barium onto the inside of the envelope to give the familiar mirrored coating.

The explosion is initiated electrically. A heating current may be directly passed a through the getter's metallic supporting structure in the case of metal-envelope valves. For glass envelope valves, the getter may be shaped as a short-circuited loop aerial. In this case, the heating current is induced via an external RF field, **Fig. 4**.

Some of the getter material is inevitably de-activated by the explosion. The remainder continues to consume gas molecules throughout the life of the valve because gas inevitably seeps into the valve vacuum. It seeps in either via the seals, where the leads leave the envelope, or by outgassing from a hot anode. Gas molecules must touch the getter to be consumed by it, and normal Brownian motion ensures this if the heater reaches operating temperature before HT is applied to the anode.

If the entire getter material is converted to oxide, perhaps because of a micro-fracture in the glass envelope, the mirrored coating turns white, so this is a clear evidence of catastrophic failure of the vacuum. Alternatively, very slightly gassy valves may still operate, but exhibit a gentle blue glow internally.

Softness

NOS valves are likely to have been sitting in a cold warehouse for at least twenty, and possibly fifty, years, so gas molecules will inevitably have permeated the envelope. In theory, the getter should consume the molecules to maintain the vacuum, but the most common cause of failure in unused valves is that the vacuum has deteriorated, or has gone 'soft'.

Valve testers. Because the grid ionization, or gas current, is a crucial parameter, and it is a known failure mode, most valve testers incorporate some means of quantifying this current. As an example, the Avo VCM163 allows the user to set anode and cathode conditions, then insert a 100μA meter in the grid circuit to measure the gas current.

Some power valves might be considered to be perfectly acceptable when passing 10μA of gas current into the grid circuit. However, 1μA would be unacceptable in a small signal valve whose grid was intended to be biased to –2V with a 1MΩ grid-leak resistor.

Results of valve tests

I recently bought a batch of twenty-nine Siemens D3a NOS valves. The D3a is a very high slope pentode. Its g_m is around 30mA/V, its maximum anode dissipation is 4W, and it is typically biased with a V_{gk} of around –2V. Strapped as a triode, its μ is 80. Together with its high g_m, this makes it an excellent candidate for a cathode follower.

I tested the batch for emission and gas at the proposed operating point on an AVO VCM163, and the raw results are shown in **Table 1**.

The results were disappointing. Five valves had very low emission, and the remainder passed a microamp or more of gas current. Worse, many valves exhibited unstable anode current. According to the date code on the envelope, namely 382, these NOS valves were made in 1982, so they were eighteen years old.

It seemed likely that the valves with very low emission failed due to manufacturing defects. The remaining valves were not quite acceptable, yet it seemed unlikely that they had gross defects, but more probable that the getter had failed to mop up leakage over the years.

Although operating the valves for a few hours might clear the residual gas, the anode current was unstable, and in that time, the fragile cathode would be bombarded, possibly reducing life expectancy.

I didn't want to risk damaging the new valves, so I decided to experiment on some expendable valves. The EF184 RF pentode can barely be given away, so I repeated the tests with a mixed batch of twenty nine NOS EF184 of differing ages and manufacturers. The results are summarised in **Table 2**.

There are well known 'fixes' for low-emission and poor heater-to-cathode resistance that will be covered later in this article, but I now had ten valves with the same symptoms as the D3a that I could afford to destroy.

Since chemical reactions double their

rate for every 10°C rise in temperature, perhaps the getter could be provoked into restoring the vacuum by heating the valves in an oven? Maximum envelope temperature is typically specified as 200°C, and domestic ovens are known to have erratic temperature control, so the gas oven was set to 'warm'.

Three hours later, the valves were removed and tested. Gratifyingly, there was a noticeable improvement, with all ten valves registering approximately half their previous gas current.

Clearly, the hypothesis was plausible, but the removal of gas seemed likely to be exponential. Reducing the residual gas and gas current to <1% of its original value would require more than 7×3=21 hours (2^7=128).

A cookery book that I consulted suggested that gas Mark 2 is around 130°C. As the oven temperature had previously been measured with a thermocouple probe as being around 100°C, this should give an eight-fold improvement without causing damage, and reduce the time required to 7 hours.

Although the AVO valve tester was capable of indicating the gas current and its improvement, it could not make an accurate measurement. So I made a break in its grid bias supply and substituted a Fluke 89 IV true RMS reading ammeter to enable a better reading. The true RMS reading DVM was necessary for comparison with the internal meter because valve testers test with half wave rectified AC, but gas current is read by a moving coil meter calibrated for DC.

All the D3a valves were added to the EF184, the oven was set to gas Mark 2 – later measured at 120°C – and the valves were left overnight for 13 hours. On removal, the valves were tested for gas current, and the results after baking were added to Table 1.

Although the gas current measurement before baking was necessarily somewhat inaccurate – 1μA on a 100μA FSD movement – the average improvement in gas current due to baking was a factor of five. While testing, I found that anode current and g_m were noticeably lower for all valves, but stable. They now agreed closely with a known good valve, suggesting that the previously high, and unstable, characteristics were a direct consequence of the gas current.

Although they showed an improvement in gas current, the very soft valves remained low emission and were set aside.

Baking tips

Valves that have no manufacturing defects, but have been in storage for many years may accumulate a little gas. It is possible to accelerate the getter and reduce gas current by a factor of five by heating the valve to 120°C in an oven for 12 hours, without any risk to the cathode.

Grid ionization current could easily be the dominant form of noise in high-resistance circuits, such as condenser microphone head amplifiers. As a result, it makes sense to routinely bake valves intended for this type of use *before* selecting for low noise.

However, because surface contamination on the glass envelope of the valve produces leakage paths that can cause noise, it is usual to clean the envelope scrupulously, and subsequently only handle the valve with cotton gloves[3]. This process should be done before baking, otherwise it might not be possible to remove any hardened contamination.

Unfortunately, as a consequence of the baking, the painted lettering on the valve shows a little discolouration, appearing as if the valve has had a few hours use.

Other valve 'fixes'

Although baking appears not to carry any risk of valve damage, the final two 'fixes' carry considerable risk...

Low R_{hk}. The consequences of low heater-to-cathode voltage depend greatly on the circuit and on V_{hk}. An EL34 output valve passing a cathode current of 70mA, with V_{hk} at +35V, and R_{hk} at 100kΩ will scarcely notice the extra 350μA leaking into the cathode. Any noise will be short-circuited by the cathode bypass capacitor.

Conversely, a cathode follower used in an active crossover requires a signal-to-noise ratio of at least 90dB. Although cathode resistance is low, it is not a short circuit. Worse, V_{hk} may well be elevated to 100V, so excellent R_{hk} is likely to be required.

Contamination causing poor R_{hk} can sometimes be burnt away on a valve tester. The tester is set to test R_{hk} with normal heater voltage and the cathode is allowed to warm to normal temperature. Heater-to-cathode resistance is then closely monitored and heater voltage is increased to 150%.

The resistance will fall, but if you are lucky, the rate of fall will slow – or even reverse. Assuming this effect occurs, immediately reduce the heater supply, and allow the valve to cool. The result should be an improved R_{hk}. Some valves cannot be recovered by this technique, and others may need repetition to make them acceptable, but the success rate is quite high.

The risk is that by deliberately overheating the cathode, some of the emissive surface may be evaporated and deposited onto the nearby grid. The grid now has its own emissive surface, and if the valve is operated close to maximum anode dissipation, it may become warm enough to emit electrons, causing thermal runaway, leading to the valve's ultimate destruction.

Low emission due to cathode poisoning. Valves that have been operated for a long time at very low anode currents are likely to develop a cathode interface resistance that effectively limits electron emission, but the 'fix' is violent.

The valve is heated with 150% heater volts, anode voltage is set to a slightly higher than normal value, and V_{gk} is adjusted until enough electrons are dragged from the cathode that the anode glows deep cherry red. The valve is left to fry for perhaps five seconds, before all voltages are removed.

With a bit of luck, the control grid has not been covered in evaporated cathode material. When tested a few minutes later, the valve might show better emission.

Rejuvenation carries a very high risk, and the results are not generally very good, so the process is only really worthwhile on picture valves. Dedicated television tube rejuvenators have been made, but the risk of destroying the tube is high. Nevertheless, tube replacement is expensive, so rejuvenation may be considered to carry an acceptable risk. ■

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2. 'Valve Amplifiers', Morgan Jones, 2nd Ed. 0-7506-4425-7.
3. 'Electrostatic cardioid microphone M7', BBC Technical Instructions (Recording), Instruction S.2 Section 5.3.

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VY electrolytic capacitors from Nichicon have working voltages from 6.3 to 100V (–55 to +105°C), 160 to 400V (–40 to +105°C) or 450V (–25 to +105°C). With a laminated case the devices have capacitances of 0.1 to 68 000µF, a tolerance of 120 per cent and allowable ripple from 1.3 to 3575mA, depending on capacitance and working voltage. They can be supplied with formed or straight taped leads, and those rated up to 100V are impervious to most cleaning solvents.
Nichicon
Tel: 01276 685393

Embedded PC with 500MHz PIII

Digital-Logic has launched the Microspace MSMP5SEV PC/104plus embedded computer. This module has a Pentium III processor with a clock frequency of 500MHz. Functions include 16Mbyte DRAM expandable to 128Mbyte, 256kbyte L2 cache and 10/100Mbit/s interface for communication in LANs. The module has a buffered 3.3 to 5V interface for LCD connection without an adapter, video input for connecting up to three cameras and a bios with APM support. Features include integrated RTC battery. EEPROM support and system interfaces for keyboard, mouse, hard disk and floppy drive. The board contains a watchdog, com1 and com2 ports, printer interface and USB interface. There is a slot into which a Compactflash can be installed as a hard-disk replacement. The SXGA-69000 video controller and 2Mbyte of video RAM provide images with 256 colours and resolutions up to 1280 x 1024 pixels on the connected displays. The module has a CPU heatsink for passive or active cooling. It can be operated with a 5V, 1.5A power supply between –25 and +60°C.
Digital-Logic
Tel: 00 41 32 681 58 00

CompactPCI controller

Radisys has launched the EPC-3322 CompactPCI system controller for telecoms and computer-intensive carrier-class applications. Powered by a single or dual Intel Pentium III FC-PGA processor configuration running at 600MHz and an Intel 440BX PCI chipset, the controller supports up to 1Gbyte of PC100 SDRAM.
Radisys
Tel: 01793 411200

Film-foil capacitors

ICW has extended its range of Alcon film-foil capacitors. The electrodes on the FF-06 and FF-12, for example, are made from aluminium foil. An extended foil winding technique is used to reduce self-inductance, making them suitable for applications up to 10kHz. They are available from 1nF to 2.25F and in voltages up to 3kV DC (1.2kV AC at 50Hz).
Industrial Capacitors Wrexham
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Vectron has introduced a surface-mount temperature-compensated crystal oscillator for frequencies from 2 to 52MHz. The TO-700 TCXO incorporates an Asic temperature compensation method and uses a sealed crystal design. The oscillator can achieve ageing of less than 1ppm/year with stability to ±1ppm at –40 to +85°C. It comes in packages down to 13.95 x 9.25 x 6mm high depending on the type of output and board mounting required.

46cm TFT LCD desktop monitor

A 46cm TFT LCD desktop monitor from Kent Modular Electronics is SXGA (1280 x 1024 pixels) compatible but will autosize any input frequency down to VGA (640 x 480 pixels). It comes in a metal desktop style cabinet that can withstand a 1.6kg pushover force. This is helped by mounting the power supply in the base of the unit and an industrial clamping mechanism at the rear of the screen. It is suitable for the process control environment and for use with a touchscreen. The TFT panel is protected by laminated safety glass. The monitor has front mounted controls and an onscreen setup display, and is available with a standard VGA or a VGA and PAL/NTSC inputs. The internal control board provides a direct connection to a PC via the VGA input. In the VGA and PAL/NTSC version, the monitor can accept the input of two signals simultaneously, for example an input from a computer and a camera or VCR. The monitor can be set to switch automatically between sources on a priority basis showing for example XGA if no PAL signal is present.

Kent Modular Electronics
Tel: 01634 830123

Available outputs are sinewave and HCMOS requiring a supply voltage of +5V DC and LVHCMOS, requiring a supply voltage of +3.3V DC. Communications applications include ISDN, ATM, Sonet, SDH, GSM, CDMA, mobile radio and network systems.
Vectron
Tel: 01703 766288

Upconverter MMIC

Agilent has released the HPMX-7202 dual-band, trimode upconverter MMIC for mobile phone handsets that operate with 800 and 1900MHz CDMA and AMPS services. It

incorporates adjustable gain on the CDMA drivers. The PCS and cellular transmit chains provide an adjacent channel power rejection of typically –58dBc 30kHz at an output of +9dBm for PCS and –55dBc 30kHz at +7dBm for CDMA. The noise floor complies with TIA98-C requirements. The CDMA drivers have output power up to +10dBm for direct interfacing with CDMA power amplifiers. The drivers are adaptively biased to reduce current consumption and extend battery life.
Agilent Technologies
Tel: 00 49 6441 92460

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Molex is to launch an RJ45 jack with electrostatic protection and integrated magnetic shielding for Ethernet 10baseT and 100baseT hub switches. The Hyperjack has an integrated electrostatic discharger letting the user connect an electrostatically charged cable during use without damaging the physical layer. This has been achieved by adding a conductive elastomeric grommet to every port, which is connected to the housing shield and insulated from the jack connectors. The plug casing discharges static through the grommet before making a connection. This is integrated into 6 by 2 and 8 by 2 stacked ganged versions. Features include integrated shielding for common mode rejection, DC insulation of cable side versus board side and possible transforming of the signal and voltage depending on the configuration of the physical layer. Integrated resistors and capacitors terminate any unwanted noise from unused pairs and handle the DC share of the common mode noise of the signal pairs.
Molex
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Radial electrolytic capacitors made by Cross Link Manufacturing are available from CPS. The S5, SS, SM, SL, SK, SH, SB, SD, SC, SF and SN devices have capacitances from 0.1 to 10 000μF and voltages from 4 to 450V. The ranges include low-impedance capacitors for SMPS, high-frequency and pulse applications, and bipolar and

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HP Frequency comb generator type 8406 - £400.
HP Sweep Oscillators type 8690 A+B + plug-ins [illegible] 18GHz also 18-40GHz.
HP Network Analyser type 8407A + 8412A + 8601A - [illegible] 110Mc/s - £500 - £1000.
HP 8410-A-B-C Network Analyser [illegible] 18 GHz - plus most other units and [illegible] set-up - 8411a-8412-8413-8414-8418-[illegible]-8650. From £1k.
Racal/Dana 9301A [illegible] 1.5-2GHz - [illegible] stock £250-£[illegible]
Racal/Dana M[illegible] Meter Type 9009-9008 [illegible] 1.5GHz - £150-£250 [illegible]
Marconi RCL Bridge type TF2700 - £150.
Marconi Microwave 6600A 1 sweep osc., mainframe with [illegible] 6.5 GHz or 6651 PI - 26.5-40GHz-£750 or PI only £600 [illegible] only £250.
Gould J3B test oscillator + manual - £150.
Marconi 6155A Signal Source-1 to 2GHz - LED [illegible]
Barr & Stroud Variable filter EF3 0.1Hz-100Kc/s + high pass + low pass - £150, other makes in stock.
Racal/Dana 9300 RMS voltmeter - £[illegible]
HP 8750A storage normalizer [illegible] N, A
Marconi mod meters type TF [illegible] - £250 - TF[illegible]5 - £1,000.
Racal/Dana counters [illegible] 9906-99[illegible]-9916-9917-9[illegible] 50Mc/s-3GHz - £100 - [illegible] fitted with FX standards.
HP180TR, HP181T, HP182T mainframes £300 - £500.
HP432A-435A or B-436A-[illegible] meters + powerheads to [illegible] - £150 - £1750 - spare heads available.
HP3586A or C selective level meter - £500.
HP86222A+B Sweep PI -01-2.4GHz + ATT £1000-£1250.
HP86290A+B Sweep PI-2 - 18GHz - £1000 - £1250.
HP8620C Mainframe - £250. IEEE £350.
HP8165A Programmable signal source - 1MHZ - 50Mc/s - £1k
HP3455/3456A Digital voltmeter - £400.
HP5370A Universal time interval counter - £1k
HP5335A Universal counter - 200Mc/s-£1000.
HP3552A Transmission test set - £350.
TEKTRONIX 577 Curve tracer + adaptor - £[illegible]
TEKTRONIX 1502/1503 TDR cable test [illegible]
HP8699B Sweep PI YIG oscillator .01 - [illegible] £[illegible]0. 86908 MF- £250. Both £500.
Dummy Loads & Power att up to 2.5 kilowatts FX up to 18GHz - microwave parts new and ex equipt - relays - attenuators - switches - waveguides - Yigs - SMA - APC7 plugs - adaptors etc. qty. in stock.
B&K Items in stock - ask for list.
Power Supplies Heavy duty + bench in stock - Farnell - HP - Weir - Thurlby - Racal etc. Ask for list. Large quantity in stock, all types to 400 amp - 100Kv.
HP8405A Vector voltmeter - late colour - £400.
HP8508A Vector voltmeter - £2500.
HP8505A Network Anz 500KHz-1.3GHz - £1000.
HP8505A + 8502A or 8503A test sets- £1[illegible]
HP8505A + 8502A or 8503A + 8501A [illegible] - £1750-£2000.
Phillips 3217 50Mc/s oscilloscopes - [illegible]
Wavetek-Schlumberger 4031 Radio communication test set

LIGHT AND OPTICAL EQUIPMENT
Anritsu ML93A & Optical L[illegible] Meter - £250.
Anritsu ML93B & Optical Le[illegible] Meter - £350.
Power Sensors for above MA96A - MA98A - MA913A - Battery Pack MZ95A.
Anritsu MW97A Pulse Echo Tester.
PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.
Anritsu MW98A Time Domain Reflector.
PI available - MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.
Anritsu MZ100A E/O Converter.
+ MG912B (LD 1.35) Light Source + MG92B (LD 0.85) Light Source £350.
Anritsu MZ118A O/E Converter.
+MH922A 0.8 O/E unit + MH923 A1.3 O/E unit £350.
Anritsu ML96B Power Meter & Charger £450.
Anritsu MN95B Variable Att. 1300 £100.
Photo Dyne 1950 XR Continuous Att. 1300 - 1500 £100.
Photo Dyne 1800 FA. Att £100.
Cossor-Raytheon 108L Optical Cable Fault Locator 0-1000M 0-10kM £200.
TEK P6701 Optical Converter 700 MC/S-850 £250.
TEK OF150 Fibre Optic TDR - £750.
HP81512A Head 150MC/S 950-1700 £250.
HP84801A Fibre Power Sensor 600-1200 £250.

