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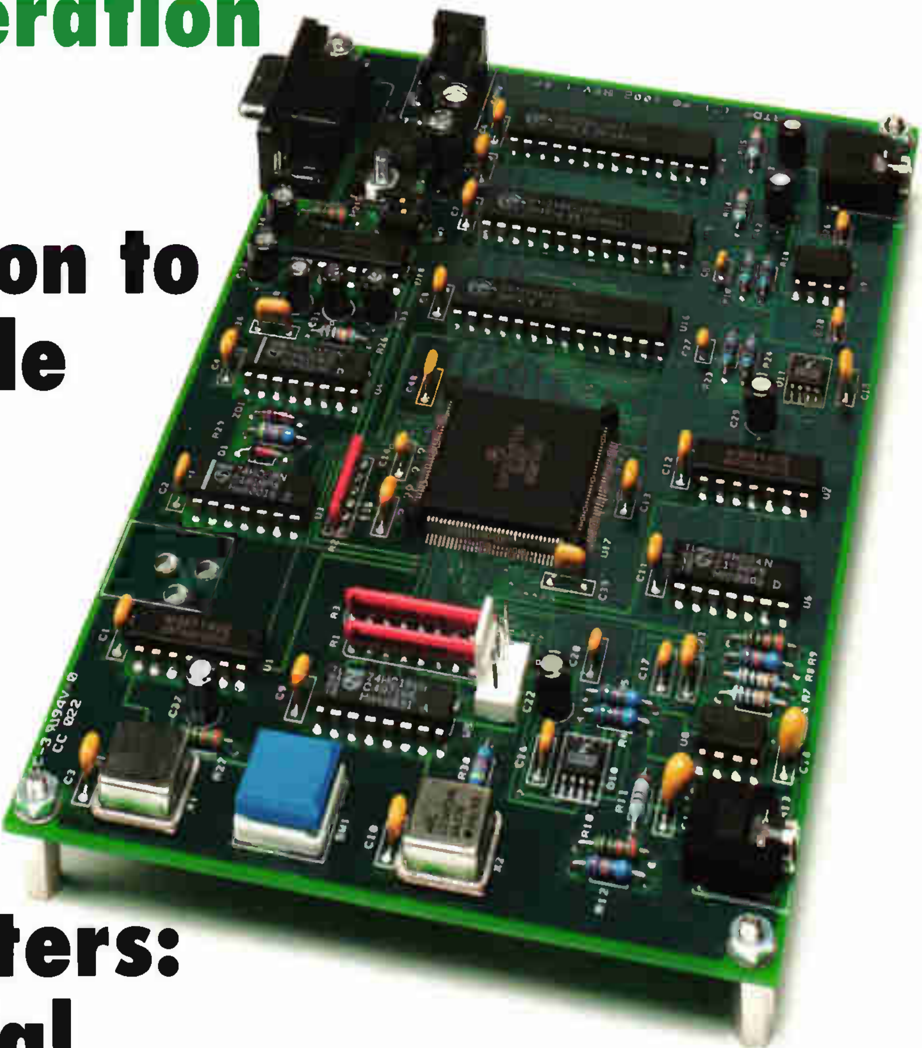

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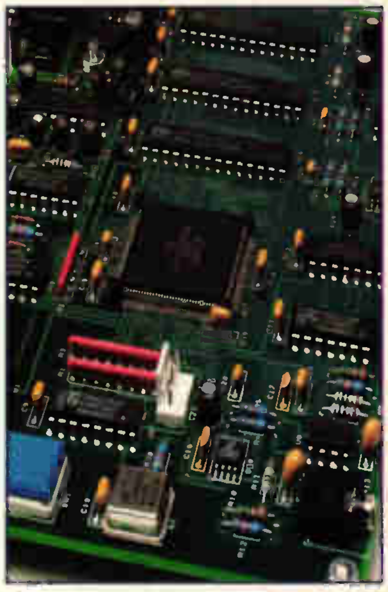
CONTENTS

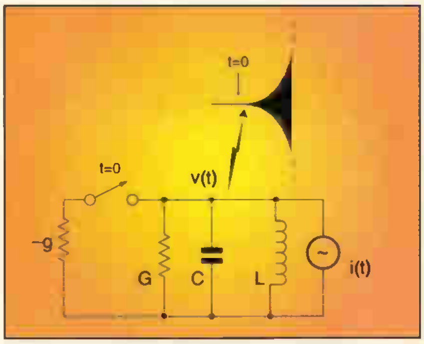
APRIL 2002 VOLUME 108 NUMBER 1792

The latest from the world of micromachining – a chain and sprocket drive – gets round the problem of misaligned gears. Read more about it on page 7.

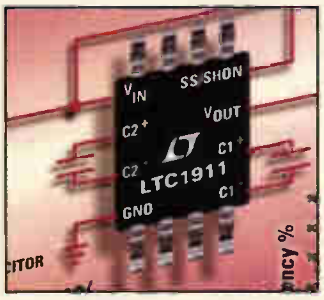
Among our extended letters pages this month are tips for designing and packaging active probes, and a discussion of Mosfet power amplifier design. Letters start on page 55.



Cover shows the real-time digital filter that Patrick Gaydecki uses to illustrate his article on digital filters, page 20.



Eddie Insam provides a rare in-depth description of super-regeneration and its application in wireless design, page 46.



This inductorless step-down DC-DC converter produces a 1.5V or 1.8V regulated output from a 2.7V to 5.5V input voltage. It'll run from a single Lithium Ion cell. Read more about this and a multitude of other new products starting on page 39.

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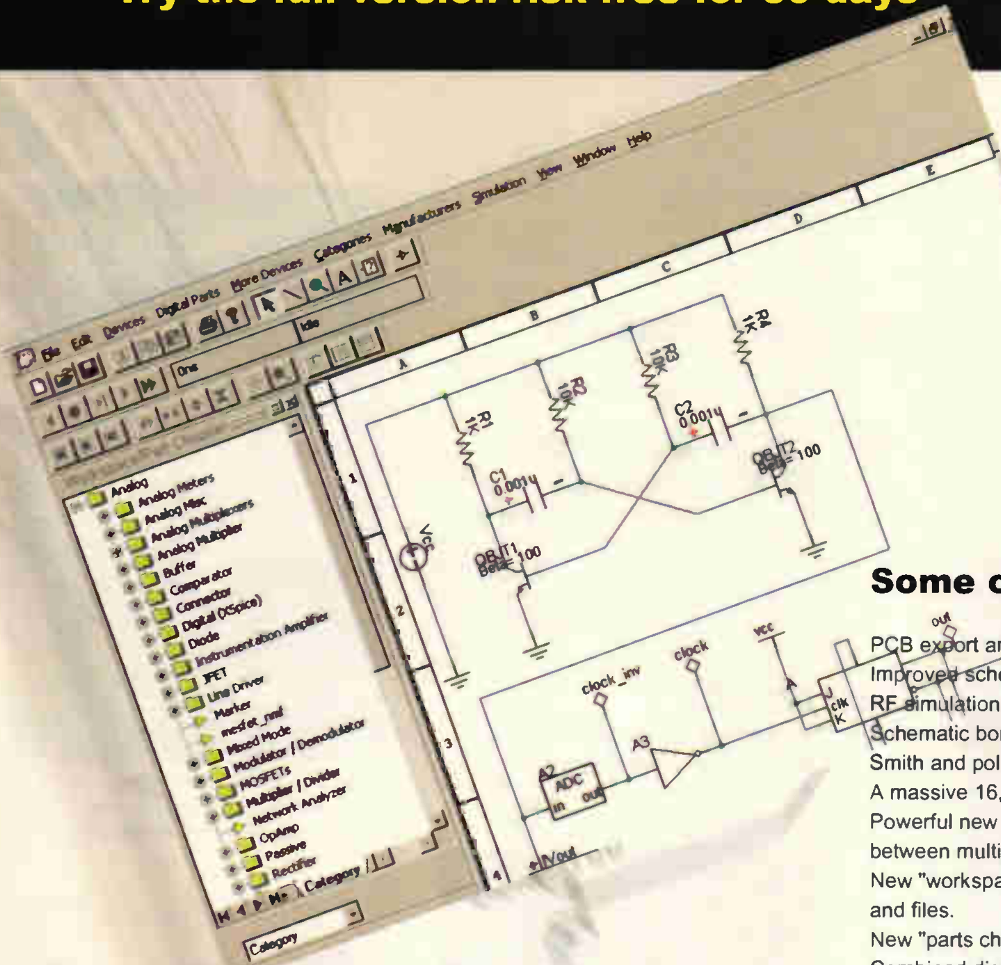

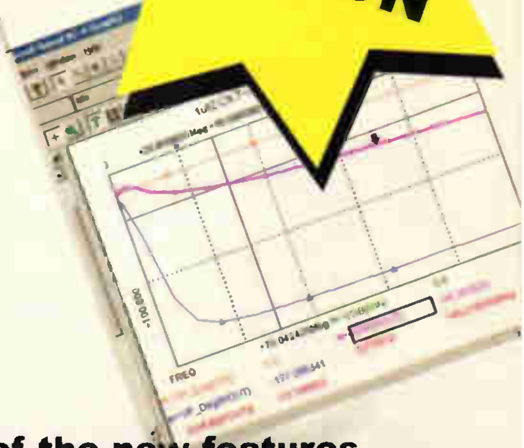

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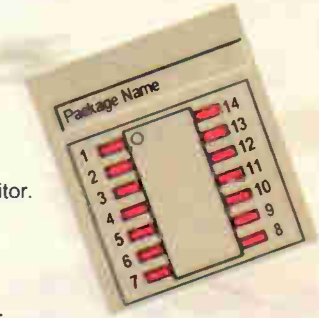

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Common sense

We need to throw out the sheep that are herded by the management consultant brigade. And we need to throw away those endless meetings.

How many man-hours have we wasted waiting for a long, urgent e-mail or file to download? Why is the UK falling behind in the provision of broadband Internet access, when high speed Internet connections are clearly desirable?

BT offers ADSL over twisted-pair cables. The old twisted-pair cables are being touted because they are in the ground; there is no other reason to use this tired old copper. In many cases the cable runs are too long and of too poor quality to support ADSL.

After all the investment in stretching the bandwidth from these old cables the customers will soon demand more bandwidth. Just as we have seen the memory fitted in the PC go from a few kilobytes to many gigabytes, the need for extra bandwidth to enable complex data and photographs to be transmitted over the Internet will certainly follow.

Cable companies are ideally placed to provide broadband modems and deployment is beginning. These operators already have last mile links to the home capable of passing many megabits. The backbone structure is built using modern fibre optic technologies. However there is a problem: many areas are not yet passed by the cable companies

Why then is the deployment of broadband so slow? Why isn't BT investing in new broadband links to every home: why are the cable companies not rolling out cable as fast as they can?

Simply, I believe, because of a shortage of money. Many of the companies that should be providing the service are in a financial mess.

Telecom company managers today are trying to run companies that are already in deep financial trouble. Many are suffering from a legacy of wasted money, fuelled by raising as much cash as possible on ill-conceived plans. There is no money left to improve networks or expand networks to new areas. I fear that very serious debt write-offs and even bankruptcies will have to precede any future broadband progress.

Electronics is an undervalued industry. The cost of the plastic case for a portable calculator is greater than the cost of the sophisticated electronics inside it. Why are we giving our skills away? Mobile phones are not valued because they are too cheap. Computers would still sell if they were to cost 50% more. The extra profit could be used to cushion the companies involved against the inevitable dips in demand.

Too many companies have the, "I want to go out and kill the competition," attitude. We have seen the result. Everyone ends up wounded. Companies should build steadily on their strengths, build expertise and charge the right price for the skills they have.

With enough cash margin, good companies could leave the over aggressive companies to give away product and burn themselves out in an attempt to buy a market. In the current climate, with companies running on borrowed money, any downturn causes panic moves that de-stabilise markets even more.

Further down the corporate structure in the telecom/cable companies, it is not the technology that causes problems to many customers: it is the lack of attention to detail and the lack of trained and caring staff. I am upset at the way that keen and competent people are driven from their jobs because they are de-motivated by office boys masquerading as their managers.

It is time to take very critical look at the telecom industry's management. There are excellent managers in the industry, but they are in the minority. We need to throw out the sheep that are herded by the management consultant brigade. And we need to throw away those endless meetings.

Good, dedicated engineers must be groomed to become managers and moved into top management positions. Engineers founded most of the successful companies in the electronics industry and continuing engineering knowledge is still of vital importance.

It is time to bring back common sense to management at all levels. If we do, the investors will return and help the telecom industry build a broadband broad-based future, and not the leaning tower of Babel that we have now. ■

Chris Swires, managing director of Swires Research

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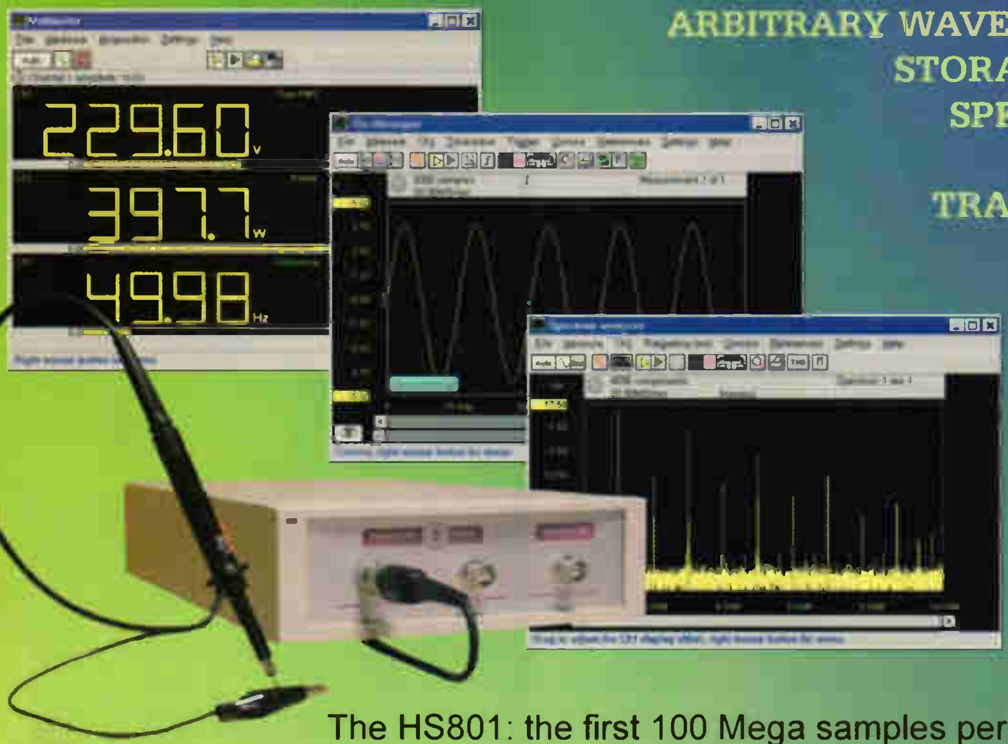

The HS801: the first 100 Mega samples per second measuring instrument that consists of a MOST (Multimeter, Oscilloscope, Spectrum analyzer and Transient recorder) and an AWG (Arbitrary Waveform Generator). This new MOST portable and compact measuring instrument can solve almost every measurement problem. With the integrated AWG you can generate every signal you want.

Reliabilty

- The versatile software has a user-defined toolbar with which over 50 instrument settings quick and easy can be accessed. An intelligent auto setup allows the inexperienced user to perform measurements immediately. Through the use of a setting file, the user has the possibility to save an instrument setup and recall it at a later moment. The setup time of the instrument is hereby reduced to a minimum.
- When a quick indication of the input signal is required, a simple click on the auto setup button will immediately give a good overview of the signal. The auto setup function ensures a proper setup of the time base, the trigger levels and the input sensitivities.
- The sophisticated cursor read outs have 21 possible read outs. Besides the usual read outs, like voltage and time, also quantities like rise time and frequency are displayed.
- Measured signals and instrument settings can be saved on disk. This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments.
- The (colour) print outs can be supplied with three common text lines (e.g company info) and three lines with measurement specific information.
- The HS801 has an 8 bit resolution and a maximum sampling speed of 100 MHz. The input range is 0.1 volt full scale to 80 volt full scale. The record length is 32K/64K samples. The AWG has a 10 bit resolution and a sample speed of 25 MHz. The HS801 is connected to the parallel printer port of a computer.
- The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3.xx / 95 / 98 or Windows NT / 2000 / XP and DOS 3.3 or higher.
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COMPUTER CONTROLLED MEASURING INSTRUMENT

Electricity from the sea floor

Microbes on the sea-floor could be harnessed to generate electricity, say researchers at the University of Massachusetts.

Scientists at the University managed to create small amounts of energy from organic marine sediments and organic pollutants.

Graphite wires sunk into an anaerobic area of sediments formed anodes, while wires formed the cathodes in the overlying oxygenated seawater. Microbes in the mud oxidise the organic compounds and provide around 16mW per square metre of electrode.

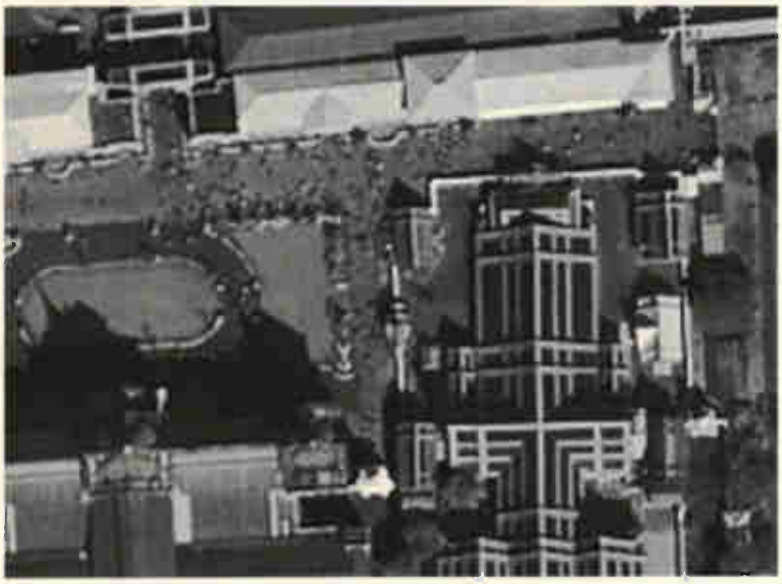
According to the scientists in the magazine *Science*, when the bacteria growing in the mud was analysed, one microbe tended to dominate – a member of the Geobacteraceae family. This is well known as an oxidiser of toluene.

Although the precise method of creating electrons is not yet known, the work could eventually lead to marine power sources and methods for reducing marine pollution.

Deep in the ice on the island of Spitsbergen, a NASA robot is melting its way down through a glacier. The cylindrical robot reached 23m by expelling hot water at its front. Instruments on the robot measure ice characteristics, avoiding having to hoist ice cores up to a lab.

Satellite pictures have a resolution of less than 61cm

Satellite images with resolution of well under a metre have started to come back from the QuickBird satellite. The black and white images down to just 61cm resolution are the best commercially available. Test images of Bangkok Palace show the detail possible from the camera. Until now such detail has been limited to military satellites. Operated by DigitalGlobe, QuickBird orbits 280 miles up in a 98 degree sun-synchronous orbit. The image swathe is over ten miles across. Full colour images reduce resolution to 2.44m.



New laser for free-space optical links

A mid infra-red laser suitable for free space optical links has been developed by scientists at the University of Neuchâtel in Switzerland.

The quantum cascade laser (QCL) emits at 9.1μm, above the worst region for water absorption in the atmosphere. Using the laser in a free-space optical link could more than double the range of existing systems to 5km or more.

The beauty of the Swiss device is its continuous wave (CW) operation at room temperature. Current CW lasers at this wavelength need cryogenic coolant systems to operate.

In addition the QCL uses a semiconductor structure, unlike the gas lasers normally needed to reach this wavelength.

Molecular-beam epitaxy is used to grow the InGaAs/InAlAs core on an indium phosphide substrate. Electrons liberate photons as they cascade through the layered structure of the device.

The issue of heat, which usually burns such devices out, is handled by burying the laser inside the InP substrate, which is very thermally efficient. In CW mode, the QCL runs at up to 312K with an output power of 3mW. When cooled to 292K it can output 17mW of optical power.

First portable music-player to feature hard-drive storage

SONICblue has unveiled its first hard drive-based portable music player – the Rio Riot. Designed to store an entire music collection, the Rio Riot is built around a 20Gbyte hard-drive that can hold over 5000 songs or 400 CDs worth of music.

The Riot also includes an FM tuner, lithium-ion battery and music management software. "To date, the most common feedback we've received regarding high capacity players is that the ability to store lots of music is great, but totally useless if you can't find the music you want to hear, when you want to hear it," said David Huffman, company v-p of marketing.

A 16Mbyte SDRAM buffer stores music during playback so the hard drive can be stopped to save power.

A 3D virtual glove has been shown off by US firm Essential Reality, which it claims will replace the mouse and joystick. The P5 is particularly suited to games with their 3D interfaces, said the firm. The glove uses a USB interface to communicate hand and finger movements to the PC. Cost is expected to be under $150 when it is released later this year.

Mass-storage flash memory density looks set to overtake DRAM

Flash memory is set to overtake DRAM as the density driver in the memory market, according to Chang-Gyu Hwang, chief executive of Samsung Semiconductor's memory division.

In a speech at this year's ISSCC, Hwang concluded that Moore's Law is set to continue for at least ten years, but that the drive to smaller, faster, cheaper memory chips will be led by mass-storage flash memory.

Flash is riding the technology curve far more aggressively than DRAM, said Hwang. If current trends continue, NAND flash densities will overtake DRAM shortly after 2005.

Hwang also pointed out that as servers, PCs, set-top boxes, games and networks demand different specifications, then the memory market will become more diversified. Bandwidth, latency, system performance, power and other metrics vary with each sector.

Other predictions made by Hwang include that servers will drive high density DRAM, 512Mbit DRAM will be widely available this year, while 16Gbit DRAM will hit the shelves within ten years, and that 1V DRAM will arrive within three years.

Technical innovations in the coming decade will include a shift to smaller memory cells in DRAM and flash, with areas of $4F^2$ (four times the minimum feature size squared) compared to today's $8F^2$ and $6F^2$ cells. Multi-bit cell will also become more common, Hwang claimed.

This children's development toy, created by Michael Cohen from Brunel University, has won the Toy Design Awards. Called the Rock 'n' Roll Ball, the toy has infra-red sensors in its 12 legs that allow it to create different musical notes when it is handled or placed on different surfaces. Fisher-Price, which organised the competition, said it is testing the ball and is planning for a Christmas 2004 release.

New micromachining first: a chain and sprocket

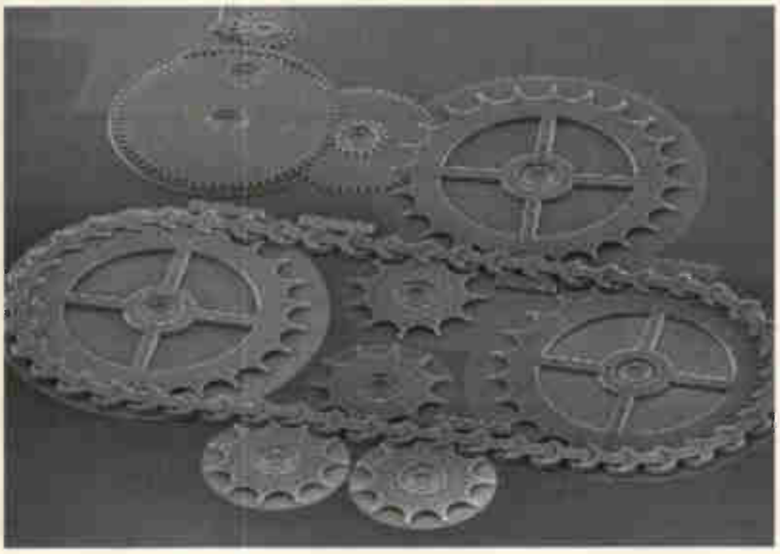
Albuquerque's Sandia Labs, famous for its micromachining stunts, has made a micro drive chain and sprocket set.

The lab is proposing chains to replace multiple motors in micro-electromechanical systems (MEMS). "All those drives take up a lot of real estate on chips," said Sandia technician Ed Vernon, who has a patent on the microchain.

Vernon fabricated a chain, rather than a belt, because belts are spring-like and produce too much torque on gears not aligned in a straight line. Each chain link can flex ±52° with respect to the preceding link without putting force on the support structure.

The longest unsupported span is 500µm. A microchain tensioner is needed to moderate longer stretches.

Sandia's Summit IV and Summit V processes were used in construction.



Quick draw mobile phones

Mobile phones of the future will come in holsters containing methane powered fuel cells, if US technology firm Manhattan Scientifics has its way.

The fuel cell specialist has been awarded a patent covering a holster style design for charging mobile phones from its methanol and water powered cell.

Manhattan reckons the technology can also be used to power other mobile equipment such as laptop PCs, camcorders and power tools. Moreover, the firm reckoned the cells can be manufactured cheaply on a plastic film in a reel-to-reel process.

Digital oscilloscope has 5GHz front end

LeCroy has introduced a new 5GHz machine right at the top end of the digital-storage oscilloscope tree.

To achieve its 10Gsample/s per channel performance, the company has had three custom silicon-germanium chips made by IBM. These are a 5GHz adjustable-gain front-end amplifier, a 10Gsample/s 8-bit flash A-to-D converter and a trigger chip.

Its four channels can act independently at 10Gsample/s each or be combined as two pairs, providing 2 by 20Gsample/s channels.

Possibly the most impressive feature of this machine is its high speed sample acquisition memory. A whopping 24Msamples per channel can be acquired, even at full speed.

To keep screen update rate high, LeCroy has created a new internal architecture that passes data from the acquisition memories to the main processor bus in packet form through two 100Mbyte/s Ethernet links.

Once on the main bus, data is passed directly to the cache memory of the Pentium III-class processor for display processing.

40 automatic measurements can be made on waveforms, and user-defined routines can be added to this list. $70 000 and it's yours.

WaveMaster specifications

4 channels
5GHz/chan bandwidth
10Gsample/s/chan. (20 in 2-channel mode)
24Mpoint/chan. memory
(48Mpoint/chan. in 2-channel mode)
5GHz trigger bandwidth
26cm 800-by-600 colour LCD
26 math tools
40 auto measurement tools
Cost – around $70 000

Still not enough electronics graduates

Numbers of electronics graduates are expected to continue to fall short of industry needs, as college entrants move to arts and media subjects.

This conclusion comes from industry representatives as the number of students taking science and engineering subjects drops, despite a five per cent rise in university admissions this year, according to figures from UCAS.

For HND electronics the figures are worst, with a drop of 25 per cent in college entrants, while those starting degrees in electronics dropped several per cent as a percentage of the total.

Similar figures afflict other engineering subjects, maths and the physical sciences.

"It's a real concern for the electronics sector that potential supply in the labour market is declining at such an alarming rate," said Anthony Parish, director general of the Federation of the Electronics Industry (FEI).

"It's particularly disappointing given that, despite tough market conditions worldwide, demand for the right skills remains high," he added.

The Institution of Incorporated Engineers' head of professional development, Katy Turff, said: "It is disappointing that there's this continued decline in people going into engineering."

According to Turff, the newly formed Engineering and Technology Board (ETB) will lead efforts to persuade more students to take maths and science subjects at school, leading to engineering at college.

"If the Board is successful we should see a reversal of this trend over the next few years," she said.

The ETB came into being this year, and was spun out of the Engineering Council to deal specifically with the promotion of engineering.

Science and engineering are suffering at the expense of 'trendy' subjects such as media studies, business and computer science. In fact six times as many people take either business management or computer science degrees as take electronics.

So, you thought that your 2GHz Pentium was fast...

Blue Gene, the most powerful number cruncher under construction, has spawned a progeny.

To be called Blue Gene/L, it will be a 200Tflop machine – the original aims at 1Pflop – jointly developed with the US Lawrence Livermore National Laboratory.

IBM designed and is funding the original BlueGene to calculate how proteins fold into the special shapes that make them biologically active.

"Our initial exploration made us realise we can expand our Blue Gene project to deliver more commercially viable architectures for a broad customer set, and still accomplish our original goal of protein science simulations," said Mark Dean, vice-president of systems at IBM Research.

Researchers plan to use Blue Gene/L, which should be finished by 2005, to simulate explosions, fires and the aging of materials, said IBM.

IBM and Livermore have worked together before, most significantly on the ASCI (Accelerated Strategic Computing Initiative) program. IBM built ASCI White, the most powerful computer yet finished, for Livermore.

"Up until now, ASCI supercomputers have been designed to address the entire spectrum of numerical simulations required of the [nuclear] stockpile stewardship effort," said David Nowak, ASCI Program Leader at LLNL. "Blue Gene/L can address an important subset of those computational problems."

Unlike ASCI White, BlueGene is more calculator than general purpose computer.

New package adds just 0.1mΩ to Mosfet on resistance

Cutting the package resistance of power Mosfets is a big driver for designers, so International Rectifier's 0.1mΩ design is something of an achievement.

IR's UK design team in Surrey carried out the development work that aims to double power densities.

The DirectFET package places the Fet upside-down inside a metal can, which forms the drain of the device. The source and gate on the top of the die – the base of the package – connect directly to the circuit board.

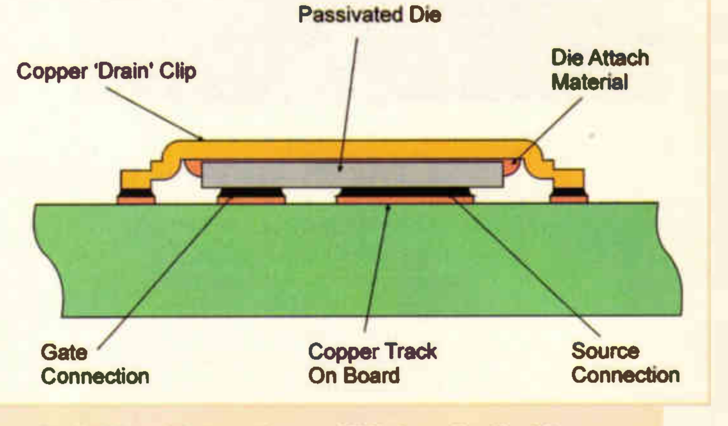
The firm claimed that a pair of the devices is enough to control a 30A DC-to-DC converter, which doubles power density over traditional SO-8 packaged Fets.

The team designed the package in this way to allow heat to be taken out of the device through the top, via a heat sink or chassis connection. "Today a lot of heat goes into the PCB and the board is saturated from the heat," said Odile Ronat, director of marketing at IR.

The large connection area to the drain substrate dramatically reduces the thermal resistance of the package to about 3°C/W. Similarly the base of the package has thermal resistance of 1°C/W or less, much lower than the SO-8's 20°C/W.

"We can remove 50 per cent more heat through the top of the device than the SO-8," said Ronat.

Four devices are available from IR in the DirectFET packaging,



two with 20V breakdown and two at 30V. One of each of the devices can source up to 28A with $R_{DS(on)}$ of 2.8mΩ and are aimed at the synchronous FET position in a converter.

The two 14A devices for use as the control FET have higher $R_{DS(on)}$ of 8 and 10mΩ, but offer better – i.e. lower – gate charge of 5nC or less.

The 0.7mm high devices contain no lead or bromide, said IR.

International Solid-State Circuits Conference report

51GHz VCO in CMOS

At this year's International Solid State Circuits Conference, a team of engineers from Infineon outlined a 51GHz voltage-controlled oscillator made entirely in CMOS.

VCOs are an integral part of communications systems, but are becoming increasingly harder to design as systems move to higher frequencies. Marc Tiebout from Infineon pointed out that to reach 51GHz would normally require an exotic process technology, such as gallium arsenide or silicon germanium, rather than standard CMOS, albeit a 0.12μm process.

Tiebout's design uses metal layers four to six to form a three winding, centre tapped inductor, while NMOS transistors form the varactor.

The 0.5 by 0.9mm chip is dominated by pads – the core is just 60 by 40μm. From a 1V supply the circuit has a power consumption of 1mW.