HP8158B ATT OPT 002+011 1300-1550 £300.
HP81519A RX DC-400MC/S 550-950 £250.
STC OFR10 Reflectometer - £250.
STC OFSK15 Machine jointing + eye magnifier - £250.

COMMUNICATION EQUIPMENT
Anritsu ME453L RX Microwave ANZ - £350.
Anritsu ME453L TX Microwave ANZ - £350.
Anritsu MH370A Jitter Mod Oscillator - £350.
Anritsu MG642A Pulse Patt Gen. £350.
System MS02A Timer & Digital Printer - £500.
Complete MS65A Error Detector.
Anritsu ML612A Sel Level Meter - £400.
Anritsu ML244A Sel Level Meter - £300.
W&G PCM3 Auto Measuring Set - £300.
W&G SPM14 Sel Level Meter - £300.
W&G SPM15 Sel Level Meter - £350.
W&G PS19 Level Gen - £500.
W&G DA20+DA21 Data ANZ £400
W&G PMG3 Transmission Measuring Set - £300.
W&G PSS16 Generator - £[illegible]
W&G PS14 Level Generator - £[illegible]
W&G EPM-1 P[illegible] Head M[illegible]wat Power Meter - £450
W&G DLM[illegible] Phase Jitter & Noise - £350
W&G DU[illegible] Data Line Test Set - £400.
W&G PS1[illegible] PM10 Level Gen. - £250.

MISCELLANEOUS ITEMS
HP 3852A Data Acquisition Control Unit + 44711A 16ch input £[illegible]
HP 4261 LCR meter - £650
HP 4274 EX LCR meter - £1,500.
HP 4951A Protocol ANZ - £500.
HP [illegible] Switch Control Unit + PI Boards - £500.
HP 75000 VXI Bus Controllers + E1326B-D[illegible] quantity.
HP 8922A GSM DCS/PCS 1805-1880MC/S converter for use [illegible] - £2,000.
HP 1630-1631-1650 Logic ANZ's in stock.
HP 8754A Network ANZ 4-1300MC/S + 8502A + cables - £1,500.
HP 8754A Network ANZ H26 4-2600MC/S + 8502A + Cables £2,000.
HP 8350A Sweeper MF + 83540A PI 2-8.4GHZ + [illegible] 12.4GHZ [illegible] - £2,500.
HP MICROWAVE TWT AMPLIFIER 489A 1-2GHZ - [illegible]B - £400.
HP PREAMPLIFIER 8447A 0.1-400MC/S - £200. Dual - £300.
HP PREAMPLIFIER 8447D 0.01-1.3GHZ - £400.
HP POWER AMPLIFIER 8447E 0.01-1.3GHZ - £400.
HP PRE + POWER AMPLIFIER 8447F 0.01-1.3GHZ - £500.
HP 3574 Gain-Phase Meter 1HZ-13MC/S OPT 001 Dual - £400.
MARCONI 2305 Modulation Meter-50KHZ-2.3 GHZ - £3,000.
MARCONI 2610 True RMS Meter - £450
MARCONI 893B AF Power Meter (out S and filter) - £250-£350.
MARCONI 6950-6960B Power Meters + Heads - £500-£900.
MARCONI SIGNAL SOURCE 6055-6056-6057-6058-6059 - FX Range 4-18GHZ - £250-£400.
RACAL 1792 COMMUNICATION RX - £500 early - £1,000 - late model with fast tuning and bite test.
RACAL 1772 COMMUNICATION RX - £400-£600.
PLESSEY PR2250 A & H COMMUNICATION RX - £500-£800.
TEK MODULE MAINFRAMES - TM[illegible] 502-503-504-506 [illegible]
TEK PI 5010-M1 - Prog Multi Interface - £250. FG Prog 20MC/S Function Gen - £400 - S1 Prog Scanner - £250 - DM Prog DMM - £400.
TEK 7000 OSCILLOSCOPE MAINFRAMES - 7603-7623-7633-7834-7[illegible]-7904A-7104 - £150-£1,000.
TEK 7000 PI's - 7A11-7A12-7A13-7A18-7A19-7A22-7A24-7A26-[illegible]-7B10-7B15-7B53A-7B80-7B85-7B92A-7D15-7D20. TEK 7000 [illegible]-7S12-7S14-7M11-7S1-S2-S3-S4-S5-S6-S61-S52-S53.
HP POWER SUPPLIES - 6621A-6623A-6624A-6632A-6652A. Qtys available. Also other types £POA.

RADIO COMMUNICATION TEST SETS
BULK PURCHASE ONLY FROM JOHNS RADIO
HP 8920A RF Communication Test Sets - Opts available 001-3-4-5-7-11-12-14 H13-K13. £1,500-£1,750.
HP 8920A with opt 002 Spectrum anz plus tracking generator plus opts. 001-3-4-5-11-12-014 available in part includes syn sig generator - digital oscilloscope distortion meter - mod meter - RF power meter etc. £2,500

SPECIAL OFFERS
MOTOROLA R2600A plus RLN4260A RF Test Set - £1,800.
MARCONI 2955 RF Test Sets-1000MC/S - £1,000 each.
MARCONI 2958 RF Test Sets-1000MC/S - £1,000 each.
MARCONI 2960 RF Test Sets-1000MC/S - £1,250 each.
MARCONI 2955A RF Test Sets-1000MC/S - £1,500 each.
MARCONI 2960A RF Test Sets-1000MC/S - £1,500 each.
ANRITSU MS555A2 Radio Comm Anz-1000MCs - £750 each.
MARCONI 2019A SYNTHESIZED SIGNAL GENERATORS - 80KC/S-1040MC/S - AM-FM - £400 inc. instruction book - tested.
MARCONI 2022E SYNTHESIZED SIGNAL GENERATOR - 10KC/S-1.01GHZ AM-FM - £500 inc. instruction book - tested.
R&S APN 62 LF Sig Gen 0.1Hz - 260KHz c/w book - £250.

WE KEEP IN STOCK HP and other makes of RF Frequency doublers which when fitted to the RF output socket of a S/Generator doubles the output frequency EG.50-1300MC/S to 50-2600MC/S price from £250 - £450 each.

SPECTRUM ANALYZERS
HP 3580A 5HZ-50KHZ - £750.
HP 3582A Dual 0.2HZ-25.5KHZ - £1,500.
HP 3585A 20HZ-40MC/S - £3,500.
HP 3588A 10HZ-150MC/S - £7,500.
HP 8568A 100HZ-1.5GHZ - £3,500.
HP 8568B 100HZ-1.5GHZ - £4,500.
HP 8590B 9KC/S-1.8GHZ - £4,500.
HP 8569B 10MC/S (0.01-22GHZ) - £3,500.
HP 3581A Signal Analyzer 15HZ-50KHZ - £400.

TEK491 10MC/S-12.4GHZ + 12.4-40GHZ - £500.
TEK492 50KHZ-21GHZ OPT 2 - £2,500.
TEK492P 50KHZ-21GHZ OPT 1-2-3 - £3,500.
TEK492AP 50KHZ-21GHZ OPT 1-2-[illegible]
TEK492BP 50KHZ-21GHZ - £[illegible]000-£4,000.
TEK495 100KHZ-1.8GHZ - £[illegible]
HP 8557A 0.01MC/S-350MC/S - £500 + MF180T or 180C - £150 - 182T - £500.
HP 8558B 0.01-1500MC/S - £[illegible] - MF180T or 180C - £150 - 182T - £500.
HP 8559A 0.01-21GHZ - £1,000 - MF180T or 180C - £150 - 182T - £500.
HP 8901A AM FM Modulation ANZ Meter - £800.
HP 8901B AM FM Modulation ANZ Meter - £1,750.
HP 8903A Audio Analyzer - £1,000.
HP 8903B Audio Analyzer - £1,500.
MARCONI 2370 SPECTRUM ANALYZERS - HIGH QUALITY - DIGITAL STORAGE - 30HZ-110MC/S Large qty to clear as received from Gov - all sold as is from pile complete or add £100 for basic testing and adjustment - callers preferred - pick your own from over six hundred - discount on qtys of five.
A EARLY MODEL GREY - Horizontal alloy cooling fins - £200.
B LATE MODEL GREY - vertical alloy cooling fins - £300.
C LATE MODEL BROWN - as above (few only) - £500.

OSCILLOSCOPES
TEK 465-465B 100MC/S + 2 probes - £250-£300.
TEK 466 100MC/S storage + 2 probes - £200.
TEK 475-475A 200MC/S-250MC/S + 2 probes - £300-£350.
TEK 2213-2213A-2215-2215A-2224-2225-2235-2236-2245-60-100MC/S - £250-£400.
TEK 2445 4ch 150MC/S + 2 probes - £450.
TEK 2445A 4ch 150MC/S + 2 probes - £600.
TEK 2445B 4ch 150MC/S + 2 probes - £750.
TEK [illegible]8B D.S.O. 100MC/S + 2 probes - £500.
TEK [illegible] 350MC/S + 2 probes - £650.
TEK 2465 4ch-300MC/S - £[illegible]
TEK 2465A 4ch-350MC/S - £[illegible]
TEK 2465ACT 4ch-350MC/S - £[illegible]50.
TEK D.S.O. 2230 - 100MC/S + 2 probes - £1,000.
TEK D.S.O. 2430 - 150MC/S + 2 probes - £1,250.
TEK D.S.O. 2430A - 150MC/S + 2 probes - £1,750.
TEK D.S.O. 2440 - 300MC/S + 2 probes - £2,000.
TEK TAS 475-485 - 100MC/S-20MC/S-4 ch + 2 probes - £900-£1.1K.
HP1744A - 100MC/S + 2 probes - £250.
HP1741A - 100MC/S storage + 2 probes - £200.
HP1720A - 1722A - 1725A - 275MC/S + 2 probes - £300-£[illegible]
HP1744A - 100MC/S storage - large screen - £250.
HP1745A - 1746A - 100MC/S - large screen - £350.
HP54100A - 1GHz digitizing - £500.
HP54200A - 50MC/S digitizing - £500.
HP54501A - 100MC/S digitizing - £[illegible]
HP54100D - 1GHZ digitizing - £1,[illegible]

MICROWAVE COUNTERS - ALL LED READOUT
EIP 351D Autohet 20Hz-18GHz - £750.
EIP 371 Micro Source Locking - 20Hz-18GHz - £850.
EIP 451 Micro Pulse Counter - 300MC/S-18GHz - £700.
EIP 545 Microwave Frequency Counter - 10Hz-18GHz - £1K.
EIP 548A Microwave Frequency Counter - 10HZ-26.5GHz - £1.5k.
EIP 575 Microwave Source Locking - 10Hz-18GHz - £1.2K.
EIP 588 Microwave Pulse Counter - 300MC/S-26.5GHz - £1.4K.
SD 6054B Micro Counter 20HZ-24GHZ - SMA Socket - £800.
SD 6054B Micro Counter 20HZ-18GHZ - N Socket - £700.
SD 6054D Micro Counter 800MC/S-18GHz - £600.
SD 6246A Micro Counter 20Hz-26GHz - £1.2K.
SD 6244A Micro Counter 20Hz-4.5GHz - £400.
HP5352B Micro Counter OPT 010-005-46GHz - new in box - £5k.
HP5340A Micro Counter 10HZ-18GHz - Nixey - £500.
HP5342A Micro Counter 10HZ-18-24GHz - £800-£1K - OPTS 001-002-003-005-011 available.
HP5342A + 5344S Source Synchronizer - £1.5K.
HP5345A 500MC/S 11 Digit LED Readout - £400.
HP5345A + 5354A Plugin - 4GHz - £700.
HP5345A + 5355A Plugin with 5356A 18GHz Head - £1K.
HP5385A 1GHz 5386A-5386A 3GHz Counter - £1K-£2K.
Racal/Dana Counter 1991-160MC/S - £200.
Racal/Dana Counter 1992-1.3GHz - £600.
Racal/Dana Counter 9921-3GHz - £350.

SIGNAL GENERATORS
HP8640A - AM-FM 0.5-512-1024MC/S - £200-£400.
HP8640B - Phase locked - AM-FM-0.5-512-1024MC/S - £500-£1.2K. Opts 1-2-3 available.
HP8654A - B AM-FM 10MC/S-520MC/S - £300.
HP8656A SYN AM-FM 0.1-990MC/S - £900.
HP8656B SYN AM-FM 0.1-990MC/S - £1.5K.
HP8657A SYN AM-FM 0.1-1040MC/S - £2K.
HP8657B SYN AM-FM 0.1-2060MC/S - £3K.
HP8660C SYN AM-FM-PM-0.01-1300MC/S-2600MC/S - £2K.
HP8660D SYN AM-FM-PM-0.01-1300MC/S-2600MC/S - £3K.
HP8673D SYN AM-FM-PM-0.01-26.5 GHz - £12K.
HP3312A Function Generator AM-FM 13MC/S-Dual - £300.
HP3314A Function Generator AM-FM-VCO-20MC/S - £600.
HP3325A SYN Function Generator 21MC/S - £800.
HP3325B SYN Function Generator 21MC/S - £2K.
HP8673-B SYN AM-FM-PH 2-26.5 GHz - £6.5K.
HP3326A SYN 2CH Function Generator 13MC/S-IEEE - £1.4K.
HP3336A-B-C SYN Func/Level Gen 21MC/S - £400-£300-£500.
Racal/Dana 9081 SYN S/G AM-FM-PH-5-520MC/S - £300.
Racal/Dana 9082 SYN S/G AM-FM-PH-1.5-520MC/S - £400.
Racal/Dana 9084 SYN S/G AM-FM-PH-.001-104MC/S - £300.
Racal/Dana 9087 SYN S/G AM-FM-PH-.001-1300MC/S - £1K.
Marconi TF2008 AM-FM Sweep 10KC/S-510MC/S - £200 Fully Tested to £300, as new + book + probe kit in wooden box.
Marconi TF2015 AM-FM-10-520MC/S - £100.
Marconi TF2016A AM-FM 10KC/S-120MC/S - £100.
Marconi TF2171/3 Digital Synchronizer for 2015/2016A - £50.
Marconi TF2018A AM-FM SYN 80KC/S-520MC/S - £500.
Marconi TF2019A AM-FM SYN 80KC/S-1040MC/S - £650-£1K.
Marconi TF2022E AM-FM SYN 10KC/S-1.01GHz - £1K-£1.2K.
R & S SMPD AM-FM-PH 5KHz-2720MC/S - £3K.
Anritsu MG3601A SYN AM-FM 0.1-1040MC/S - £1.2K.

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Johns Radio, Whitehall Works, 84 Whitehall Road East, Birkenshaw, Bradford BD11 2ER. Tel: (01274) 684007. Fax: 651160

Please quote *Electronics World* when seeking further information

40V switching regulator

The TC2574 regulator is a monolithic IC for step-down switching circuits. Available from DT Electronics with a choice of nominal 3.3, 5 or 12V outputs, the device can operate with a DC input from 4.75 to 40V and needs four external standard components to complete its function. The output voltage is within ±4 per cent and the 52kHz internal oscillator within ±10 per cent. An adjustable version is available with an output voltage of 1.23 to 3.7V. The output switch includes cycle-by-cycle current limiting and thermal shutdown. Current limit protection is included and there is a standby mode with 60μA typical current consumption. In most cases the copper tracks on the PCB are the only heatsinks needed. Applications include step-down regulation, pre-regulator for linear regulation, on-card switching regulation, positive to negative conversion, negative step-up conversion and power supplies for battery chargers.

DT Electronics
Tel: 024 7643 7400

HTN LCDs

HTN LCDs from Anders come with options from glass alone to a module with support electronics such as driver and backlight. Operating range is between –40 and +85°C at an optimum multiplex rate between 0.25 and 0.125. Molecular twist is 110°. Operating voltage is 2.5V, making them suitable for hand-held applications, such as a blood glucose meter from Hypoguard, which operates at 2.9V and 0.25mux.

Anders Electronics
Tel: 0207 388 7171

Custom connectors

Samtec's custom connector capabilities include design and development from simple modifications of standard board-to-board connectors to custom IC-to-board connections. Applications include custom and modified pins and insulators for board stacking, high speed, power, one-piece and card-to-board connectors. Connectors can be developed for automotive and other applications. In-house plating and mould-making operations can be used to supply platings or make minor changes to tooling. Stereo lithography and wire ADM are used to develop prototypes for design and mechanical studies.

Samtec
Tel: 0236 739292

SIM connector

ODU designs and makes connectors for mobile phone SIM card readers and chip card readers. Units include sliding contacts with PCB pressure contacts, in-cover contacts and SMT contacts. Other products for mobile communications and instrumentation include connectors for speakers, LCDs and miniature board-to-board products.

ODU
Tel: 0116 230 1457

Switch-mode power supply

Approved to international standards, a switch-mode power

4-bit micros target telecoms

Flint has introduced two 4-bit microcontrollers for telecoms applications. The Winbond W921 and W742 have dual clock operation, stop mode and power management features. Typical uses include line-powered feature-phones, hand-held instruments, metering and logging systems. Both provide a programmable DTMF tone generator, configurable for clock sources from 400Hz to 4MHz. The W921 has between 21 and 64 lines of I/O, some of which may be used to sink up to 15mA per port. The W742 provides between 24 and 32 I/Os. Included is a clock-synchronous serial port. The W921 provides beep tone generators and multiple timer channels, which can directly produce PWM outputs via an arbitrary waveform generator circuit. For LCD applications, the W742 can drive displays up to 40 x 16 pixels and has an MFP output for modulated output waveforms.

Flint Distribution
Tel: 01530 510333

Please quote *Electronics World* when seeking further information

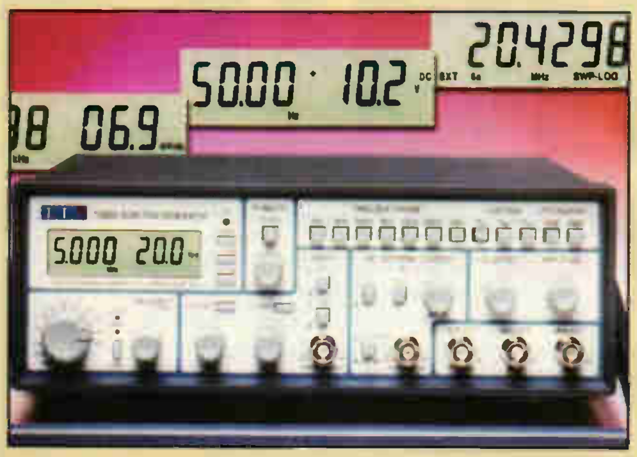
supply from Oxford Electrical Products produces 36W from a 110 by 50 by 20mm package weighing 200g. Ripple is down to 1 per cent. For portable devices such as laptops and mobiles, the supply comes in eight versions with outputs from 5 to 48V. Each has continuous short circuit protection and EMI suppression, and meets UL, CSA, TUV and CE standards. Universal input is 90 to 264 V AC.
Oxford Electrical Products
Tel: 01993 883214

Narrow-band 433MHz radio modules

Providing 34 channels, narrow-band transmitter and receiver modules from Quasar let designers use the licence-exempt 433MHz band. For one-to-one and multinode applications requiring radio links at up to 20kbit/s up to 1km, the QBT37 transmitter and QBR37 receiver operate between 433.075 and 434.725MHz in 50kHz steps. They can operate even in the presence of wideband 433MHz radio equipment. Applications include building security, panic attack alarms, wireless networking, in-vehicle telemetry systems, car security and industrial process monitoring. Each comes in an SIL package and operates from a 5V supply. Maximum transmitter current is 22mA and data input can be a 5V digital signal or a 3V peak-to-peak analogue signal. The receiver has ceramic filters to improve adjacent channel rejection and a CMOS-compatible output that can directly drive external decoders. Adjustments for transmit and receive frequency and squelch are built-in.
Quasar
Tel: 01993 883214

Front-end chipset for digital satellite

Insight has launched a front-end chipset for digital broadcast satellite applications. The Conexant CX4108 silicon tuner and CX24110 QPSK demodulator are for set-top box and PC receiver card applications that operate from a 5V supply. It is for use with the CX22490/1 interactive TV decoder. The demodulator has integrated forward-error correction. The chipset's 5V operation lets manufacturers design products that attach directly to the PC's internal PCI bus without the need for a DC-to-DC converter.
Insight Memec
Tel: 01296 330061

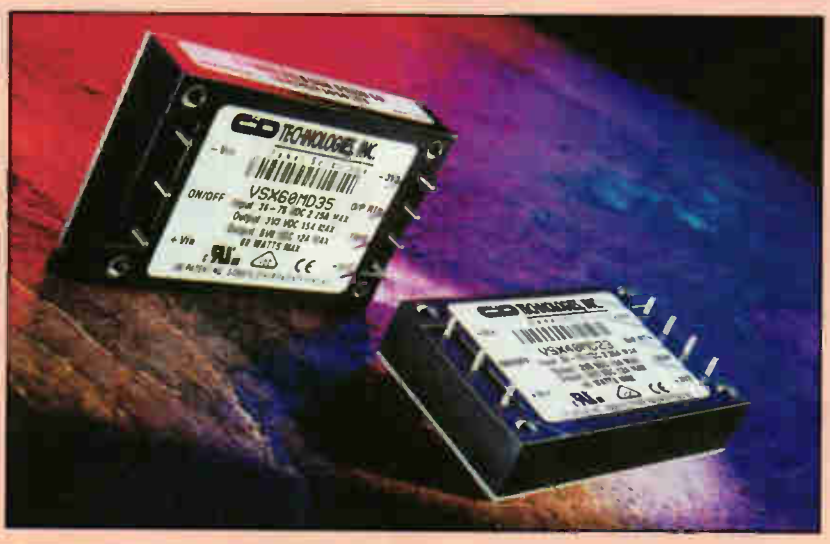


5MHz generator features dual display

The TG550 from Thurlby Thandar is a 5MHz function generator that provides simultaneous digital display of frequency and output. Auto-ranging reciprocal frequency measurement gives four-digit resolution with rapid update down to 2Hz. Accuracy is within ±1 digit (±0.02 per cent at full scale). To maintain a fast update below 2Hz, the measurement mode is changed to one based on the VCO control voltage. The output level display can be selected to show any of three values - peak-to-peak amplitude, RMS amplitude or DC offset. The decimal point and units are changed automatically, resulting in a display of the true amplitude regardless of the attenuator setting. It generates sine, square, triangle, pulse and ramp waveforms from 0.005Hz to 5MHz. A built-in sweep generator provides linear or logarithmic sweeps with periods from 20ms to 20s. AM capability is incorporated, with a choice of internal or external modulation from 0 to 100 per cent. A digital frequency locking system improves frequency stability. Once the frequency has been set, pressing the lock key engages a measure-and-correct circuit that compensates for thermal and mechanical drifts. The frequency is compared with a crystal controlled reference and is maintained to an accuracy of better than one digit on the display.
Thurlby Thandar Instruments
Tel: 01480 412451

60W dual-output DC-to-DC converter

C&D Technologies has introduced a 60W dual-output DC-to-DC converter measuring 5.84 by 3.81 by 1.27cm. The VSX60 includes input filtering, remote on-off with positive or negative logic options, output voltage trim, output current limit and built-in non-latching protection against undervoltage, overvoltage, short-circuit and over-temperature conditions. The dual outputs are 3.3 and 5V with a power trading facility letting 60W be available from either output. Efficiency is 90 per cent. There are two versions with inputs of 18 to 36V or 36 to 75V. Baseplate operating range is –40 to +100°C and input to output isolation 1.5kV DC.
C&D Technologies
Tel: 00 353 61 474133



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Zetex
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Burr Brown
Tel: 01923 233837

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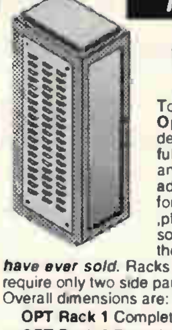

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by 600 or 1024 by 768. Through an add-on digital daughterboard, it accepts full-screen PAL, NTSC and SECAM composite inputs. Another daughterboard provides connectivity to TDMS and LVDS panels. Measuring 179 by 140 by 19mm, it is compatible with 3.3 and 5V LCDs. It uses up to 8-bit per colour (16.7 million colours, including on 3 by 6-bit panels).

Digital View
Tel: 0208 236 1112

IR transceiver

Agilent Technologies has announced an IrDA-compatible infra-red transceiver for PDAs, pagers and mobile phones. The HSDL-3202 can operate with logic levels down to 1.8V, so it can work with microcomputer chipsets and custom Asics that operate with 1.4V or higher signal levels. It draws 20nA in shutdown mode and 100μA in standby mode. When transmitting data, an integral 32mA current source drives the LED light source, which lets the LED be powered directly from 2.7V. The transceiver works at up to 20cm when communicating with other IrDA low-power devices and at 30cm

with standard IrDA devices. It is 2.5mm high and is eye safe under IEC825-1 class one specifications.

Digital View
Tel: 01344 366666

Wireless LANs

From Nighthawk Electronics are wireless LAN products from No Wires Needed with speeds up to 11Mbit/s for banks and retail organisations. They come with Airlock security software - a 128-bit public key encryption. Indoor wireless LANs and outdoor LANs for communicating between two or more buildings are available.

The indoor version is a pico cellular system that operates off the back of any Ethernet and provides on-site mobility. The outdoor system provides wireless bridges between buildings. It can be used to connect networks in different buildings to one LAN at up to 10km. Point-to-point and multiple bridge setups are supported. They comply with IEEE802.11.

Digital View
Tel: 01799 540881

Tactile switches

From Deltron Roxburgh are four SM tactile switches.

Standing 1.5mm off the PCB, the PTS525 is rated for 300 000 cycles. Available in two profiles, 2.5 and 4.3mm, PTS635 devices measure 6.5 by 3.5mm. The PTS645 measures 6 by 6mm with a height of 4.4mm.

Deltron Roxburgh
Tel: 01724 281770

2.5Gbit/s framer in standard CMOS

Exar has announced the XRT95L51 Sonet OC-48 2.5Gbit/s framer in standard CMOS. It is an ATM POS physical-layer processor with integrated Sonet OC-48 and STM-16 framing controller. It can transmit and receive an OC-48c or STM-16c signal or compose/decompose four OC-12 or 12c signals. ATM direct mapping and cell delineation are supported as are PPP mapping and frame processing. ■

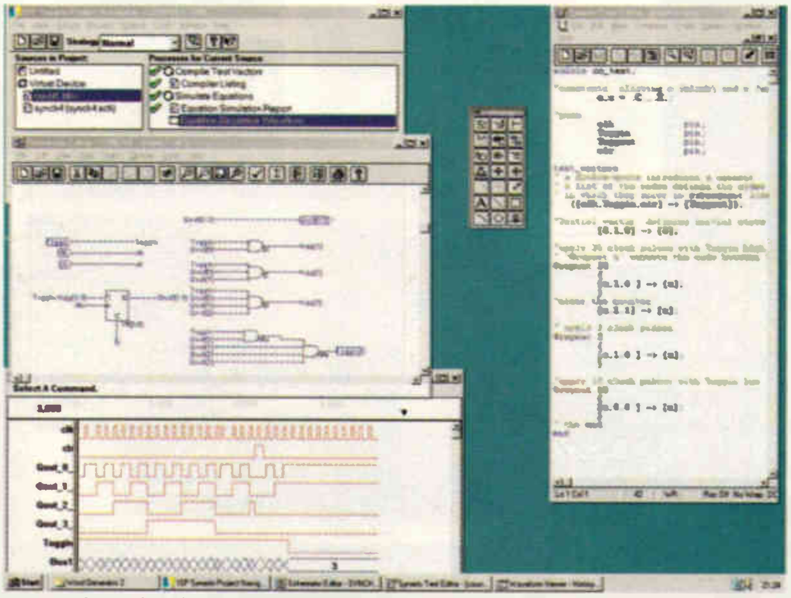
Exar
Tel: 00 31 1 3458 7676

Versatile stimulus for DIGITAL TEST

When troubleshooting digital systems, a generator capable of producing a stream of predetermined logic levels can be invaluable. Colin Attenborough's PLD-based generator produces a byte-wide stream of between 64 and 8192 words whose logic levels are predetermined on a PC via a GUI. Having a PLD at its heart, Colin's generator is relatively easy to implement.

An encounter with a digital simulation package set me wondering what was the best way to specify a bit stream or group of bit streams. The result is this word generator, which you should find useful for testing digital system designs.

Receiving its data from an LPT port, the generator can be set up intuitively via a menu system. This menu runs on a PC under Windows 95 or 98. On Windows NT and Windows 2000, Mr Gates spoils all the fun by making it impossible to control the printer port directly from the C programming language.



Screen photo showing a simulation of a counter used in the word generator design. The free PLD programming package's simulation facility can save you a lot of prototyping time – and money.

What does the generator do?

The word generator generates a sequence of eight-bit output words, from 64 to 8192 words long. The sequence can be free-running or stopped at a selected point. Once a sequence has been defined, it can be stored as a file and recalled later.

A Visual Basic interface shows an oscilloscope-like display of eight traces over a period of 64 time slots. Click on the screen to highlight a time period and a trace. The chosen trace is shown by labels to the left of the screen.

Menus allow the clearing, setting, or toggling of the trace for a selectable number of time slots from the selected point. Additionally, a settable number of clock cycles can be inserted. Right-clicking the screen toggles the chosen trace at the selected point. Clicking/dragging on the screen allows a logic state to be lengthened or shortened.

The 64 time slots on screen are a window on a possible 8192 time slots; the window can be 'slid' through the 8192 available time slots. The horizontal cursor keys shift the window by one time slot; if shift is pressed, the window shifts by a screen width of 64 time slots.

On pressing the 'Free run' button, data can then be read out from RAM at one of eight selectable rates from 10MHz to 78.125kHz, set by a 20MHz crystal oscillator. Manual and external clock drives are also available.

How it works

The state of the screen display is stored in an 8192 element array of bytes, which is updated every time the screen changes.

Pressing 'Free run', 'Run to cursor' or 'Manual' transfers the data represented on screen to a static RAM, via the print-

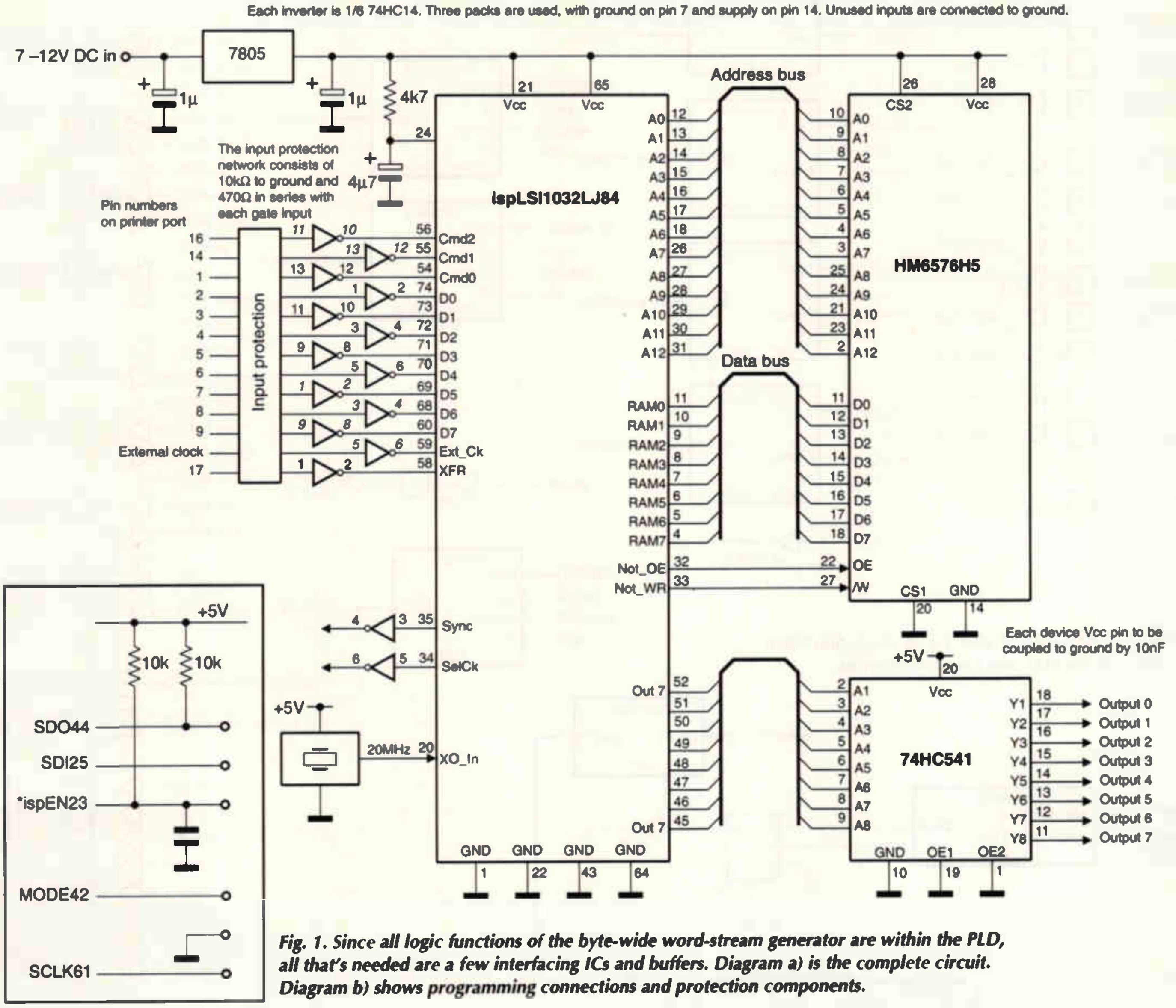


Fig. 1. Since all logic functions of the byte-wide word-stream generator are within the PLD, all that's needed are a few interfacing ICs and buffers. Diagram a) is the complete circuit. Diagram b) shows programming connections and protection components.

er port. This demands a dynamic link library written in C, as Visual Basic cannot address memory locations directly.

What you need. A PC running Windows 95 or 98.

What you don't need. You don't need the Visual Basic and Visual C languages used to write the software. The software on the CD contains files that will let the program run on any machine. However, source files *are* included for those of you who have access to the languages and want to examine the code.

The software for creating the programmable logic device used in this design can be obtained from Lattice, whose details are given later. I will be supplying ready-programmed PLDs, but I hope that this article will stimulate you enough to make you consider using PLDs in your own designs.

This article serves two purposes. It not only describes a useful piece of digital test gear but it also serves to demonstrate how useful PLDs are in simplifying a digital design's hardware.

You will need the Lattice software in order to get the full benefit of the article. Much of the description relates directly to how you use the software to develop the PLD needed. But don't worry – the software is available free of charge from Lattice.

Circuit details

In the circuit, **Fig. 1**, signals from the PC's printer port are buffered by 74HC14 Schmitt triggers before reaching the PLD. A RAM holds the byte-stream data. Signals from the RAM are buffered before reaching the output.

The PLD 'wraps up' the other circuit functions. It contains an address counter to select the RAM contents sequentially, and registers to hold cursor position data and other control signals.

A counter divides the 20MHz oscillator to give a choice of eight clock rates; a multiplexer selects one of the clock rates, and a second multiplexer selects between internal, external and manual clock sources.

For the 'Run to cursor' mode, a comparator compares the RAM address counter with data representing the cursor position. The comparator output sets a latch which stops the RAM address counter.

Communicating with the PC

Two output ports are available on the printer connector – an eight-bit main port addressed at 378_{16}, and a four-bit control port at $37A_{16}$.