Organic LEDs break new ground

An organic LED microdisplay was the subject of a paper presented by display company eMagin and chip designer Tality at the International Solid-State Circuits Conference (ISSCC) in San Francisco.

Organic LED (OLED) displays, as soon as the technology is reliable enough, will be a competitor to emissive colour LCDs.

The eMagin display is a silicon backplane device with a chip supporting the display material and providing drive and control logic.

Like all LEDs, organic LEDs have to be driven with a fixed current to get a fixed brightness. This differs from just about every other display technology, where brightness is dependent on pixel voltage.

This means that OLED display makers have to make up their own drive schemes as they go.

eMagin was faced with a particularly difficult drive problem because its extremely small pixels run at around 1nA – too low for reliable delivery by traditional MOS current sources operating in their saturation region, said the conference speaker.

To get over this the companies developed a sub-threshold driver where current sources operate in the sub-threshold region where they are almost off.

Controlling their gates in this mode is a tricky business and the engineers developed an ingenious self-adjusting sample-and-hold scheme to set the current.

The final driver reliably delivers currents between 250pA and 25nA per pixel.

Overall the display has 853 by 600 colour pixels covering 16 by 14mm.

The 10-million transistor chip consumes 200mW and is made with conventional 0.35μm 3.3V CMOS.

Each column of pixels has a driver like this. Rows of pixels are loaded simultaneously. A current signal depending on the required pixel brightness sets the current mirror on the left hand side of the diagram to draw an appropriate current down the column bus.

In the sample phase, transistors M_2 and M_3 are on, and M_4 is off. As all the other rows are off, this configuration forces the column current to be draw through gate capacitor C and current source M_1. Capacitor C charges until the gate voltage of M_1 sets the current source exactly to the bus current. The circuit has automatically set itself to supply the right current.

In hold mode, M_2 and M_3 turn off, fixing the gate voltage and therefore the current in M_1. Transistor M_4 turns on and directs the current to the pixel.

To get the settling time right in hold mode, it was found that 100 times the pixel current had to be set in M_1. A step voltage, the right height to divide the source output by 100, is applied to the other side the capacitor during hold to compensate for this.

CMOS sensor replaces film in SLR cameras

A CMOS image sensor, aimed at replacing film in standard, unmodified SLR camera bodies, has been developed by STMicroelectronics in Edinburgh. The firm hopes the device will allow it to break into the high performance digital stills camera market, at present dominated by CCD sensors.

Described at ISSCC, the sensor combines a 1808 by 1120 pixel array with an RGB colour filter. On-chip processing allows three frames a second to be output by the device.

However, operating in a conventional camera constrains the sensor to work with the camera's shutter, thus its only knowledge of when a picture is being taken is when the shutter opens.

Rather than waiting for the shutter to open and then starting to integrate, the ST sensor continuously integrates, resetting on a regular basis if no light is present. During the time the camera's shutter is open, the reset pulses are inhibited.

To detect the opening of the shutter and inhibit reset pulses, 16 pairs of photo-transistors are arranged around the edge of the array. In case no light hits the edge, regular samples are also taken from various points on the array itself. This is possible because the read from a pixel is non-destructive.

While the shutter is closed, the device's power consumption is under 250μW.

Because the sensor has to mimic 35mm film it is much larger than normal – the array itself is 36 by 22mm. Therefore several die are stitched together, the array being formed from two halves, while the analogue readout, shift registers and pads are made on another die. Only metal lines run across the boundary between die.

Single-inductor, multiple-output switching PSUs

A development of a power supply topology first aired at last year's ISSCC was the subject of a paper from two Hong Kong universities at this year's conference.

The topology is called single-inductor, multiple-output (SIMO) and is a way to save cash and space by sharing an inductor, switching it between outputs one at a time.

SIMO regulators have a major shortcoming. Unless carefully controlled, a variable amount of current is left in the inductor at the end of an output's cycle. This gets passed to the next output and causes output interaction.

One method to prevent this interaction is to ensure current is always zero when the inductor is passed on.

This is called discontinuous mode operation, and unfortunately increases system noise and losses as higher peak currents are needed in the inductor to transfer a given amount of power.

In the new circuit, the extra switch is added across the inductor of a discontinuous-mode SIMO regulator.

This is used to short out the inductor and maintain a fixed current after the output has finished using it. The same current is always passed on, so there is no output interaction, and current always flows in the inductor, so noisy and lossy discontinuous operation is avoided.

A 550mW regulator was made on 0.5μm standard 3.3V CMOS. Power efficiency is 85 per cent at 310mW with 1.2-2V inputs and outputs at 2.5 and 3V with 20mV and 25mV ripple respectively. Operating frequency is 1MHz.

What's that you're wearing?

Poma, the first wearable consumer computer from Xybernaut, got its public airing at the 2002 Consumer Electronics Show in Las Vegas. Based on Hitachi's SuperH 32bit 128 MHz Risc processor, it includes a Compact Flash slot, USB port, 32Mbyte RAM, 32Mbyte ROM, a custom optical mouse and a removable internal rechargeable battery. The 300g device runs Windows CE and includes instant on/off Internet, said Xybernaut. Hitachi contributed to development, has licensed the concept, and will also be making wearable computers. Poma costs $1499 in the US. For this you get the computer itself and the head-mounted display.

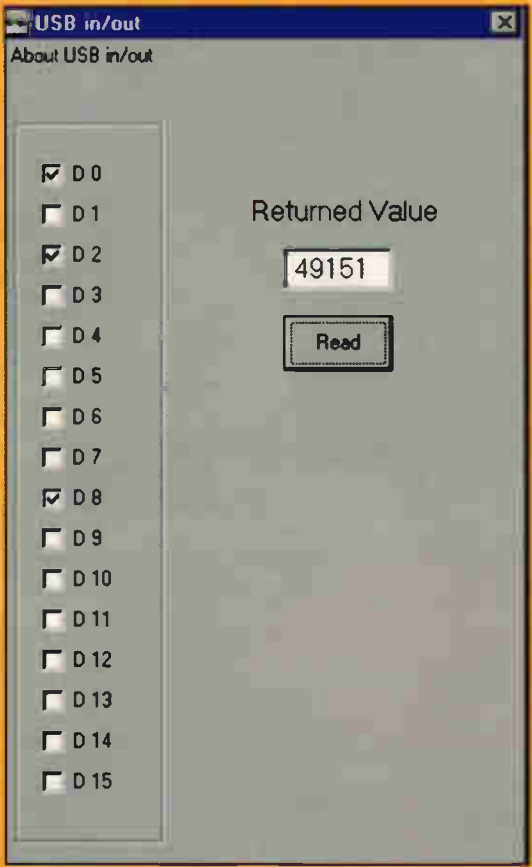
Researchers patent new organic LED enhancement

A new manufacturing process for organic light-emitting diodes, or OLEDs, has been developed and patented by Princeton University.

The technique extends the organic vapour-phase deposition (OVPD) process used to manufacture OLEDs by using low pressure gas. Atmospheric pressure OVPD can lead to rougher surfaces than desired, while the low pressure version produces smoother surfaces, essential for flexible substrates.

US flat panel display firm Universal Display Corporation (UDC), which developed the flexible OLED pictured, is expecting to use the process. ■

16-bit i/o via USB



Colin Attenborough has abandoned his love affair with the printer port and cosied up to USB instead. Here he shows how data can be received and sent via a USB port. Colin demonstrates how, with the aid of a programmable logic device and a USB interface module, you can make a 16-bit input and output port, controlled using Visual Basic.

Why did I move over to USB for interfacing? Well, the newer Microsoft operating systems don't allow programmers to write to the printer port using _outp(<printer port address>, <data>). That spoils the fun of those of us who like to get data into and out of PCs on the cheap.

Of the alternative i/o methods available, USB is an attractive approach from the user's point of view. But is complex from the interface and hardware viewpoint – and remember that it's not available under Windows NT.

An easy way into USB

The FT8U245AM integrated circuit from Future Technology Devices International (http://www.ftdichip.com) contains the core circuitry for a USB interface. This IC is at the heart of the DLP-USB1 module made by DLP Design (http://www.dlpdesign.com/), and available through FTDI.

Costing a little less than £30 at the time of writing, this module allows the transfer of data through the USB port using a first-in-first-out-like structure. Its connections are on a 0.1in grid, allowing use by those of us lacking 20:20 vision.

Note that the pins of the module are sized for perforated boards rather than IC sockets. This article uses the module rather than the chip.

Software options

FTDI offers two software approaches; the device can be addressed as if it were attached to a serial port, if 'virtual com port', or VCP, drivers are used. Alternatively, a dynamic-link library (DLL) allows direct drive.

Anticipating that you may want to use Visual Basic in conjunction with the module, and mindful of the fact that the cheaper versions of Visual Basic don't allow control of the serial ports, I've chosen to use the direct drive approach. FTDI also provides royalty-free USB port drivers for Windows 98 and Windows 2000.

Programmable logic: an overview

I've dealt at length with Lattice programmable logic devices in two earlier articles in the February 1999 and November/December 2000 issues. For those of you who are unfamiliar with programmable logic, they represent a very attractive way of getting from logic idea to logic hardware.

All you have to do is enter the circuit using free software from Lattice, compile a file from the circuit, and download the file to the programmable logic device. Though free, the software is large – several tens of megabytes – and you'll need to apply for a licence. This can be done automatically and painlessly at the Lattice web site: http://www.latticesemi.com/.

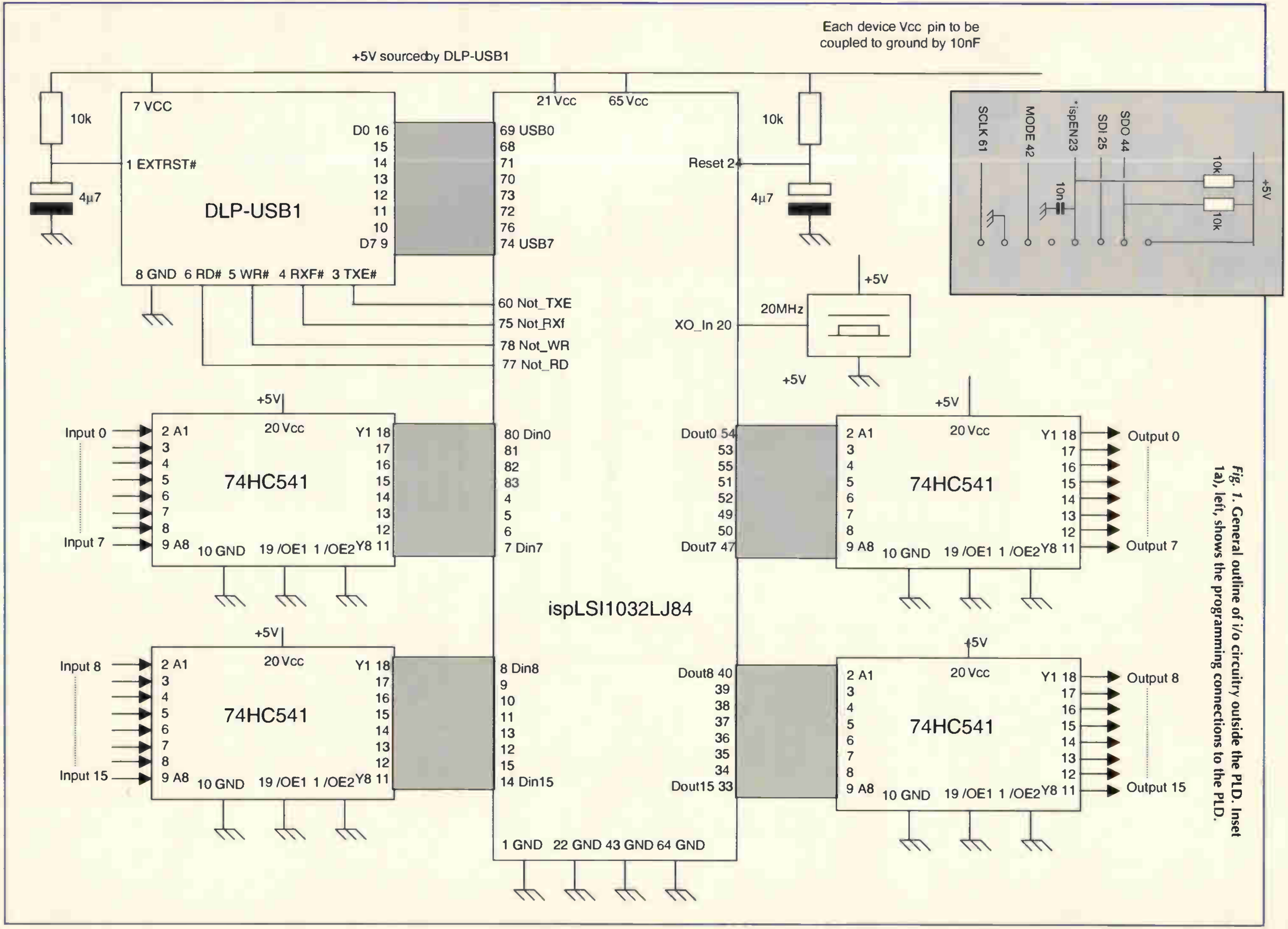


Fig. 1. **General outline of i/o circuitry outside the PLD. Inset 1a), left, shows the programming connections to the PLD.**

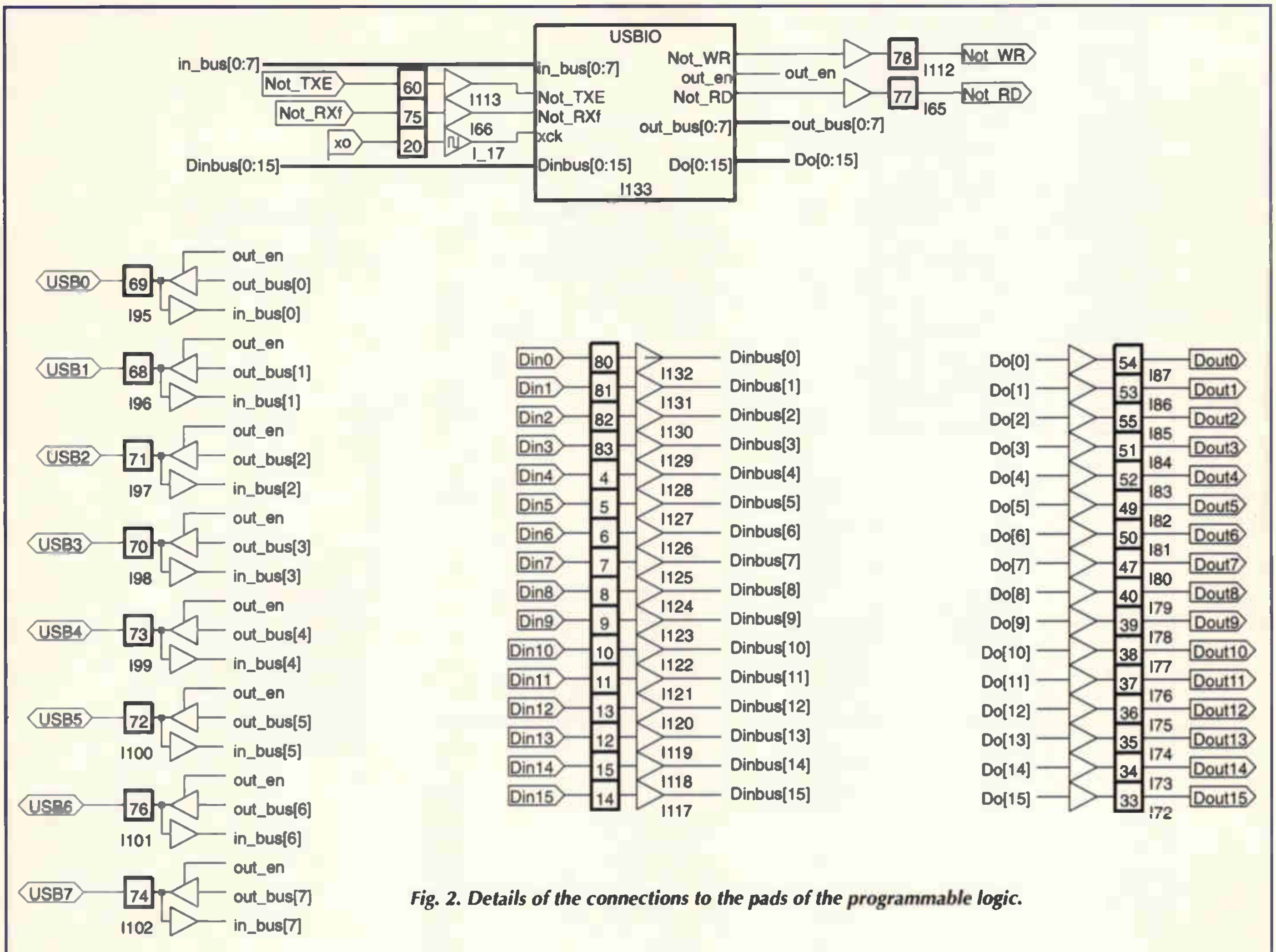


Fig. 2. Details of the connections to the pads of the programmable logic.

DLP-USB1 module connections

The module accepts a standard USB cable, and provides 5V and data signals to the computer.

Power to peripheral. 5V and 0V are provided, derived from the PC's power supplies; up to 500mA can be supplied to a peripheral.

Reset. The EXTRST# input resets the device when taken low.

Data i/o. Eight tristateable lines carry data in and out of the module. Activity on these lines is controlled by the four following lines.

Data from PC ready. RXF# is an output that goes low to show that data are present in the first-in-first-out register's 128-byte receive buffer.

Acknowledge data from PC. When RXF# has gone low, data may be read from the data i/o lines of the module by taking the RD# input low for at least 50ns. The data i/o lines then become outputs, and data can be read from them. They return to the high impedance state when RD# is taken high again; RXF# returns to logic 1 in response to RD# rising.

Read data from peripheral. The module's WR# input is normally high. When taken low, it reads data on the module data i/o lines into the first-in-first-out register's 384 byte transmit buffer.

Transmit-buffer full. When the transmit buffer is full, the TXE# output goes low. This must be used to inhibit the writing of further data to the module.

Using the module

Having sung the praises of the module, I must admit that the interface provided is most appropriate to a system using a microprocessor. I set out to provide simple 16-bit input and output without recourse to a microprocessor, using a programmable-logic device (PLD) instead.

Writing from the PC to the peripheral. When RXF# goes low, the PLD line holds RD# low for four clock cycles. Data from the USB module are latched during the third of the four clock cycles.

We want 16 output bits, but the bus from the module carries only eight. To get the 16 bits in the correct positions, I've chosen to send four words, with the lower nibble (four least-significant bits) as the data proper. The most-significant nibble is an address field that determines which of four four-bit latches stores the data proper.

Output from the four-bit latches can't be used directly. This is because the four-bit sections arrive sequentially rather than simultaneously. To ensure simultaneous changes in all output bits, a 16-bit register follows the four-bit registers, and is updated after data have been latched into the last of the four-bit registers.

Writing from the peripheral to the PC. To read from the peripheral, the PC must make the module's WR# input go low, once for each byte to be read. Take care to avoid the tristate lines of the PLD, and of the module,

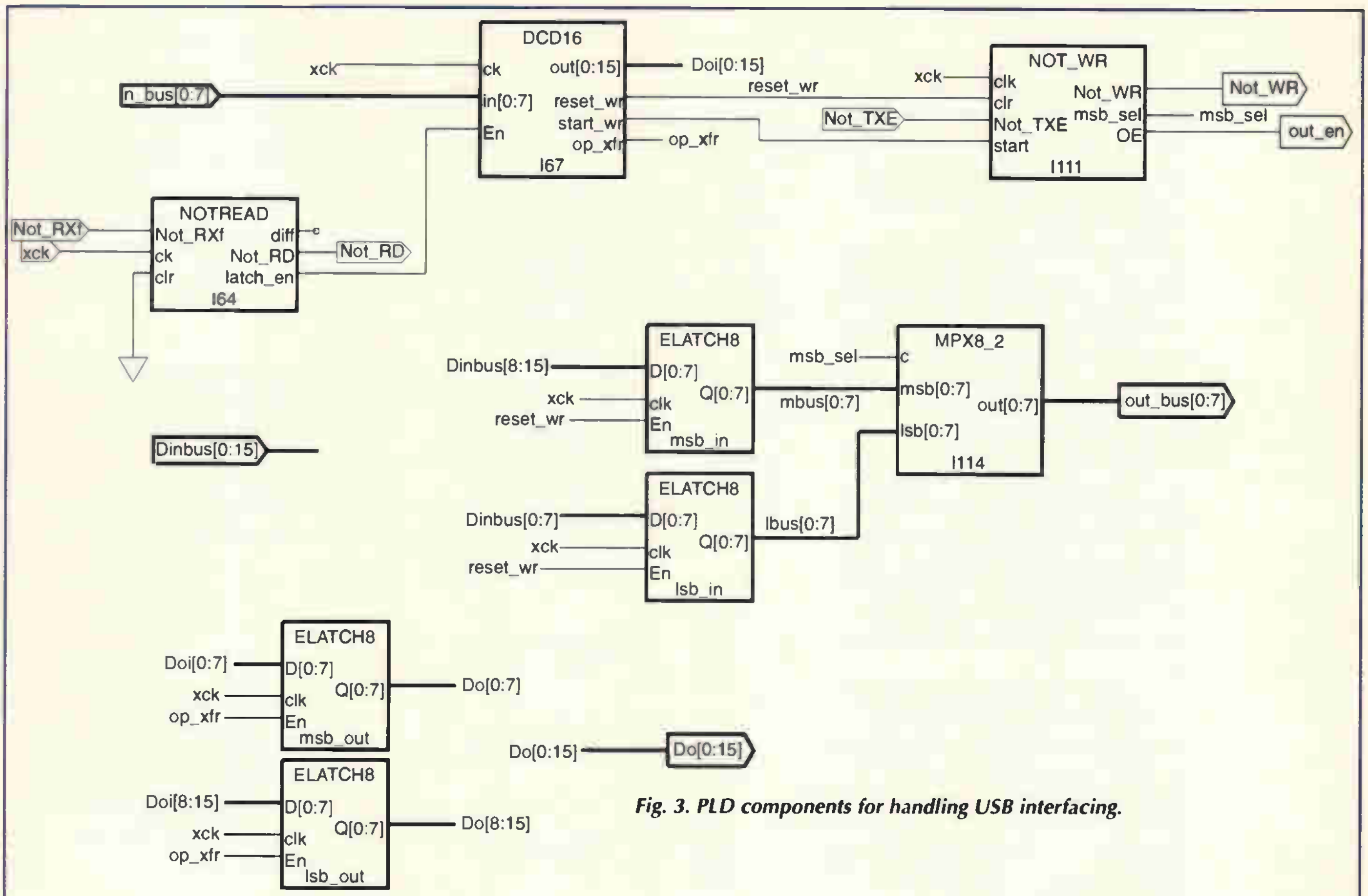


Fig. 3. PLD components for handling USB interfacing.

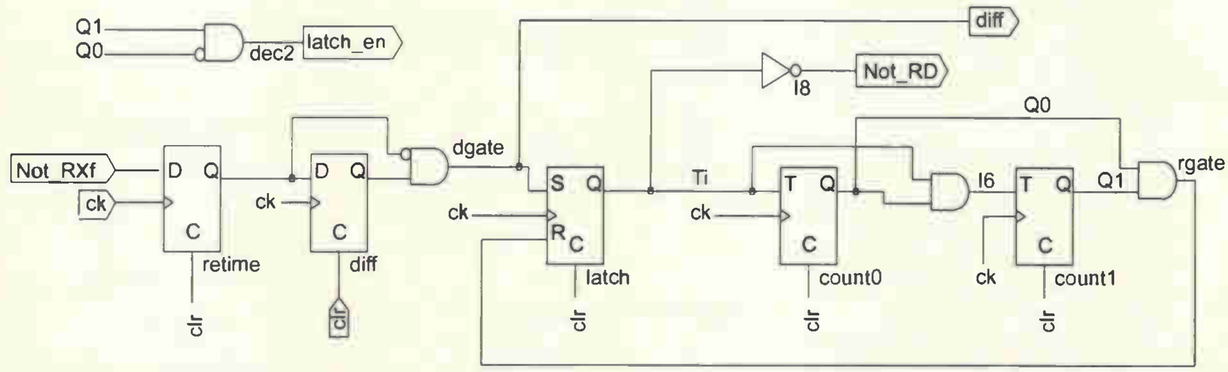


Fig. 4. The 'NotRead' area of the PLD. This circuit section responds to a low state on RXF# by generating a low level on RD# and providing a pulse to enable latches to catch data on the data bus.

both being in the output state at he same time.

A series of D-type flip-flops gives the necessary delay between sending the read instruction and the peripheral putting data on the module's i/o bus. At the start of the process, the 16 bits are latched into a register.

An eight bit wide multiplexer selects the bits eight at a time for feeding to the module's output; two read cycles (logic 1 to logic 0 transitions on WR#) are needed to transfer all 16 bits.

Circuitry outside the PLD

Figure 1 shows the general outline of the circuitry outside the PLD: Fig. **1a)** shows the programming connections to the PLD.

Power for the interface comes from the PC via the DLP-USB1 module. A 20MHz oscillator module provides a clock signal for the PLD.

On the assumption that, in the case of mishap, it's better to frazzle something cheap rather than something expensive, the inputs and outputs of the interface reach the PLD through 74HC541 buffers.

Details of the PLD

I've used a Lattice Semiconductors ispLSI1032E device; it has the necessary number of inputs and outputs. Also, when inserted into an appropriate socket, it has a 0.1in pin-out like the module.

Figure 2 illustrates the 'Top_pads'. This level of the circuitry shows only the connections to the pins of the device. Representing the complexities of the circuit as the next layer down in the hierarchy allows a clearer view of the function of the circuit without the distraction of the (many) input and output connections – at least I think so...

All but one connection to the USB i/o module emerges through input and output pads to the pins of the device. The exception is the 'out_en' signal, which controls whether the data connection to the USB module is an input or an output.

Signals USB0 to USB7 are the bidirectional data lines to the DLP-USB1; Not_TXE, Not_RXf, Not_WR and Not_RD are the data control lines. Lines Din0 to Din15 and Dout0 to Dout15 are the interface's input and output lines respectively.

Figure 3 shows the USB i/o section. This circuit shows the top-level circuit elements and their interconnections. The function of the 'NotRead' circuit area, **Fig. 4**, is to respond to a low state on RXF# by generating a low level on RD# and providing a pulse to enable latches to catch data on the data bus.

The D-type flip-flop labelled 'retime' guarantees a pulse at least one clock pulse wide at the input of 'diff'. A one clock pulse wide signal is generated by 'dgate' at the falling edge of the 'diff' output. This sets 'latch', outputting the Not_RD signal and allowing the four-state counter 'count1' and 'count2' to run.

Now that Not_RD is low, valid data are output on the module's bus. To catch the data while it's valid, the signal latch_en is generated by gate 'Dec2', which decodes state 2 of the counter (numbering the states 0 to 3). State 3 of the counter is decoded by gate *rgate* and used to reset *latch*, sending the Not_RD signal back to logic 1.

In the DCD16, **Fig. 5**, data on the bus are latched by the latch_en output of NotRead. Gates 'dec1' to 'dec5' decode the upper four bits of the data on the bus in(0:7).

During the latch_en pulse, one of the four four bit latches 'I1' to 'I4' is enabled, capturing at its output the data the lower four bits on in(0:7). For example, if the four most significant bits are 0010_2, this is decoded as 2 by gate 'dec2', and latch 'I2' is enabled.

The Visual Basic software that drives the interface is configured so that bits 0 to 3 of the 16 bit output are stored in latch 'I1', bits 4 to 7 in 'I2', bits 8 to 11 in 'I3', and bits 12 to 15 in 'I4'.

Gate 'dec4_1' provides the pulse which transfers the outputs of 'I1' to 'I4' to the 16-bit output latch. Gate 'dec5_1' provides, via a series of D-type flip-flops to provide delay, clear and start pulses to the not_wr area of the circuit.

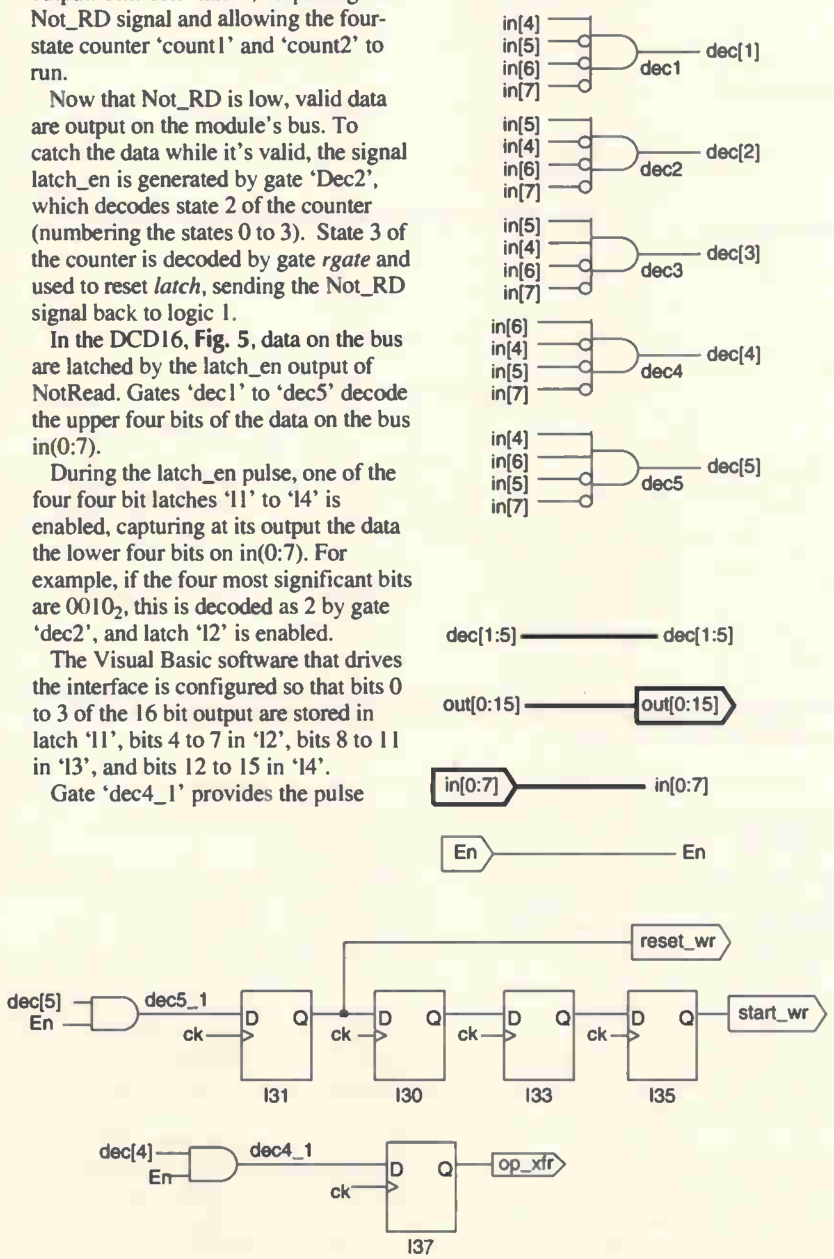
In the Latch4 circuit element, shown in **Fig. 6**, there are four D flip-flops with enable inputs. A two-input AND gate feeds the enable input, so that the flip-flops are enabled when the latch_en pulse and the decoder gates 'dec1' to 'dec4' output logic 1s.

There are four locations where the Elatch8 circuit of Fig. 6 is used – to latch the 16 bits of data into the PLD, and to clock out the four nibbles of data so that all data at the PLD output change simultaneously. The circuit of this element is simply a group of eight D-type flip-flops with enable and data inputs. These have common clock and enable signals.

The last section of Fig. 6 is Mpx8_2. To select which of the bytes of input data from Elatch8's 'msb_in' and 'lsb_in' inputs is fed to the USB module, an eight-pole two-way switch is needed. This function is provided by Mpx8_2. It consists of eight two-to-one multiplexers with a common control line.