Data are held in one of five registers in the PLD; three of the control port lines are decoded to define the register in which data from the main port will be stored. Data from the

The cmd bits are from hex37A - decoder is twisted to allow for inversion of bits in pc, and in external inverters

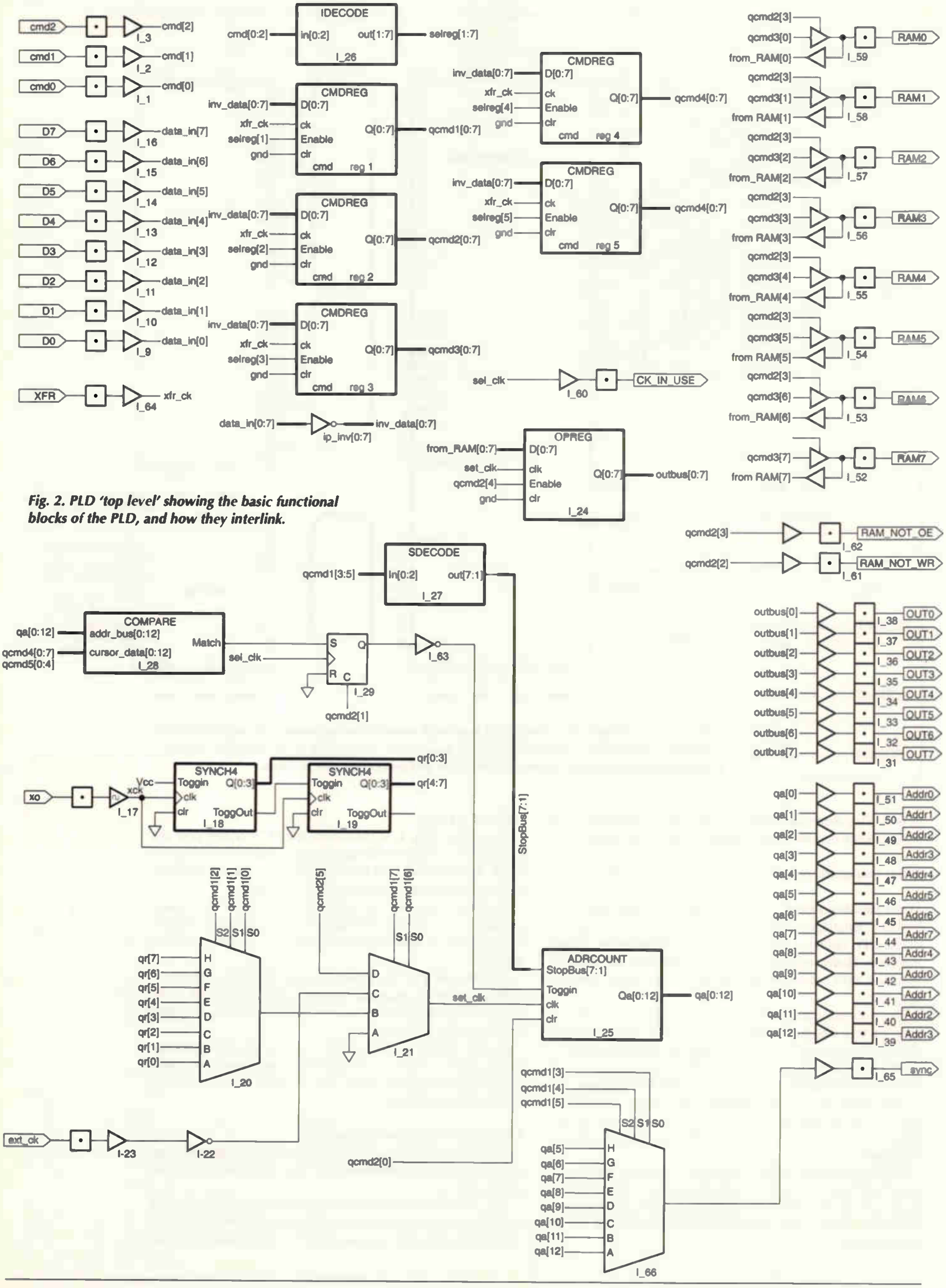


Fig. 2. PLD 'top level' showing the basic functional blocks of the PLD, and how they interlink.

main port are transferred to the selected register when the fourth line of the control port is pulsed.

Register 1 stores the clock rate, the sequence length and the source of the clock drive, the options for which are no source, a divided version of the crystal oscillator, an external source, or a signal from the PC.

Register 2 stores various control signals:

Bit 0 clears the RAM's address counter.

Bit 1 clears a latch that is set when the address counter reaches the cursor position in the 'Run to cursor' mode. This bit is set in the 'Free Run' mode, to ensure that the address counter runs continuously.

Bit 2 is an active-low write-enable signal to the RAM, used when entering byte-stream data into the RAM.

Bit 3 turns on tristate outputs, allowing data from the PLD to enter the RAM.

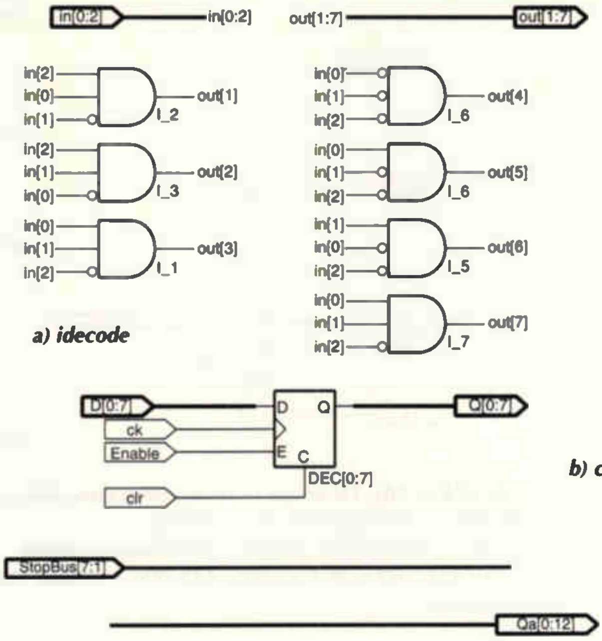
Bit 4 enables the output register. The output register used to retime data from the RAM is inhibited during writes to the RAM.

Bit 5 is the clock from the PC, used to clock the RAM address counter during data writes and manual clocking.

Register 3 stores data which will be written to the RAM.

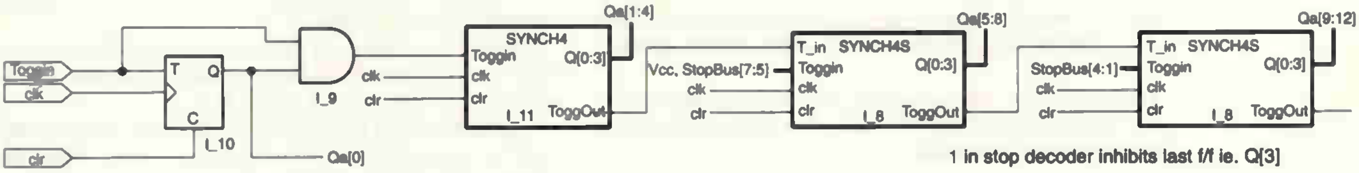
Register 4 stores the eight least significant bits of the cursor position.

Fig. 3. Breakdown of the PLD's various functional blocks:



a) idecode

b) cmdreg



c) adrcount

Register 5 stores the five most significant bits of the cursor position.

Elements of the PLD circuit

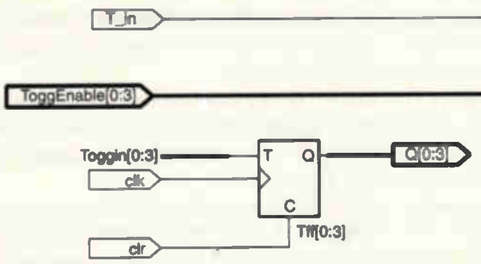
Figure 2 shows the top level circuit of the programmable logic device. To be able to benefit from the the following, you will first need to obtain and load Lattice's PLD programming package.

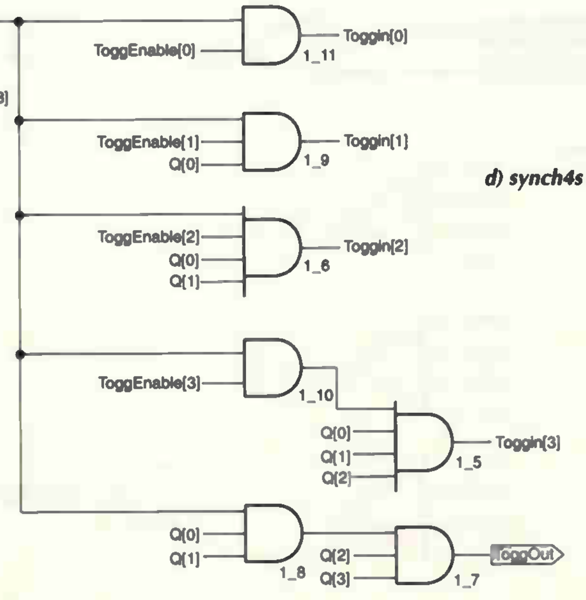
Idecode, Fig. 3a), decodes the instruction word from the control port, selecting the register in which data on the main printer port will be stored. The '0' output is not decoded, as I anticipated that it would not be used. Although states '6' and '7' are not used in this design, they have been left in for possible future applications.

The sharp-eyed among you may object that the decoding is addled, and that for example state '1' does not correspond to binary 101. This apparent aberration follows from the inclusion of inverting buffers between the PC and the PLD, and from the fact that bits 0 and 1 of the control port are inverted in the PC.

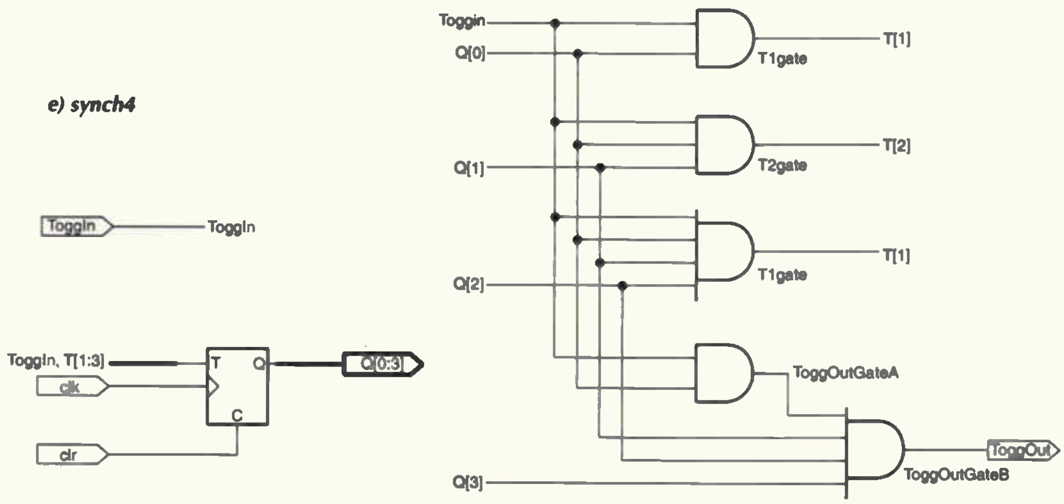
Cmdreg – Fig. **3b)** – is an eight-bit wide parallel-in, parallel-out shift register. Data from the main printer port, inverted inside the PLD to compensate for the external buffers, are transferred to the *Q* outputs on the positive edge of the clock signal if the Enable input is at logic 1. There are five instances of *Cmdreg*, one for each of the registers mentioned above.

Adrcount – Fig. **3c)** – defines the address of the RAM. It is 13 bits long, to accommodate the 8192 (2^{13}) states of the RAM. It is made up of two four-bit synchronous counters,





d) synch4s



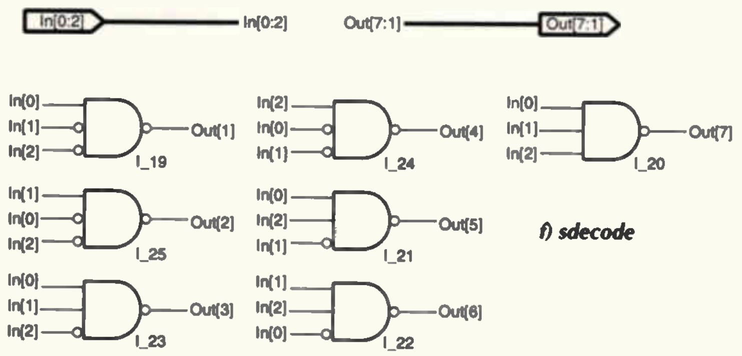
e) synch4

synch4s, of Fig. **3d**). These can be truncated to three, two or one bits by control signals on the 'StopBus'.

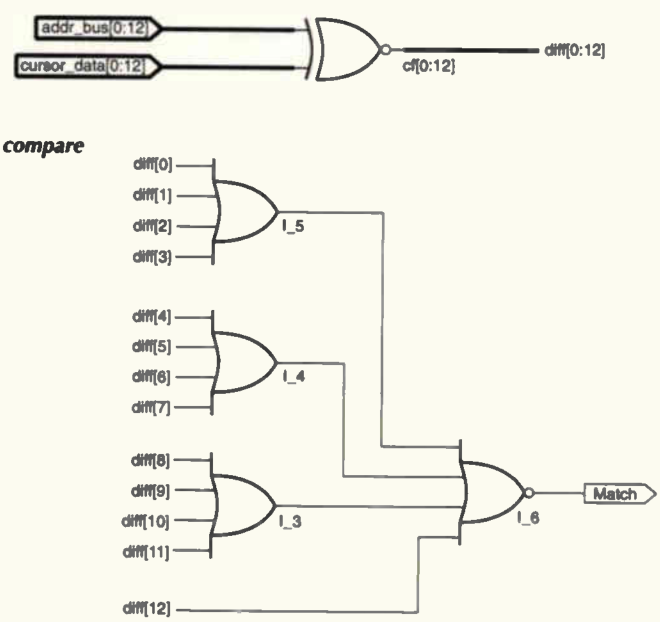
A simpler fixed-length four-bit synchronous counter – *synch4* of Fig. **3e**) – and a toggle flip-flop complete the address counter.

Sdecode – Fig. **3f**) – decodes the three bits specifying the sequence length, providing seven active-low outputs to inhibit toggling of the appropriate stages of the address counter.

Compare – Fig. **3g**) – compares data representing the current cursor position and the address counter outputs, giving a logic 1 when they match.



f) sdecode



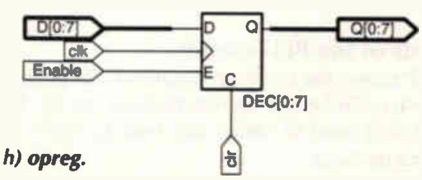
g) compare

Opreg – Fig. **3h**) – retimes the output from the RAM. An inhibit input prevents the outputs changing while data are written to the RAM.

An 8:1 multiplexer selects the most significant active bit of the counter to generate a synchronisation signal. The multiplexer is needed because when the count is shorter than its maximum, the later outputs are held static; the multiplexer is controlled by the 'StopBus'.

The synchronisation output is a cycle ahead of the outputs associated with the zero state of the address counter. This is because of the one cycle delay in *Opreg*.

A clocked latch holds the comparator output and stops the address counter in the 'Run to cursor' mode when the address counter and cursor position are equal. That's a white lie, actually; the presence of the output latch means that the address counter is stopped at *one less than* the cursor position.



h) opreg.

Busses and iterated components: ways to make life easier

If you compare this design with the earlier one – namely my logic analyser – you'll see that much more use is made of busses and iterated components.

Compare Fig. 3h), the *Opreg* function described elsewhere, and **Fig.** 4 – the same function realised without bus notation. Bus notation is far more compact, and allows a group of connections to be made at one move.

It's sometimes useful to make up busses from signals from different sources. **Figures 5** and **6** show a serial-in, serial out register implemented with and without bus notation. It is important to think clearly about which end of such a register is which.

Because the flip-flops are *Dff*[0:7] with outputs *Q*[0:7], it follows that the input data go to the *D* of *Dff*[0], the *D* of *Dff*[1] goes to the *Q* output of *Dff*[0], and so on. That's why the bus feeding the *D* terminal is defined as D_{in},Q[0:6]. Output *Q*[7] is the last *Q*, and feeds only the output. Unlike the other *Q*s, it doesn't feed a *D* input.

As the clear input is common to all the flip-flops, it is not a bus, and is labelled simply clr. The same reasoning applies to clock input. **Figure 7** shows how bus notation simplifies the block symbol as well as the circuit.

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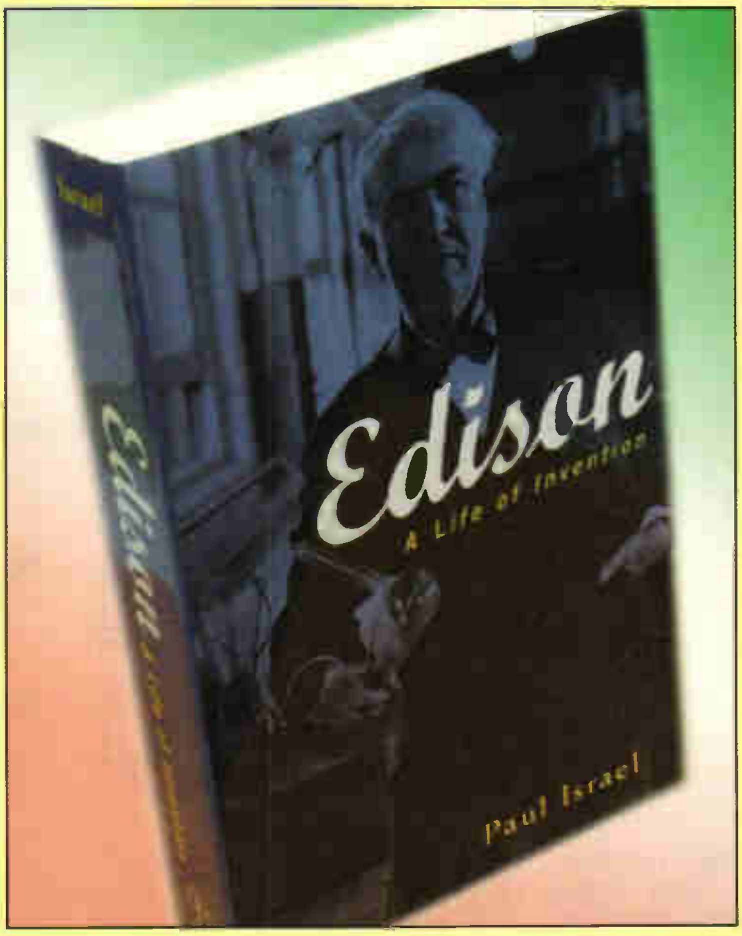

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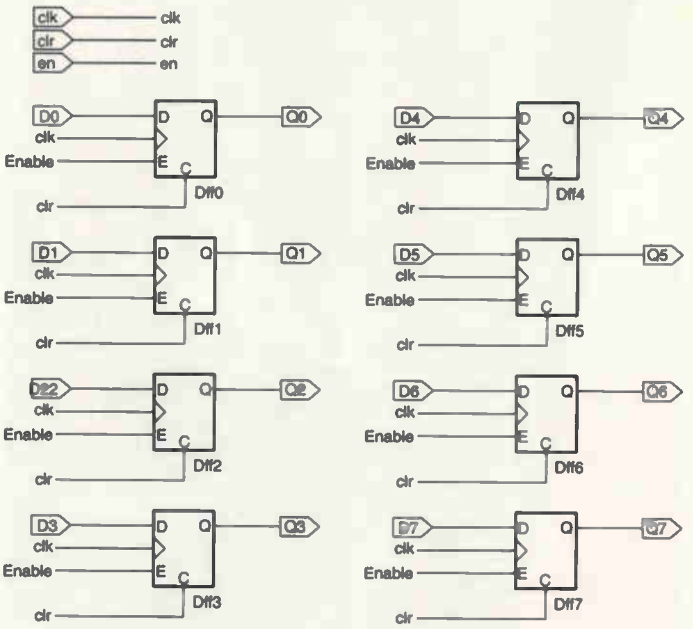


Fig. 4. Logic block 'opreg' with no bus. Compare this with the bussed version, Fig. 3h).