Sending data to the PC

Writing the data from the peripheral to the USB module is controlled by the not_wr area of the circuit, **Fig. 7**. Writing data with the most-significant nibble set to five generates two pulses. The first goes to the clr input and the



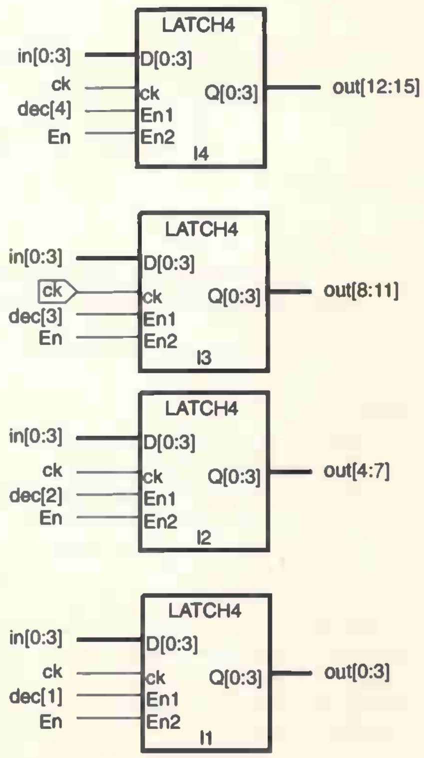


Fig. 5. In the DCD16 section of the PLD, data on the bus are latched by the latch_en output of NotRead. Gates 'dec1' to 'dec5' decode the upper four bits of the data on the bus in(0:7).

second to the start input of not_wr.

The clr signal resets not_wr; it also latches input data into the Elatch8's 'msb_in' and 'lsb_in' inputs . The start input sets 'latch', allowing the counter composed of 'count1'. 'count2' and 'count3' to run.

Setting 'latch' also puts a logic 1 on the OE output, which sets the tristate gates connected to the USB module's data lines in the output state. The connection to Not_TXE inhibits the counter if the transmit buffer overflows.

The second clock pulse to the counter – after it has been enabled – generates a falling edge on the Not_WR output; this is fed to the WR# input of the USB module, reading data on its data lines into its transmit buffer. At that time, the data are the least significant byte of the data being read, fed through the mpx8_2 circuit.

Two further clock pulses return Not_WR to logic 1, and also send msb_select high. This drives the mpx8_2 module, putting most significant byte of input data on the USB module's data lines. After two further clock pulses, Not_WR again falls, and data are written to the USB module.

When the counter reaches the all-ones state, the next clock pulse resets 'latch', inhibiting the counter and returning the Not_wr circuit to its initial state.

Software

There are three items of software to be considered; the USB drivers, the dynamic link library, or DLL, and the Visual Basic user-interface software. The DLL is called 'Ftd2xx.dll'. It adds the USB related functions, which can be called from Visual Basic.

The first two items can be downloaded from FTDI's web site.

How to obtain the software

A copy of the necessary Visual Basic and PLD software can be obtained by free of charge by e-mailing electronics.world@ntlworld.com. This does not include the source code. Bear in mind though that the files are 2.2Mbyte. Alternatively, you can receive a CD containing these files by sending £7 to 16bit i/o, Electronics World, Cumulus Business Media, Anne Boleyn House 9-13 Ewell Road, Cheam SM3 8BZ. The source code is available on CD for £25 fully inclusive. All remaining software is available from the web site: http://www.ftdichip.com

The DLL is embedded in the object-code version of the software; in the source code, it is already provided and in the correct position relative to other files. The USB driver must be downloaded from the ftdi web site.

The Visual Basic section doesn't so much draw heavily on the code sample on the FTDI web site; to paraphrase Terry Pratchett, it more scribbles all over it.

On the output side, the Visual Basic user interface consists of 16 check boxes to decide the state of the 16 output lines. Changing any of the 16 check boxes sends a series of four bytes out of the PC's USB interface.

Data are represented as a series of characters. As stated in the 'Details of the PLD' section, under DCD16, the first byte, latched into 'l1', has bits 0 to 3 as its least significant nibble, and 1 as its most significant nibble. Similarly, the second, third and fourth bytes, to be latched into 'l2', 'l3' and 'l4' have respectively bits 4 to 7, 8 to 11, and 12 to 15 as their least significant nibbles, and 2, 3 and 4 as their most significant nibbles.

The value representing the 16 input lines is read and displayed in a label when a command button is pressed. This process is initiated by sending a byte with 5 as its most significant nibble- see the Not_wr under the heading, 'Details of the PLD'.

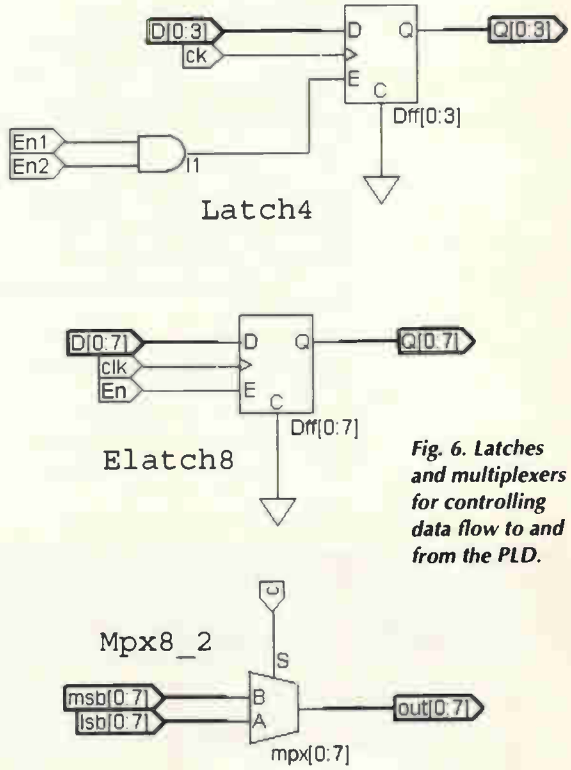


Fig. 6. Latches and multiplexers for controlling data flow to and from the PLD.

Four functions are involved in the process of sending and receiving bytes via the USB module. These are called from the DLL provided by FTDI.

Firstly, FT_Open(0, lngHandle)

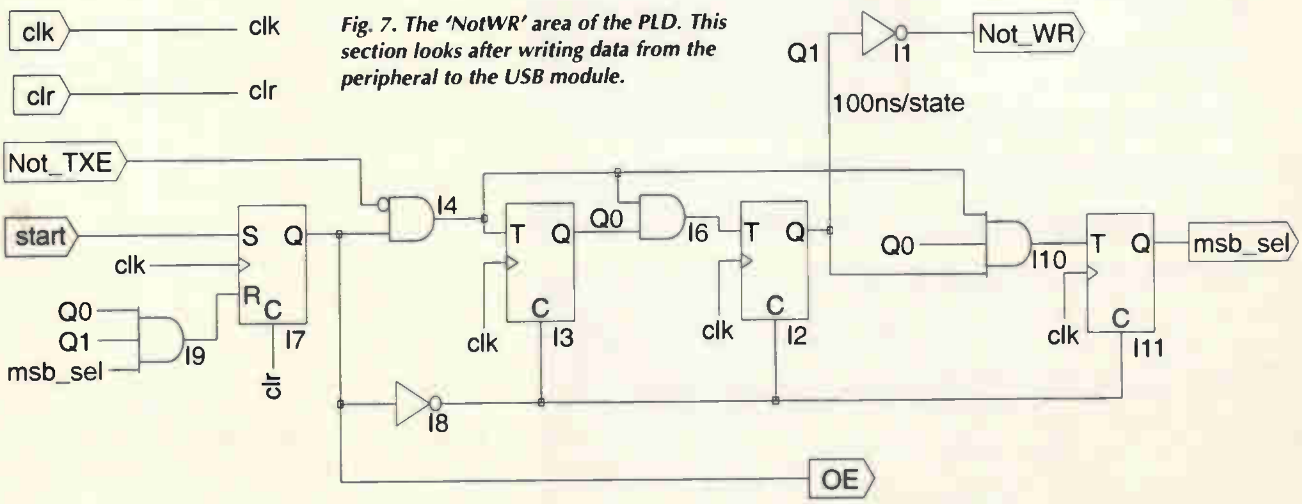


Fig. 7. The 'NotWR' area of the PLD. This section looks after writing data from the peripheral to the USB module.

opens the device, and allocates a handle called 'IngHandle' for use in subsequent communication. The actual write to the device is performed by FT_Write(IngHandle, strWriteBuffer, Len(strWriteBuffer), IngBytesWritten). Here, strWriteBuffer is a string holding the characters representing the bytes to be sent.

The third parameter of the function is the length of the buffer, defining the number of bytes to be sent; the fourth parameter is the number of bytes which have been sent.

The device is closed by FT_Close(IngHandle). All these three functions return the value FT_OK if successful; failure is detected and used to open a warning message box.

Reading from the device uses FT_Read(IngHandle, strReadBuffer, n, IngBytesRead) where 'strReadBuffer' is a string that receives the data that's been read, 'n' is the number of bytes to be read, and 'IngBytesRead' is the number of bytes that have been read.

Bear in mind that when many bytes are being transferred rather than a few, there's something. The USB module sends data at 1ms intervals. Therefore, where there's a lot of data to be transmitted, it's important to send multi-byte strings rather than individual characters for best data transfer rate.

Whether or not you have Visual Basic on your machine, you'll need to install the USB drivers, which are for Windows 98 and Windows 2000 systems.

Extract the zipped driver file – downloaded from the ftdi web site – to a directory of your choosing (and remember where you put it…) When you first connect the interface to the computer, you'll get a "new device found" dialogue box. Browse your way to the directory where you've expanded the driver file; the system will find the necessary files and install the driver.

In summary

Those of you with long memories or good filing systems may find yourselves referring back to the April 1997 edition of *EW*, where I achieved 16 bit input and output from the printer port, using just four CMOS ICs. The design presented here achieves the same result, but demands greater complexity. On the other hand, the output rate is higher.

Yes, maybe it's a bit like taking Concorde down to the corner shop, but this article will have achieved its purpose if it brings the DLP-USB1 module to your attention, and shows how it can be used. You may find yourself using it in designs that exploit its ability to transfer large quantities of data quickly; I certainly have something in mind.

I'm grateful to my employers, Cambridge Consultants, for permission to publish this article. FTDI must take the credit for the good bits of the software associated with this article; I take the blame for the bad bits. ■

Reader offer:

Unique debugging monitor for RS232 comms: save £25

Laplace is offering its Earwigg serial comms debugging monitor to EW readers at £25 discount. Normally priced at £110.00 plus VAT (£129.25) in the UK, the Earwigg is available to you at a discount of 20% – a price of £88.00 plus VAT (£103.40). This represents a saving of over £25. Offer closes 7 June 2002.

This serial communications development and debugging tool provides the most powerful and easiest means to develop and debug any RS232 serial Link. It features two modes, namely standard terminal mode and 'earwigging' mode.

Earwigging mode provides non-intrusive monitoring of traffic on a serial link. The unit copes with 'unprintable' characters and allows display formats including hex, ASCII and decimal to be selected.

Unique 'Earwigging' allows you to monitor and record the traffic on a serial link, to indicate traffic content and direction, without imposing any timing problems or loading on the target link. This feature ensures that the target system operates as normal, avoiding the potential for faults to be masked by timing delays imposed by the test system.

Earwigg's software is exceptionally easy to use. You can control all the comms parameters, display formats and control strings at the click of a mouse. The 'Earwigging' hardware is provide with a 9-way D-type connectors for simple connection into the target link and host PC.

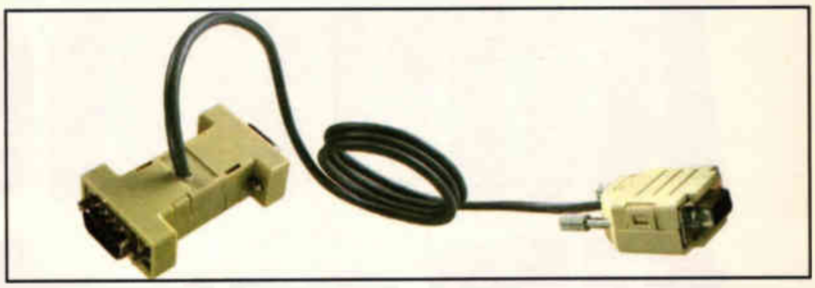
Earwigg is unique in its ability to 'listen-in' to serial traffic without affecting the target system. This makes Earwigg the ideal tool for all those involved in the development/design/test and debugging of serial comms links in industrial, communications, computing, scientific, instrumentation and laboratory environments.

Earwigg...

Intercept and display RS232 traffic
Data direction identified
Non-intrusive technique.
Copes with all standard baud rates.
Fully Windows compatible software.

Benefits:
See the data: If you can see it, you can fix it!

Supplied complete with a hardware adaptor, user manual and PC software. This software run on Windows 95, 98 or 2000.



Earwigg's specifications

Ports supported	Com 1 through to com 8.
Bit rates	150, 300, 600,1200,2400, 4800, 9600, 19200, 38400, 57600, 115200
Parity	None, odd, even, mark, space
Data bits	5, 6, 7, 8
Stop bits	1, 1.5, 2
Display format	ASCII, hex, decimal
End of line character	User choice
Time out delay	User choice
Transmission format:	
Bytes	Decimal, hex, ASCII
ASCII strings	Decimal and hex arrays
Controls	DTR, RTS
Capture	Free running or triggered on match
Display	Data in selected format
RS232 handshake lines	CTS, DSR, DCD, RI, DTR, RTS

Hardware details:

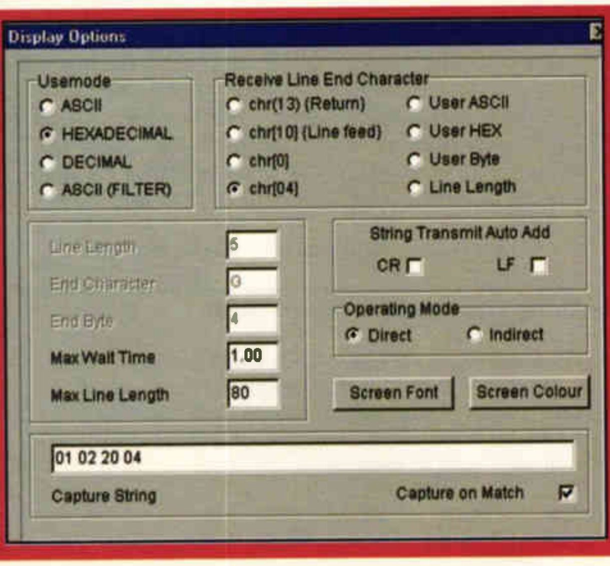
Throughput delay	None
Imposed Load	100kΩ per line
Power	From host PC
Connections:	
Target link	9 way serial plug/socket
To host	1m lead with serial socket
Weight	80g

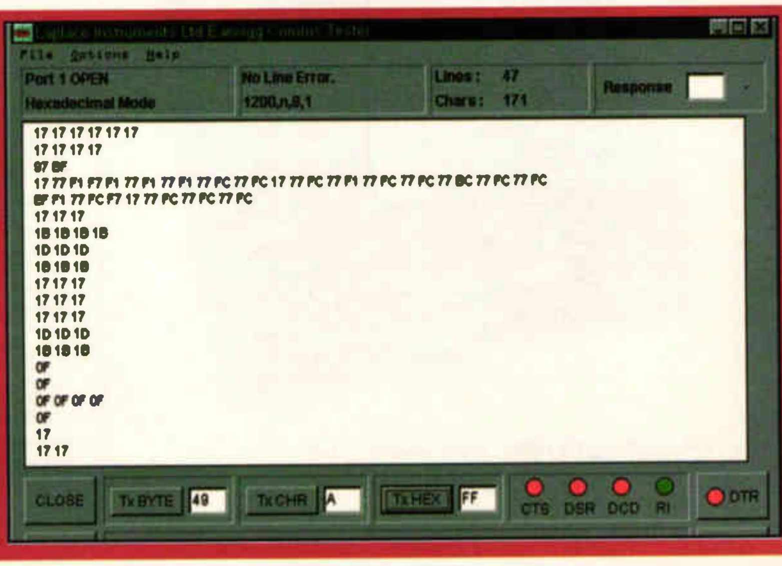
System requirements: 486 or above with Win 95, 98 or NT. At least one free RS232 port, 1.44 MB 3.5in disk drive.

Payment details:

By cheque – payable in Pounds Sterling on a UK bank – for £103.40. By credit card, (VISA or Mastercard). Card details may be sent by post or telephoned to 01692 500 777. Shipment will be made same day as payment received.
Non-UK customers please add £30.00 to cover the additional shipping costs. E-mail, tech@laplace.co.uk, web www.laplaceinstruments.com

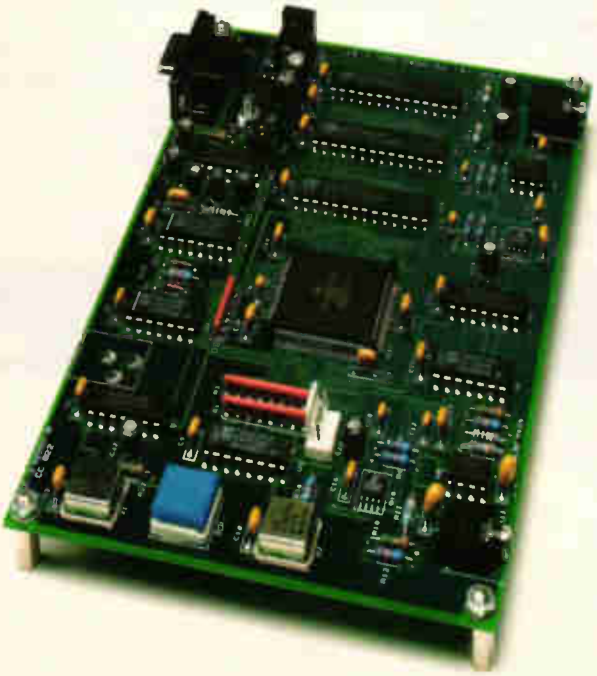
Laplace Instruments Ltd, 3B Middlebrook Way, Cromer, Norfolk, NR29 9JR UK
Telephone +44 (0) 1263 51 51 60 fax to +44 (0) 1263 51 25 32





Advanced filter design

Using the RTDF as an example, Patrick Gaydecki explores the benefits of digital filters over their analogue counterparts. Along the way, he describes how you can use a digital filter to make any old violin sound like a Strad.



The RTDF digital filter – more details on page ???

One of the most interesting points of commonality between technology and the natural world is the number of things whose behaviour can be described by 'linear system mathematics'. A few from a myriad of examples include the behaviour of a loudspeaker, the reverberations in a concert hall, the body resonance of a violin, the action of an analogue filter or the echoes of whale songs in the Antarctic ocean.

Now to qualify for the definition of 'linear', the characteristics of a system must conform to some pretty strict mathematical rules. In general though, they all boil down to a few simple properties, chief of which are stability, causality, superpositionality and uniqueness of output.

Stability is self-evident; causality means that the output must depend only on present or previous inputs. Superpositionality means that if an input consisting of the sum of a number of signals is applied to the system, the output is the sum, or superposition, of the system's responses to each of those signals considered separately. Uniqueness of output means that the system will always produce a unique output to a unique input.

We could go on like this, but you probably have the picture by now.

Just because a system is linear, it does not necessarily follow that its behaviour is simple; indeed, many of the finest brains in physics are still trying to understand why an acoustic guitar sounds the way it does. But, no matter how complex linear system become, they are *all* united by a single equation that encapsulates their response. It is called the convolution integral, and looks like this:

$$y(t) = \int_{-\infty}^{\infty} h(\tau)x(t-\tau)d\tau \qquad (1)$$

where $y(t)$ is the system's output, $x(t)$ is the incoming signal, τ is a time-shift operator and $h(\tau)$ is the 'impulse response' of the filter†. What this means is that the output is obtained by mixing the incoming signal with the impulse response of the system.

What is the impulse response? Very simply, it is the way the system rings, or responds, to a very short excitation pulse – such as a voltage spike in the case of an electronic filter, or a sharp tap in the case of a bell.

Electronics engineers call such transients 'impulse functions',

Dr Patrick Gaydecki is a Senior Lecturer at UMIST

† For the sake of brevity, we are here working with the one-dimensional definition of the convolution integral, although the argument holds true for n-dimensional systems. Also, here the independent variable t represents time, although it could equally be any other parameter, such as distance.

and the important point here is that they should be short time events in comparison to the impulse response.

It is also worth mentioning at this stage that the impulse response can be obtained by calculating the inverse Fourier transform of the system's frequency response.

In practice, the solution to equation (1) can take many forms. It is often done digitally. In this case, it is necessary to replace the integral with a summation, to get an equation of the form,

$$y[n] = \sum_{k=-\infty}^{\infty} h[k]x[n-k] \qquad (2)$$

To use equation (2) in practice, you can either use a finite impulse response, or FIR, solution, of the form,

$$y[n] = \sum_{k=0}^{M} h[k]x[n-k] \qquad (3)$$

Or you can make use of the infinite impulse response, or IIR,

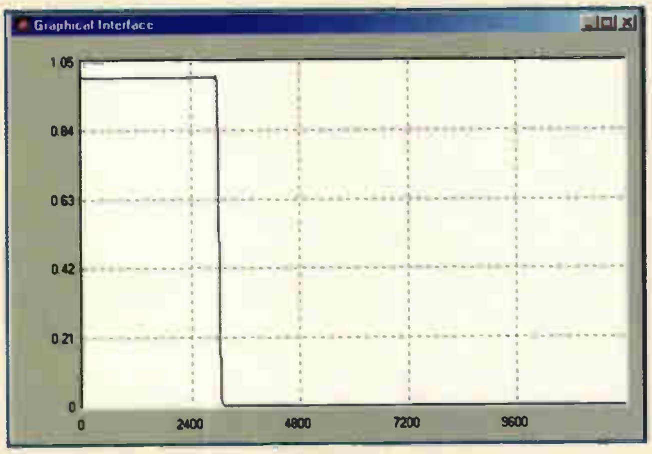


Fig. 1. Brick-wall filter (linear-linear plot).

Fig. 2. Signal masked by broadband noise starting at 3.2kHz.

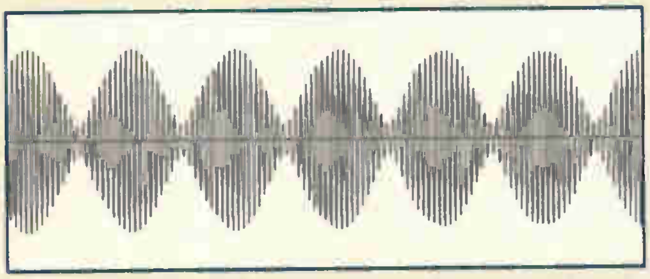


Fig. 3. Signal shown in Fig. 2 recovered after filtering using a brick-wall type, with no phase error.

which is expressed as:

$$y[n] = \sum_{k=0}^{N} a[k]x[n-k] - \sum_{k=1}^{M} b[k]y[n-k] \qquad (4)$$

In this article, I will be using the RTDF real-time digital filter to illustrate the techniques discussed. This system can be described as a digital filter, but such a simple description limits our understanding of what can be done with it.

More appropriately, the RTDF is a universal linear systems emulator, since, by exploiting equation (3), it can be programmed with arbitrary impulse or frequency responses to behave in a particular manner, within the limits set by the calculation speed of its processor.

What is more, it can do these things either in real-time, or it can perform the processing off-line, on signals already stored on the computer, such as in WAV file format. It's important to bear this in mind when reading the examples given below.

Because it uses an FIR approach, the chosen response will always by unconditionally stable and have no phase distortion – something which is significant when designing high-order filters. OK, that's the background; let's see what we can actually do with it.

Brick-wall filtering

To specify a low-pass brick-wall filter using the RTDF is very simple; just enter the cut-off point, select the number of taps and click on update. The number of taps determines how sharp the filter is.

Figure 1 for example, shows a 3kHz low-pass brick-wall filter using 901 taps, and a sampling frequency of 24kHz. To see how effective this is, look at **Fig. 2**. Buried in this seemingly random data is tone-burst signal; in this case, the noise was broadband, starting at 3.2kHz.

Using the brick-wall filter as designed, it is possible to recover the tone burst trace, shown in **Fig. 3**. This comprises a mix of two sinusoids at 2.85kHz and 3kHz.

Although the recovery looks impressive, digital filters do this job very easily. Since the bandwidth of the noise does not encroach on the signal bandwidth, total signal restoration is possible as long as the filter is sharp enough with pure phase linearity.

Standard analogue filter responses

You may want to design a filter with a predictable analogue roll-off response – a Butterworth, for example. Again, a DSP system like the RTDF can handle this without difficulty.

What's more the output will always be stable and phase-error free, regardless of order. This argument applies to any filter type you like – Chebyshev, Bessel etc, but let's stick with the Butterworth for the moment.

The frequency response of this kind of filter is given by,

$$\frac{v_{out}(f)}{v_{in}(f)} = \frac{1}{\sqrt{\left[1 + \left(\frac{f}{f_c}\right)^{2n}\right]}} \qquad (4)$$

where f_c represents the cut-off point. This kind of filter has a roll-off of $6n$ decibels per octave, where n represents the order, or number of poles, of the filter.

Now, say you want to design a low-pass 18-pole Butterworth filter, i.e. with a roll-off of 108dB per octave, and a cut-off frequency of 4kHz. All you need to do is to enter equation (4) in a spreadsheet such as Excel or, if you're adventurous, code it as a

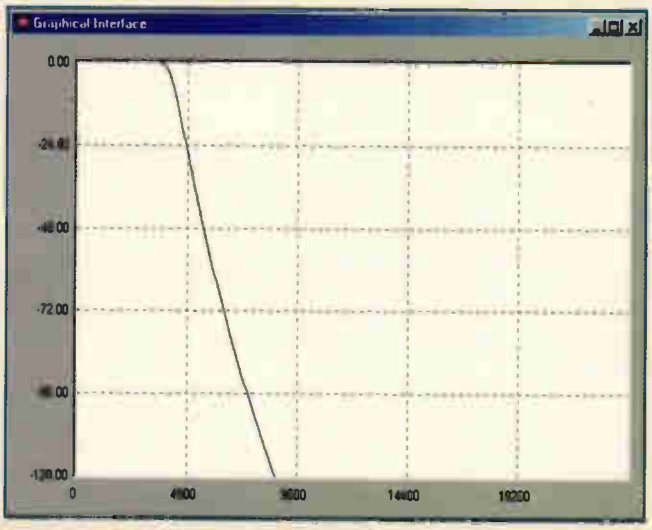


Fig. 4. Frequency response of an 18-pole Butterworth low-pass filter, designed by the RTDF – y-axis in decibels.

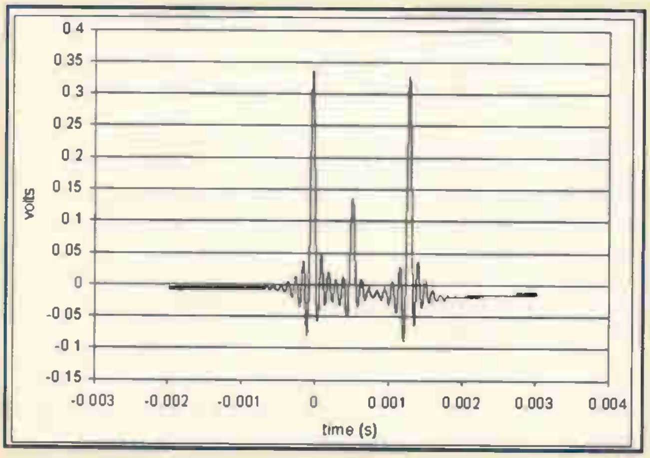


Fig. 5. A test signal sent to the RTDF by a sound card, comprising a group of three impulsive values. Ripples are due to the waveform synthesis software and sound card.

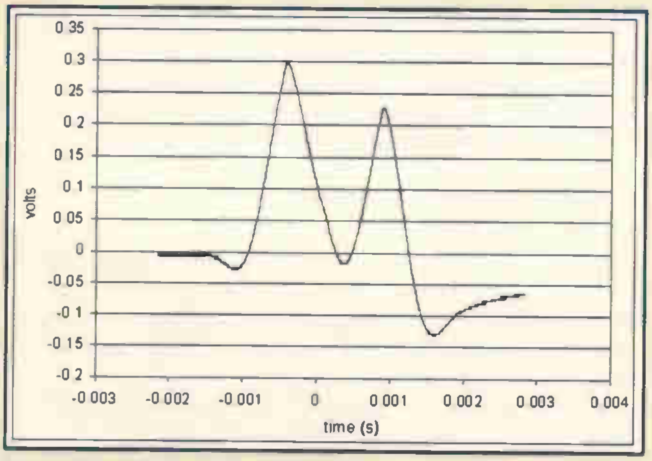


Fig. 6. Signal of Fig. 5 after degradation by a low-pass filter.

program. Now save the output as an ASCII file, which represents the frequency response of the filter.

Next, read in the file to the RTDF using the 'import' facility. The RTDF software will automatically calculate the impulse response, which can then be used for both real-time and off-line processing.

Figure 4 for example, shows the above filter within the RTDF design window. Note that the *y*-axis is specified in decibel, but the *x*-axis is shown as linear, which is why the trace looks slightly flared.

Close examination reveals a –3dB point of 3.984kHz and a –108dB point at 8.016kHz. Incidentally, what is most important here is that the hardware will follow the transition zone roll-off, not the final stop-band value. This is because the RTDF uses 18-bit converters. Any value less than –108dB coming out of the processor is simply too small for the D-to-A converter to see, i.e. it produces 0V.

The RTDF will happily handle much higher orders of filter – 50, 60, 70 poles or more – without complaining. In addition, you can actually design and use 'fractional' order filters, for example, a 7.3-order system; there is no law stating that *n* in equation (4) has to be an integer value. Digital systems can handle this because the impulse response calculation, and the filter implementation, is strictly a matter of software.

Moreover, using the above methodology, you can design any kind of filter you like – for passive as well as active circuits. A tuned *LCR* filter has a frequency response given by,

$$\frac{v_{out}(f)}{v_{in}(f)} = \frac{j\omega RC}{j\omega RC - \omega^2 LC + 1} \qquad (5)$$

In this case, all you need to do is import the magnitude of the response, ignoring the phase. The RTDF will take care of everything else.

Signal shape recovery

A digital filter like the RTDF has the ability to import a frequency response, invert it, take its inverse Fourier transform and use the resulting (rather strange) impulse response as an 'inverse filter'. This may sound a rather esoteric function, but in fact it has real practical use.

For instance, imagine a signal that started life as a nice crisp waveform, but during its travels, due to parasitic capacitances and inductances, it arrived at its destination in a somewhat smeared form.

Figure 5 for example, shows three pulses generated by a sound card (the ripples are a function of the impulse generation software and the frequency response of the sound card). **Figure 6** shows what happens when the pulses pass through a system with a low frequency cut-off point of around 1kHz.

The signal has been degraded so badly that it is impossible to distinguish the middle pulse, and the two remaining ones are spread out in time.

Figure 7 shows the restored signal; not only has the temporal relationship between the three values been precisely restored, but so too in large measure have their amplitudes. This was achieved by feeding the degraded signal into the RTDF and processing it with a real-time filter that represented the inverse of the frequency response of the degrading system.

In practice, inverse-filter restoration is a pretty fiddly business, and it often requires tweaking of offsets to achieve the desired result. There are entire scientific journals devoted to inverse problems.

The RTDF gives complete control over all the adjustment parameters, so a user can tailor a solution to his or her needs.

From shoe box to Stradivarius

From the above discussion you have probably gathered that you can import any arbitrary frequency response you like into the RTDF, and it will reproduce it faithfully. You could even sketch out a signal by hand, convert it to ASCII and the software does the rest.

So what use is this? Plenty. Think for a moment about the way musical instruments work – take a violin for example. Whatever you may have heard, a violin is a linear system, whether it's the most expensive Strad or the cheapest school squawker.

The only non-linearity comes from the player, and how he or she moves the bow across the strings. When the bow is drawn across a string, it sticks to the hairs on the bow and is displaced. If the bow is moved at a constant velocity than the displacement is uniform with time.

At a certain point, the displacement force is overcome by the restoring force and the string flies back to its original position. Then the whole thing starts up again: this is known as the slip-stick regime. Thus the force acting on the bridge is actually a simple saw tooth – something you can prove by placing a transducer on the bridge and feeding its output to an oscilloscope.

Now a saw tooth sounds horrid and nasally, but that is not what you hear. The saw tooth – in other words the input signal – is mixed (scientists use the word convolved) with the impulse response of the bridge/body system, giving rise to the characteristic violin sound.

So if you can obtain the frequency or impulse response of the bridge/body, then you can dispense with the violin. Just keep a frame with strings and a bridge to mount the transducers, and program the RTDF as a virtual Strad. This sounds almost too good to be true, but it actually works.

Figure 8 for example, shows the frequency response of an acoustic violin, obtained by tapping the bridge with a delicate instrumented hammer (these weigh a few grams), and recording how the violin sang in response.

The frequency response is then converted by the RTDF into an impulse response, and can then be used to filter a saw tooth recorded from the bridge. If you visit the RTDF web site (www.umist.ac.uk/dias/pag/rtdf.htm), you can listen to part of Bruch's Violin Concerto #1 in G minor, recorded from the bridge, and then to the same sound having been processed by the RTDF, programmed with an acoustic violin frequency response. The effect is pretty convincing.

In truth, to get a perfect reproduction you also need to take into account near and far acoustic sound fields. As it does with most instruments, the frequency response of violins varies with position. Nevertheless, linear system mathematics still apply and in any case, most people listen to a recording of a violin, taken at a specific point.

Real-time and off-line processing; filter transpositions

As I mentioned earlier, the RTDF generates filter coefficients that can be used to process data both in real-time and off-line.

Off-line processing has its uses. For example, your data may already have been recorded and you need to filter it. In this case the RTDF is applied to design a filter, specifying a sample rate equal to the rate at which your data were sampled. Then export the filter coefficients as an ASCII file, and use them in an appropriate convolution algorithm (i.e., Equation (3)).

The RTDF design interface has eleven preset sample rates (since the hardware must be set to one of these), ranging from 48kHz down to 1.5kHz. If you want to design a filter outside of this range, you can of course do so – there are no restrictions – and export the coefficients as described. However, you can only download filters to the hardware that have been designed using one of the eleven presets.

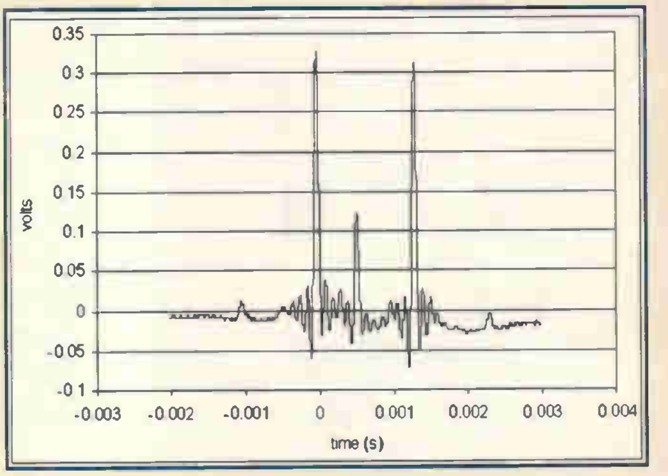


Fig. 7. Result of applying an inverse filter to the signal shown in Fig. 6. Although some noise is now present, precise temporal information of peaks has been restored.

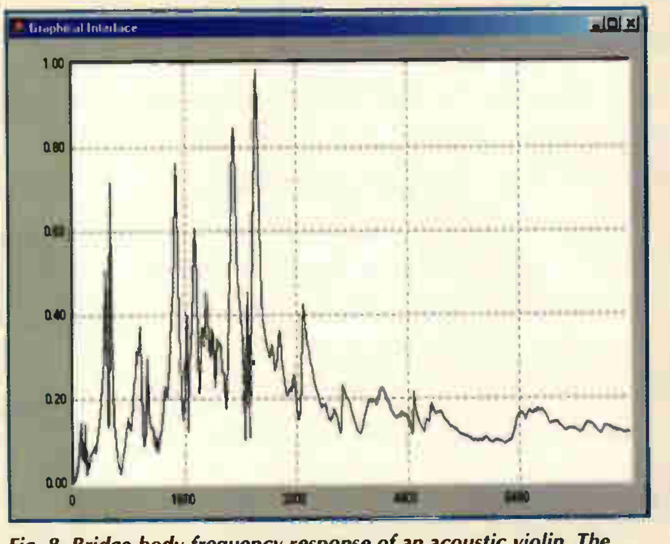


Fig. 8. Bridge-body frequency response of an acoustic violin. The RTDF will happily reproduce this if programmed appropriately.

Talking of presets, it is quite possible to change the sample rate of the hardware during run time with a single mouse click. This means that if a filter is running, it is instantly transposed to a different bandwidth.

The future

A host of new features are being planned for the RTDF, including musical effects such as all-pass filters, standard analogue emulations, echo, reverberation, virtual surround sound and so on. All of these software modules will link to the existing design interface. In the meantime, there is still a world to explore with the current system.

Finally, my special thanks go to Jim Woodhouse from the Engineering Department at Cambridge University for supplying the violin frequency response discussed in this article. ■

Andrew Malcolm looks at the evolution of programmable logic devices.

Programmable logic – an introduction

Programmability goes to the heart of electronics and computing. Computers, microprocessors and microcontrollers are all programmable, allowing the user to specify the action of the machine in terms of a program.

Computers such as these were at one time built from small logic units in the form of 'and' gates, 'or' gates and flip-flops. These units themselves were units built from vacuum tubes or transistors, and more recently integrated circuits.

Most of you will be familiar with the 74 series of TTL logic devices. Their life has stretched the last 25 years or so. Since then, we've seen all manner of high levels of integration, from the very first four-bit microprocessors, such as Intel's 4004 to today's multi-million gate devices, such as Intel's Pentium 4 and Motorola's PowerPC.

My career in electronics has encompassed most of that 25 years. In that time, I have worked with many microprocessor and logic families. At the start of that

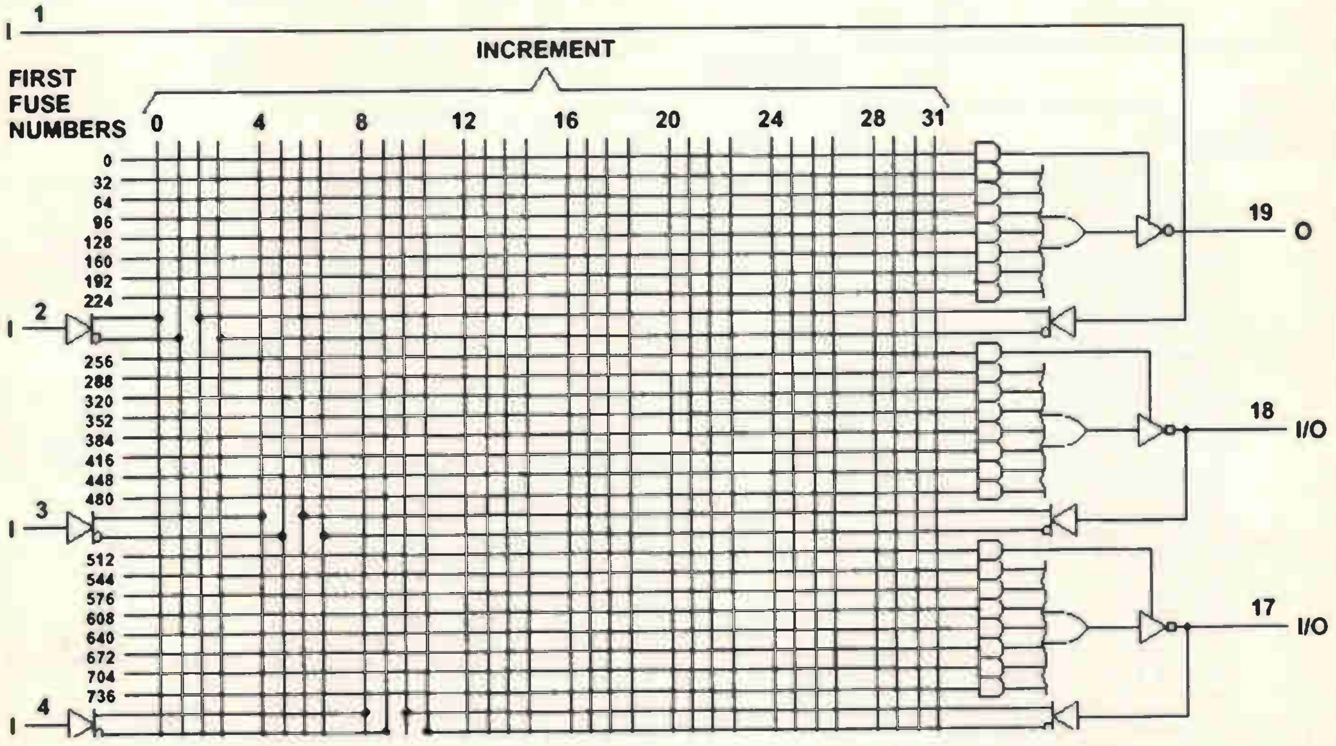


Fig. 1. A combinatorial programmable array-logic device, or PAL.

period, supporting logic for microprocessors, bit-slice processors and memory systems was typically created using the available TTL devices. Gates, flip-flops and counters from these families were combined to produce the required function, such as address decoders, video RAM controllers or bus buffers.

The original design for the PC by IBM, for instance, had a printer port constructed from two TTL octal latches and a handful of other TTL gates. This approach still has a place in small designs and 'glue' logic today. Many examples can be found in the pages of the technical press. Such an approach is somewhat disparagingly known as 'random logic'.

Many functions implemented in this way do of course have a great deal of structure to them. This fact led many to realise quickly that programmable devices would lead to more efficient and flexible solutions.

Programmable, read-only memories

The first programmable devices I encountered were PROMs, or programmable read only memories. These were the forerunners of today's EPROMS, EEPROMS and FLASH memories.

Implemented in bipolar technology, PROMs were as fast as the corresponding bipolar TTL parts, and were compatible with those logic families. Typical device densities were 64, 128 and 256 bytes, packaged in 14 to 20 pin DIL packages.

Some of these are still available today. Used as look-up tables, these devices could implement any Boolean function of '*n*' inputs and '*m*' outputs, were '*n*' is the number of address lines and '*m*' the number of data bits.

The snag was that no development tools were available then. As a result, many engineers resorted to writing out the PROM contents long-hand from their Boolean equations and the typing them into the PROM programmer.

Of course, the devices were one-time programmable, so any mistake meant a useless device. As technology progressed, the devices became larger and faster.

Some gained 'registered' outputs, meaning that the outputs of the ROM array first passed through a register – i.e. a bank of flip-flops. This innovation allowed the engineer to implement Mealy or Moore state machines.

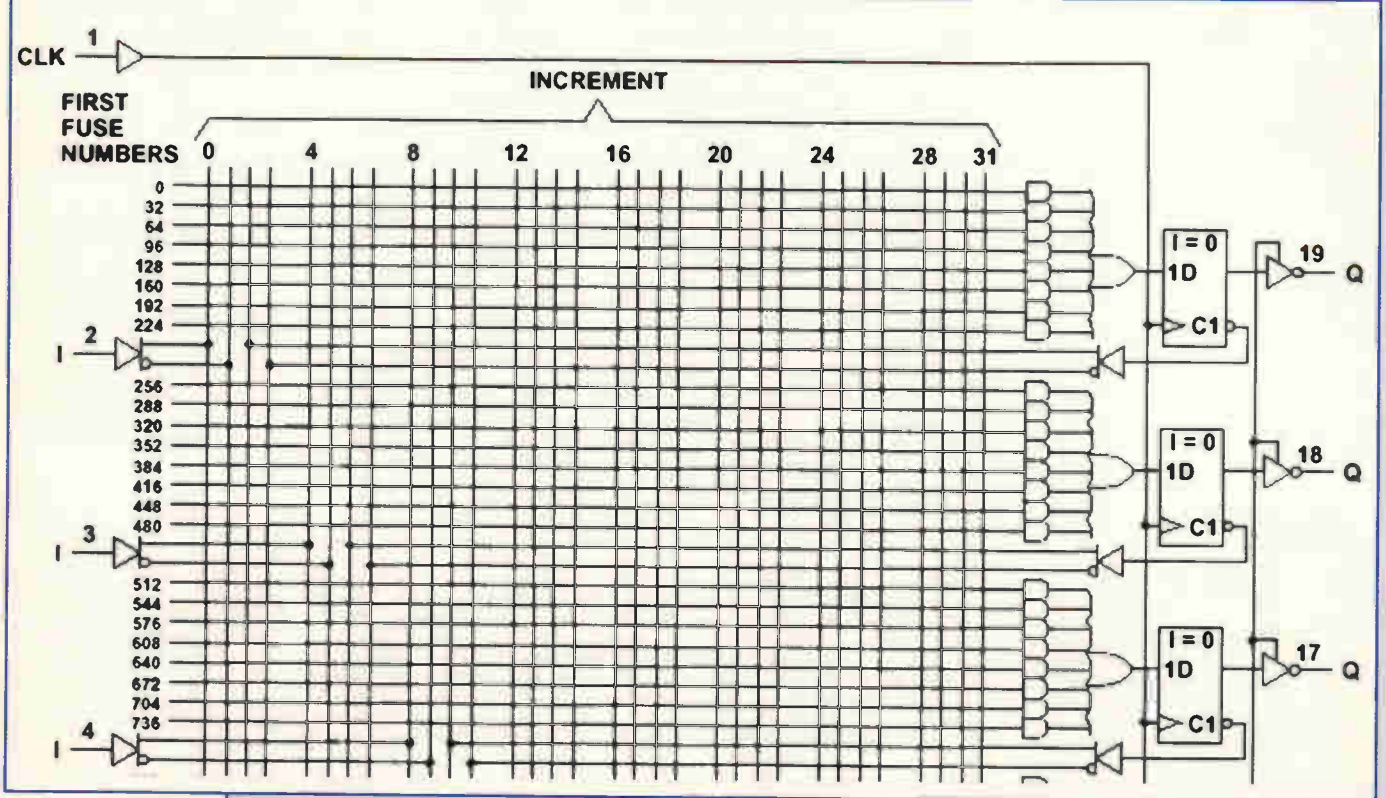
Programmable array logic

Around this time a company called MMI introduced the PAL, or programmable-array logic. These devices allowed the implementation of an 'and-or' array, sometimes with registered outputs.

Inputs to the device passed into a fuse array, the outputs of which went to several 'and-or' clusters. **Figure 1** shows an extract from a data sheet for such a PAL while **Fig.** 2 shows a registered version. Often, an engineer would photocopy this page from the data sheet, and mark on it the fuses he required to remain intact.

Programming the device involved removing the unwanted fuses, leaving the required Boolean function intact. Device programmers initially displayed this fuse map on a CRT, and the user could edit the required fuses and 'burn' the device. Again, the first

Fig. 2 A programmable array-logic device featuring registers.



devices were one-time only.

These devices were still difficult to use. However, the added flexibility that they offered, coupled with the more intuitive and-or structure, ensured that they quickly became very popular. A whole family of devices evolved, ever increasing in speed, date density and feature sets.

The PAL concept survives today in various manufacturers' 'GAL', or generic-array logic, families. These families offer many features beyond those of the original PAL devices, including re-usability. Some can be erased using UV light, while others are electrically erasable. More recently, devices with in-system programmability have started to appear.

Perhaps the most popular and widely used architecture of this type is the 22V10 – a device which added asynchronous set and reset functions and many other features to the basic PAL architecture, **Fig. 3**.

The 22V10 introduced another important innovation. This innovation – the macrocell – has survived and been expanded in today's more sophisticated devices. The macrocell replaced the simple flip-flop that registered some or all of the outputs of a PAL device, and was itself programmable.

Fuses in the 22V10 allowed greater flexibility in the selection of signals destined for their associated output pins. The signal could be the direct output from the and-or array, a registered version of this, or an inverted version of the first two choices. Additionally, it was possible to feed the combinatorial or registered version of the input back into the fuse map.

Figure 4 shows the macrocell's structure.

Complex PLDs

It soon became apparent that the limitation of these devices was the restricted number of I/O pins. The largest package used by GAL devices in through-hole form had 24 pins and a 0.3in wide body.

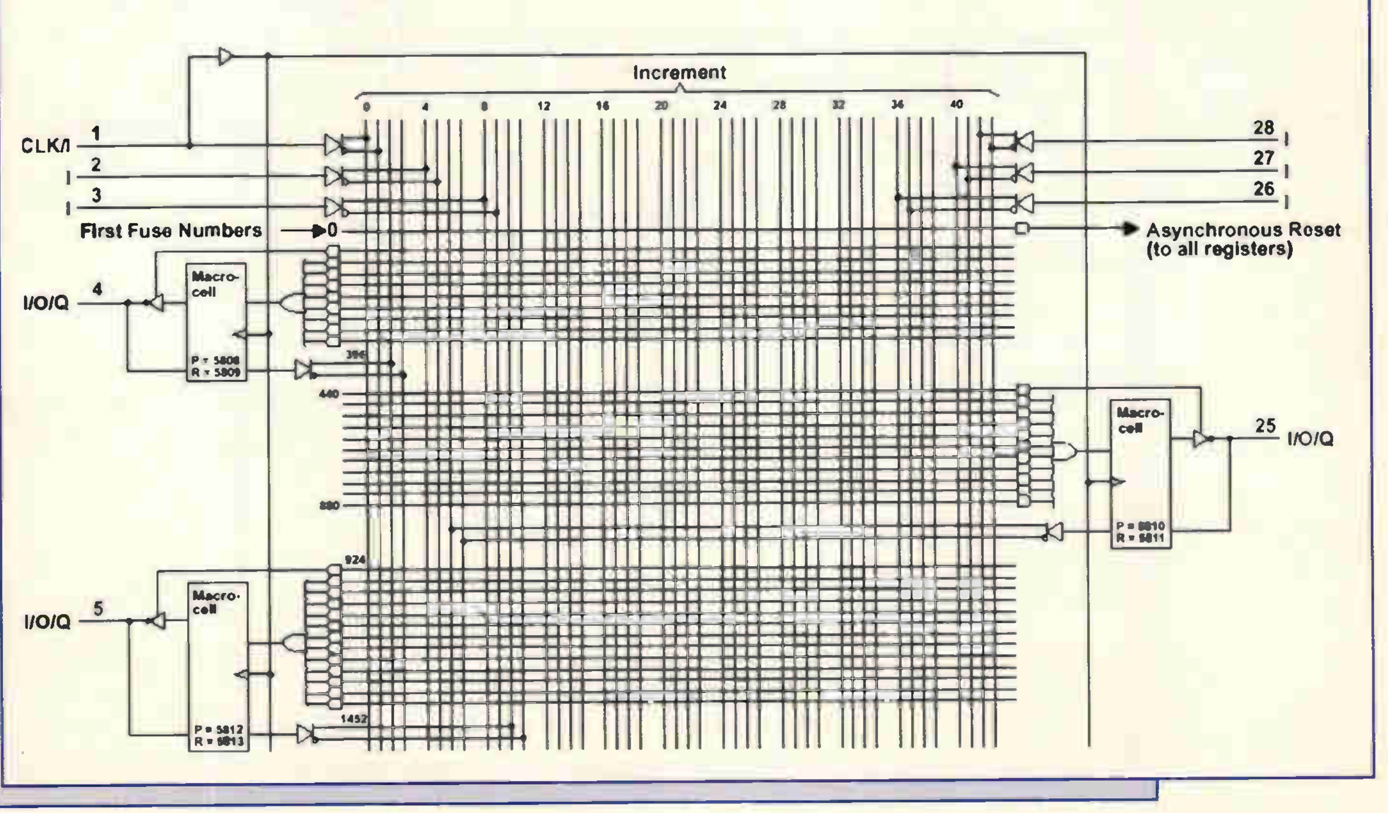
Some designs required more than one device and communications between a group of devices. An example is the state register in a state machine. In such designs, these signals soon began to eat into the inputs and outputs required for the core application function.

At this point, two things happened. Firstly, the advent of surface-mount technology, or SMT. This made packages with pin counts of 28, 44, 68 and 84 on PLCC (plastic, leaded chip carrier) possible without consuming large amounts of PCB area.

Secondly, device designers divorced the macrocell from the I/O pin to produce the 'buried' macrocell. This made possible the introduction of devices capable of supporting much more complex designs.

Set against the backdrop of ever increasing gate densities in the memory and processor products, the resulting devices became highly functional, and the complex programmable logic device, or CPLD, was born. These devices typically have hundreds, or perhaps thousands, of programmable gates and registers, and operate at clock rates of several hundred megahertz. Vendors include Lattice, Texas Instruments and National.

Fig. 3. Perhaps the most popular and widely used programmable array-logic architecture – the 22V10.



Field-programmable gate arrays

The field-programmable gate array shares many characteristics with the CPLD, but its history is somewhat different.

In parallel with the developments described above, many manufacturers offered alternatives to full application-specific integrated circuits, or ASICs. This was achieved by applying programmability at the photolithography stage of chip manufacture.

Devices were designed with a 'sea of gates' – vast arrays of uncommitted gates and flip-flops. These devices were manufactured up to the final metallisation stage, the gates and flip-flops remaining unconnected.

A specific customer's design was implemented by producing custom lithography masks for the final metallisation layers, wiring up the device as required. This 'gate array' technology was very popular with high-end customers, but clearly required a great deal of investment both in time and technology. It was unsuitable for small production runs.

Designs had to be carefully validated in simulations before committing to silicon, as the consequence of failure could be expensive – maybe even resulting in a product's market window being missed.

To solve this problem, designers took a leaf out of the CPLD device's book, and replaced the photolithography with electronic programmability.

The resulting FPGA, or field-programmable gate array, tends to be RAM-based, that is its programmability is volatile. As a result, the required function has to be reloaded into the device at power up.

Because of their history, these devices use state-of-the-art semiconductor processes to produce truly awesome gate densities. Leading device vendors include Altera and Xilinx.

A typical Xilinx device from the company's Virtex-II family, has six million gates, four megabits of user RAM, as well as embedded high-level functions such as 18-by-18 multipliers and configurable I/O pins.

Ball-grid array packaging takes device pin counts up beyond one thousand. Again, they operate at clock rates of several hundred megahertz. It is perfectly possible to program these devices as complete 64 bit microprocessors.

What's next?

In the next article in this set, I will be looking more closely at today's CPLD and FPGA architectures, as well as at practical applications and development tools. I will review some of the available software and hardware development kits, and introduce the industry standard language for describing FPGA functionality, VHDL. ■

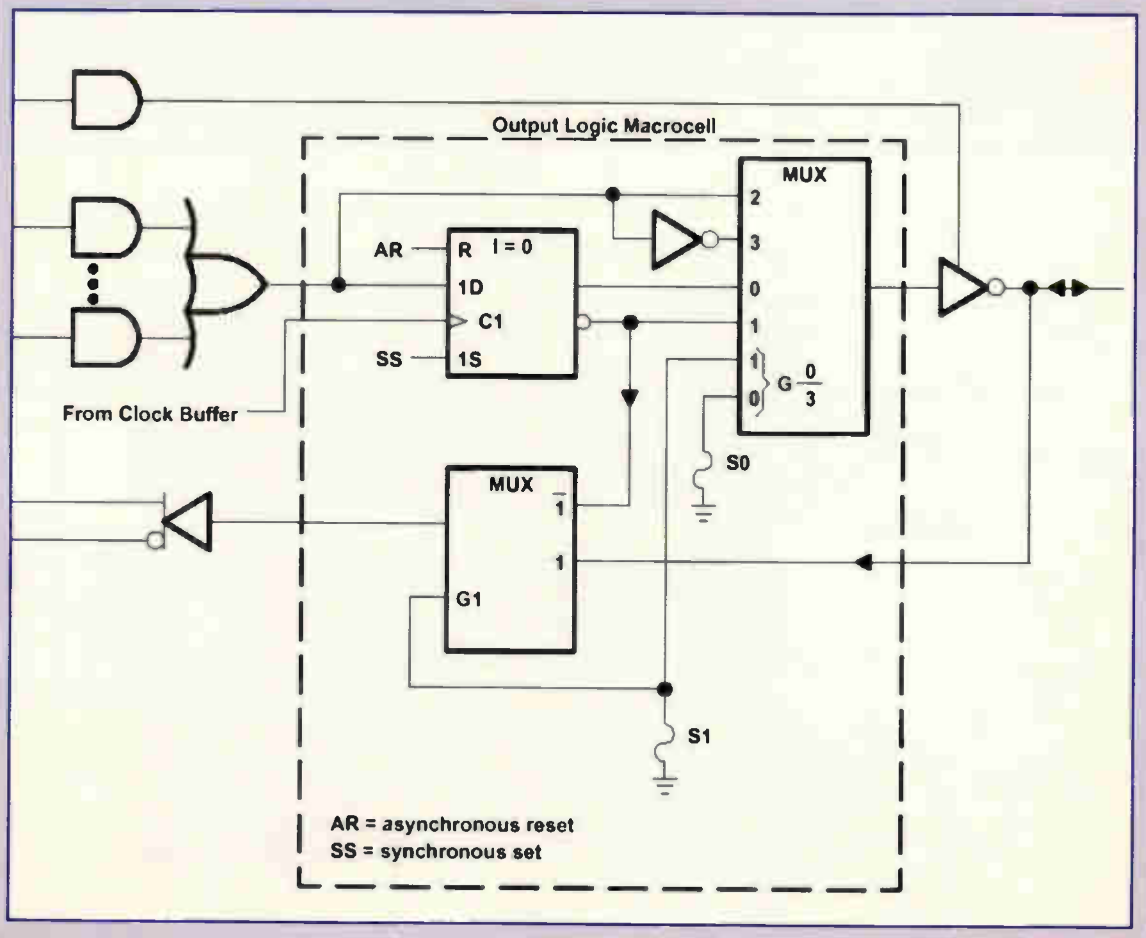


Fig. 4. An important innovation in the 22V10 was the 'macrocell'. This cell replaced the simple flip-flop that registered some or all of the outputs of a PAL device.

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POWER AMPLIFIERS: FETs less linear than BJTs; Distortion in power amplifiers 1-8; Distortion residuals; Trimodal part 1, 2; Load-invariant power amp
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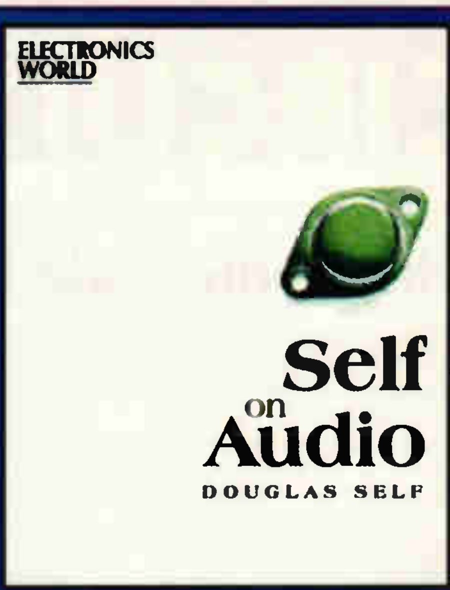

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Electronic code lock

This is a simple but efficient code lock. First the user enters the correct code of eight digits. When the 'Enter' button is pressed, the relay activates and the lock will open. If an incorreect code is entered, the circuit will reset.

At the heart of the circuit is 4017 decade counter. At power up, the counter will be at zero count and only pin 3 of the counter will be high.

If switch S_1 is pressed momenterally it will provide a clock pulse for the counter which will increment the count. After receving the first pulse now the counter is now at 1, which means this time only pin 2 will high.

Only by pressing S_2 can the next clock pulse be generated. Hence by pressing S_1 to S_8 sequentially, the high level on the pins shifts from pins 3, to pins 2, 4, 7, 10, 1, 5, 6 and finally pin 9. While pin 9 is high, pressing the 'Enter' button will activate the relay, which will in turn open the lock. Any mis-entry will reset the sequence.

The circuit can be intentionally reset by pressing the 'Reset' button, S_{10}. One or more then one parallel connection switches of S10 can be placed in between the switches S_1 to S_8 to increase the complexity of the code.

Sandeep Gaurkar
Karnatak Regional Engineering College
Karnatak
India
(F95)

Components breakdown

Resistors

R_1	1k
R_2	100k
R3	68k
R4	68k
R5	2.2k

Capacitors

C_1	2.2μF
C_2	1μF
C_3	2.2μF

Transistor

T_1	SL100 (n-p-n))

Diodes

$D_{1\text{-}8}$	1N4148
D_9	1N4001

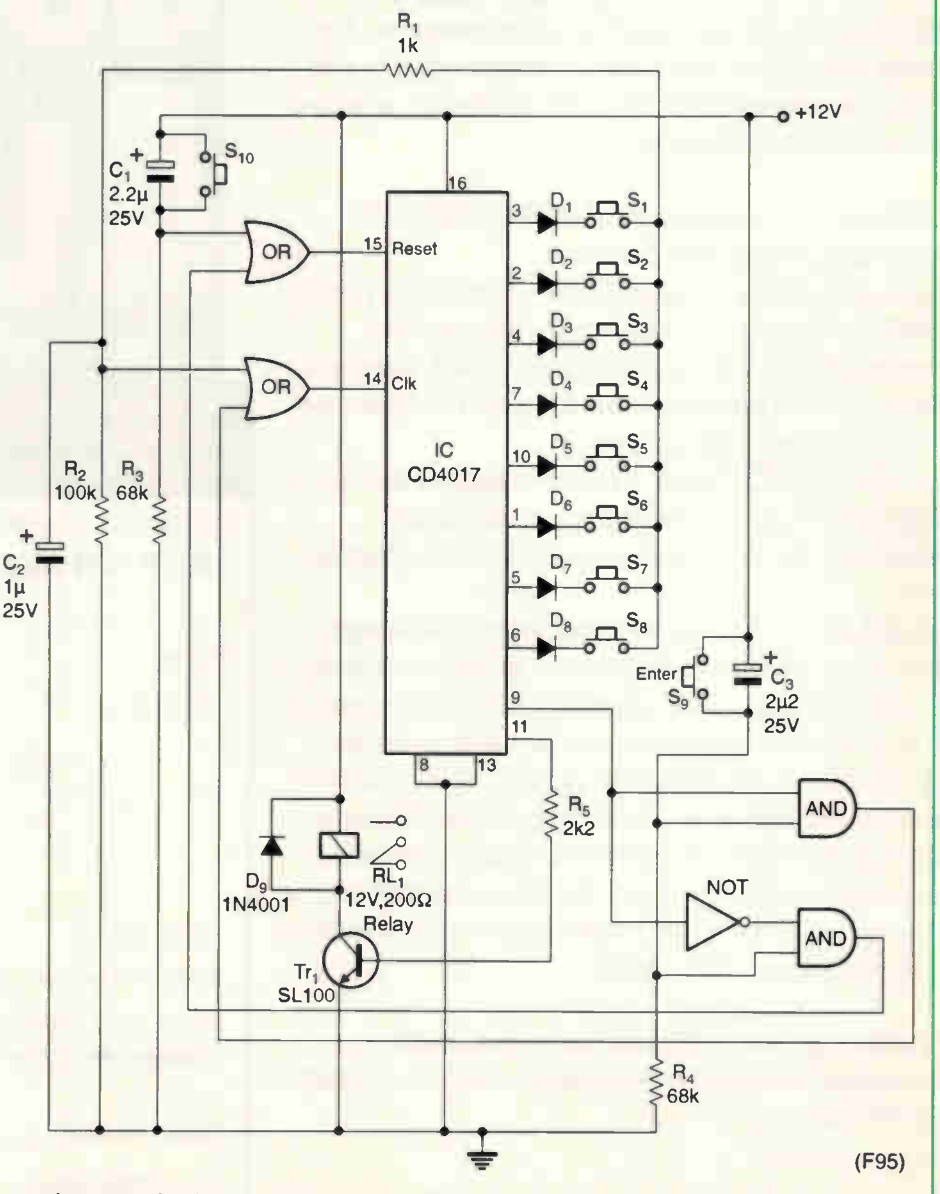
Misc

Relay 12V, 200Ω

S1 to S10 push buttons

ICs

IC_1	CD4017
IC_2	CD4071B
IC_3	CD4081B
IC_4	CD4069UB



To alter the code of this electronic combination lock, you simply rewire the switches. There is no unique code as such. Rather, this lock relies on each button being pressed sequenctially, starting with S_1.