Bus tools

What tools are needed to exploit busses and iterated components? To tell the system that a component is to be repeated, use the 'Name Instance' tool, **Fig.** 8, and enter a name of the form *Dff*[0:7].

The [0:7] section implies eight occurrences of the component. The *Q* outputs can then be labelled *Q*[0:7] using the 'Label net' tool. The wire will be wide, to show that it's a bus.

If you find an occasion to call the component MyComp[42:37] the system won't object; iterations don't have to start at zero, or to have increasing indices.

What *does* matter, though, is that you label nodes consistently with the component iteration. If you label a component *Dff*[0:7] and label the *Q* outputs *Q*[7:0], then you're on your own – you and your bottle of migraine tablets. If you label the *Q* outputs [0:6], then the system will give an error message when you check the circuit via the tick symbol at the top of the schematic editor.

Another possible source of errors is when you connect, say, Output[7:0] from one block to Input[7:0] on another with bus MyBus[0:7]. Think carefully! It's simplest to keep the directions of the indices the same on both busses.

There's a simple way to label the individual wires from (or to) a bus that you've labelled already, if you need to do so. Select the 'Label net' tool and click on already-named bus. The cursor will pickup, say, MyBus[0:7]; press the *right* mouse button, and the text

PLD basics

I'll start by summarising the main points of my earlier article[1], which described a logic analyser using a Lattice Semiconductor programmable logic device.

Programmable logic devices let you draw the circuit you want, using appropriate software, and generate a file which can be downloaded into the device. Early PLDs had to be programmed in a programmer but current versions can be programmed – and rapidly re-programmed – while in-circuit.

Downloading the PLD's firmware needs only a simple piece of hardware, as described in my earlier article. It connects between the PC's printer port and the device being programmed. The device then implements the circuit that you drew on your PC in the free PLD development package.

The function of the programmable logic device remains even if you turn off the power. If you get the design wrong, correct the drawing and download the new file. No UV eraser is needed and there is no erase process.

Subcircuits can be stored and used in later designs and duplicated as necessary. You can choose which pin on the device is which input or output; if the requested pinout cannot be fitted, you'll get an error message.

You can also choose to tell the software to reserve the programming pins to avoid using them for inputs or outputs. That allows you the delightful simplicity of reprogramming the device without disconnecting any of the other devices in the circuit.

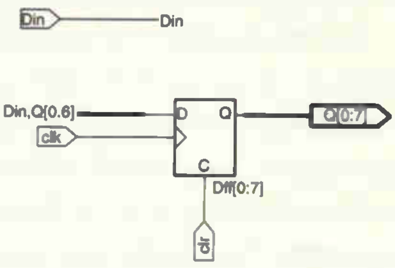


Fig. 5. Serial-in, serial-out register implemented using bus notation.

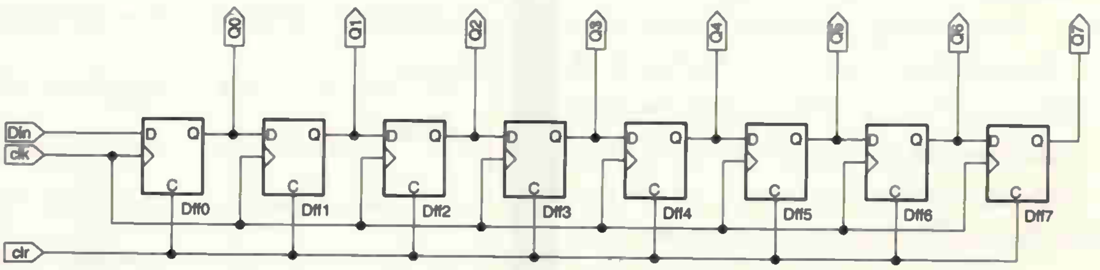


Fig. 6. The same serial-in, serial-out register as in Fig. 5, but this time implemented without using bus notation. Note the increase in visual clutter.

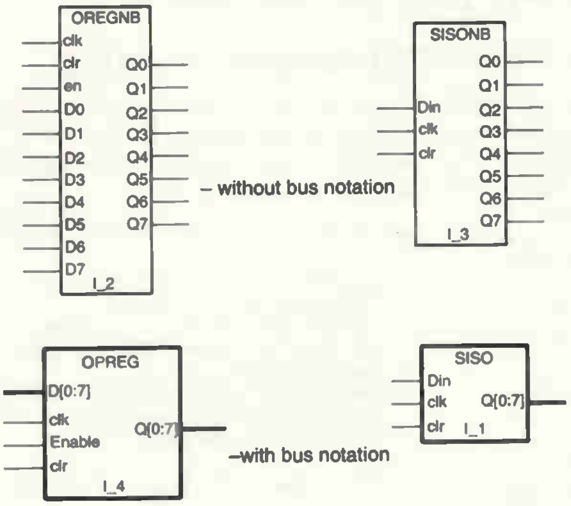


Fig. 7. Bus notation simplifies block symbols as well as busses.

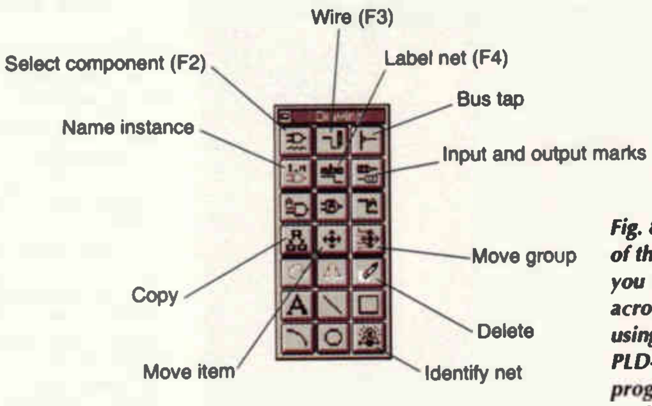


Fig. 8. A 'toolbox' of the type that you will come across when using Lattice's PLD-programming package.

will change to MyBus[0]. Click on the net that is to bear that name; the net takes that name, and the cursor will change to MyBus[1]. Continue this process till all the bus elements are named.

There's a similar trick for simplifying the naming of other nets. If there's a net called 'MyNet' that you want to connect to a distant node, select the 'Label net' tool and click on the MyNet label; the cursor changes to MyNet, and can be applied to the net to be labelled.

When nets are connected this way, you may sometimes want to identify everything connected to the net. To do this, select the 'Identify net' tool and click on one piece of the net; all pieces of the net will be highlighted.

Where you need to split a bus, for example to take the bussed output of a counter to output pads, use the Bus Tap tool. Select the bus, and drag a connection to the desired point.

Have I designed what I think I've designed?

The software lets you define input stimuli for a circuit and observe its response. Under the "Source/New...", select 'ABEL Test Vectors', and enter the simulation commands.

The photo at the beginning of the article is a composite screen shot of the stimulus file (count1.abv), the circuit to which it is applied (a four-bit synchronous counter using bus notation and iterated components) and the results presented as byte streams.

Comments in the .abv file explain the syntax. When the .abv is completed and saved, select it in the left-hand panel of the Project Navigator, and double-click 'Equation Simulation Waveform' in the right-hand panel. The Waveform Viewer window will open.

Signals can be added using the 'Show' option under the 'Edit' menu; this also allows busses of signals to be defined and examined. The view can be zoomed in or out; when the view is sufficiently zoomed in, signals presented as busses show the value represented by the bus.

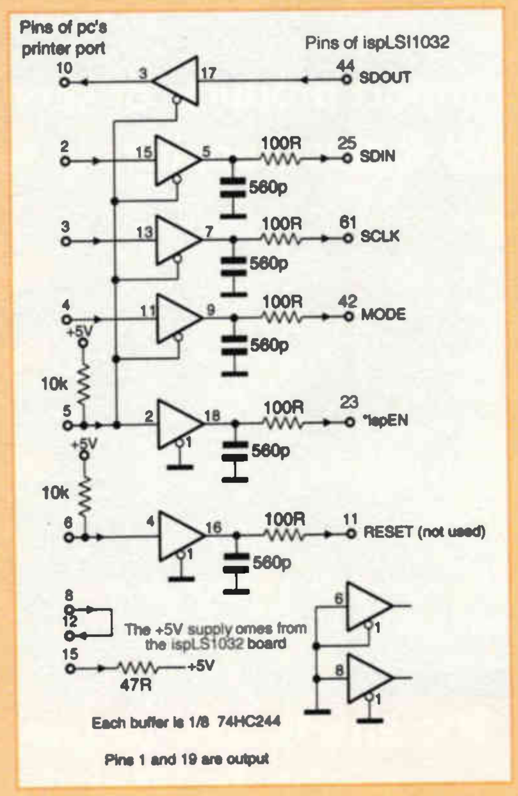
Save the stimulus files on disk. They might be hard or even impossible to read from the picture.

Pin allocation – a pictorial method

My earlier article[1] showed how to define what signal was connected to what pin by allocating pin numbers on the *I/O* pins on the circuit diagram. There's another way though,

Programming the PLD

This is the buffer circuitry needed to program an ispLSI1032 via your pc's printer port. Take great care with directions and power supplies here – particularly if your printer port is an integral part of your mother board. It may be wise to use an add-on i/o card to give you LPT2 and probably two extra COM ports. These are very cheap. Note that this interface handles 5 and 3.3V devices in an isp or JTAP programming chain, but not a mix of isp and JTAG parts. A ready-made cable is available for this task.



that lets you see the position of the allocated pin on a picture of the package. This is an advantage when you're considering how associated components will connect to the PLD.

The relevant program is 'ispDS+'. You will find it under 'Lattice Semiconductor' in Windows' Start/Programs – assuming you've installed the software of course. I've used it while generating the PLD for this design; you can find the relevant files on the CD.

First, draw your circuit and compile it under ISP Synario as usual, then start up ispDS+. Under the 'Project' menu, select 'New'; browse your way to the directory in which you've built your design. Double-click the .tt2 file which you'll find there.

Next, select 'Device' under the 'Assign' menu, and choose a device. Then click 'Pin Locations', also under the 'Assign' menu. You'll see a list of pins on the left, and a representation of the device you chose on the right. Click the 'Zoom In' icon to suit; you can now start assigning inputs and outputs to pins.

Double-click a pin on the representation of the device, and the top-most i/o from the list of i/os will attach itself to the pin. If you'd rather allocate them in a different order, click your chosen i/o in the list, and then double-click the pin.

Continue until all i/os are allocated, then select 'Compile' in the 'Tools' menu. If you haven't given the system an impossible task, the i/os will be allocated to pins in the way you've requested. At stages of the process, a window at the bottom of the ispDS+ area gives information, error and warning messages as appropriate.

You can now program the device from within ispDS+; you'll find 'Download' under the 'Tools' menu. Under the 'Configuration' menu of the download window, select 'Scan board'.

Assuming that the board is powered, is connected to the download cable, and contains a correctly connected device, you'll get a confirmation message. Browse your way to the '.jed' file which has been written for your design.

Be careful to go to the lower level directory generated by ispDS+, under the original ISP Synario directory. You can then select "Run" in the download menu, and the device will be programmed.

Writing DLLs Visual Basic

There are several subtleties to writing a DLL that can be called by Visual Basic; here's a guide through the minefield.

Start Visual C++, and select File/New/Projects tab. Enter the name and directory of the DLL, and select Win32 Dynamic Link Library; press OK. On the next screen, select 'A DLL that exports some symbols' and press 'Finish'.

The system now generates the necessary files for you to customise. A file called MyDLL.h lists the functions that the DLL contains; the file MyDLL.cpp contains the functions themselves.

Before writing your functions, two changes need to be made under 'Project/Settings'. Under the C/C++ tab, select 'Code Generation' in the list box, and change the calling convention to '_stdcall'. That's the calling convention that Visual Basic expects.

Under the Link tab, tick 'Generate mapfile'. You'll need to generate a map file because Visual C++ 'decorates' the names of functions. As a result, you need to tell Visual Basic what the decorated function name it finds in the map file will be.

> **A word of caution**
>
> The version of PLD design software I used has one subtle shortcoming – it expects the the Windows 3.x convention for file names. So limit file and path name lengths, and don't use spaces.

When you've written and debugged your code, open the map file. It will be in the 'MyProject/Debug' or 'MyProject/Release' directory, depending on what you've selected under Build/Set Active Configuration.

You may find it convenient to associate '.map' files with Wordpad. Look for the name of your function and then use the decorated version in a 'Declare' statement in the module of your Visual Basic project. For example, the init function of the Word Generator project is declared by the line,

```
Declare Function init Lib
"WfmGenDLL.dll" Alias "?init@@YGHXZ" ()
As Integer
```

where the name `?init@@YGHXZ` is to be found in the map-file as the decorated version of init.

Where do I put my DLL?

Finally, where does the DLL go when you've made it?

You can specify a complete path name in the 'Declare' line inside Visual Basic, but that causes problems if the DLL is moved. The simplest way is to put it in the directory with all the VB files; alternatively, put it in the Windows/system directory.

One nasty little feature: if you put the DLL in the same directory as the VB files, it will only be found if the VB project is started by double-clicking the VB project file. It will *not* be found if you start VB and navigate your way to your project. Even so, I think that putting the DLL with the VB files is the best way.

There are three things that may trip the unwary. Firstly, remember that C/C++ always deals in functions, while Visual Basic has functions and subroutines. If you need a *subroutine* in your DLL, write a *function* that returns a dummy constant; in Visual Basic, use

```
Dummy%=MySub
```

Secondly, the process of setting the calling convention only applies to the current configuration. If you've set it for the debug, you'll need to set it again for the 'Release' convention.

Thirdly, "But it hasn't made my DLL," I hear you say. Oh yes it has; you've set View/Folder Options/View to "Do not show hidden or system files".

My thanks go to my employers, Cambridge Consultants Limited, for permission to publish this article. ■

Reference

1. Attenborough, C, 'Five-chip logic analyser', *Electronics World*, Feb. 1999, p. 93.

More detail about the word generator files on the CD will appear next month.

To obtain the Lattice software, register at http://www.latticesemi.com. You will be sent a CD. While at the site, you might find a glance at the programmable analogue devices worthwhile – Ed.

CIRCUIT IDEAS

Fact: most circuit ideas sent to *Electronics World* get published

The best circuit ideas are ones that save time or money, or stimulate the thought process. This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

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Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

£50 Winner

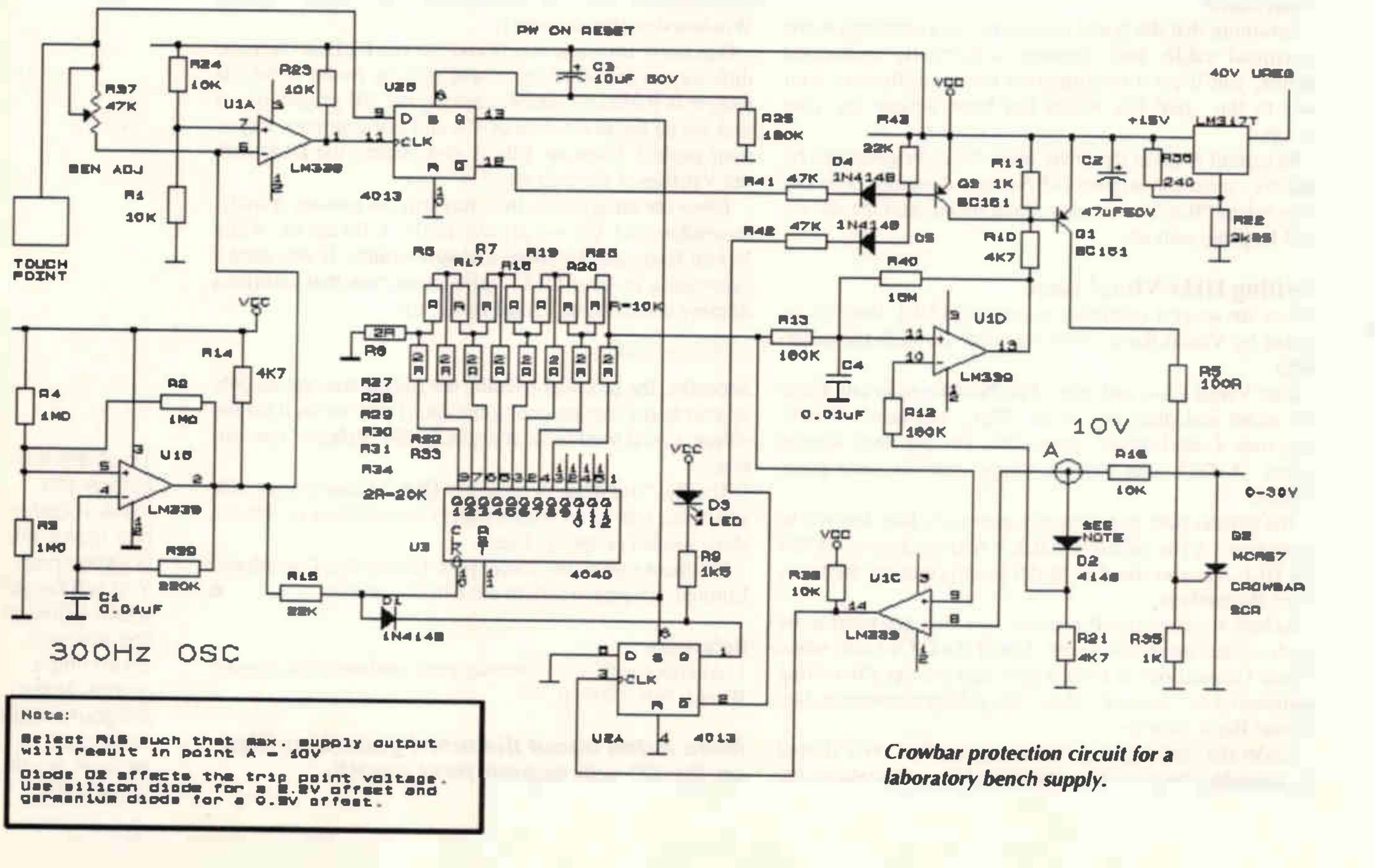
Crowbar protection power supplies with variable output

In a regulated power supply, the traditional crowbar circuit is designed for fixed voltage output only, as the trip point must be hard wired. The lack of over voltage protection on a variable-voltage output power supply is a hazard when connecting circuits to be tested or repaired to such power supply.

The circuit shown was developed to provide crowbar protection on a regulated power supply with 0-30V output, using only two CMOS ICs and one voltage comparator IC.