Solar-powered battery charger works on cloudy days

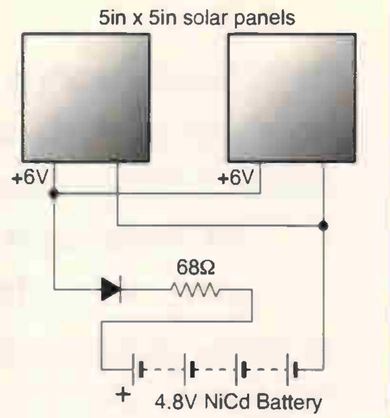
Since the late righties, Maplin Electronics has been making a range of solar-powered battery chargers. I have been an owner of several of these devices.

The first was a little bigger than a cassette box, with an AA NiCd battery holder inside. The later charger was about the size of a human hand. Batteries could be put in a holder in the back or attached externally. Both of these circuits worked well in sunlight, but sunlight isn't something we have very often in the UK.

What actually happens to a solar panel in dim light is that the output voltage doesn't fall very much, but the internal resistance increases. A 100mA, 4. 5V panel has an internal resistance of 45Ω. On an overcast day this internal resistance rises to, typically, 900Ω, reducing the output to 5mA, or, on an even duller day, to 4.5kΩ, reducing the output to 1mA. This is almost totally useless for charging anything.

A bigger solar panel has a lower internal resistance. By using two larger 6V 150mA panels, approximately the size of CD boxes, connected in parallel, currents of four or five times as much may be generated under cloudy conditions.

A typical overcast day can now produce 20mA, and even a relatively dull cloudy day produces 5mA – enough to be of some use. Placing the apparatus on the windowsill to get plenty of light, a set of four NiCd AA batteries can be charged over a few days in the typical British weather.



A solar charger designed for British weather!

This works for eight months of the year. In winter, daylight is not bright enough for the charger. Your eyes may notice little difference, but winter daylight is actually much dimmer.

This design creates another problem. What happens when the sun comes out? Two 150mA panels will produce 300mA of current, enough to overcharge the batteries as they are only meant to be continually charged at 50mA. A resistor is used to limit the current to 50mA.

Panels that are nominally 6V typically produce 9V and the NiCd batteries are usually around 5.6V when fully charged. This makes the potential difference between batteries and panels 3.4V, and the resistance should be 68Ω.

The power dissipation of the resistor is only a sixth of a watt. It doesn't need to be high a power type. The circuit is incredibly wasteful in sunlight, producing only 40-50mA from quite large panels, but it wasn't really meant for sunlight.

A reverse-blocking diode was added to prevent current flowing back from the batteries into the solar panels during the night, when the battery voltage is higher than the voltage of the solar panel. With silicon solar panels there is actually only a very small current of a few microamps under these conditions, but I thought that this might be harmful to the solar cells.

Malcolm Lisle
Gateshead
Tyne and Wear

Tuner scanning add-on

£50 winner

With so many radio receivers available on a chip these days, it takes little time to build a receiver for any application. But when faced with the ability to provide a scanning facility for an airband monitor it looked liked using a micro and PLL with display.

This add-on can be used with any IC radio chip with built-in local oscillator and employed a small piece of Vero about 60 by 40mm with miniature LEDs mounted above each preset pot in two rows of four, spaced behind the front panel with 6mm spacers at the corners.

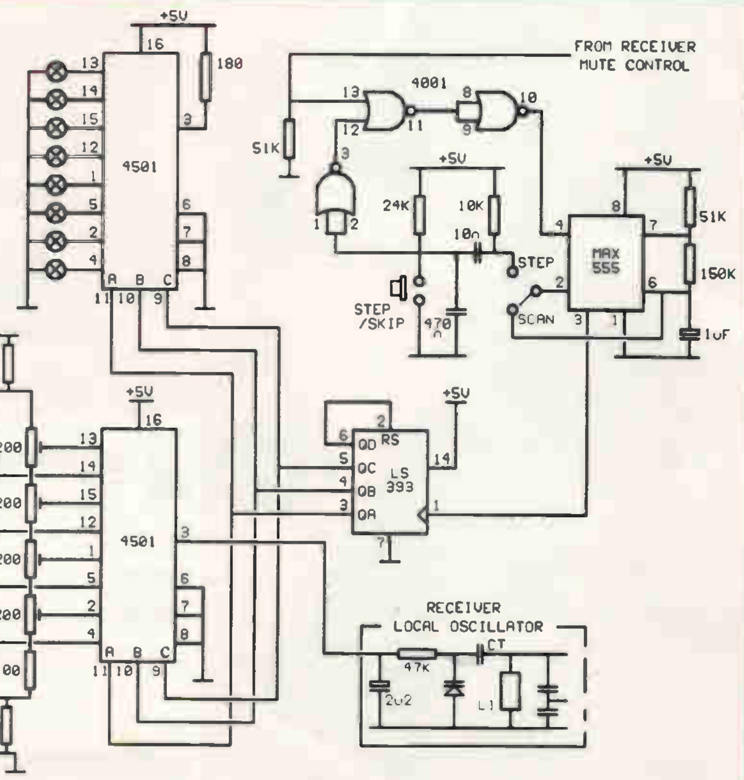
The dual linear 4.7kΩ potentiometer was mounted to the right hand side of the LED assembly to provide fine tuning by an equal amount irrespective of the preset selected and is normally left in mid position when setting up.

When using the scan mode the push button may be used to skip the active channel during which the 'mute' signal inhibits the scan.

Placing the switch in 'step' mode allows single stepping and setting up. The bandwidth of interest is set by capacitor C_T and restored to mid-band by re-adjustment of L_1

Ron Wheatley
Birmingham
(G11)

This digital add-on for scanning a number of channels work with a variety of IC radio chips.



Dynamic diode checker and matcher

I developed these two circuits some years ago to check for balance of characteristics of diodes, dynamically. The idea is based upon the fact that if two diodes, connected m a bridge circuit, are identical and fed via a repetitive ramp of voltage through identical value resistors, forming the bridge, then the instantaneous value of voltage across each diode will be identical.

The four resistors used for the differential amplifier and also the top resistive arms of the bridge, should be selected from batches of resistors, to match to 0.1% or better. The voltage drops across the diodes are compared using a true differential amplifier, as shown.

If the diodes are identical in performance, there is theoretically zero output from the differential amplifier. If the characteristics are different, then there is an output, as the bridge is then unbalanced. An oscilloscope will show this difference quite easily, and also in which direction the difference lies.

It is also possible to detect the difference using a meter, but the oscilloscope shows more information. If a double-trace scope is available, then the output from the comparator, pin 7 of IC_1, should be used to lock the oscilloscope time-base. The other channel will be used to monitor the output from the differential amplifier or the following amplifier.

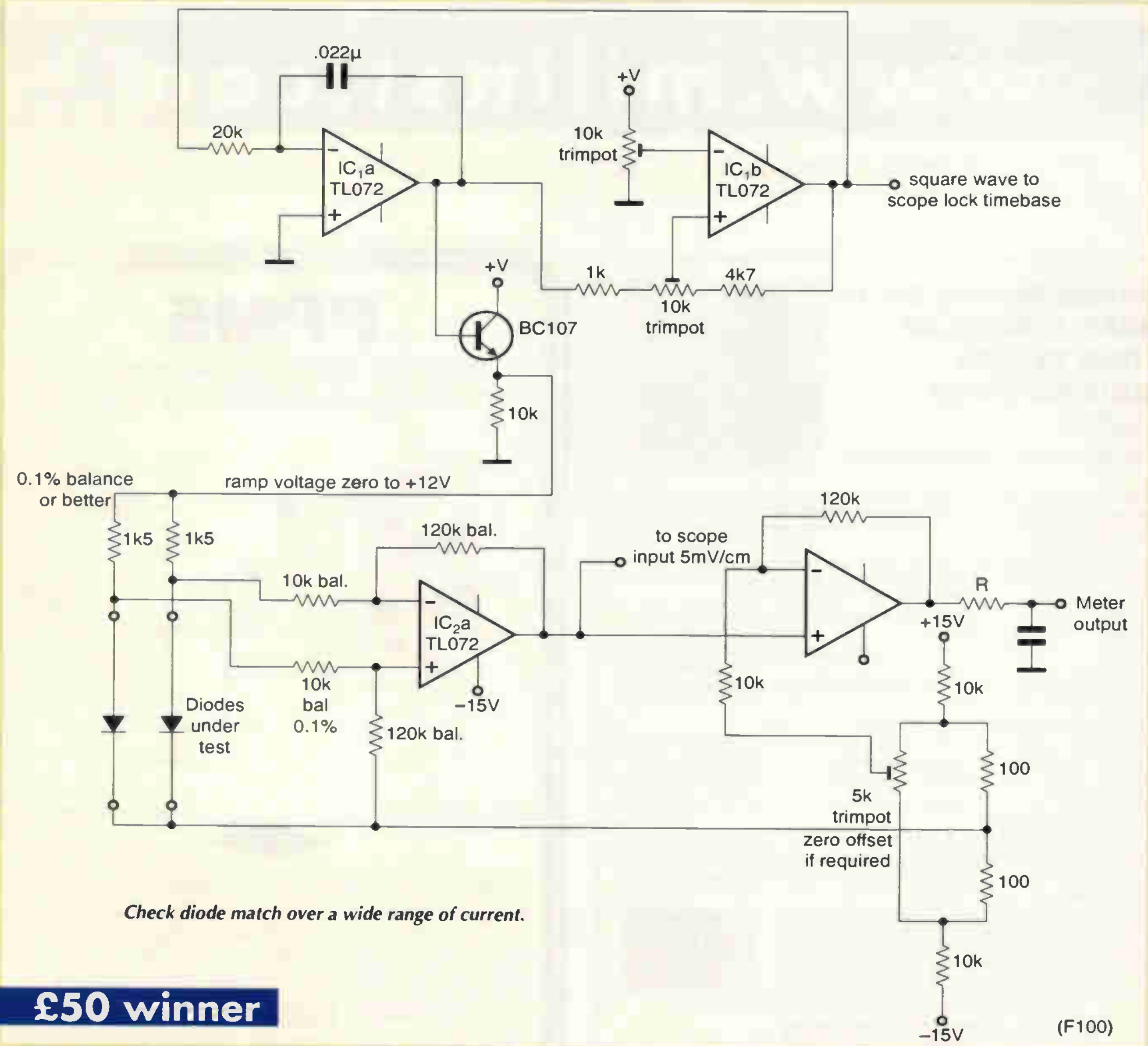
The system is so sensitive that the changes caused by temperature may easily be seen when one of the diodes is touched with a finger, the difference being seen quite dramatically at the output of the differential amplifier.

There are two parts to the circuit; the zero voltage-referenced ramp generator of 12 volts peak, and the bridge circuit. With component values as in the diagram, the frequency of the ramp is of the order of 1200Hz. The ramp is quite linear, as the input resistor feeding the integrator sees a virtual earth.

Balanced diodes are required for various types of circuit, e.g. balanced mixers for radio detector/modulator circuits. The circuit may also be used to match resistors and capacitors, over limited ranges.

The idea may be modified with extra circuitry to allow the selection of dynamically balanced transistors and FETs. It is important in this case that the ramp is adjusted to run from zero, at the emitter of the BC107 ramp output buffer transistor, using the two trim-pots iteratively.

Ivan J Eamus
Aylesbury
Buckinghamshire
F100



Check diode match over a wide range of current.

£50 winner

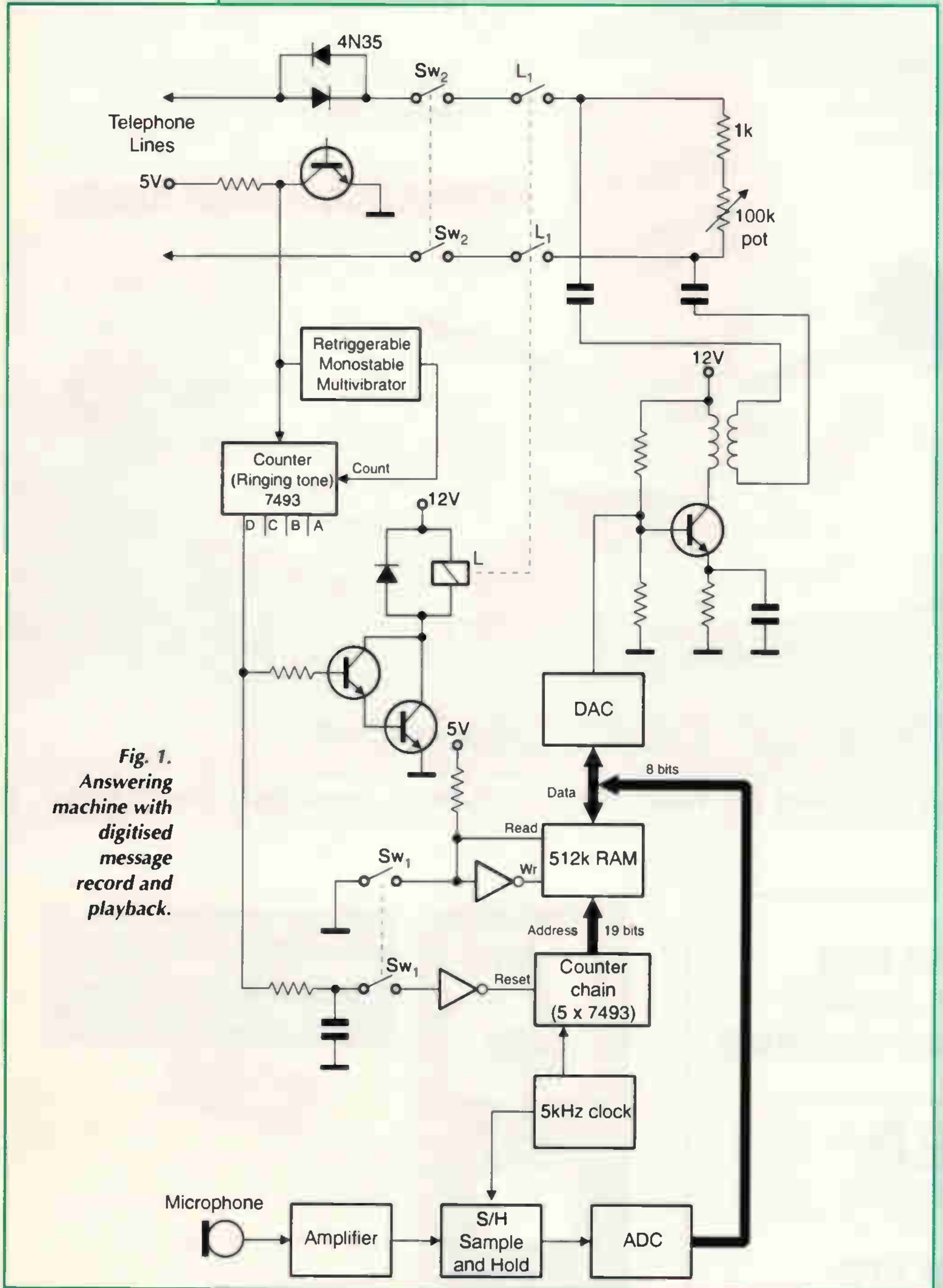


Fig. 1. Answering machine with digitised message record and playback.

Answer-back unit for home telephone

In this addition to the telephone set, a recorded voice message is returned to the caller as soon as he or she completes dialling and ringing starts to originate from the exchange.

Referring to **Fig. 1**, an opto-coupler connected to one arm of the incoming telephone line cable produces impulses when ringing tone arrives from the exchange. When eight pulses are counted, the most significant bit (*D*) of the counter becomes '*1*'. This operates the relay '*L*' which closes the contacts L_1 to connect the telephone lines to a passive circuit (resistances).

The passive circuit simulates the conditions of taking the telephone transceiver from the cradle. Ringing stops at this moment. The passive circuit is disconnected when the user operates switch Sw_2 so that normal calls can be received.

A voice message for answer-back is digitised at 5kHz sampling rate and stored at 512k RAM. The RAM is addressed by a binary counter chain that is reset by the '*D*' output of the ringing tone counter.

The recorded message at the RAM is delivered automatically to the digital-to-analogue converter, which in turn produces the voice signal for the amplifier. This voice signal is relayed to the caller through the telephone circuit closed by the contacts L_1.

After about a minute – a period

One-transistor electro-luminescent flasher

Some thirty years ago, the LED flasher was state-of-the-art, superseding the neon flasher. The neon flasher was in fact far more efficient than the average LED flasher, but disappeared with the demise of high-voltage valve radio batteries.

Only recently has a new technology emerged, making possible the electro-luminescent flasher. The present circuit flashes a 5cm by 6cm electro-luminescent sheet at about 1.5 candelas' brightness, at a rate of just over 1Hz.

Although not as visible as a LED during the day, this flashes very brightly at night, producing some three times as much light as a super-bright LED flasher, yet drawing little more than 2mA.

A unijunction transistor, or UJT, produces 9V pulses across the series-connected secondary of a small 230V to 5-0-5V mains transformer. The pulses are determined by the monostable multivibrator – ringing tone counter is reset, causing relay L to open the contacts L_1.

In recording mode, the user operates switch Sw_1 manually. This causes the address-generating counter to enter count mode from the reset status and puts the RAM into write mode. In this case, voice message of the user from the microphone passes is sampled and held then passed to the a-to-d converter. From there, the digitised voice is sampled into successive locations of the RAM.

The sampling rate is kept at 5kHz to allow only up to 2.5kHz of the voice frequency to minimise the storage space needed. After recording is over, the user returns Sw_1 to its answer back mode position.

An audio-cassette recorder and player could be used for recording the answer back message and injecting into the lines whenever a call originates. This alternative is shown in **Fig. 2**.

In this case, three relays are used to control various operations, L for closing the line contacts L_1 to the passive circuit, P for playing the message (P_1) and RS for rewind and auto stop (RS_1).

While in record mode, the user operates switch Sw_1 to extend the connection for recording from the starting position of the tape. In answer-back mode, the counter operates the L relay to close L_1 and the P relay for playing the cassette.

At the end, the monostable multivibrator operates RS relay for rewind and stop.

K Balasubramanian and H Camar
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(F97)

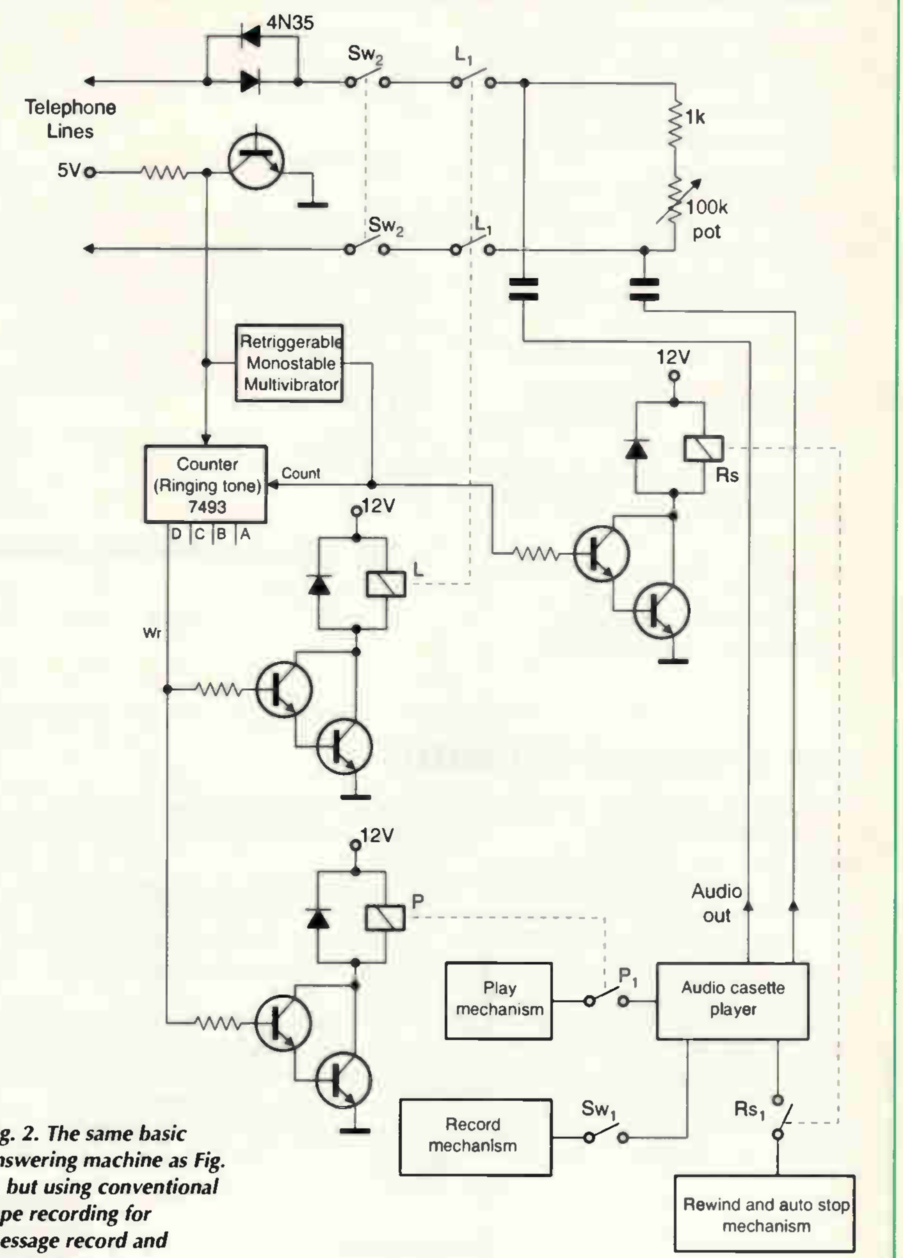


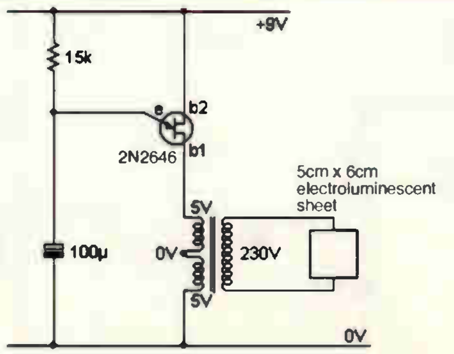
Fig. 2. The same basic answering machine as Fig. 1, but using conventional tape recording for message record and playback.

stepped up by the transformer to a voltage suitable for pulsing the electro-luminescent sheet, which typically requires 115V AC.

A relatively large value is chosen for C, so that sufficient current is available across the transformer's secondary windings. Should the electro-luminescent sheet fail to flash brightly, try other transformers, since the different characteristics of transformers may affect brightness.

A red 5cm by 6cm electro-luminescent sheet is manufactured by Seikosha, and is available from RS Components, stock No 267-8762. Various colours are available, including white. ■

Rev. Thomas Scarborough
Cape Town
Republic of South Africa



This electro-luminescent sheet is highly visible at night, yet only needs 2mA of drive current.

A method of measuring Q at VHF and UHF

Instrumentation to measure directly the Q of an *LC* resonant circuit at VHF and UHF frequencies hardly exists whereas instruments that measure return loss (reflection coefficient in decibels) at these frequencies are readily available.

The return loss measured at the input to an inductive loop coupled to a resonant circuit or line provides information from which Q_0 – unloaded Q – is easily calculated. The method is well suited to measurement of resonant lines, **Fig. 1**.

The procedure consists of two steps:

1. Adjust the coupling of the loop to obtain maximum return loss (±30dB) at the resonant frequency.
2. Measure the return loss bandwidth (*B*) at several levels, e.g. 5, 7, and 10dB. **Fig. 2**.

The effective Q of the coil or line is then calculated from $Q_0=f_o/B$ where:

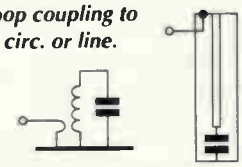
F_o = resonant frequency (MHz)
B = return loss bandwidth (MHz)
A = a parameter depending on the level at which B is measured. The relationship between return and loss and the parameter is:

$$\text{Return loss} = 10\log\frac{a^2}{a^2+4}$$

Some values are given in the table gives . The calculation can be repeated for several different decibel levels and a mean value for Q_0 obtained. The result is independent of the impedance of the return loss bridge, whether 50, 75Ω or another value (step 1 above takes account of bridge impedance.)

C J Hall
Giubiasco
Switzerland
(G10)

Fig. 1. Loop coupling to resonant circ. or line.



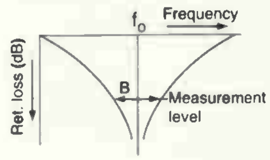


Fig. 2. Return loss graph.

Table. Values of return loss versus parameter A.

Return loss (dB)	5	7	10	15	20
Parameter *A*	1.360	1.00	0.667	0.361	0.201

Low drop-out regulator

This circuit was developed to fill a need for a regulator, with low drop-out characteristics, that could be used over a wider range of voltages and current than conventional monolithic regulators.

The circuit shown above is for a 5V, 2.5A regulator running off a transformer driven from 190V to 264V. A standard regulator would go into thermal shutdown at full load at the upper end of this input range.

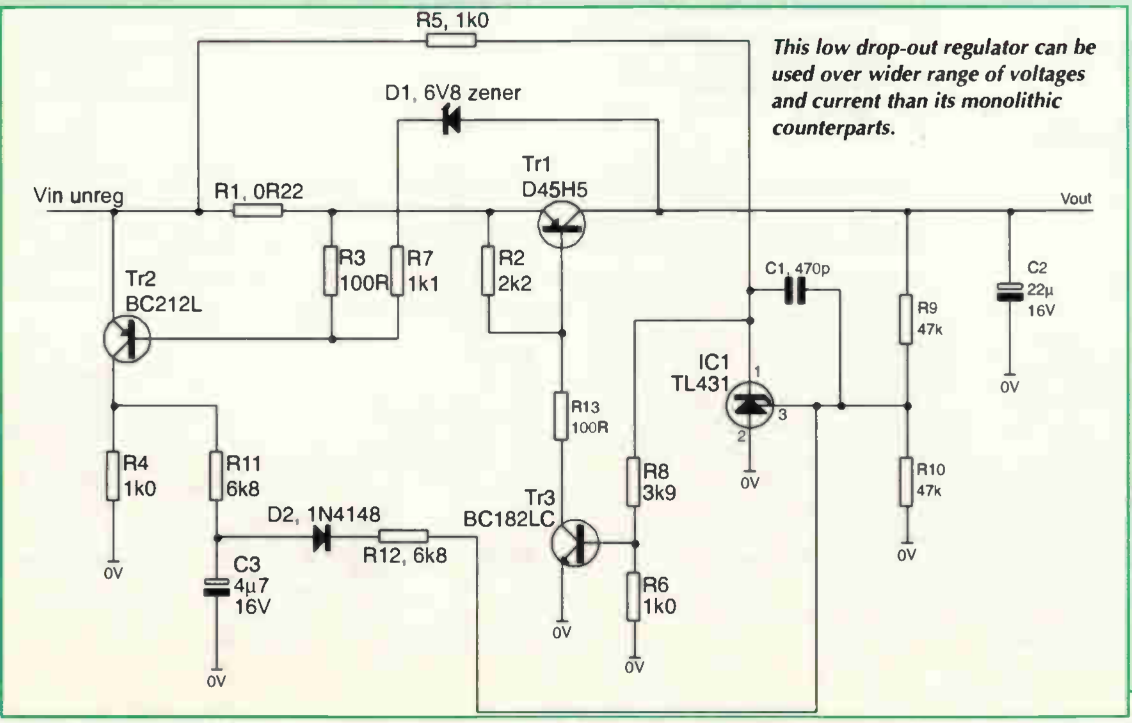
Output voltage is potted down by R_9 and R_{10} and compared with the internal reference voltage of IC_1, a TL431 shunt regulator. Standing current for IC_1 is supplied by R_5. Capacitor C_1 keeps the loop stable.

Output from the shunt regulator is amplified by Tr_3 to drive the series pass transistor Tr_1. This transistor specified is a high gain device with a low saturation voltage.

Current is sensed by R_1. When the voltage across R_1 is sufficient Tr_2 will begin to turn on, feeding current into IC_1. This is seen by IC_1 as an increase in output voltage, causing the control loop to reduce the regulator output.

Diode D_1 and R_7 are fitted if a foldback current limit is required. The values of these will need adjusting for a different input voltage range. Components R_{11} and C_3 keep the input ripple out of the control loop under current limit conditions. Diode D_2 isolates the error amplifier from the filter network under normal, non-limit, conditions.

John Vincent
Caterham
Surrey



This low drop-out regulator can be used over wider range of voltages and current than its monolithic counterparts.

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Bluetooth baseband controller is 1.1 compliant

Ericsson Microelectronics has introduced a Bluetooth baseband controller IC pre-qualified to Bluetooth version 1.1. The controller is based on a scalable Ericsson Bluetooth 1.1 compliant core combined with an ARM7 TDMI processor and an AMBA bus. It supports up to 7 slaves, making it suitable for both point-to-point and point-to-multipoint applications. The PBM 990 90 offers 64kbytes of on-chip RAM, an external bus interface, variable input frequency and integral data security. Host controller interface (HCI) connectivity is provided through UARTs or a USB port. One UART supports HCI connectivity up to 921kbit/s while a second can be used at up to 230kbit/s for production and debugging purposes. The USB port is a self-powered, USB 2.0 compliant, full-speed device interface.

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Tel: 01793 488300
www.ericsson.com

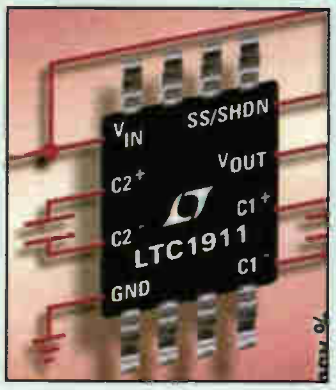
Ferroelectric memory cell based on single transistor

Ramtron has introduced what it claims to be the first 1-transistor, 1-capacitor ferroelectric random access memory (FRAM). The FM24C256-SE is a 256kbit two-wire, serial FRAM product that uses a one-transistor, one-capacitor memory cell. According to the supplier, the architecture lays the groundwork for memories that can combine the density of DRAM and the low power, equal read/write speed of SRAM with the non-volatile storage capability of flash or EEPROM. The architecture uses a single capacitor as a common reference for each column in the memory array, effectively cutting the required cell area in half compared to Ramtron's existing 2 transistor and 2 capacitor Read and writes are performed at bus speeds up to 1MHz. It operates from a 5V power supply and draws 75µA for reads and writes at 100kHz. The FM24C256 offers 10 years of data retention.