An LM317T three-terminal regulator connected to the main rectifier smoothing capacitor provides the circuit with a regulated 15V supply. The common 7815 is not applicable because of its low maximum input voltage.

The circuit consists of an oscillator, touch-operated switch, digital sample-and-hold circuit, trip voltage comparator and crowbar SCR. The touch point is made to the metal knob or the metal spindle of the variable



Crowbar protection circuit for a laboratory bench supply.

resistor that operator adjusts to change the regulated supply output voltage.

If the variable resistor is mounted on a conductive panel, the resistor case and spindle must be insulated from the panel for the touch sensor to function properly.

Oscillator U_{1B} generates a 300Hz square wave that feeds the digital sample-and-hold circuit and the touch switch. Clock polarity is inverted via U_{1A} and applied to the U_{2B}'s clock input. The 4013 bistable multivibrator samples the data input 'D' during the rising edge of the clock.

When the case where the metal knob is not being touch, because of the delay cause by U_{1A}, 'D' will always be low while clock rises. When the operator touches the metal knob to change the power supply output voltage, the human body introduces capacitance. This results in delay to the signal applies to 'D' and a change of state occurs at the 4013 outputs.

While the operator adjusts the supply output voltage, the touch switch operates. Bistable device U_{2B} changes state and sets U_{2A}. Transistor Q_3 conducts and trip circuit Q_1 is disabled.

Counter U_3 is now forced into reset with all its outputs, $Q_{1\text{-}8}$, low and no voltage appears at the R-$2R$ ladder network. Green LED D_3 turns off to signal protection circuit is now out of commission.

After the correct output voltage is set, the operator releases the metal knob and touch switch resets. Diode D_4 becomes inactive and reset line of 4040 returns to low.

The 4040 starts counting up. When the output at the R-$2R$ network matches that of the regulated supply voltage at point A, U_{1C} operates to enable U_{2A} to change state. This results in bypassing clock pulses from 4040 and turns on the LED.

Simultaneously, the trip circuit is armed. However, as the sampling point of the trip comparator U_{1D} pin 10 is taken after diode D_2; a voltage offset exists between the R-$2R$ network voltage and the pin 10 voltage. It is this offset voltage that gives the ceiling of the over voltage protection.

As the input is sampled via R_{16} and R_{21} voltage divider, this translates to 300% of the diode voltage drop added to the external voltage that would result in a trip. Choice of D_2 between germanium or silicon types gives different trip ceiling voltage.

Select R_{18} such that point A is at +10V at maximum output voltage. The test results of the actual tripping voltage on my prototype were measured as in the **Table**.

Table. Measured test results for the variable-voltage power supply over-voltage trip protection circuit.

Power supply nominal o/p (V)	Trip voltage with germanium D_2 (V)	Trip voltage with silicon D_2 (V)
5.0	5.9	7.2
10.0	11.0	12.3
15.0	15.9	17.3
20.0	21.0	22.3
25.0	25.8	27.2
30.0	30.8	32.3

Lim Chung
Haywards Heath
Sussex
E30

Simple inductance meter handles RF coils

This circuit measures inductances down to 10nH and is useful for RF coils. Measurement is easy and readings are accurate. They can be carried out at the actual working frequency of the coil. This avoids errors which can be serious – especially when measuring ferrite-core inductors.

The meter is based on a resonant VCO built around a Motorola MC1648 IC. The oscillating frequency,

$$F_{osc} = \frac{1}{2\pi\sqrt{LC}}$$

is determined by the inductance to be measured and an electronically variable capacitor, namely a potentiometer-controlled varactor diode. The oscillator output is displayed on a spectrum analyser to measure its frequency. Alternatively, a 200MHz frequency meter, or other counter with prescaler, can be used.

After connecting the unknown inductor to the test terminals, the potentiometer is adjusted until the oscillating frequency equals the inductor's intended operating frequency. Then the value of the inductance is obtained from,

$$L = \frac{1}{4\pi^2 f_{osc}^2 C}$$

To calibrate the potentiometer dial in pF use a 10cm square loop with 0.8mm copper wire, Fig. 2, which has an inductance of exactly 0.44μH. By measuring this inductance at different test frequencies the capacitance in picofarads as a function of potentiometer setting can be obtained from,

$$F_{osc} = \frac{1}{2\pi\sqrt{LC}}$$

where F_{osc} is in megahertz.

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Barcelona
Spain
E29

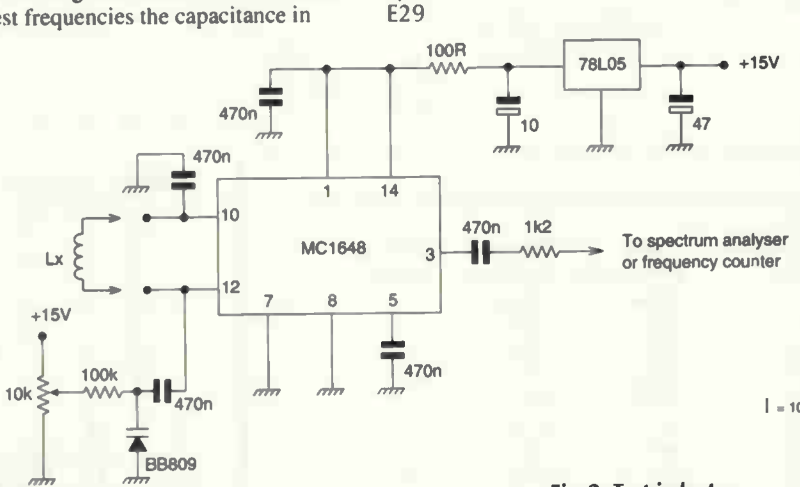

Fig. 1. Simple inductance meter, primarily for inductors used in RF designs.

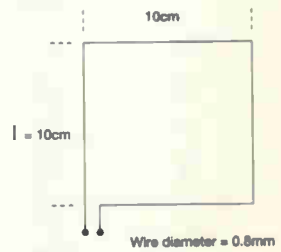

Fig. 2. Test inductor:

$$L_{test} = \left(\frac{2\mu_o l}{d}\right) \times \ln\frac{2l}{d} = 0.44\text{mH}$$

Use these test frequencies together with the test inductor to calibrate the meter.

F_{osc}(MHz)	98	76	62	53	43	38	34
C (pF)	6	10	15	20	30	40	50

Power supply noise-reduction circuit

Periodic radiated and conducted electrical noise at a spot frequency can be reduced in power by spreading the spectrum of the noise signal. The technique was used on an adapted power supply circuit with good results.

This circuit alters the switching period on a per pulse basis. Due to its complexity, it may find use as a final line of defence against electrical noise where the addition of further filtering and screening had proved ineffectual.

Where the power supply period of operation is set by the charging of a capacitor through a fixed resistance, this circuit feeds an error current into the capacitor. The error current changes per power supply cycle.

Gate drive from the power supply is used to clock a pseudo random noise generator formed by a shift register with tapped feedback. The weighted output of all 8 taps is fed to the power supply timing capacitor circuit. Attenuation of the this error signal gives reduced deviation of the power supply switching frequency.

On this circuit 5V power was taken from the regulated 5V output of the power supply switching *IC*, (UC3843 50mA output max.). The advantage is the generation of the power on reset signal to the shift register.

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E7

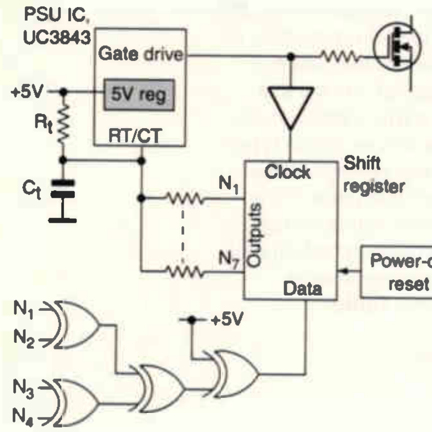

Fig. 1. Block diagram of circuit for reducing electrical noise from a switching power supply.

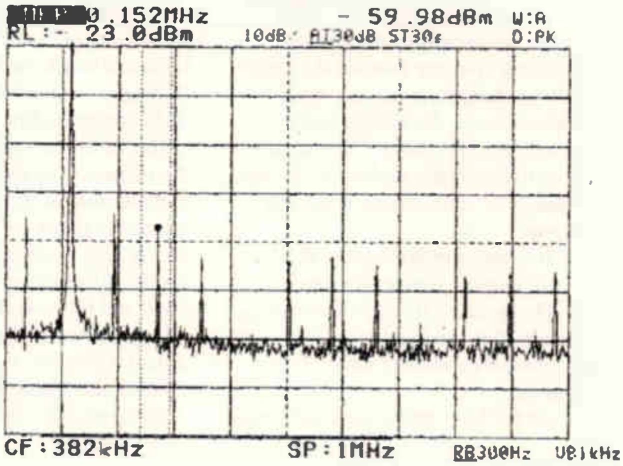

Fig. 3. Spectrum of power supply output, 5V in this instance, circuit disabled. DC – i.e. 0Hz frequency – is shown for reference.

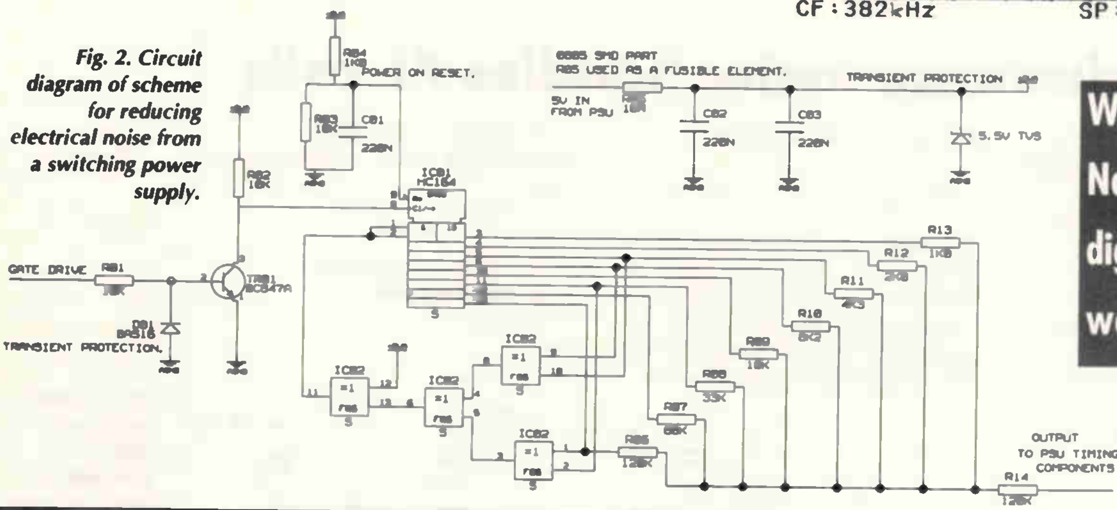

Fig. 2. Circuit diagram of scheme for reducing electrical noise from a switching power supply.

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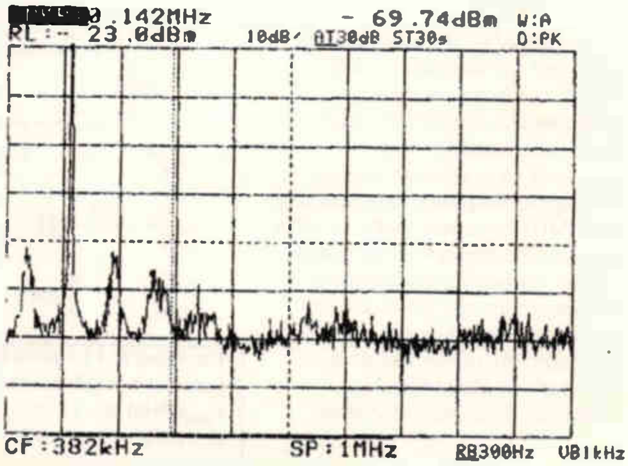
Fig. 4. As Fig. 3, circuit enabled.



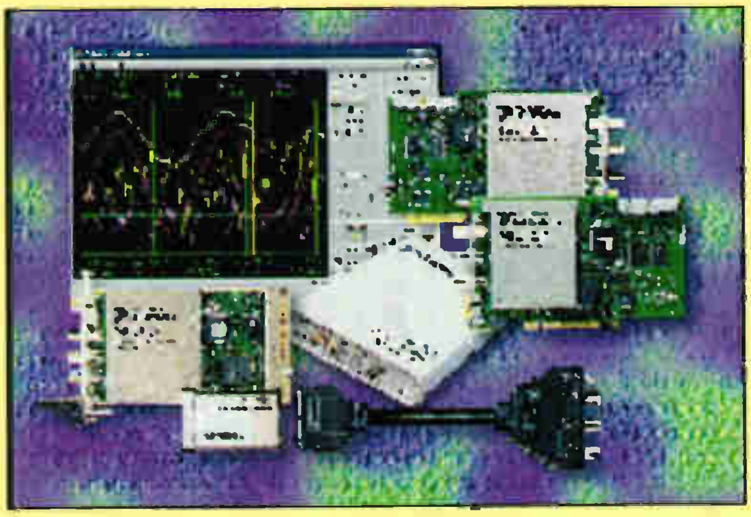
Comparison of results taken with and without the power supply noise reducer in circuit.

Harmonic	Frequency (kHz)	Power unmodified (dBm)	Power modified (dBm)	Change (dBm)
	76	–57	–65	–8
2	152	–60	–70	–10
3	228	–66	–74	–8
4	304	<–80	<–8	
5	380	–67	<–80	–13
6	456	–67	–78	–11
7	532	–68	<–80	–12
8	608	<–80	<–80	
9	684	–70	<–80	–10
10	760	–69	–79	–10
11	836	–69	<–80	–11

National Instruments sponsors Circuit Ideas

National Instruments is awarding over £3500 worth of equipment for the best circuit ideas.

Once every two months throughout 2000, National Instruments is awarding an NI4050 digital multimeter worth over £500 each for the best circuit idea published over each two-month period. At the end of the 12 months, National is awarding a LabVIEW package worth over £700 to the best circuit idea of the year.*



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National Instruments offers hundreds of software and hardware products for data acquisition and control, data analysis, and presentation. By utilising industry-standard computers, our virtual instrument products empower users in a wide variety of industries to easily automate their test, measurement, and industrial processes at a fraction of the cost of traditional approaches.

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*All published circuit ideas that are not eligible for the prizes detailed here will earn their authors a minimum of £35 and up to £100.

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The NI 4050 is a full-feature digital multimeter (DMM) for hand-held and notebook computers with a Type II PC Card (PCMCIA) slot. The NI 4050 features accurate $5^1/_2$ digit DC voltage, true-rms AC voltage, and resistance (ohms) measurements. Its size, weight, and low power consumption make it ideal for portable measurements and data logging with hand-held and notebook computers.

- DC Measurements: 20mV to 250V DC; 20mA to 10A
- AC Measurements: 20mV rms to 250V rms; 20mA rms to 10A rms;
- True rms, 20Hz to 25kHz
- Up to 60 readings/s
- UL Listed
- $5^1/_2$ Digit Multimeter for PCMCIA

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- Rapid application development
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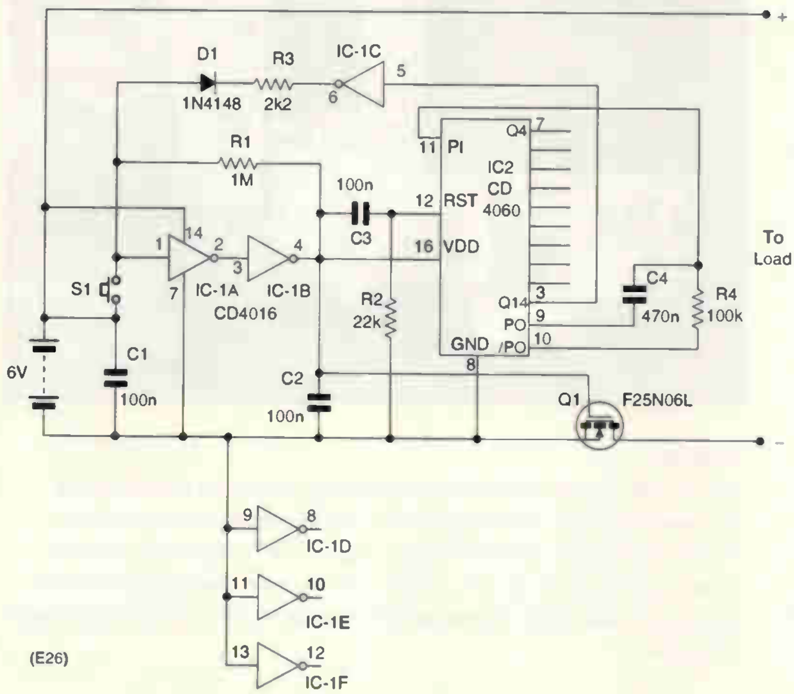
Delayed turn-off button



This is a delay switch for a battery powered system. Power to the load can be turned on by pushing S_1. The switch will turn off automatically after a certain amount of time.

Gates IC_{1A} and IC_{1B} form a bistable multivibrator. When S_1 is pushed, the output of IC_{1B} becomes high and stays high after S_1 is released. This high status turns on the MOSFET Q_1 and powers IC_2, which is a 14-bit binary counter with a built-in oscillator. The transition of the low-to-high status resets the counter through C_3 and R_2.

Counting frequency is determined by C_4 and R_4. When IC_2's Q_{14} output goes high, IC_{1C}'s outputs goes low, forcing the bistable off through R_3 and D_1. When C_4 is 0.47μF, the delay time is 1 minute 27 seconds, 11 minutes 6 seconds and 1 hour 46 minutes if R_4 is 10kΩ, 100kΩ and 1MΩ respectively.

When the switch is off, power consumption is less than 0.1μA.

Yongping Xia
Torrance
California
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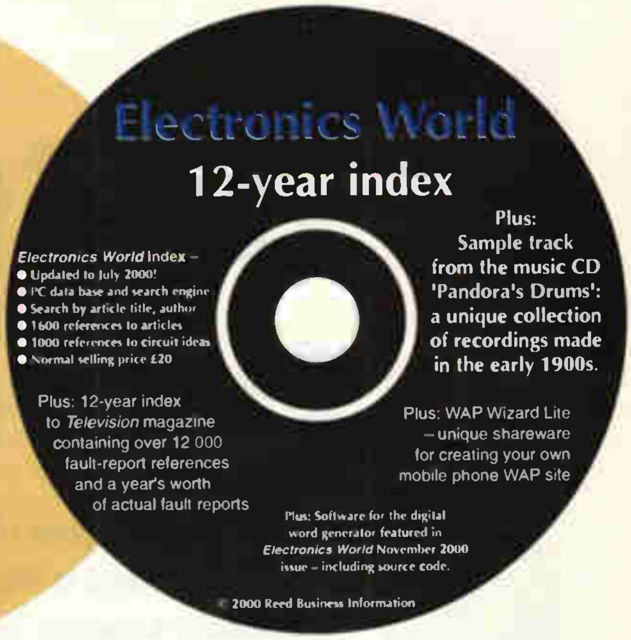
E26

Free cover CD



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PLUS: All software – including source code – for Colin Attenborough's digital word generator featured in this issue.