Ramtron
Tel: 001 719 481 7000
www.ramtron.com

DC-DC converter is efficient without inductors

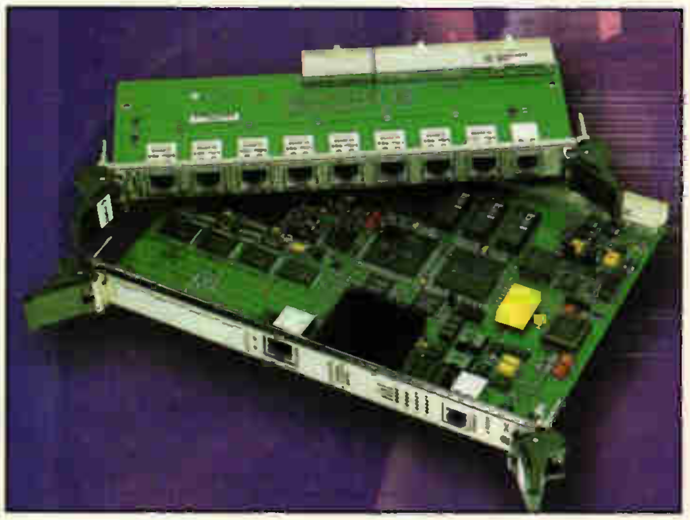
Linear Technology's inductorless step-down DC-DC converter produces a 1.5V (LTC1911-1.5) or 1.8V (LTC1911-1.8) regulated output from a 2.7V to 5.5V input voltage. It is designed for applications that use a single cell Lithium Ion or multi-cell NiCd/NiMH batteries to power low voltage DSPs. The device uses switched capacitor fractional conversion to achieve its efficiency over the input voltage range. Efficiencies up to 85 per cent can be realised without an inductor. An internal switch network modifies the step-down conversion ratio to optimise efficiency as the input voltage and load conditions vary. The device can provide up to 250mA of output current while minimising operating current to 180µA. Input current is being further reduced to 10mA in shutdown mode. A constant frequency architecture provides a low noise regulated output as well as lower input noise than conventional charge pump regulators, thereby minimising any potential interference with wireless transceivers. High frequency operation (fosc=1.5MHz) simplifies output filtering and allows the use of small external ceramic capacitors to minimise



Gigabit ethernet switches to PICMG2

Performance Technologies is offering its first Gigabit Ethernet products which are compliant with the bus standard designed to extend the life of PCI-based systems by moving traffic from a shared bus architecture to an embedded Ethernet backplane. The Ipnexus CPC5400 is a Layer3 Gigabit Ethernet switch which is compliant with the interface standard ratified by the PCI Industrial Computer Manufacturers Group (PICMG). Features include eight Gigabit Ethernet ports, wire-speed layer2/layer3 switching and a number of availability features for carrier-grade operation. Other features include support for 802.3AD link aggregation, 802.1q VLANs and 802.1p QoS/CoS that is designed to ensure propriety traffic receives the necessary handling. The switch can be deployed in a fully redundant, non-blocking network.

Performance Technologies
Tel: 01491 410066
www.pt.com



Please quote *Electronics World* when seeking further information

the footprint and profile, said the firm.
Linear Technology
Tel: 01276 677676
www.linear-tech.com

Voltage regulator is rad-hard to 300krad

Intersil has announced a TO-39 can version of its 117-type linear voltage regulator for space power applications with electrical performance tested and guaranteed to 300krads. Designated the HS-117RH, the 1.2 to 37V linear regulator has an 0.5A peak output current and a 40V maximum input voltage.
Intersil
Tel: 01344 350250
www.intersil.com

Quartz crystal for Bluetooth transceivers

Epson's latest surface mount quartz device is the FA-248 AT-cut crystal which oscillates in fundamental mode in a range between 12MHz and 32MHz. Maximum overall frequency deviation is $\pm 20 \times 10^{-6}$ including first year aging. Aging or long-term stability reaches its highest value in the first year's use of the quartz. The value of $\pm 2 \times 10^{-6}$/year means a lower level in the following years. Better overall stability values are available on request. The series resistance ranges from 20Ω for 32MHz to 70Ω for 12MHz.
Epson
Tel: 0049 891 4005 277
www.epson-electronics.de

Test probe with easy tip replacement

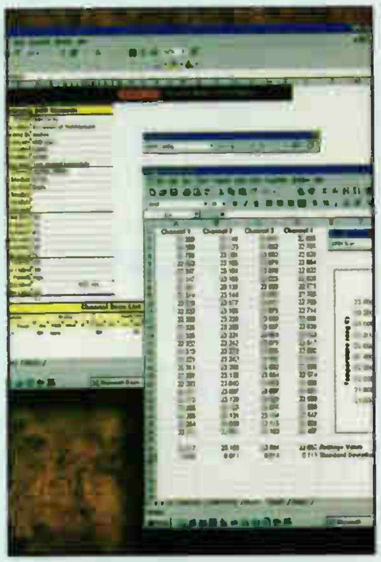
Pomona Electronics has introduced a range of test probes offering speedy tip replacement without tools. The SMD Microtip probe tips are designed for making contact on high density chip leads. It has a quick-change chuck that allows the user to change tips and customise the probe for different uses, eliminating the need to keep many different types of probes on hand, said the firm. The probe is designed for use with a variety of DMMs.
Pomona Electronics
Tel: 0031 40 2675 212
www.pomonaelectronics.com

Output diodes run at high temperature

International Rectifier has announced a 70A, 200V fast diode, with a 175°C operating temperature rating. It is for high frequency switching and output rectifier applications such as welding, switch-mode power supply (SMPS) and power conversion. Switching characteristics are optimised for high frequency applications. For example, the diode has a reverse recovery time of 26ns at full-rated current.
International Rectifier
Tel: 020 8645 8003
www.irf.com

Data acquisition and analysis software excels

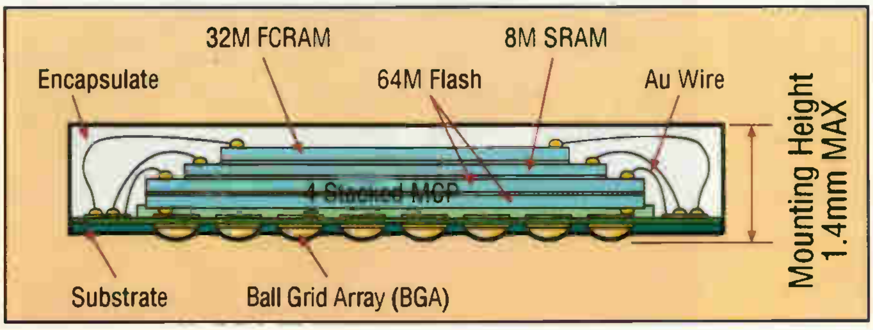
Keithley Instruments has added to its line of ExceLINX software products which collect, analyse, and chart test data from within Microsoft Excel, using the firm's



Model 2700 and Model 2750 DMM-based IEEE-488 benchtop instruments. Test data is imported into Excel in real time, without programming, separate applications, or importing and exporting data files. It enables Excel to control test hardware, and log data directly to a file, spreadsheet or graph. It will provide more power than start-up software often bundled with instruments and plug-in boards. The company says the software package will be suitable first time and occasional users who want to begin acquiring and viewing data quickly, without investing time in programming.
Keithley
Tel: 0118 957 5666
www.keithley.com

Multi-chip device is well stacked with memory

Fujitsu has introduced its first four-stacked multi-chip package (MCP) incorporating a pair of 64Mbit NOR dual-operation flash memory chips, a 32Mbit mobile FCRAM with asynchronous SRAM interface and 8M SRAM. In addition, a two-stacked MCP incorporating 64Mbit FCRAM has also been announced, representing what the company terms the highest mobile FCRAM on the market in a 9 x 9mm package. Samples of both are currently available with volume production expected during Q2. The package combines the memory used in two existing MCP products in a single four-chip stacked MCP with a 103-ball (10.0 x 9.0 x 1.4mm) plastic BGA package. Combining these chips into one package reduces the total mounted surface area to about 63% of the previous combination of packages, said the company. According to the firm, it is the introduction of 2.5G and 3G mobile phones next year in Europe which is increasing demand for MCP memory that is physically smaller and runs on less power, but with more capacity and speed. The MCP includes two 64Mbit NOR-type dual-operation flash memory chips for storing programs and data. This flash memory, which was fabricated using of 0.17-micron process technology, runs faster than previous NOR-type flash products, performing sector erase operations in 0.2 seconds (one second for previous models), and has access speeds of 70ns.
Fujitsu
Tel: 0049 6103 690257
www.fme.fujitsu.com



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Viterbi decoder core is 802.16 compliant

Altera has released an enhanced version its Viterbi decoder core, claimed to run over 185Mbit/s throughput. The Viterbi decoder version 3.0 is up to 50 per cent smaller and twice as fast as previous. It is powered with the firm's MegaWizard plug-in technology, enabling users to parameterise various features such as constraint length (3 to 9) and number of soft decision bits (2 to 16). It is fully compliant with the IEEE 802.16 specifications for air interface of fixed point-to-multipoint broadband wireless access (BWA) systems. These require a Viterbi block decoder with constraint length of 3, traceback length of 32, and minimum throughput requirement of 44.8Mbit/s.

Altera
Tel: 01494 602000
www.altera.com

"Hole-less" Mosfets up the power

Ixys has introduced three Trench Power Mosfets for low voltage/high current applications. The IXUC100N055, IXUC200-N055 and IXUC160N075 have current ratings of 100A, 200A and 160A respectively. The IXUC100N055 and the IXUC200N055 are rated at V_{DS} (drain-source voltage) equal to 55V while the IXUC160N075 has a V_{DS} rating of 75V. The devices are housed in what the company terms a "hole-less" package conforming to the TO-220 outline with an electrically isolated mounting tab. The elimination of the screw hole makes room for a larger die, which can carry higher current than the regular TO-220 package. Features include 2,500V (rms) isolation and maximum operating junction temperature rating of 175°C.

Ixys
Tel: 001 408 982 0700
www.ixys.com

High-definition TV format conversion IC

Gennum is supporting HD-SD format conversion for the European broadcast market with its latest IC products aimed at broadcast, projector, plasma/LCD display and other professional applications. The GF9330 de-interlacer is a VDSP engine that performs de-interlacing of interlaced digital television signals and supports both SDTV and HDTV signal formats, including 1080P60. It includes edge directional adaptive filters, a motion adaptive vertical filter and a motion adaptive inter-field filter. The GF9331 motion co-processor can be used in conjunction with the GF9330 to implement a motion adaptive de-interlacer for both SDTV and HDTV digital television signals. The GF9331 contains pixel-based adaptive edge detectors that cover a number of different edge detections and a motion detector that covers special motions in the vertical direction. The GF9320 scaling processor performs scaling of video as well as graphic images up to 2k by 2k pixel resolution and provides pan, scan and zoom/shrink abilities. A fully programmable output matrix provides for YUV to RGB conversions, gain and hue controls. In addition, it offers colour background and output TRS insertion. Typical applications range from high-end broadcast through to professional display applications.

Gennum
Tel: 01252 747001
www.gennum.com

2.8GHz clock dividers

ON Semiconductor has introduced two low-voltage clock chips that generate, divide and distribute synchronous output clocks from a single master clock running at up to 2.8GHz. Both the MC100EP139 and MC100LVEP34 were designed specifically for low-

GSM test set simplifies AM modulation

IFR Systems is offering an AM signal modulator for use with its IFR2935 GSM radio test set. The aim is to replace the separate signal generators previously required for adjusting GSM mobiles for AM suppression and speed up test time by reducing the need continually 'make and break' RF connections. It features two modes of operations - Direct and AM. In Direct Mode, the GSM mobile is connected to the test set and can be left for the majority of the testing and aligning operation. In AM Mode, the signal is amplified and modulated for updating and operated for just a few seconds while the mobile's AM suppression is adjusted.

IFR Systems
Tel: 01438 772087
www.ifrsys.com

Planar transformer

Pulse has announced a series of 30-watt planar transformers which it claims can offer a size and weight reduction of up to 60 per cent in DC power supplies or in discrete power supply systems. Primary applications for the new transformer series include telecom power systems, industrial control systems, or automotive and heavy equipment applications where high package density is required. Based upon planar technology, the transformers are formed from multi-layer printed circuit boards (PCBs) sandwiched between ferrite cores. EMC compliant, with operating frequencies from 150kHz to 750kHz, and various turns ratios to suit a wide range of product applications, the transformers are available in both tray, and tape and reel packaging.

Pulse
Tel: 0033 384 350448
www.pulseeng.com

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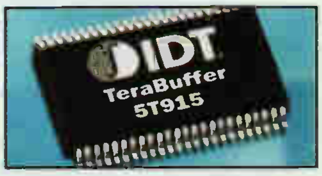
skew and low-jitter applications, said the company. Both parts feature a master reset for synchronization of multiple chips and a synchronous enable/disable function.
The MC100EP139 can generate four output clocks from a single master clock. This device gives designers the choice of programming a divide-by 2/4 on one pair of outputs, and a divide-by 4/5/6 on the other pair.
The 3.3V to 5V EP139 can accept input clock frequencies up to 1.3GHz. Typical output skew is a mere 50ps. This device is offered in two 20-pin package options - the low profile TSSOP and the standard SOIC.
ON Semiconductor
Tel: 0033 0534 611390
www.onsemi.com

Testing SoCs for set-top box and xDSL designs

Schlumberger Semiconductor has extended the capability of its EXA3000 system-on-chip (SoC) test system with a broadband channel that expands multi-site analogue capabilities when testing SoC devices in DVD, set-top box, HDTV, xDSL and cellular baseband applications.
One broadband channel option can source and measure two channels in parallel with greater than 86dBc SFDR at 1MHz.
This, with 100MHz of usable bandwidth at 14 bits of resolution, is designed to improve analogue test coverage.
Each card has two independent channels embedded.
MPEG video decoders, used in set-top box and DVD designs, are the first targeted application for the Broadband Channel, said the company.
Schlumberger
Tel: 0033 1474 66943
www.slb.com

Clock buffer has over 40 I/O combinations

IDT's latest family of fan-out clock buffers is for dual and single data-rate clock distributions. Called the TeraBuffer, the device can offer more than 40 I/O combinations. The devices are non-phase-locked loop (PLL) clock buffers and feature the industry's output-to-output skew specs. They are claimed to be the first clock management devices providing signal and voltage translation capabilities that include 1.8V TTL. They accept single-ended or differential 2.5V TTL, 1.8V TTL, HSTL, eHSTL and LVEPECL input signals, and 2.5V LVTTL, 1.8V LVTTL, HSTL, and eHSTL output signals in various combinations.



IDT
Tel: 01372 366112
www.idt.com

PCB design tool with all the angles

Zuken's Cadstar radial design environment for printed circuit boards enables the design of radial boards, those with identical blocks of design centred around a single device or area.

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This can improve design efficiencies by introducing a degree of automation and conformance to pre-defined manufacturing rules.
Zuken
Tel: 01454 207800
www.zuken.com

Schottky diodes for current protection

Schottky diodes from Shindengen provide secondary rectification in DC-DC converters and protect portable terminals against battery reverse current and short circuit conditions. These may be caused by incorrect connection of an AC adapter or battery charger. The M-series diodes for DC-DC converter applications are suitable for use in step-down, boost and inversion circuit topologies up to 30V maximum reverse voltage. Forward voltage drop is 0.4V, minimising in-circuit losses. The M1FM3 packs 3A continuous forward current handling into the industry standard M1F surface mount package measuring 1.8mm x 3.9mm x height 1.4 mm M1F. The 5A D1FM3 in industry standard D1F packaging measures 2.5 x 5.0 x height 2.0mm, and the M2FM3 for forward current up to 6A is packaged to M2F standard measuring 3.75 x 5.1 x 2.0mm. The device handles continuous forward current 1.5A in the small M1 outline, with the D1FH3 rated for 3A current and available in the D1F package. The M2FH3 in the M2F package protects equipment drawing up to 6A in normal operation.
Shindengen
Tel: 01707 332992
www.shindengen.co.uk

Digital TV conversion platform

STMicroelectronics has introduced a hardware and software platform it claims can bridge the gap between PAL analogue television and digital-TV technology. The CTV100 platform comprises two chips (STV2310 and STV3500) and associated software for 100Hz or progressive scan TVs. It is designed to provide digital conversion for analogue TV input standards – PAL, NTSC, SECAM – and analogue RGB and YCrCb input formats which are converted into a digital CCIR 656/601/square pixel stream (D1 format). A VBI slicer is embedded and the STV2310 is programmable through an I^2C interface. ■
STMicroelectronics
Tel: 01628 890800
www.st.com

Designing super-regens

Super-regenerative radio designs have been around for well over 60 years, and hundreds of circuits have been published – usually with little or no description of how they work. Eddy Insam dispels the myths.

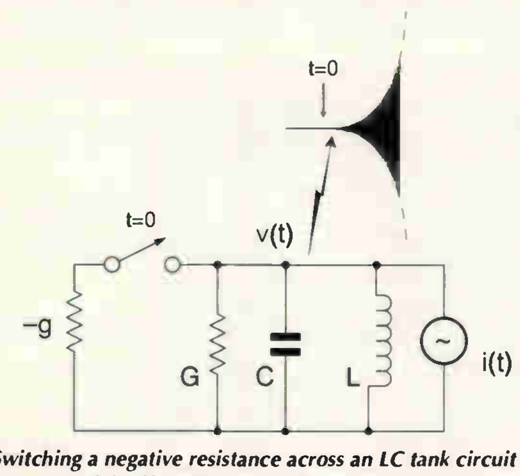

Fig. 1. Switching a negative resistance across an LC tank circuit causes positive exponential self-oscillations to be generated at its natural resonant frequency. The start-up time is a function of initial conditions – namely the tiny RF currents induced in the coil from an antenna. The resulting wave growth can be easily measured by external circuitry. The negative resistance is generated in practice by the active component used: transistor, FET, or other.

A few years ago, I was doing a short project for a cell-phone manufacturer. During lunch break, a number of the company's RF engineers were sitting around the table looking at a piece of paper with a little circuit drawn on it. "How does it work?" they asked me.

It was obviously a super-regenerative receiver, it said so underneath. It had one transistor, a few resistors, and some capacitors. "All this RF expertise, and not one of you knows how this works?" I said, saving myself from giving a direct answer, as I looked for the door.

Some of us have a box in the back of our minds where we put all those concepts that we cannot quite grasp. One might label this box, 'things that work by magic.' This circuit was certainly in there somewhere.

I decided I would find out how it worked. After all, the circuit only had one transistor. It can't be that difficult to crack, or can it? Soon, I found that most of the articles and manufacturers' literature I looked at were particularly short on theoretical descriptions. "Adjust R_4 for best sensitivity," was the best I could get. I had to go quite a way back in time to find any publications that covered the subject in detail. I was also surprised to find how little had been published since then.

Nowadays, with so many alternative solutions available, super-regenerative radios appear to be outmoded and irrelevant. On the plus side, the concept has virtues; such radios are simple and consume little power. A better understanding of how they work may widen their applications into new areas. I would not be surprised if they make a come-back every so often under different guises – especially as new components such as stable hi-Q resonators are developed.

Super-regen theory

One always gets a good feeling when the operation of an intricate physical process is made clearer by the use of simple maths. Equivalent circuits are a good case in point. Look at the circuit in **Fig. 1**. This describes the very basic operation of a super-regenerative receiver.

The circuit is a simple LC tank that includes a conductance G representing all the resistive losses in the circuit. I'll be using conductances rather than resistances as it simplifies the maths and the final analysis. To all extents, $G=1/R$, where R is the loss resistance.

A small RF source $i(t)$ is also shown in parallel with the coil. This represents the external RF radiation that is somehow induced into the tank, maybe directly with the coil acting as an aerial, or coupled from an RF amplifier (not shown). Just assume for the moment that this signal has very small amplitude, and has a frequency similar to that of the self-resonance frequency of the tank:

$$i(t) = A \times \sin(\omega t) \quad \text{where,} \quad \omega \cong \sqrt{\frac{1}{LC}}$$

Before t=0, the switch is in the off position. The only voltage across the tank is that resulting from the RF excitation. I am calling this quiescent voltage V_o, which is basically $V_o = i(t)/G$ at resonance. So far, so good.

The interesting bit happens when the switch is closed. At this point an extra conductance is added across the tank, and the total load is now the sum of the quiescent conductance G plus the new conductance just switched in.

To simplify the analysis, imagine replacing the two conductances by a single one that switches its value from G at time t<0, to a new value $-g$ at t≥0. Notice that this new conductance is negative. In other words, at time t=0 and beyond, the LC tank has a negative conductance $-g$ applied across it. You cannot buy negative resistors at the shops. However, a negative resistance can be emulated with an active component, such as a transistor or FET. More on this later.

The corresponding equation for the circuit is:

$$C\frac{dV}{dt} + gV + \frac{1}{L}\int V dt = i(t) \qquad (1)$$

And the solution, in terms of the resulting voltage across the tank, is of the general form:

$$v(t) = v_o \times k \times \exp\left(+t\frac{g}{2C}\right) \times \sin(\omega_o t) \qquad (2)$$

where:

$$\omega_o = \sqrt{\left(\frac{1}{LC}\right) + \left(\frac{g}{2C}\right)^2}$$

Note the positive sign in the exponential. The output is an exponentially-rising, high-frequency oscillation. Figure 1 shows the shape.

Mathematical types among you will immediately notice that such an output requires non-zero initial conditions. This is provided in our circuit by the minute RF excitation. If the input RF signal had zero amplitude, there would be no drive, and the circuit would take an infinite time to start oscillating. This is not really an option. In practice, small disturbances such as thermal noise are enough to start the oscillations.

One peculiar characteristic of exponential waveforms is that two similar functions of different amplitudes look like the same two functions but time delayed. **Figure 2a)** shows the effect where two input signals of different amplitudes, $e_1(t)$ and $e_2(t)$, are shown as apparent time delayed versions of each other. This time delay is a logarithmic function of the ratio of amplitudes, i.e.:

$$e_2(t) = \text{k} \times e_1(t) \times \exp(t\frac{g}{2C})$$

$$e_2(t) = e_1(t) \times \exp\left((t+T)\frac{g}{2C}\right)$$

where:

$$\text{k} = \exp\left(\frac{g}{2C}\right) \quad \text{a constant}$$

What this means is that, no matter how small our input RF signal V_o is, sooner or later there will be a corresponding exponentially rising waveform generated across the coil. This waveform will always have the same 'shape' and amplitude, but its 'rise time' or delay – if we could define such a term – will vary. It will also be dependent on the amplitude of our minute source signal V_o.

Left to themselves, the oscillations would carry on increasing in amplitude for ever. In practice they will level out as nonlinearities in the associated circuits come into effect. The system then becomes a steady oscillator.

We are not interested in this however. We are only concerned about detecting the time between the switch closure and the time at which the self-oscillations rise to a specific level or threshold. Measure these times, and you have a radio receiver. Figure 2b) shows the principle.

One very important point is worth noting: The input RF excitation provides the starting conditions at time t=0 and thereabouts. The input signal plays no part in the process much before, or after this time. In other words, this is a RF sampled data system.

Another important fact is that the resulting self-oscillations are at the natural frequency of the LC tank as given by equation, and not at the frequency of the input RF signal.

The detector

How does all this translate into radio receiver designs? Going back to Fig 2b), you can see the principles of a possible detector scheme. A voltage-level comparator

Further reading

Most of the original work on super regenerative receivers was done in the early forties and during WW2. To my knowledge, there has been amazingly little published in terms of in depth coverage since then. The most interesting articles are:

Frink, F., 'The Basic Principles of Super-Regenerative Reception', *Proceedings of the IRE*, Vol. 26, No 1, January 1938.

Bradley W.E., 'Super-regenerative Detection Theory', *Electronics*, Vol 21, September 1948.

Strafford F.R.W., 'The super-regenerative detector: an analytical and experimental investigation', *Proceedings of the IRE*, September 1948.

MacFarlane G.G, and Whitehead J.R., 'The theory of the super-regenerative receiver in the linear mode', *Journal IEE*, Vol. 95, pt III, pp. 143-157, May 1946.

Recent articles and relevant literature on oscillators and associated components:

Ian Hickman, 'Practical RF Handbook' 2nd Edition, 1997. Newnes Publications ISBN 0 7506 34472

Ian Hickman, 'Superregen or superreplacement', *Electronics World*, February 1999. A recent article on receiver design that mentions super-regenerative receivers.

Staff feature, 'Low Noise Microwave Oscillators', *Electronics World*, November 1990. A short note on the design of low noise oscillators.

Martin Eccles, 'Applying Crystals', *Electronics World*, August 1994. Useful information on the workings of crystals and associated oscillator circuits.

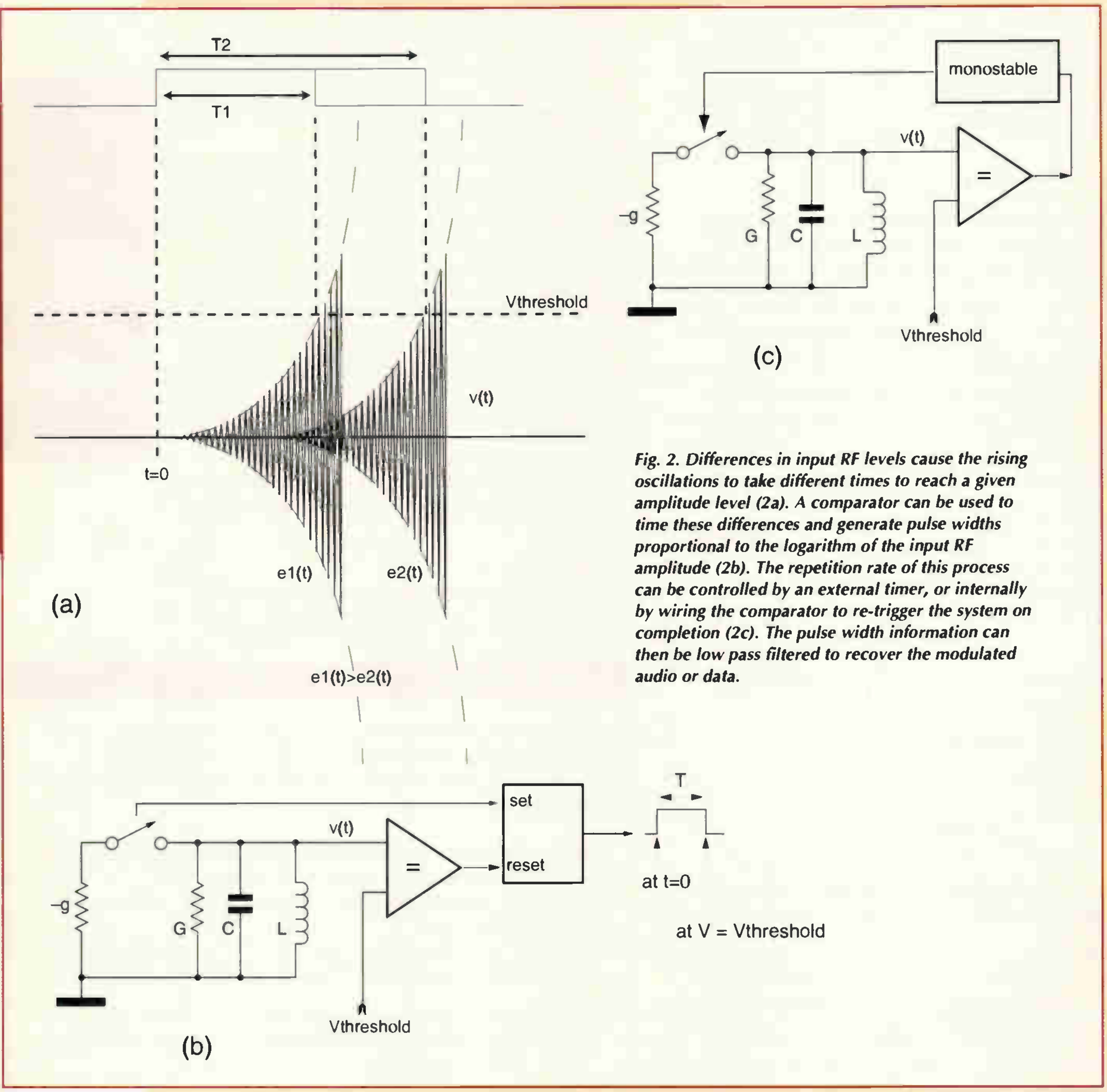

Fig. 2. Differences in input RF levels cause the rising oscillations to take different times to reach a given amplitude level (2a). A comparator can be used to time these differences and generate pulse widths proportional to the logarithm of the input RF amplitude (2b). The repetition rate of this process can be controlled by an external timer, or internally by wiring the comparator to re-trigger the system on completion (2c). The pulse width information can then be low pass filtered to recover the modulated audio or data.

senses voltage peaks across the coil. The detector triggers when the first peak reaches some fixed threshold. The actual value is not too important, as long as it occurs before saturation.

At time $t=0$, a latch flip-flop is set at the same time the switch is closed. Some time later, the comparator resets the latch when the first positive cycle of $v(t)$ hits the threshold. We then take the pulse width generated by the latch. The general relationship between this width and the original RF input amplitude V_o is:

$$T = \log(kv_o)$$

That is, the resulting width is proportional to the logarithm of the input signal amplitude. In other words, linear microvolts are converted to logarithmic milliseconds. This logarithmic response is what gives super regenerators their intrinsic automatic gain control, or AGC, behaviour.

One thing may not be very clear at this point. The detector is a 'once only' device. It takes one sample of the input RF signal, generates a single pulse width, and then stops. In order to detect more samples, the switch has to be opened, oscillations across the tank have to be allowed to settle or decay to zero, and the operation has to be restarted by closing the switch again. The process is then repeated *ad infinitum*.

Quenching

The traditional name given to the process of stopping and restarting the detector after each sample is 'quenching'. It comes from the fact that the *LC* tank needs to be 'quenched', or stopped from oscillating, after it has reached full swing. Quenching rate is another word for sampling rate.

When the switch is opened, the self-oscillations remaining across the tank will start to decay. The decay rate will be given by the original components in Fig. 1, that is:

$$v(t) = v_1 \times \exp(-t\frac{G}{2C}) \times \sin(\omega_o t)$$

Note that the conductance is now G, the original quiescent loss resistance of the tank, as opposed to the negative add-on $-g$, which has now been disconnected.

This waveform must be allowed to decay to as near zero as possible, and to less that the expected input RF signal level. If it isn't, the receiver may sample remainder signals from previous oscillations in preference to our wanted RF input, reducing the sensitivity.

This decay can take a considerable time – especially in low loss tank circuits. This places a limit on the receiver's repetition sampling rate.

Going back to the detector of Fig. 2b), a simple way to provide quenching is by driving the switch on and off using an external, fixed-rate, repetitive signal. The repetition rate of this external source must be low enough to allow sampling, and also to allow for full discharge of the *LC* tank. The final output from the detector will be a fixed rate, variable pulse-width modulated (PWM) signal. This method is traditionally known as external quenching.

The advantages of external quenching are that it is easy to implement, and produces predictable results. The disadvantages are that a large part of the pulse width cycle is wasted, resulting in low conversion gain when decoding pulse widths into analogue signals.

An alternative solution is one where the sampling is restarted as soon as the pulse width terminates. That is, we make the end of the detection pulse activate the next start. This is known as self-quenching.

Figure 2c) shows the concept. The falling edge of the comparator's output triggers a monostable that produces a pulse wide enough to discharge the *LC* tank, and is used to re-set the switch. The detector output is now a pulse train of variable width and variable frequency. No external repetitive source is required, as the circuit will self trigger. Self-quenching generally produces higher conversion pulse width recovery rates.

Quench rate

Like any sampled data system, the quench or sample rate will place a limit on the recovered bandwidth: do not confuse this with radio selectivity. This rate must be at least twice that of the wanted recovered bandwidth. For example, for audio work, quench rate should be at least 10kHz.

There are three parts to the quench cycle, Fig. 3. These are RF rise, RF decay, and the device's relaxation time, which overlaps it. The RF rise is due to the effect of the negative resistance during sampling. RF decay is the time taken for the tank to drain its oscillations, and the relaxation time is the time taken for the circuit itself to 'recover'.