Using the CD

Each piece of software is in a folder.

The folders 'EwIndex' and 'TvIndex' for Windows 3.1, 95, 98 and NT contain a .TXT text file containing a description of the software and installation details.

To load the WAP Wizard Lite shareware for 95, 98 or NT4-SP4, run the 'Setup.exe' program within the folder. There's an HTML file in the folder containing instruction on how to use the program.

You can use your audio CD player as normal to play the audio CD. Your player should ignore the data, but as a precaution, at first playing, start with the volume turned low. Note that if you are using an audio CD player or CD player on the PC, these tracks appear as audio tracks 2 and 3. Track 1 is blank. On the Mac, they're tracks 1 & 2.

All software relating to Colin Attenborough's article on page 880 is in the folder called 'Wordgen'. This folder includes the PC controlling program and PLD software.

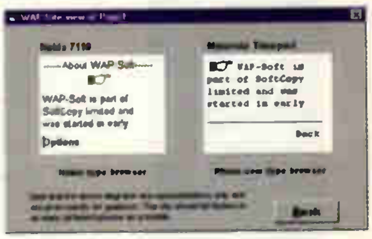
Object code and Protel PCB files for Emil Vladkov's 8-input video selector are in the folder entitled, 'VidMPX'. The PC-based control software is included. Copy the contents of the two folders to two floppies then run 'setup.exe' on the first floppy.

The folder 'Lights' contains the software for the Christmas lights 'Circuit Idea' in the December 1999 issue.

Two free tracks from the CD 'Pandora's Drums'

The Volunteer Organist and The Olio Minstrel, both recorded in 1913. For more details, see page 879.

Plus: WAP Wizard Lite



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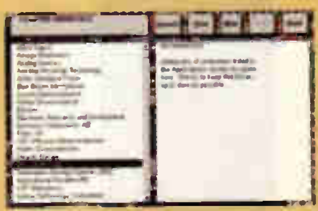
Electronics World index



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Digital thermometer

Roy Atkins' digital thermometer uses a transistor as a temperature transducer and a PIC controller to keep the hardware simple. This design comes complete with PIC source code.

Many a-to-d converters use an internal current charging a capacitor to create a ramp generator. The digital output is taken from a counter used to measure the time taken for the capacitor voltage to reach the input voltage.

In a dual-slope system, the capacitor is charged for a fixed time by a current proportional to the input voltage. It is then discharged using an accurate current, the digital output again being taken from a counter, but this time used to measure the discharge time.

This method eliminates the necessity of having an accurate capacitor. It also eliminates errors due to variation in the comparator threshold and the clock frequency of the timer.

Microchip, makers of the large range of PIC micro controllers, have an application note[1] that describes how a dual-slope a-to-d converter is used to measure resistance. If a temperature-dependent resistor is substituted, the converter can also measure temperature.

In this circuit, a capacitor is charged through the resistor connected to the PIC's V_{DD} supply pin. This resistor's value is converted to digital form.

The capacitor connects to one of the PIC input ports so that the port is toggled when its threshold is reached. At the same time as the capacitor begins to charge, the PIC starts a software timer.

When the port threshold is reached the timer stops, its value is stored, and the capacitor is discharged. The process now repeats, but this time the capacitor is charged through a known resistor. When the threshold is reached the timer stops and its value is again stored.

From the value of the known resistor and the two times, the digital value of the unknown resistor is calculated. This method eliminates most errors, for example, variations in the input port threshold voltage, the supply voltage, and also does away with the necessity for an accurate capacitor.

Getting the temperature involves firstly calculating the unknown resistor, which is fairly complex to do in the assembler language. Secondly, it involves converting this to a temperature in a look-up table containing the resistance versus temperature data for the particular thermistor used.

There are, of course, many temperature transducer ICs you could use. For example, there are zener-like devices whose breakdown voltage varies with temperature, like the LM135, or transducers that produce an output voltage proportional to

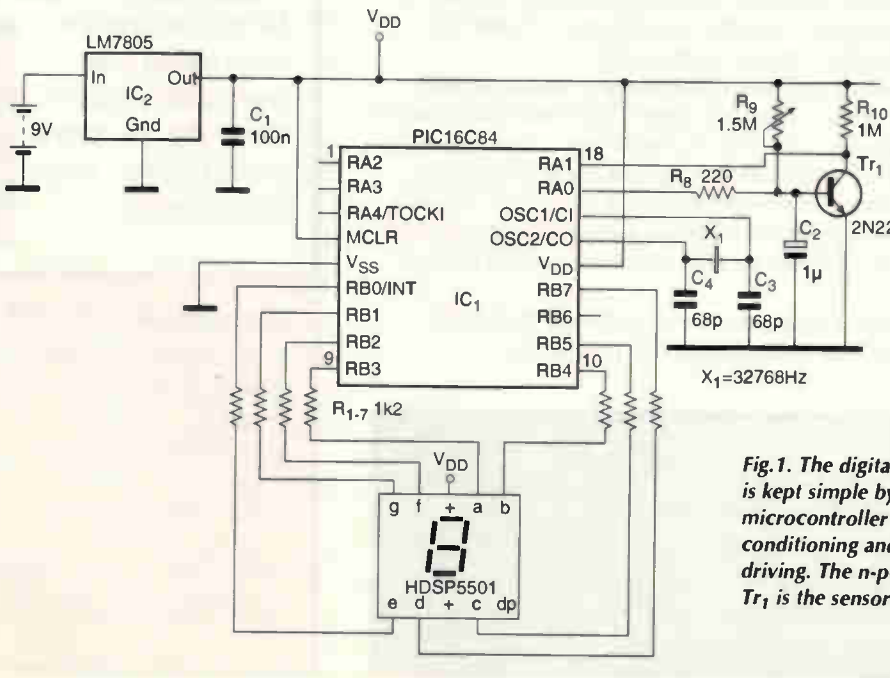


Fig. 1. The digital thermometer is kept simple by using a PIC microcontroller for signal conditioning and display driving. The n-p-n transistor Tr_1 is the sensor.

temperature, like the LM35.

The problem with these is that their outputs need amplification and conversion to digital form. This requires at least an op-amp, a-to-d converter, and display driver, before the temperature can be displayed.

The circuit described here, **Fig. 1**, uses a low-cost bipolar junction transistor for the temperature sensor, and a PIC micro-controller to do the control, processing, and display driver functions, resulting in a simple and cheap digital thermometer.

An n-p-n bipolar transistor's base-emitter voltage, V_{be}, has a temperature coefficient of about –2.5 mV/°C. By measuring the V_{be} required to produce a certain collector current, you can determine the junction temperature of the transistor.

The advantage of using V_{be} to measure temperature is that it varies very nearly linearly with temperature[2], which means that the temperature can be determined without the use of complex calculations or look-up tables.

How it works

Capacitor C_2 is charged via R_9 from V_{DD} when RA0 changes to high impedance at the beginning of the measurement cycle. Input port RA1 on the PIC monitors the transistor collector so that the exact time taken for the capacitor to charge up can be measured.

Capacitor voltage rises more or less linearly with time over a small part of its charging curve. As a result, the base voltage – and hence the junction temperature – can be accurately determined by the time taken for the capacitor to charge.

This time also varies linearly with temperature. If the capacitor charging rate and the software loop counting rate correspond, the number of counts in a measurement cycle will change by –1 for every 1°C rise in temperature; the minus sign is there because V_{be} has a negative temperature coefficient. All that is needed is for the charging rate, and a software constant, to be set correctly, and the thermometer will display the temperature in degrees directly after each measurement cycle.

Figure 2 is graphs of the base and collector voltages of the transistor during charge cycles at two different temperatures, taken from PSPICE simulations. When the base voltage of the sensor, Tr_1, reaches about 350mV, at +50°C, it conducts and its collector voltage falls to about half V_{DD}, about 86ms after the capacitor C_2 begins charging up.

At –10°C conduction occurs at a base voltage of about 480mV, and after a charging time of about 115ms. Thus a temperature difference of 60°C gives rise to a charging time difference of about 29ms

In turn, this translates to a difference of 60 iterations of the software timing loop, giving the required one count per degree. Consequently, the temperature can be displayed directly without any calculation being necessary other than a single subtraction.

The 5V regulator, IC_2, is used since a constant charging current is required. The PIC is used in the low power, 'LP', mode because this offers minimum current consumption and excellent frequency stability.

Resistor R_9 sets the charging current, R_8 limits the discharge current of C_2 when RA0 switches to low impedance, before the start of each measurement cycle. Resistors $R_{1\text{-}7}$ are current limiting resistors, one for each LED display segment of the HDSP5501.

The only critical component is C_2, the integrating capacitor. Its absolute value is not important and it does not have to be very stable with temperature, since the circuit is calibrated. However, it is important that its value has a linear dependence on temperature if variations are to be compensated for.

The best capacitors for temperature stability are the polyphenylene-sulphide types. In practice though, even the much cheaper and more readily available polyester types give good results – less than 1°C error over a 60° range – because their capacitance versus temperature curves do not deviate too much from a straight line.

Software for the PIC

The object code given below may be input directly into a suitable programmer and programmed into the PIC. Note that the program will not initially give the right results, but the calibration procedure does require a programmed PIC to be used.

I suggest that the hexadecimal code be used in the PIC at first, to be replaced later with a new version compatible with the particular components used, after calibration. This procedure is described later.

The PIC micro-controller program works as follows:

1. Capacitor C_2 is discharged by means of a low impedance port (RA0).
2. Port RA0 changes to high impedance, and the capacitor starts charging via R_9. A software counter/timer is started. The PIC monitors the collector voltage and tests whether the RA1 port threshold voltage has been reached.
3. When this happens, the timer/counter stops, its value is saved, and the capacitor discharge starts via RA0, which changes to low impedance. If the counter overflows, the display shows 'E' and the program starts another measurement cycle.
4. The timer value is subtracted from a constant to yield the temperature in binary form. If the result is negative, the absolute value is taken, and a flag is set.
5. The absolute binary value is checked to make sure it is less than 100. If not, the display shows 'E', and the program starts another measurement cycle. Otherwise, it is converted to decimal, as two variables: 'tens' and 'units'.
6. The temperature is displayed by a single seven-segment display. If negative, the sign '–' is displayed first, then tens, assuming that the temperature is 10°C or more, then units, with 250ms pauses between successive flashes.
7. The cycle begins again.

By using the same display for three values – sign, tens, units – circuit complexity is minimised. The duration of each value is long enough for easy reading. If the temperature lies outside the range –20 to 99, then an error occurs, and 'E' is displayed.

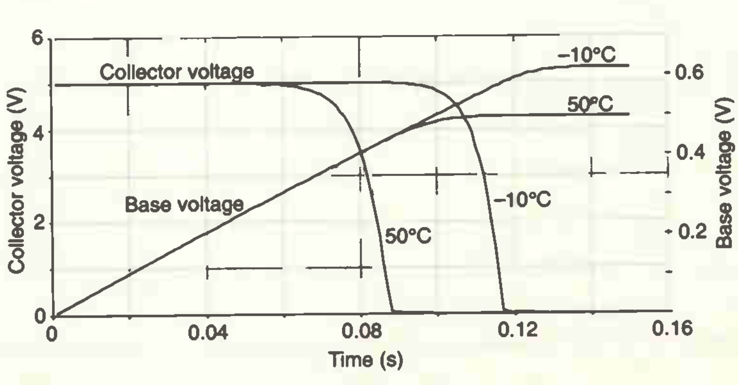
Setting up and calibration

There are three quantities that are not sufficiently well controlled to allow the circuit to be used without calibration, i.e. using off-the-shelf components.

First is the base-emitter voltage V_{be}. Different transistors would give up to 5°C variations in displayed value. Next is C_2's value and temperature coefficient. The error will be about 2°C per 1% capacitance change. Finally, the PIC's input-port threshold voltage plays a role.

Simulations in PSPICE indicate that the PIC-input thresh-

Fig. 2. Simulation of the n-p-n sensor transistor's base and collector voltages versus time.



old voltage would also result in variations of several degrees over the specified tolerances. This is because the collector current and hence V_{be} also vary with collector voltage.

The calibration procedure given below will allow all these quantities to be compensated for; in fact any transistor can be used, even though the temperature coefficient of V_{be} is not the same for all n-p-n silicon transistors.

In this set-up procedure the capacitor charging rate is first approximately adjusted, mainly to ensure that the software counter and display routines do not overflow at any stage. After this, two temperature tests must be done, if possible near the extremes of the range over which the thermometer will be used. For example, –15°C and +50°C.

If this setting-up procedure carried out properly, the thermometer will not need further testing, as both the capacitor charging rate – set by R_9 – and the software constant will have been set once and for all.

Approximate set up of capacitor charging rate

The charging rate of the capacitor is calculated as follows:

$$\frac{dV}{dt} = \frac{dV_{be}}{dT} \times \frac{dT}{dN} \times \frac{dN}{dt}$$

where V is voltage on capacitor, T is temperature in °C, N is the number of cycles of software timing loop, V_{be} is the base-emitter voltage at which Tr_1's collector voltage toggles the PIC input port and t is time in seconds.

Now,

$$\frac{dV_{be}}{dT} = -2.50\text{mV/°C}$$

for the 2N2222,

$$\frac{dT}{dN} = -1°C/\text{count}$$

PIC source code for the digital thermometer.

```
;Transistor junction thermometer, based on Vbe
dependence on temperature.
;label   op-code operand
         LIST    C=132
         LIST    N=63
;**      ************************  PIC16C5X Header
IND      equ     0
RTCC     equ     1
PC       equ     2
STATUS   equ     3
FSR      equ     4
OPTION1  equ     01h        ;actually 081h, but bank 1
                            ;selected in STATUS
TIMR0    equ     01h
PortA    equ     5
trisa    equ     5
PortB    equ     6
trisb    equ     6
C        equ     0
Z        equ     2
rp0      equ     5
rp1      equ     6
;
same     equ     1
W        equ     0
;*********************File register allocations
time     equ     0ch
result   equ     0dh
tens     equ     0eh
degrees  equ     0dh
flags    equ     010h
const    equ     011h
;********************I/O allocations  PortA
disch    equ     0
sense    equ     1
minus    equ     0
zero     equ     1
;*****************************************DISPLAY INFO
PortB
;SEGMENT PORT    PIN
;a       RB3     9
;b       RB4     10
;c       RB5     11
;d       RB7     13
;e       RB0     6
;f       RB2     8
;g       RB1     7
;*****************************************Discharge &
sense PortA
;FUNCTION       PORT      PIN
;disch    RA0   17
;sense    RA1   18
;**************************************************SET
UP
         org    00h
begin    clrf   time
         clrf   PortA
         movlw  0ffh        ;|clear display
         movwf  PortB       ;|Display has common anode;
high = OFF
         bsf    STATUS,rp0 ;bank 1
         movlw  b'00000010'
         movwf  trisa       ;RA1 = I/P, all others O/P
;RA0 (discharge) is low to discharge cap.
         movlw  b'00000000'
         movwf  trisb       ;PortB drives display;
                            ;set as O/P
         movlw  b'11010001'         ;|OPTION set for
clkout,
                            ;and 1:4 prescaler
         movwf  OPTION1     ;|
;        bcf    STATUS,rp0 ;setup for bank 0 in future
;**************************Mainloop
mainloop clrf   time
         clrf   flags
         clrf   tens
         movlw  b'00000011'         ;|
         bsf    STATUS,rp0 ;bank 1
         movwf  trisa       ;|open cct disch pin of
                            ;PortA so cap can charge
         bcf    STATUS,rp0 ;setup for bank 0 in future
count    incf   time,same  ;increment time counter
         btfsc  PortA,sense        ;has sense gone low?
         goto   count       ;NO, increment again
         movlw  b'00000010'        ;|YES
         bsf    STATUS,rp0 ;bank1
         movwf  trisa       ;|disch pin O/P to dischage
cap
         bcf    STATUS,rp0 ;bank0
         movlw  193         ;|193 in constant
         movwf  const       ;|
         movf   time,w      ;time in w
         subwf  const       ,w      ;|w=constant-time
         movwf  result      ;|copy to result
         btfsc  STATUS,C    ;has carry ocurred ("not
borrow" low)?
         goto   cont        ;NO, continue
         comf   result,same        ;|YES, invert, add 1,
set minus flag
         incf   result,same        ;|
         bsf    flags,minus        ;|
```

which is what we want to achieve,

$$\frac{dN}{dt}=\frac{1}{488\mu s}$$

which is determined by the PIC clock frequency and software. So,

$$\frac{dV}{dt}=-2.5\text{mV/°C}\times(-1)\times\frac{1}{488\mu\text{s}}$$
$$\approx 5.1V/s$$

This value may be used as a starting point, to be set up using an oscilloscope. Alternatively, if an oscilloscope is not available, R_9 should be set by ohm-meter to 0.95MΩ. This will result in the charging slope being reasonably close to the desired value, provided the value of C_2 is reasonably close to 1µF.

Final setting up is done by empirically, as follows.

Setting up accurately

In this procedure, variable resistor R_9 is adjusted, and the new software constant calculated. The display is used to monitor the effects of changes that are made.

When measuring temperature, the PIC first measures the time required to charge the capacitor as explained above. This is a number equal to the number of times the program executes the timing loop. It then subtracts this from the programmed constant, and displays the result, with a minus sign if negative.

During calibration, the thermometer will be subjected to two different environmental temperatures, which must be measured as accurately as possible, preferably using a laboratory grade thermometer.

Ideally a climatic chamber that can be set to maintain a constant temperature, should be used. If one is not available, you can use a refrigerator for the cold setting, T_1, and an oven for the hot setting, T_2.

```
cont                          ;positive number in result,
                              ;minus flag set if necessary
                              ;first check result <100
                              ;(max displayable
temperature)
        movlw   99            ;|check result-99 non-
negative
        subwf   result,w      ;|answer in w, result
                              ;unaffected
        btfss   STATUS,C      ;|did borrow occur,indicating
                              ;result>99?
        goto    bindec        ;|YES, "not borrow" is lo. Go
                              ;on to bin/dec conversion.
        movlw   b'01110000'           ;|NO.  Load "E"
segments in w. Lo=ON segment.
        movwf   PortB         ;output
        call    del500        ;display 500ms
        movlw   b'11111111'           ;|
        movwf   PortB         ;|switch off display
                              ;(common anode)
        goto    finished
;Now do binary/dec conversion
bindec  movlw   10            ;10 in w
        incf    tens,same     ;increment tens
        subwf   result,same           ;result=result-10
        btfsc   STATUS,C      ;has carry ocurred?
        goto    bindec        ;NO, do it again
        addwf   result,same           ;YES.
result=result+10,
                              ;decrement tens
        decf    tens,same
;NOTE : At this point tens and units are ready to be
displayed.  minus and zero flags ready.
        btfss   flags,minus           ;is temperature
negative?
        goto    dispten       ;NO display temperature
;display minus sign           ;YES, display minus sign
        movlw   b'11111101'           ;middle segment ON
        movwf   PortB
        call    del500        ;display for 500 msec
        movlw   0ffh          ;|turn off display
        movwf   PortB         ;|
        call    del250        ;pause 250 msec
dispten movf    tens,w
        btfsc   STATUS,Z      ;is tens = zero?
        goto    dispunit      ;YES, suppress zero, go
straight to display units
        call    dispn         ;NO, display what is in W
        call    del500        ;keep on for 500 msec
        movlw   0ffh          ;|turn off display
        movwf   PortB         ;|
        call    del250        ;pause 250 msec
dispunit movf   degrees,w     ;units to W
        call    dispn         ;display number in W
        call    del500        ;display 500 msec
        movlw   0ffh          ;|switch OFF display
        movwf   PortB         ;|

finished call   del500        ;wait before starting again
        call    del500
        call    del500
        goto    mainloop
;****************SUBROUTINES*****************
del250  movlw   01h           ;preset timr0 so doesn't exit
                              ;immediately
        movwf   TIMR0
        movf    TIMR0,w       ;|
        btfss   STATUS,Z      ;|has overflowed?
        goto    del250+2      ;NO
        retlw   00h
del500  call    del250
        call    del250
        retlw   00h
dispn                         ;display number in W
;                             ;first get pattern, then copy
                              ;to PortB
        call    pattern       ;display number in W
        movwf   PortB
        retlw   00h
pattern andlw   b'00001111'           ;limit to 15
        addwf   PC,same       ;n is in W, PC advances n
                              ;steps further. lo=ON segment
        retlw   b'01000010'           ;"0"
        retlw   b'11111010'           ;"1"
        retlw   b'01100100'           ;"2"
        retlw   b'01000101'           ;"3"
        retlw   b'11001001'           ;"4"
        retlw   b'01010001'           ;"5"
        retlw   b'01010000'           ;"6"
        retlw   b'11000111'           ;"7"
        retlw   b'01000000'           ;"8"
        retlw   b'01000001'           ;"9"
        retlw   b'01110000'           ;>9, display "E"
        retlw   b'01110000'           ;>9, display "E"
        retlw   b'01110000'           ;>9, display "E"
        retlw   b'01110000'           ;>9, display "E"
        retlw   b'01110000'           ;>9, display "E"
        retlw   b'01110000'           ;>9, display "E"
        end
```

Executable PIC object code for thermometer.

```
:100000008C018501FF308600831602308500003OA8
```

Note that the oven should be switched off when hot enough, for example 50°C, and allowed to cool slightly before using. Make sure the calibration thermometer's bulb is very close to the transistor so as to minimise the effect of any temperature gradient present in the oven or refrigerator.

Detailed procedure

1. Place thermometer in cold environment, for example 0°C, and note the temperature, T_1 and display, D_1. If the display does not show a number, but the letter 'E', then this is an error condition caused by the software timing counter overflowing, and the value of R_9 must be decreased.
2. Place in hot environment (e.g. 50°C), and note new temperature, T_2 and display, D_2.
3. While keeping the thermometer at the same temperature, T_2, adjust R_9 to obtain a display reading of,

$$D_2' = K_1 - \frac{(K_1 - D_2)(T_2 - T_1)}{D_2 - D_1}$$

where D_2' is the new D_2 displayed value at temperature T_2 after R_9's adjustment, K_1 is the initial software constant programmed into the PIC, obtained from the assembler source code, T_1 and T_2 are the cold and hot environment temperatures respectively and D_1 and D_2 are the displayed values at T_1 and T_2 respectively, prior to adjusting R_9.

4. Calculate a new value for the software constant, K_2

$$K_2 = \frac{(K_1 - D_1)(T_2 - T_1)}{D_2 - D_1} + T_1$$

5. Edit the assembler source code, and change the value of the constant to the new value. The code must be re-assembled and programmed into the PIC. For this reason a re-programmable PIC must be used, either an EEPROM version, such as the PIC16F84, or a ceramic window-ed version, such as the PIC16C54A-JW.

The thermometer should be accurate to within 1°C over a wide temperature range – at least from –20°C to 50°C. This circuit is a very cheap way to make a digital thermometer using standard components. ■

References

1. Microchip Technology Inc., *Implementing Ohmmeter/temperature sensor*, AN512
2. Verster, TC, 'p-n junction as an ultralinear calculable thermometer', *Electronic Letters*, 3 May 1968 Vol. 4, No 9, pp. 175-176.

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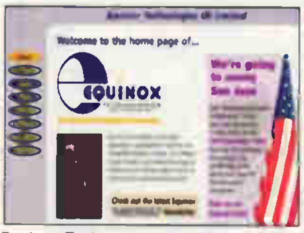

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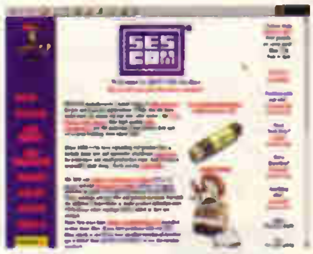

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Following on from the newsgroup discussion last month there is a UK Email group for TV technicians where you can send an Email to everyone in the group. There's just over 30 people in the group at present. For more details and how to register look at the egroup home page. Just a general comment though - you do have to be careful who you give your Email address to so that you can avoid "spamming" - that is getting lots of unwanted Email about dubious Russian site (amongst others).

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Wide range capacitance meter

Built from little more than a quad CMOS gate, Ian Hickman's educational capacitance meter also represents a useful piece of test gear, capable of measuring from 1µF down to around a picofarad – or perhaps even less.

This month's circuit involves no more than a couple of integrated circuits and a score of passive components, but nonetheless builds into a useful laboratory instrument. This is made possible by using that common or garden item of laboratory test gear – a digital voltmeter – as the read-out.

Due to the very simple design, you cannot expect the ultimate in accuracy, but the answers it provides will be correct to a few percent at worst.

The arrangement uses a CD4070, a CMOS quad exclusive-OR gate, also known as an XOR gate. Each of the four gates accepts two logic inputs and produces a 'high' or logic 1 output, usually +5 volts, only if one input is high and the other 'low', i.e. logic 0, ground or zero volts. It doesn't matter which of the two is high, but if both inputs are high, or both low, then the output will be low.

Capacitance meter circuitry

Figure 1 shows the circuit diagram together with the pin-outs of the CD4070 and 78L05. Gates IC_{1a} and IC_{1b} form a low frequency 'relaxation oscillator'.

In this arrangement, IC_{1a} is non-inverting, as its unused input is tied to ground, i.e. logic 0. Its output relaxes from saturation at a logic 1 output, due to negative feedback from the other gate, until both gates enter the active region. Its output then switches to saturation at logic 0, the feedback round it via the 6.8nF capacitor being positive.

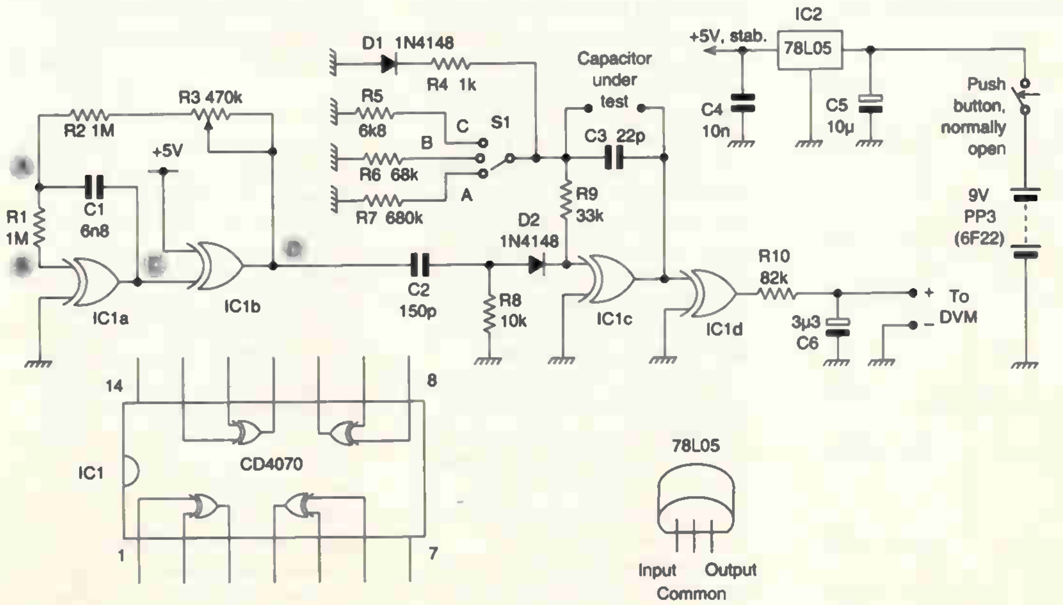

Fig. 1. Wide range capacitance meter, made simple since it relies on a DVM as a read-out device.

This changes the output state of IC_{1b}, which acts as an inverter as its other input is tied to logic 1. Voltage at the input of IC_{1a} now 'relaxes' from a negative extreme, back towards the active threshold of about +2.5V, and the sequence repeats.

Figure 2 should clarify the sequence of events. Note that the waveform at B is similar to that at A except that the tips have been clipped off by internal protection diodes. These are built-in at each gate input. Note that the drawing is not quite to scale.

The frequency of the relaxation oscillator is adjustable by means of potentiometer R_3, which is used as a variable resistor. This allows you to calibrate the completed instrument.

Once set up, the calibration will hold regardless of the battery voltage, thanks to the supply stabiliser, IC_2, with its +5V output. Using a push-button instead of an on/off switch prevents the battery being run down due to inadvertently leaving the instrument switched on.

How the meter works

The capacitor under test is the period-determining element in a monostable circuit formed from IC_{1c}. This monostable is repeatedly triggered at a constant frequency by the relaxation oscillator.

Consequently, the proportion of the roughly 18ms pulse repetition period of the relaxation oscillator, for which the output of the monostable is high, is directly proportional to the capacitance at the test-capacitor terminals.

Network C_2/R_8 differentiates the output of $IC1_b$, giving narrow 5V spikes, alternately positive and negative going. Negative-going spikes have no effect, but each positive-going spike appears via diode D_2 at the input of IC_{1c}, causing its output to switch to logic 1 level, +5V.

Positive feedback via C_3 and the test capacitor hold the input positive even after the spike has decayed, leaving D_2 reverse biased.

One of three resistors selected by S_1 then discharges the left-hand plates of C_3 and the capacitor under test from +5V towards 0V, until IC_{1c}'s threshold is reached. That causes its output voltage to start falling, and positive feedback causes the output to drop abruptly to 0V.

This drives the input to about –2.5V. But the gate input voltage returns to ground in a short time, as the resistor selected by S_1 discharges C_3 and the capacitor being tested. Thus by the time the next trigger pulse is applied via D_2, IC_{1c}'s input has settled at 0V ground. So unlike the case of IC_{1a}, when IC_{1c}'s output switches to +5V, its input rises from zero to +5V also, and the internal protection diodes do not turn on.

Output from the monostable is buffered by IC_{1d} and smoothed by R_{10}, C_6, to provide a dc level that is measured using a digital voltmeter.

Scaling is such that a 199.9mV reading on a 'three-and-a-half' digit DVM corresponds to a capacitor under test reading of 1999pF, 19.99nF or 199.9nF on ranges A, B and C respectively.

The nitty-gritty

To be useful, a capacitance meter needs to cover a wide range of capacitance. This creates practical problems at both ends of the range; large capacitances and also the smaller ones. Practical electronics is all about overcoming problems, and finding solutions that work – even if they are not particularly elegant.

Range B is designed to produce 200mV output to the DVM when the capacitor being tested is 20nF. Range C produces the same output voltage with a capacitor under test of 200nF, but capacitors larger than this can be measured by selecting the 2V range on the DVM.

In principle, this provides a range up to 2μF, but accuracy suffers. This is because the voltage at the input of IC_{1c} does not have time to settle to 0V at the end of the monostable's pulse, before the next trigger pulse via C_2 arrives.

Components D_1 and R_4 help in this respect, by returning the left hand end of C_3 to 0V more quickly following the negative-going edge at the output of IC_{1c}. Usable performance is thus obtainable up to about 1μF, providing a useful extension to the instrument's range.

The problem at the other end of the range, very small values, is that the capacitor being tested is in parallel with C_3 plus an unknown amount of stray capacitance due to the circuitry. How to cope with this is dealt with below.

Calibration

The instrument can be used as it is, without calibration, for matching pairs of capacitors or measuring their ratio. Clearly though, a DVM readout giving the capacitance of the unknown capacitor directly is generally much more useful.

This requires a capacitance of known value, for use as a calibration standard. The exact value doesn't matter, provided it is known, and much greater than one hundred times C_3; the reason for this will become clear in a minute.

Polystyrene capacitors are readily available in ±2½% tolerance, and, not quite so common, in ±1%. Let's assume you have found an 18nF 1% capacitor. This can be used to calibrate the instrument, on Range B.

Unless you are unusually lucky, the DVM reading will not be the desired 180.0mV. With luck, adjusting R_4 will enable it to be set to this value. If the desired setting is out of range, it will be necessary to increase or decrease the value of C_1, which was probably only a 20% tolerance type anyway.

Providing that R_5, R_6 and R_7 are all 1% tolerance types, then having set up the instrument on range B, the other ranges should also be – very nearly – correct. In principle, if R_6 is at one extreme of the 1% range, and R_5 and R_7 both at the other, there is obviously a 2% error, but you would be distinctly unlucky if matters were that bad.

Range C, and its extension to measure capacitors larger than 200nF, has already been mentioned. Range A covers capacitors up to 1999pF, or 1999.9pF, if using a 4½-digit DVM. With a resolution of 1pF – or even one tenth of a pF – it is necessary to make allowance for C_3 and circuit stray capacitance.

If there were no stray capacitance, and the monostable would work with a C_3 of zero capacitance, things would be fine, but that is obviously not practicable. Capacitor C_3 is needed to provide a defined minimum period for the

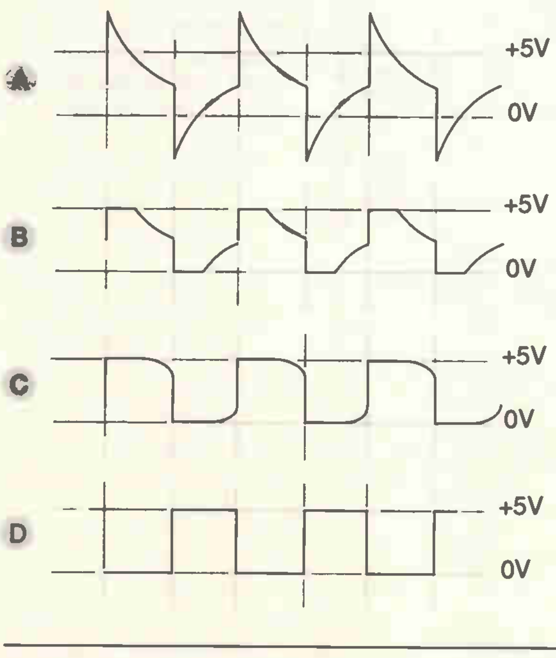

Fig. 2. Waveforms associated with the relaxation oscillator section of Fig. 1, based on $IC_{1a,b}$. Clean square output D feeds the next stage – namely the monostable multivibrator.

monostable, greater than the brief period for which its output would be high, due to the trigger pulse.

Why range A is different

A finite minimum pulse width, corresponding to C_0 – i.e. C_3 plus circuit stray capacitances – is unavoidable. So the next best thing is to know the value of C_0. If you know it, when using Range A, the value can be subtracted from the measured value to give the true value.

Capacitance C_0 can be determined with the aid of another standard capacitor: 100pF 1% is a convenient value.

Output voltage V, measured by the DVM, is proportional to the monostable period t. Further, t is proportional to capacitance C. So let the DVM read V_1 without the 100pF capacitor connected, and V_2 with. Then,

$$V_1 = K \times C_0$$

and

$$V_2 = K \times (C_0 + 100\text{pF})$$

Here, K is an unknown constant. As we have two independent equations, K can be eliminated, and C_0 determined.

A little elementary algebra yields,

$$C_0 = \frac{V_1 \times 100\text{pF}}{V_2 \times V_1}$$

This method of coping with C_0 on Range A can rightly be called 'practical, if inelegant'. Since Range B will be used for measuring capacitors greater than 2000pF, C_0 is less relevant, becoming insignificant on Range C.

Elegant, but impracticable

This capacitance meter circuit is not original to me. I saw something similar in one of the American controlled-circulation* electronics magazines. It used a most ingenious arrangement, with a second 'dummy' monostable circuit identical to IC_{1c}. This had the same C_{3b}, but no duplication of the provision for a capacitor under test of course.

Both monostables were triggered at the same time, and their outputs were connected to another XOR gate. The output of this XOR gate switched high only after the dummy monostable timed out, returning low after IC_{1c}'s output went low. Thus that part of the period due to C_0 was neatly subtracted from the answer, in hardware.

There is just one problem with this scheme; it is impossible to measure very small capacitors of a few picofarads or less. With a $3^1/_2$-digit DVM, on Range A the least significant digit corresponds to 1pF, while a $4^1/_2$-digit instrument can resolve to one tenth of a picofarad. But with the ingenious arrangement just described, this is all to no avail.

When a very small capacitor is measured, IC_{1c} should time out only microseconds after the dummy monostable – if that. But the negative-going edge of the dummy monostable output gets capacitively coupled into the input of IC_{1c}, all the active devices being in the same IC package, causing it to terminate early, in synchronism with the dummy monostable.

Often in electronics, as in life generally, when the elegant solution to a problem turns out to be impracticable, one has to settle for the practical solution – even if a trifle inelegant. ■

*A 'controlled-circulation' magazine is a freebie whose revenue comes from advertising only. Ed.

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