The minimum value for the quench rate is bound by the rise and decay times of oscillations in the tank circuit. It helps to visualise these quench times by relating the figures to the Q or merit figure of the circuits as follows:

$$\text{min quench } T = T_1 + T_2$$

$$T_1 > \frac{2Q_1}{\omega_o} \quad \text{where,} \quad Q_1 = \frac{\omega_o C}{g}$$

$$T_2 >> \frac{2Q_2}{\omega_o} \quad \text{where,} \quad Q_2 = \frac{\omega_o C}{G}$$

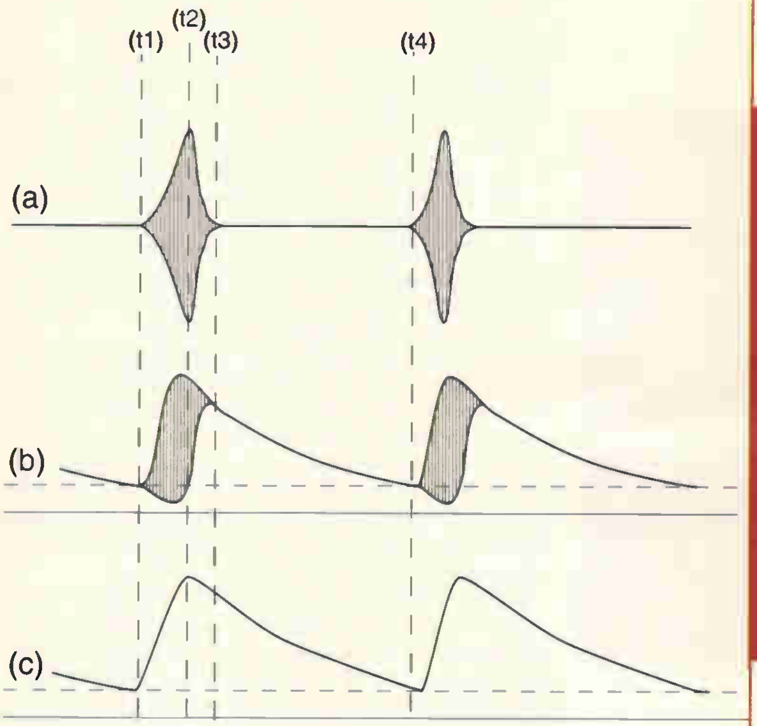

Fig. 3. Self-quenching action for the circuit in Fig. 4. Section a) is the voltage at the collector, b) at the emitter, and c) at the R_1/C_2 junction. The horizontal dotted line represents the threshold voltage below which the transistor comes into conduction (about 1.4V). During conduction, i.e. between t_1 and t_2, the transistor operates in class C, building up a charge that eventually causes the transistor to switch off suddenly at t_2. A slow discharge via R_1/C_2 allows oscillations to start again at t_4.

Later on, I'll mention that best performance is achieved when the negative conductance is kept as low as possible around $t=0$. However, this would make T_1 in the equation above too large.

A design compromise calls for a circuit in which g is kept very low at the start, but somehow is sharply increased in value after this time. Remember that the receiver sampling action only occurs around $t=0$.

Also, note that T_2 is required to be much greater than the natural Q of the quiescent tank. You want the remaining self-oscillations built up in the tank to drop to as near zero as possible: The definition of Q is associated with a drop of only $1/e$ of amplitude. This is the main reason why it is difficult to use high-Q devices such as quartz crystals in super-regenerative receivers. For example, a 30MHz crystal may place a quench rate limit of less than 400Hz.

Self-quenching in practice

One of the reasons why super-regenerative receivers were so popular in the past was that a single device – a valve or a transistor – could be used to perform detection and self-quenching all in one, and at the same time, thus saving on components.

The principle of self-quenching in active devices is based on the relaxation principle. The two operations, namely self-oscillation and self-quenching, are also linked. The active device behaves as a non-linear negative resistance device, converting DC power into AC power, stored as dynamic energy within the *LC* tank.

Design of the circuit is such that as energy in the *LC* tank is 'absorbed' and the wave amplitude increases, the average DC quiescent operating point of the active device shifts. Most active devices only support negative resistance over a short operating range. Therefore, as the DC operating point shifts, there comes a point at which the active device cannot provide any more negative resistance.

A normal oscillator will settle just below this operating point and provide constant oscillations. A relaxation system cannot, because it will have extra components added to prevent this.

When the limit operating point is reached, the active device will 'starve' and oscillations will just stop. The only option left is to discharge the circuit to its original DC conditions before starting all over again.

A common super-regenerative design uses the rising self-oscillation peaks to pump-charge a capacitor, taking advantage of some diode-like nonlinearity such as the base emitter junction of a transistor. This charging results in the increase of some related dc parameter such as base bias.

During build-up, the RF cycle peaks are always higher and higher, causing conduction of the transistor on the tip of every further cycle, and ahead of the rising DC bias, resulting in a form of pumped Class C device.

At some point – collector saturation for example – the rate of rise will flatten, causing no more pumping cycles. This leaves the transistor heavily negatively biased, and therefore switched off. The transistor can now only wait for the bias capacitor to discharge slowly, before it resumes conduction, allowing the cycle to start all over again.

Most of the cunning designs seen in super-regenerative receivers mainly relate to methods and techniques for self-quenching using the least of components. As I'll be mentioning later, there is usually a clash between circuit simplicity and ideal design objectives.

Time-varying conductance

In my simple analysis of Fig. 1, I have assumed that the negative conductance is applied suddenly. In practice, this is not the case.

A transistor, coming into active mode will present a load that varies with time. In other words, g is really a function of time $g(t)$. Equation 1 needs to be written as:

$$C\frac{dV}{dt} + g(t)V + \frac{1}{L}\int V dt = i(t)$$

As the solution of this equation is a bit more complex, I am going to make some assumptions. The only area of interest is what happens at around $t=0$. You can also assume that $g(t)$ is a simple function of time, for example a linear slope, and that the input RF signal is a short duration burst at around $t=0$.

A simplified solution for the resulting waveform at time T, where T is a short time after $t=0$ is:

$$v(t) = v_o \times k \times \exp\left(\frac{1}{2C}\int_0^T g dt\right)\sin(\omega_o t)$$

Compare this with equation 2.

Some extra terms should be included because the RF input signal is present during the whole of the time. A more comprehensive result is as follows (still not including all the terms!):

$$v(t) = \left(v_o \times k\sqrt{\frac{1}{|g\Phi|}}\right) \times \left(\exp\left[\frac{C(\omega-\omega_0)^2}{|g\Phi|}\right]\right) \times \left(\exp\left[\frac{1}{2C}\int_0^T g dt\right]\right) \times \sin(\omega_0 t) \qquad (3)$$

where:

$$g\Phi = \frac{dg}{dt}$$

The significance of some of these terms will be covered in the following sections.

Comments on sensitivity

It may be a good idea to start by defining what is meant by sensitivity. Normally, this is the signal level a receiver must have at its input, in order to generate an output which can be recorded as a signal to noise ratio, or signal+noise

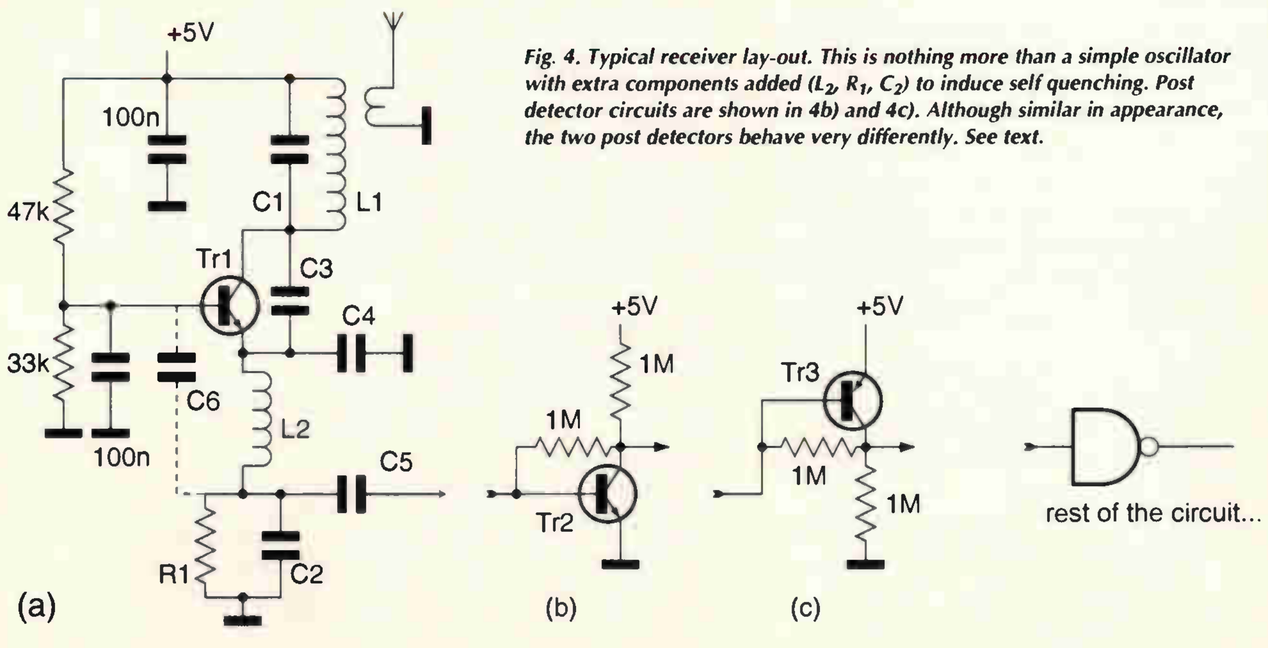

Fig. 4. Typical receiver lay-out. This is nothing more than a simple oscillator with extra components added (L_2, R_1, C_2) to induce self quenching. Post detector circuits are shown in 4b) and 4c). Although similar in appearance, the two post detectors behave very differently. See text.

to noise ratio. In practical terms, this is determined by total gain, and by the various sources of noise that impart a floor to performance.

In a super-regenerative receiver, the 'instant' sensitivity goes through a maximum value at $t=0$, and falls off rapidly on each side. It is primarily a function of the negative conductance and its slope at this point.

During this time, the receiver has a very large gain. So large in fact, that noise level is the only limiting factor. You can work the gain out from equations 1 and 2. You will recall that self-oscillations are always generated, no matter how small the initial conditions, whether from a RF signal or from noise.

The main source of noise comes from the active component. A transistor operating as a negative resistance generator can be a very noisy device. A main design objective is to employ a circuit that generates the minimum amount of noise.

I am not going to cover details on how to design low-noise oscillators, except to say that this is a very important design consideration regarding sensitivity. The further reading matter mentioned in the separate panel contains some further information on this.

A number of other factors also contribute to sensitivity. Firstly, there is the necessity to couple the RF signal into the tank coil with the minimum of loss. In order to avoid radiation, the antenna should not be connected directly into the tank. A separate RF amplifier is preferable.

The RF amplifier does not need to have much gain; only enough to improve the signal to noise ratio and to provide for some RF isolation. As an aside, and to conserve power, the RF amplifier only needs to be powered during sampling time.

Secondly, there is the efficiency of the detector in converting the pulsed modulated train to useable analogue voltages. I have already covered the marginal advantages that self-quenching can provide. To maximise conversion, the PWM detector must have a wide dynamic range of mark to space ratios to RF input levels.

By means of an example, I'll be discussing in the next section how design choices in this area can have a large effect on performance.

Thirdly, a super-regenerative receiver can be made as sensitive as the noise floor level of the device will allow. It may appear that it can be as sensitive as you want it to be.

The fact is that a super regenerator does not 'capture' the entire incoming signal, but only samples a small part of it. In other words, the input RF signal is sampled for a very short time, whereas in a normal receiver the input signal is available all the time, and forms constructively in the tuned circuit.

Lastly, and because of the logarithmic performance nature of super-regenerative receivers, a small signal will produce about the same quieting levels as a loud one. A plot of signal-to-noise ratio against signal input will show a much flattened curve. This can make the 'perceived' sensitivity much less than it really is.

And bandwidth?

Bandwidth can be nominally defined as the corner frequency points at which some useful response – such as a signal level, or sound out from a loudspeaker – drops by a certain value, usually 3dB.

In the case of an instantly switched, fixed value conductance, g, the equivalent circuit for bandwidth calculation purposes is just that of a LC tuned circuit in parallel with the total conductance:

$$Q = \omega_o \frac{C}{g}$$

and,

$$bw = \frac{f_o}{Q} = \frac{g}{2\pi C} \qquad (4)$$

A more relevant situation is of course, when the conductance varies with time. One term from equation 3 will give the selectivity factor:

$$\exp\left[\frac{C(\omega-\omega_O)^2}{|g'|}\right]$$

where, $g' = \frac{dg}{dt}$ (at or around $t = 0$)

The resulting frequency response 'shape' is rounded, and not too different from a gaussian bell curve. You can derive a measure for bandwidth using:

$$bw = \frac{1}{\pi}\sqrt{\frac{g'}{C}}$$

This bandwidth is specified at one neper down – i.e. approximately 8.7dB. Note the dependence of bandwidth on conductance slope at $t=0$. This bandwidth is significantly narrower than that defined in equation 4.

In order to maximise bandwidth, a design goal is to make the conductance slope as shallow as possible at $t=0$. You have some control over this slope in that you can adjust the loop gain of the oscillator. Circuits with low loop gains produce shallow slopes; circuits with high loop gains produce sharp slopes. Unfortunately with simple transistor circuits, it is not always possible to arrange for them to perform the way we want, so compromises will be necessary.

Finally, I should say that the above analysis does not fully answer the bandwidth question. Super-regenerative receivers have a logarithmic amplitude response, and the arguments above should only apply to very low level signals. At higher RF input levels, the bandwidth response will be flattened, and the perceived bandwidth will be much wider.

Features of super-regeneration

I am not going to use the word 'problems' as this is not a fashionable word to use when talking about properties that do not work in our favour. Instead, I'll be positive, and refer to them as interesting challenges.

RF Radiation. The situation is simple: an antenna directly connected to an LC tank will radiate. This is because during those short moments after sampling, where the self-oscillations build up, the tank has considerable RF energy.

This energy will escape up into the antenna. In other words, the receiver is also a pulsed RF transmitter with narrow bursts of RF modulated at the quench repetition frequency. The radiated frequency spectrum will be spread over a bandwidth around w and at twice the quench frequency.

An obvious way to minimise radiation is to use a separate RF amplifier. Another effective way of capping the problem is to keep oscillation amplitudes and power consumption to a minimum. Most transistors will work quite happily with only microamps of current. Using higher quench frequencies, and keeping the oscillation bursts very narrow also helps.

Lack of frequency stability. Simple designs normally use single *LC* coils with inherent limitations of mechanical construction and manufacturing spreads. Stability can be sharply improved by using crystals, ceramic resonators, SAW or mechanical resonators.

Circuit stability. This is the most difficult design topic to address. Ideal super-regenerator performance needs critical negative-conductances requirements. This cannot be met by simple one-transistor circuits using components with normal tolerances such as resistors and capacitors.

In practice, circuits need to be individually tweaked or adjusted during manufacture for the right bias conditions. One possible solution is to use substrate level negative resistance generators – gyrators for example – that can be accurately controlled and programmed.

And now, some practical circuits

The circuit in **Fig. 4** is typical. This is an otherwise standard feedback oscillator that has been modified for self-quenching by the addition of L_2, R_1 and C_2.

Here is how it works: The base voltage is fixed at some value, around 2V in this case; and because of the decoupling capacitor, at ground potential for AC purposes: Ignore C_6 for the moment.

Lets start at time point t_1 in Fig. 3a), where collector self-oscillations have just started and are increasing exponentially with time. The RF voltage at the emitter is derived from this via capacitor voltage divider C_3, C_4. The clamping effect of the base-emitter junction also causes the DC average at the emitter to rise with time, resulting in the compound waveform shown in Fig. 3**b)**.

The transistor conducts only for a short time at the very peak of every negative cycle. This causes the pumping action at the collector, which in turn energises the *LC* tank with ever rising oscillations. Note that the waveform at Fig. 3**c)** is just a 'filtered' version of Fig. 3b) at the 'cold' end of L_2, and that Fig. 3b) minus Fig. 3c) equals Fig. 3a).

These self-oscillations cannot carry on rising forever. At some point, t_2 in Fig. 3a), the negative swings will cause the collector voltage to go below the base voltage causing the transistor to conduct in the reverse direction.

In a normal oscillator, this would cause signal into the emitter to be reduced, causing less collector current to flow, and making the whole system settle into a fixed amplitude level oscillator. However, the inclusion of L_2 and C_3 changes this drastically. As RF is fed from the collector to the emitter, the DC average at the emitter is slowly charging to ever increasing levels.

The transistor only conducts because each oscillation pulse cycle is always slightly larger than the previous, and enough to bring the emitter into conduction, slightly ahead of the DC rising level. This is between t_1 and t_2 in Fig. 3a). At the instant the collector oscillations start to clamp or level off at t_2, the base emitter junction stops conducting. This starts the stalling effect.

The higher than normal DC voltage at the emitter (about 3V in the circuit in question) ensures the transistor remains off from then on, and the oscillations stop completely. The transistor will remain completely switched off, and now can only wait for R_1/C_2 to discharge, between t_2 and t_4 in Fig. 3a). This discharge has to go down to the base voltage minus 0.6V to re-start conduction, i.e. to about 1.4V in this circuit.

Note that when the transistor switches off, the RF self-oscillations accumulated in the tank will also start to decay in their own time: This is between t_2 and t_3 in Fig. 3a). This time is much less than the relaxation discharge of the transistor, allowing the tank to discharge fully.

The real action happens now when the transistor begins to conduct again. This is now equivalent to $t=0$ in Fig. 1. At t_4 in Fig. 3a) – which is effectively the same as t_1, as this is a forever repeating cycle – the transistor slowly comes into conduction. As it comes into conduction, it starts presenting a negative resistance to the tank.

Self-oscillations will now start at a time depending on the tiny amount of signal RF present in the coil. You can see this on an oscilloscope connected to R_1. The time difference between t_1 and t_4 will be seen to vary depending on input signal level.

What do the components do? The base biasing resistor and decoupling capacitor are only there to provide a fixed bias: I'll cover C_6 later on. Anything from 1.5 to 3V will do.

Adjusting the base voltage is a convenient way of controlling quench rate, so you can replace the resistors with a 100kΩ trimmer potentiometer divider for development purposes.

Capacitors C_3 and C_4 form a voltage divider to feed a proportion of the collector signal back to the emitter. The values are more or less the same as would be used for an oscillator at the same frequency, but tweaked to keep loop gain as near unity as possible, consistent with reliable starts.

Capacitor C_3 should be made as small as possible; 2-10pF are common values for the 30-200MHz range. For the same reason, C_4 should be made as high as possible, and 30-50pF values are common.

Resistor R_1, with C_2, determines the quench rate which is given very approximately by $t=0.5R_1C_2$. This resistor also determines the quiescent DC current for the transistor, which needs to kept as low as possible. Values of 10-100kΩ, or even more, are common for low power solutions.

Typical values for C_2 at a quench frequency of 30kHz – useful for audio work – are between 2-5nF. The purpose of L_2 is to present a high impedance at the radio frequency and a low impedance at the quench frequency. Too large a value will cause extra ringing on the waveform at the R_1 junction. Too small a value will reduce the loop gain of the circuit. Typical values are 4-50μH.

I have made no mention of the tuning tank components or the transistor. The respective values will, of course, depend on the frequency you want to operate at. I normally use a BFY90 for development, as it is good up to 800MHz, it is relatively cheap, and has decent gain characteristics at the lower frequencies. For 418/433MHz work, the coil is a one inch diameter half turn loop, and the tuning capacitor a 3-5pF trimmer. Capacitor C_3 is 1pF, and C_4 about 5pF.

You can check that the radio works by connecting C_5 to an audio amplifier and adjusting bias and component values until you hear a loud hiss. Now, you only need a modulated signal generator as a RF source.

A spare key fob alarm or car door transmitter can be a good source of 418/433MHz signals. Be careful when using these, as some models use sequential encryption techniques which may render them out of sync with the car, if pressed too many times.

Detecting the output

At the junction of R_1, the signal is a ramp wave of about 1-2V in amplitude at the quench rate, Fig. 3c).

Depending on various base bias settings and component values, you will notice the dependency of quench frequency on RF signal level. Sometimes you will get a very small change in quench frequency with the signal, sometimes a much larger one.

Notice how the most sensitive point needs very critical adjustments. A simple low-pass filter could be used to recover the modulation. Alternatively, you could subject the signal to a digital amplitude threshold and process the pulse widths as numerical values.

Figures **4b)** and **4c)** show two seemingly similar variations on a simple one transistor 'detector amplifier' theme. The interesting part is that one circuit works and the other does not.

To understand why, go back to Fig. 3a), at time t_1. With the circuit in Fig. 4b), the quickly rising wave at the emitter has to charge not only C_2, but also C_5. This is because the n-p-n transistor Tr_2 is conducting all this time.

Such a situation is undesirable, as C_5 would be generally too large – of the order of microfarads. Similarly, as R_1/C_2 discharges, the extra combined effect of C_5 and the base 1MΩ resistor may drive Tr_2 into non conduction prematurely. Result? A pretty inefficient detector.

Consider now Fig. 4c. As the voltage in R_1/C_2 rises quickly, Tr_3 will be driven out of conduction, and C_5 is not included in the charging cycle. More importantly, as R_1/C_2 starts to discharge, C_5 will slowly be brought into conduction as Tr_3 starts to conduct. This has interesting effects.

The overall rate of decay of the emitter waveform slows down, and Tr_1 comes into conduction more slowly, resulting in a smoother introduction of negative resistance. (Remember when I mentioned the advantages of low slope negative conductance). Result; the circuit in Fig. 4c actually increases the sensitivity of the detector.

Many other tricks have been used to improve detector performance. The general aim is to control the loop gain in order to keep it as small as possible at the onset of oscillations.

Capacitor C_6 in Fig. 4a) is such an example. This is used to apply a small compensating 'pull' or bootstrap to the base voltage during the fast rise at t_1–t_2 in Fig. 3a). Other variations are possible.

In summary

The super-regenerative concept is certainly not dead. The principle is sound, although super-regeneration has had a bad image due to the use of simple designs and unstable components. By using more accurately defined components, and stable resonators, new circuits and applications are possible. ■

In a second article, Eddy looks at applications for super-regeneration.

The Author

Dr Eddy Insam is a consultant in innovative applications of telecommunications and specialises in graphics and signal processing. He can be reached on *edinsam@eix.co.uk.*

PCBs for Class G

Circuit boards for Doug Self's Class-G amplifier, detailed in the December 2001 and January 2002 issues, are available. These PCBs are double-sided with full solder masks and roller-tinning. Full component identifications are also included. Their size is approximately 190mm by 175mm each. To order a pair of these boards, send a cheque or postal order for £43.50 to Jackie Lowe, Class-G PCBs, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey SM3 8BZ. E-mail electronics.world@ntlworld.com for details of overseas postage. You can also fax your credit-card details – name and address of card holder and card type, number and expiry date – on 01782 878233 (+44 1782 878233). Please make cheques payable to *ELECTRONICS WORLD.*

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5. Dialogue For Three, Flute, Oboe and Clarinet, 1913
6. The Toymaker's Dream, Foxtrot, vocal, B.A. Rolfe and his orchestra, 1929
7 As I Sat Upon My Dear Old Mother's Knee, Will Oakland, 1913
8 Light As A Feather, Bells solo, Charles Daab with orchestra, 1912
9 On Her Pic-Pic-Piccolo, Billy Williams, 1913
10 Polka Des English's, Artist unknown, 1900
11 Somebody's Coming To My House, Walter Van Brunt, 1913
12 Bonny Scotland Medley, Xylophone solo, Charles Daab with orchestra, 1914
13 Doin' the Raccoon, Billy Murray, 1929
14 Luce Mia! Francesco Daddi, 1913
15 The Olio Minstrel, 2nd part, 1913
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17 Auf Dem Mississippi, Johann Strauss orchestra, 1913
18 I'm Looking For A Sweetheart And I Think You'll Do, Ada Jones & Billy Murray, 1913
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20 A Juanita, Abrego and Picazo, 1913
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Letters to the editor

Letters to "Electronics World" Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam Road, Surrey SM3 8BZ
e-mail j.lowe@cumulusmedia.co.uk using subject heading 'Letters'.

Homopolar reactions

In his article "Faraday's homopolar generator", Dom Di Mario makes a number of claims that I feel it necessary to take issue with. Homopolar machines are fascinating things, but their science is nowhere near as unexplained as he suggests.

Two of Di Mario's references cite information on the internet, at http://depalma.pair.com, relating to the late Bruce DePalma's work. I had a look at this site, and now quote from the home page:

"We designed this page to disseminate knowledge and accurate data relating to the Pre-Existent Primordial Field of the Universe – a sea of 'free energy' which permeates all. The N-1 Homopolar Generator – invented by Bruce DePalma – is an example of the type of device which is able to "plug in" to this Free Energy and eliminate the "need" for the continued use of fossil fuels and the consequent destruction of our only home – Earth. This device – and many others like it – have been proven to possess "over-unity" characteristics, ie: the power output is more than 100% of the input.

It is our hope that in the near future Free Energy will enable mankind to progress from a state of dependence to one of abundance."

For most *Electronics World* readers, I expect that says enough. The proponents of the miracle Faraday machine want us to believe that when the magnet rotates with the rotor, rather than being fixed to the stator, a homopolar machine can generate electrical power without the need for any torque to be applied to the rotor – other than that required to overcome the friction losses, mainly in the brushes.

The forces acting on a current carrying conductor in a magnetic field depend on the intensity of the current, the intensity of the magnetic field, and their directions. Movement of the magnetic field does not enter into it.

The strength and direction of the magnetic field are completely unaffected by rotation of the magnet, so rotation of the magnet cannot affect the forces acting on the conductors – regardless of whether or not rotation of the magnet results in rotation of the field. All this is conventional theory.

The voltage generated by the machine is the sum of its rotor and stator voltages. If the voltage generated in each part is calculated from the amount of flux cutting the conductors, the stator voltage depends directly on the difference between the angular velocities of the stator and the magnetic field. The same applies to the rotor voltage, except the polarity is reversed.

When the two voltages thus obtained are added to get the total, the voltage components in the two parts due to the angular velocity of the magnetic field itself cancel exactly. As a result, the angular velocity of the magnetic field has no effect and all that matters is the difference between the angular velocities of the rotor and stator. All this is conventional theory.

Rotation of the magnetic field does not result in any change to the flux density anywhere, or to the amount of flux passing

Amplifier output inductors

I have just got round to reading, somewhat belatedly, Doug Self's dissertation on 'Crosstalk in amplifier output inductors'. This arises from the coupling between the coils and he says "I think I'm pretty safe in saying there is no published work on this..."

There is a short paragraph in Morecroft's "Elements of Radio Communication" (Editions I & II). It discusses the problem in connection with triode radio frequency amplifiers. Not only does he show the same orientations of coils as Mr Self, but he also shows that there is a critical angle between coils, depending on their shape, which eliminates coupling.

I cannot confirm it but I think the suggestion may have been due to Hazeldine and was probably described in an early Proc. IRE.

Stan Brown
Via e-mail

The magnetic coupling between the coils of the several tuned circuits can be effectually eliminated by either of the arrangements indicated in Fig. 146. In the scheme shown in (*a*) the three coils, of approximately the form shown, can be arranged to have zero mutual inductance if their axes are placed at an angle of about 58° to the common base line. The exact angle will depend upon the form of coil used.

In the arrangement shown in (*b*) the three coils have their axes respectively at 90° to each other. Neither of these arrangements will eliminate the electric (capacitive) coupling between the different circuits. One method of eliminating this, at the same time as the magnetic coupling is eliminated without either of the schemes shown in Fig. 146, is to enclose the coils entirely in individual,

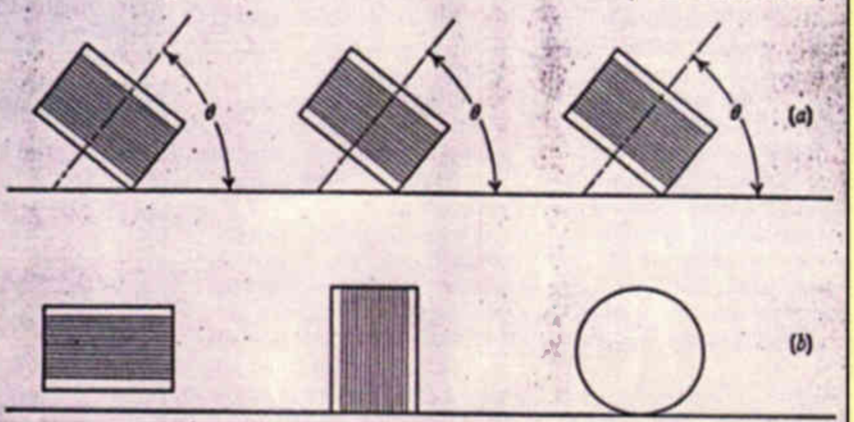


FIG. 146.—Two arrangements for eliminating mutual induction between three coils placed close to one another.

grounded, copper cans. These cans must not fit too closely around the coils as they would spoil the tuning qualities of the circuit if they did.

through any closed conducting loop, so it shouldn't be at all surprising that rotating the magnet makes no difference to the voltage generated by the machine.

If there were a difference, the amount of flux entering or leaving the loop via the rotor conductors would have to be different from the amount of flux leaving or entering the loop via the stator conductors. Again, all this is conventional theory.

However you look at it, conventional theory predicts that rotation or otherwise of the magnet in a homopolar machine is a complete non-issue. But Di Mario still manages to think that conventional theory can only explain the case where the magnet is fixed to the stator.

What he calls the "fundamental difference" between the Faraday and ordinary versions of the homopolar machine is actually nothing more than a gigantic red herring.

Di Mario also describes attempts to use a form of "magnetic screening" which could not possibly have the effect apparently intended – an "AC generator" which is in fact not a homopolar machine at all. It would be grossly inefficient due to eddy currents induced in the disc.

He also seems to think that the voltage generated is due to the non-uniformity of the magnetic field at the contact point. Rot, rot, rot, all of it.

Interestingly, Di Mario gets another thing wrong. If it really were true that a Faraday generator produces electricity without the need for any more input power than that needed to overcome friction losses, superconducting magnets and specially designed brushes would not be necessary to make a perpetual motion machine.

Larger homopolar machines than the ones he has been experimenting with can generate quite a few volts. Long before the arrival of cheap rectifiers they were used in the electroplating industry, where their low-voltage high-current outputs were well suited to the job, and their lack of commutator sparking was a useful advantage.

These machines were not Faraday machines, because they used excitation coils fixed to the stator, but that was only because it was the easiest place to put them.

Alan Robinson
Holgate
York

Fifteen years ago I was doing some research on homopolar machines and came across what at first glance appears to be a paradox involving the Faraday homopolar machine.

After some thought (more than I would care to admit) I came up with a method of viewing the machine which indicates that there is in fact no paradox and its behaviour

Probing thoughts

I feel we should all thank Cyril Bateman for his articles on scope probes, which are a long awaited replacement for the usual probe sets.

However, at the risk of starting an internecine conflict I also feel that we should thank Ian Hickman for bringing the MAX4005 to our attention in his May 1996 article. As a result of trying to use this with his 500MHz scope add-on (March 2000) we had an exchange of mail and I was quickly shown that the input capacitance of the scope is the problem.

The –3dB bandwidth is given by,

$$f_{(bw)} = \frac{1}{2\pi RC}$$

With a 75Ω termination at the scope source resistance 37.5Ω and an input capacitance of, say, 30pF this gives a bandwidth of 141MHz.

To improve on this it was suggested that rather than tapping the terminating resistance, which would further increase the probe loss, a current feedback op-amp with a gain of two could be used at the scope end to give a unity gain probe.

A search of the catalogues and perusal of many data sheets (may I suggest trying www.bgs.nu/sdw/a.html) convinced me this was easier said than done. One thing most of the sheets agree on is that driving capacitive loads results in frequency-response peaking and possible oscillation. The cure is a resistor in series with the output, which of course is just what we are trying to avoid!

Data sheets all seem to give slightly different information so it is not easy to make direct comparisons between op-amps. Most show a marked drop in bandwidth as the output voltage is increased and some show aberrations in the output as the rise time of the input decreases.

In the end I settled on the CLC449, which has a 1.1 GHz small-signal bandwidth, and Cyril Bateman's tried and tested circuit layout; despite the many recommendations in the data sheets.

I used Ian Hickman's input arrangement for the MAX4005 and a 27Ω resistor in the CLC449 output operating at x1 gain. The waveform from a 40MHz crystal oscillator module is as shown in the screen capture. This module is quoted as having a 6ns rise/fall time so the waveform seems reasonable.

It would be nice if someone could show how to improve the rise/fall times to 1ns or less for more rigorous testing though. Unfortunately I do not have the use of any high-quality test gear that would enable me to determine the bandwidth and frequency response of this probe arrangement.

I had planned to apply gain to the signal before inputting to the a-to-d converter but the output from the 500MHz adaptor has a lot of noise and as it stands this idea does not work well.

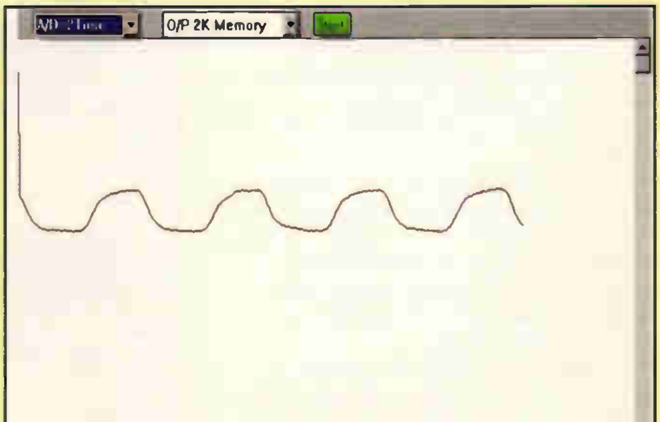
On the subject of noise, I should like to enquire if it is not the MAX4005 that determines the noise output of the probe as presumably this is amplified by the terminating gain stage. But I can find no figure quoted for this op-amp.

On the question of housing, I have found that 22mm copper tubing makes a good alternative. End caps are available and it is easy to work - photo attached.

Many thanks for an interesting series of articles even if I do have to increase my dose of Beta-blockers to deal with the surface mount devices!

John Kanaar
Via e-mail

Copper pipe makes a convenient probe housing.



Waveform from a 40MHz crystal oscillator using Hickman's input arrangement for a MAX4005.

is identical to a normal homopolar machine.

Pretend that you, the viewer are sitting on the rotating magnet. What do you see?

You see stationary flux lines extending vertically above you.

If you look over your head while you are rotating, however, you will see a conductor (the wire from the voltmeter or load) passing through the stationary magnetic filed. The Faraday voltage is generated in the wires connected to the load!

Any torque measured would affect these conductors. There is no mystery. The machine would behave identically to a normal homopolar machine.

As an aside, the US military had intense interest in homoplar machines.

If a homopolar generator is rotated at high speed and then suddenly stopped by a mechanical brake, huge currents may be placed through a load. These currents were to be used to vapourise a conducting fusible link, which would generate plasma that would accelerate a non-conducting projectile along a conducting rail to tremendous velocities.

In the vacuum of space these projectiles were to be used to shoot down enemy missiles and satellites.

Martyn McKinney
Seneca College
Toronto Canada

Superconductivity researchers at Philips in Eindhoven, working in the IIRC, developed a novel homopolar generator to test gauss limits.

The conceptual configuration was a lead disc, like a pizza, with a coil connected between the axis and the rim on one face. A rod magnet was 'stirred' around the axis on the other face, causing a monotonic increase in current and flux.

To shut down, they ran it in reverse until current was insignificant... No slip rings, no contacts, no intermittent resistances...

Topologically, it was but a toroid. The key seemed to lie in the change of topology around the 'legs', as seen by the current.

Nik Kelly
Via e-mail

Dom Di Mario's article 'Faraday's homopolar generator' in the October 2001 issue makes some claims that, for the sake of scientific rigor, merit further investigation.

When it is claimed that the measured voltage cannot be "generated in the wire connected to the brush", it is not sufficient to observe that "this happens regardless of its orientation, its size, whether it is reduced to a needle – even with the wire magnetically shielded, except for the contact point."

In fact a physicist could well complain that you shouldn't look at the voltage induced in each single wire, but rather at the voltage induced around the loop area embraced by both wires.

In other words, before discounting this simple and conventional-physics-complying explanation one should at least try these confirmation experiments:

- Repeat the experiment with the loop area reduced to the minimum feasible, by twisting the two wires together and, for the two short segments in proximity of the sliding contacts, running them very close and parallel to the disc surface (magnetically shielding each wire individually, as shown in Figs 4 and 5, cannot be effective as it does not screen said loop area). Verify whether the generated voltage is hence dramatically reduced.

- A second experiment could consist of allowing the voltmeter, sensing wires and contacts to rotate together with the magnet-disk assembly. Verify then that no voltage is generated (of course, measures will have to be taken to exclude the effect of the Earth's magnetic field on said loop area, which will result in an alternating voltage).

Only when both conventional-physics-complying explanations fail, shall one feel encouraged to seek more complex explanations.

Luca Ghislanzoni
Italy

Star-point grounding

I read the article 'Shielding and EMI' in the December issue of *Electronics World* with a feeling of growing disbelief. Here we are in the 21st century, and the concept of the star point ground is still being advocated.

This concept was in vogue at the time of valve radios. The valves were mounted on an aluminium chassis. To avoid ground loops, all the return conductors for the chassis-mounted components were taken to a single terminal bolted in the middle of the chassis.

It didn't work. There was a perceptible hum from the speaker, traceable to interference from the valve heater circuits. If the star point ground is ineffective at 50Hz, then what is its value at 50MHz?

To implement the star point ground, it is necessary to hypothesise that all signals are processed with reference to a terminal, which is assigned a name such as 'zero volt reference'. There is no mention of this concept in any textbook on electromagnetic theory.

The idea for the star point comes from the desire to avoid ground loops in the circuit – a desire based on an irrational concern.

It is standard practice in the design of electronic instruments to enclose the circuitry in a conducting metal case, and to carry signals from one instrument to the other using coaxial cable. The screen of the cable is firmly connected to the metal case of each instrument.

In a set-up where there are several items of equipment, interconnected by a network of cables, it is inevitable that there will be many ground loops, caused by conducting paths along cable screens and instrument cases. The fact that such a practice provides a high degree of immunity to interference should be sufficient to allay any fear of ground loops.

The reason is not difficult to explain.

Skin effect plays an important role in separating the desired signal from interference. Most of the return current in the coaxial cable flows on the inner surface of the screen.

Any transient interference caused by an external field will cause current to flow on the outer surface of the screen. This current flows in what is effectively a ground loop. If the cables are widely spaced, the inductance of this loop is high, and so the induced current will be relatively low.

Since the only coupling between signal and interference is caused by current in the resistance of the screen, since the resistance of the screen is relatively low, and since the current is relatively low, then the voltage induced by external interference in series with the screen will also be low. A reasonably high value of common-mode rejection is achieved.

Inside any instrument, most of the circuitry is mounted on printed circuit boards. Most of the return current of any signal flows in the ground plane immediately adjacent to the signal track. There is some stray field, and this is 'mopped up' by the creation of transient currents in the inner surface of the instrument case. These currents flow in a loop – a ground loop.

The existence of ground loops, both inside and outside the instrument cases, helps to minimise interference.

The star point concept is a thoroughly bad idea, and is based on a needless concern. If implemented, it will guarantee a set of intractable interference problems.

Ian Darney
Bristol

Train safety

I refer to the article on Train Safety by Wilf James on page 832 in the November 2001 issue, and I would like to offer the following comments.

While Mr James, as an outsider to the industry, has some valuable insight into the problem, there are statements in his article that deviate from the facts. I must declare I am in no way criticising the author in person, but facts must be set forth so that readers are given the opportunity to judge for themselves.

Mr James suggests that trains stop slowly, in the order of 0.3m/s/s (I'll be using the metric system throughout this letter). This, I quote "uses 19th century technology". I believe this is plainly incorrect.

In fact, the law of physics rooted much much earlier than 19th century, of which both the train and a long distance coach are one of its many applications. I agree that a bus can stop closer to 0.8g (approximately 7.8m/s/s), but at this braking rate, passengers without a seat belt would be thrown through the windscreen.

Train passengers do not usually have the luxury of a restrainer – in a crowded commuter train, there are lots of people standing. The train uses a much lower rate of deceleration for the comfort of the travellers instead of an older generation technology.

The modern electric brakes and disc brakes on a high-speed electrical multiple unit (EMU) are designed to brake slowly, although they can do much better.

Furthermore, trains runs on a steel-to-steel interface – wheels on rail – so the adhesion is much much lower than that of rubber tyres on asphalt.

Finally, I think it is comparing apples to oranges rating an over 500-tonne train against a 10-tonne coach.

Mr James is very insightful in his "nine-camera" system to check obstacles ahead of a train's path. However, this will only transfer the burden to the train driver, which is the exact opposite of what an automatic train control system will do.

In many studies on ergonomics (Railtrack and London Underground have commissioned similar studies), the same conclusion has been drawn. The concept was to relieve the train driver from decisions that would affect many people. Instead, such decisions are made by 'systems'. Humans, unfortunately, tend to make more mistakes under pressure than computers.

Installing nine monitors in the driving cab is, in my humble opinion, ergonomically flawed. Would the train driver's primary focus be on the screens instead of on the tracks ahead?

The transmission of video through the GSM network would create another problem: security. There have been reports on hacking into the FBI network, I believe it is much much easier on GSM, even with additional security measures. Imagine what might happen if some teenage hacker replaced the dynamic screen with nine completely unoccupied tracks.

Another ergonomics issue is the repetitive action – every 20 seconds, as Mr James suggested – of cancelling the traction hold-off. This is exactly the problem with the Advanced Warning System (AWS).

I suggest that Mr James refers to the Honourable Sir Anthony Hidden QC's inquiry into the Clapham Junction collision, in which he clearly and repeatedly suggests that Automatic Train Protection (ATP) must be implemented within five years. That was in 1988. One can only argue that the then British Rail, and now Railtrack, have been too slow to act, but the direction is clear.

I understand that spending £3bn is a high price, but if one put Mr James' proposed solution with the appropriate security and safety measures into practice, the price would not be far off. Being a railway practitioner myself, I owe the public a duty of care to explain the subtleties with which Mr James may have unintentionally misled the readers.

H Cheung
Hong Kong

Engineering esteem

With reference to Lawrence Jones's 'Comment', in the March 2002 issue, if he was a doctor in South America he would be called Doctor Jones, if a lawyer he would be called Abogado Jones, and as an engineer, Ingeniero Jones. It's probably much the same in many other countries. Which again serves to illustrate the lack of esteem with which engineers are regarded here.

David Bridgen
Camberley

For a stronger DAB signal...

On reading Richard Barnes' letter 'DAB debate' in the February 2002 issue, I wondered if he bought the same make of five-element DAB antenna as me?

My Videologic tuner reported six sixteenths of full-scale signal strength even though I can almost see the Hannington antenna. While looking for the coordinates of other transmitter sites I noticed that DAB signals are vertically polarised. My antenna was waterproof only when used horizontally.

I exchanged it for a single-element vertical antenna and I can now receive many more channels – mostly at thirteen sixteenths full scale.

P. Robinson
Via e-mail

Quiet Outlook

I've been using Microsoft Outlook' as an appointments reminder/calendar for about three years. It has an 'alarm' built in to it which is extremely useful, but only gives an on-screen indication of an imminent appointment. And, since I don't spend my whole life in front of the screen, an external alarm would help.

I'd find the ability to output the 'alarm' to a serial or USB port really useful, as I could then drive external logic with it. I'm a 'hardware man', so a few chips, a bit of PCB and I'm away! But Microsoft tell me it can't be done, so that must be true, mustn't it?

Can any of you tell just how I can get at the guts of the 'Outlook' program so that I can prove them wrong?

Trevor Wiltshire
Via e-mail

RIAA equalisation

Designing low-noise RIAA amplifiers for moving-magnet cartridges is an interesting noise optimisation problem. It comes down to minimising the total noise level over a wide band with a frequency-dependent source impedance, a frequency-dependent amplifier gain and a frequency-dependent response of the ears of the listener.

Some of the people who responded to Dr Thagard's article about a RIAA amplifier simplified their equations by calculating at a single mid-range frequency, but none of them seems to know what frequency gives the most representative result.

I recommend 3852Hz. The proof is rather too long to put in this letter, but it can be shown that optimising the RIAA- and A-weighted integral noise of a RIAA amplifier is equivalent to optimising the spot noise at about 3852Hz, provided that the following assumptions are valid:

- The magnitude of the source (cartridge) impedance can be modelled with a simple L-R series network.
- The noise of the amplifier itself can be modelled with two noise sources, a white equivalent input noise voltage source in series with its input and an uncorrelated white equivalent input noise current source shunted across its input.

These assumptions are usually close enough to the truth to make the 3852Hz rule a useful approximation.

For a typical moving-magnet cartridge with 0.5H inductance, the effective source impedance is approximately 12kΩ. The noise current produced by a 47kΩ shunt resistor then adds about 7nV/√Hz to the total noise.

Cartridges produce more thermal noise than could be expected from their DC resistances, because their effective series resistances increase substantially over the audio frequency range.

My former colleague Richard Visee made measurements of a Shure V15 III cartridge several years ago, and found that the ESR increased from 1338.8Ω at 0Hz to about 30.1kΩ at 20kHz. For this particular cartridge, the weighted average thermal noise of the cartridge itself and of the 47kΩ resistor were roughly equal. If one is foolish enough to actually play records on one's record player, the noise coming out of the cartridge

MOSFET power

As noted by Dr White[1], over the years there has been a number of designs based on the Hitachi MOSFET power amplifier, and I wonder about the rigour of the design process used for many of them – including the original.

For example, I note that the Maplin version applies Miller compensation to both halves of the second long-tail pair, whereas the original applied it only to one half. Another published design omitted it entirely – inviting the risk of fierce driver stage oscillation at 8 to 12MHz – no names, no pack drill!

It is interesting to note that the very similar all-FET design[2] that accompanied the Hitachi original had compensation on both halves of the driver, but of unequal value – *why*?

Dr White asks the question "can the design be enhanced": I think the problem is that there is such a wealth of possibilities... but he has chosen the area where the design really feels a bit too lean. I wish he had gone further.

In particular, the absence of emitter degeneration resistors on the input pair, results in a linear range of approximately 50mV ($=2V_T$). This is somewhat at odds with the quoted input sensitivity of 1.28V rms in the modified amplifier. I would have expected some degeneration to be provided, to guard against the possibility of transient overload of the input stage[3].

Of course the anonymous writer of the application note had only to worry about the static distortion reading, which would no doubt have worsened somewhat, had degeneration been provided! It is hazardous to compare dissimilar designs, but in my experience, a very different design with similar static distortion performance[4] is audibly cleaner than the Hitachi original. I suspect that this is due in part to the exemplary dynamic overload performance of the Blomley design.

As regards current sources, it may be appropriate to consider an alternative, which deserves to be better known,[5,6] **Fig. 1**. According to Baxandall this offers a 50MΩ dynamic resistance at 1mA, but more importantly "The effect of collector-base capacitance in Tr_1, which shunts the output [in a simple constant current source] is degenerated... and output capacitance values of well under 1pF are obtained."

The significance of this is that the reactance modulator effects described by Cherry[7] can be kept to a very low level. Furthermore, complementary versions of the circuit may be connected together to form a genuine two-terminal device, **Fig. 2**.

It might be worth remarking that application notes vary a lot in quality. In most cases though, the objective is to open up a market by demonstration of a device's capabilities.

Inclusion of features that any sensible user would take for granted is secondary. It follows that the most worthwhile modifications are additions. One particular addition that is missing from most versions of this circuit is comprehensive overload protection.

Remarkable faith in the ruggedness of MOSFETs does nothing to protect either the amplifier or the speakers. Depending on the rating of the fuse, fusing the output line will result in either compressed dynamics (albeit at a low steady state distortion level – if the fuse is within the feedback loop!), or inadequate protection. Of course you have to be clear on whether it is the amplifier which is to be protected, or the speakers, or both.

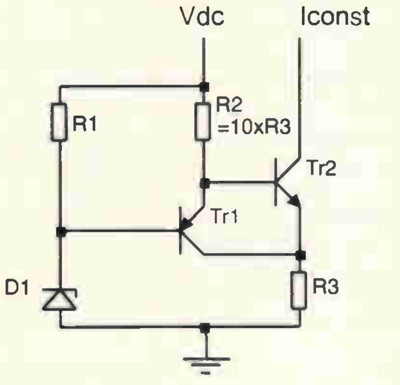


Fig. 1. Alternative current source. In this version, the effect of collector-base capacitance in Tr_1 is degenerated.

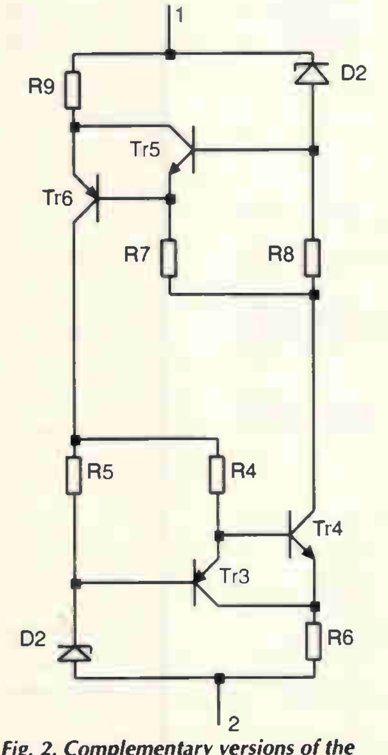


Fig. 2. Complementary versions of the circuit of Fig. 1 may be connected together to form a genuine two-terminal current source.

Equipment reviews have thrown up numerous examples of protection, which has affected the sound quality of commercially produced amplifiers; protection is not just a simple addition, but an integral part of any finished amplifier design.

For my part I would not consider this design for a new project, because of the somewhat rudimentary output stage design, which limits the value of improvements elsewhere. Other output stage designs[4, 8, 9, 10, 11] are capable of achieving open-loop distortion well below 1% at similar overall cost, and would therefore be preferable.

In summary, thank-you for a worthwhile attempt to re-shoe an old war horse, which I feel owes its popularity mainly to the relative ease of uprating it, rather than to anything very special about the sound quality of the basic design. A little work may well revitalise an amplifier that has been languishing in the loft for the last 10 years, because despite the detailed criticisms above, the original design compromises seem to have stood the test of time.

Edward Kessler
Via e-mail

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increases even more.

I intend to write an article about two RIAA amplifiers I designed and built, both featuring the 'electronic cooling' (double feedback loop) technique, and about the derivation of the 3852Hz-rule. I hope that it will be accepted.

Marcel van de Gevel
Harlem
The Netherlands

I think that Mr Underwood, in his February 2002 issue letter, has confused equalisation with tone control. Both carry out the same function of adjusting the frequency response. In general, they both use similar circuit configurations.

For this discussion I define equalisation as fixed frequency adjustment and tone control as variable frequency adjustment.

The RIAA equalisation curves – and the equivalent IEC-60098 published in Europe – were devised to overcome two problems in the recording medium. These were to preserve groove integrity at low frequencies and to overcome noise at the higher frequencies.

The curves have remained unchanged since first issued in the early sixties – apart from one change at the extreme low frequency end to IEC-60098 only. As a result, it is fair to assume that the curves are about right and should not be altered.

Incidentally, IEC-60098 gives the tolerance for the recording curve as ± 2dB of the appropriate characteristic. It specifically excludes any tolerance on the replay characteristic adding the comment that, 'record reproducers usually contain a tone control enabling the frequency characteristic to be varied over a wide range'.

However, in audio recording nothing is ever perfect. The frequency adjustment that Mr Underwood observed seems a very reasonable action to take to overcome acoustic problems or minor imperfections in the recording chain at the recording venue.

At the replay end, the average living room provides a less than ideal listening environment and the listener would probably make further tonal adjustments when playing the disc at home.

What is most sad is the number of musicians and vocalists who now feel the need to use sound enforcement during their performances. Much modern music is generated electronically so that it can only be heard via the imperfections of a speaker system. These two facts seriously reduce the opportunities to hear truly live music so the number of engineers who have a wide experience of what music actually sounds like must be diminished.

I suspect that the 'golden-eared' engineers referred to by Mr Underwood are a dying breed.

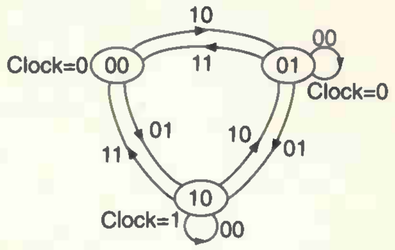
Chris Gardiner
Chelmsford
Essex

Speed with direction

In the January 2002 Circuit Idea for a speed and direction indicator, the circuit diagram should use NOR gates as indicated in the text. The state diagram is as shown.

Also note that jitter of the X input with Y high will produce output pulses all indicating the same direction. ■

John Crabtree
Via e-mail



Excitation diagram for the speed and direction

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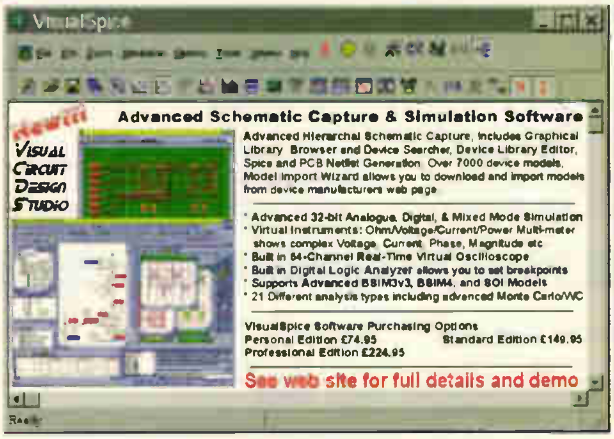

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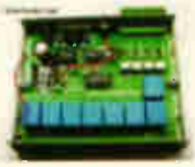

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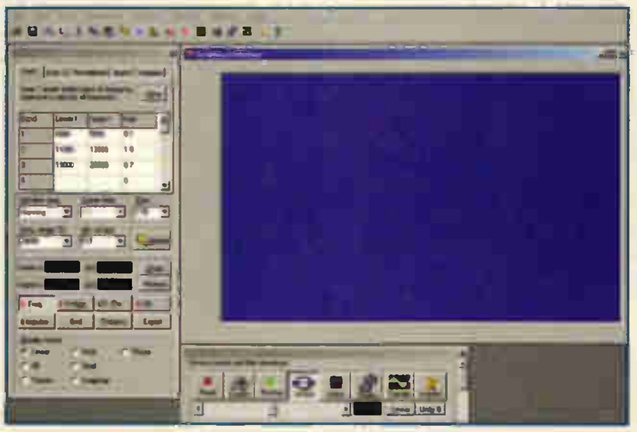
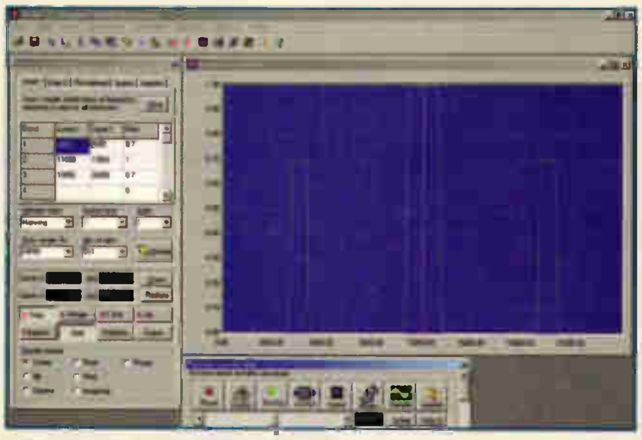
The RTDF filter system includes hardware based on an advanced digital signal processor, low-level firmware that implements the filtering operations, and a high-level PC-based software interface that designs the filter according to your requirements. Once a filter is designed, the software interface is used to download the filter to the hardware system via a serial link, where it is executed on demand.

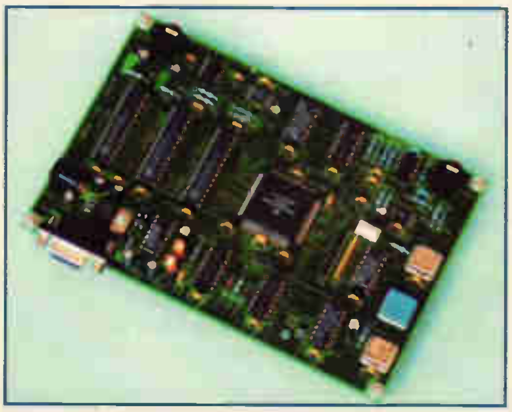
You don't need to know about digital signal processing theory or the mathematics associated with digital-filter design. But if you're a filter expert, you won't find yourself restricted by



Step 1 – Enter the characteristics of your filter, such as cut-off points and sharpness, in the Filter Design Interface Window on the left of the display.



Step 2 – Click on the Update button to view your result in graphical form, download it to the hardware module and run the filter from the hardware control panel – done!



Filter hardware: a DSP56002 board with RAM and 16/18bit oversampling analogue converters for stereo i/o at up to 48kHz sampling.

RTDF - key features

- Runs under Windows 95, 98 or ME
- Generates FIR filters with a maximum of 1024 coefficients.
- Multiple pass, stop or arbitrary filters.
- Lower –3dB frequency 3.7Hz at 48kHz sample rate and 1.2Hz at 12kHz sample rate.
- Filter operates in single or dual channel modes.
- Import mode – ASCII import of any frequency response.
- Hardware module holds up to 16 filters, instantly selectable with one mouse click.
- Zero-phase distortion in the pass, transition and stop bands, ignoring input and output coupling.
- Choice of rectangular, Bartlett, Hamming, Hanning, Blackman or Kaiser Windows.
- Virtual control panel allowing run-time changes to filter gain and sampling rate.
- Includes frequency and time domain plots of filter performance.
- Frequency response plotted as linear, dB, square, root, real, imaginary or phase.
- Impulse, frequency and phase response exportable in a variety of formats (dB, power etc) as ASCII files for incorporation into standard spreadsheets.
- 18-bit resolution in single, 16-bit in dual-channel mode.
- Normal or turbo speed, software selectable.
- User selectable sample rates of 48kHz, 24kHz, 16kHz, 12kHz, 9.6kHz, 8kHz, 6kHz, 4.8kHz, 4kHz, 3.2kHz or 3kHz.
- Maximum input and output level 4V pk-pk

For more information, visit: http://www.umist.ac.uk/dias/pag/rtdf.htm

the easy-to-use interface. If you want to do it the hard way, you can even design your filter in long-hand then download the filter's frequency response as an ASCII file to the RTDF's control program!

The RTDF is a total filter solution. Due to its flexibility, it is particularly well suited to processing audio signals in real time. High-quality analogue signal conditioning and a dual-channel 16/18-bit resolution analogue-to-digital converter and digital-to-analogue converter provide a resolution sufficient for the most demanding applications.

In short, the RTDF brings the power of digital signal processing to any audio-bandwidth domain that requires high-performance electronic signal filtering. Applications include sensor linearisation, audio signal processing, signal analysis, vibration analysis, education and research in electrical, electronic and other physical sciences.

Low-pass, high-pass, multiple band-stop / band-pass filters may be combined to produce very complex filters for frequencies up to 24kHz.

The software can accept measured responses to define a filter template. This can be used for measurement equalisation or to search out signal signatures in noisy environments.

Since the filters are implemented using a symmetrical finite impulse response (FIR) method, no phase distortion occurs in the filtered signal – no matter how sharp the filter is. Because the processing module is so fast, it is possible to design filters with responses far beyond what is possible with traditional analogue techniques.

Windows software

The control program runs under Windows and provides a user-friendly filter-design tool that de-mystifies the process of specifying the filter. The filter design process simply becomes one of describing the desired frequency response.

The design package indicates the response that will be produced and any deviations from that specified. User designs may be stored for re-use and actual responses may be entered from measurements for simulation or equalisation purposes. Once designed, filters are calculated and downloaded to the hardware within seconds.

The software designs the filter according to the user's specifications. The filter is expressed as a set of FIR, zero-phase distortion coefficients or taps. Collectively, the coefficients of a filter are known as the impulse response.

The system's gain and sampling rate can be adjusted while the filter is running and you can switch instantaneously between a filtered or non-filtered signal.

Most importantly, the software includes a flexible filter design section. A window displays the frequency or impulse response of the realised filter in various formats. A hardware control section downloads a filter and provides for general communication with the filter module.

The DSP module

The hardware DSP module connects to the PC via a standard COM port using the cable supplied.

The 16/18-bit over-sampling dual-channel sigma-delta a-to-d and d-to-a conversion system can easily be set to any one of eleven sample rates, ranging from 48kHz down to 3kHz.

So how fast can the filter operate, and how many filter coefficients can it practically employ? The maximum number of taps at the highest frequency range of 24kHz, in single channel mode, is 527. At this range, the system is sampling at 48kHz. This represents a very sharp filter indeed.

Using a frequency range of 12kHz – sample rate of 24kHz – in single-channel mode, the system can operate a filter with a maximum of 937 taps. At any range below this, it can operate a filter with a maximum of 1024 taps. The performance of a 1024-tap filter is so extremely sharp that it is quite unlikely that you would ever need to use it.

In dual-channel mode, the maximum number of taps permissable at the highest frequency range of 24kHz is 191. With a frequency range of 12kHz , it is 397. Full details concerning tap numbers, frequency ranges and operating modes are provided with the on-line documentation. ■

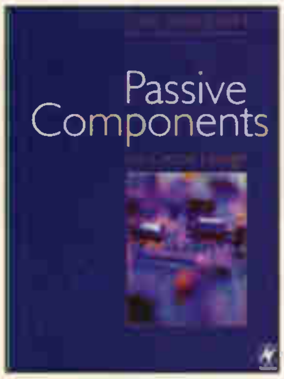
Minimum system requirements

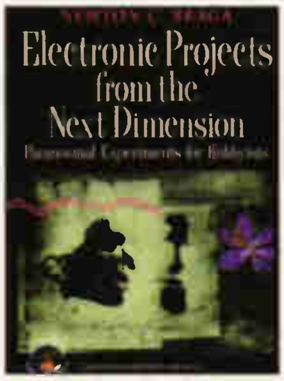
100 MHz PC running Windows' 95, 98 orME
SVGA display (800X600 pixels) or higher
CD ROM drive
10 Mbytes of free hard disc space
Serial port, capable of 19200 bit/s

System includes:

- Filter DSP board
- Windows filter design software on CD plus demonstration-filters
- Fully-worked help files – featuring tutorial
- Installation instructions
- Analogue i/o cables
- RS232 COM port download cable
- Power supply

BOOKS TO BUY





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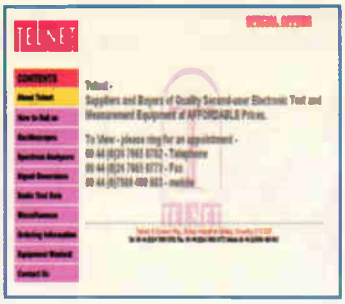

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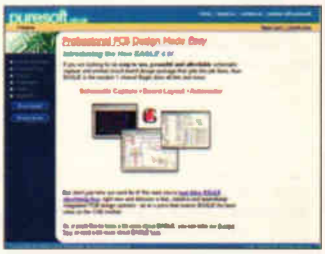

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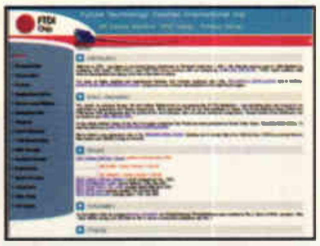

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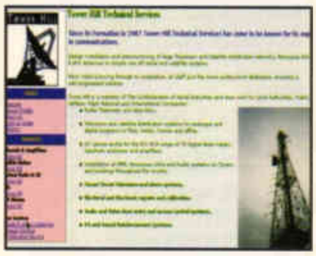

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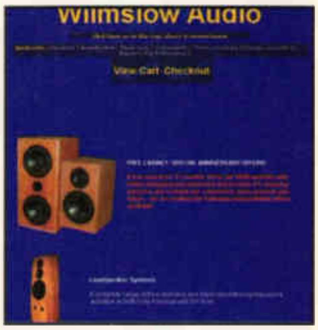

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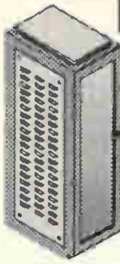

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