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Available commercially as the VRS 8x4, **Emil Vladkov's**, audio and video router system allows you to select one of the eight video inputs and one of eight stereo audio channels using either a keypad or a PC.

432 TECHNOLOGY THAT'S ALL TALK

After decades of research into speech



recognition, the technology still seems to have very few serious applications. **David Manners** reports.

434 PC INTERFACING VIA ETHERNET II

In this second article on connecting external devices to your PC via a cheap Ethernet card, **Eddy Insam** describes a development device that can get you going in no time at all.

442 DESIGNING WITH DSP

Using a design example incorporating audio data converters, **Patrick Gaydecki** reveals that the DSP hardware needed for communicating with a PC, memory addressing and codec interfacing is not too difficult to implement.

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- Detect 3-phase faults
- PC-based step motor driver
- Single IC oscillator/doubler
- Load-sensing and switching

EXCLUSIVE OFFER - PAGE 472

This is believed to be the lowest cost smart-card reader/writer and SIM adaptor kit available - at just over £20 inclusive to EW readers.



473 100KHZ IMPEDANCE METER

Designed for checking electrolytic capacitor health, **Cyril Bateman's** impedance meter has a variety of other uses.

480 WEB DIRECTIONS

Useful web addresses for the electronics designer.

483 LC RESONANT TANKS

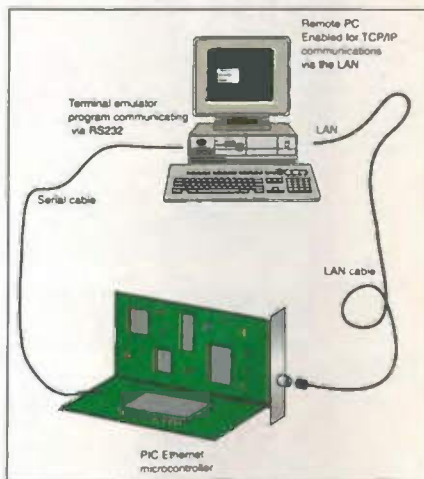
Joe Carr takes a look at how LC tank circuits work and describes how to get the best from them in radio designs.



Illustration: Hashim Akib



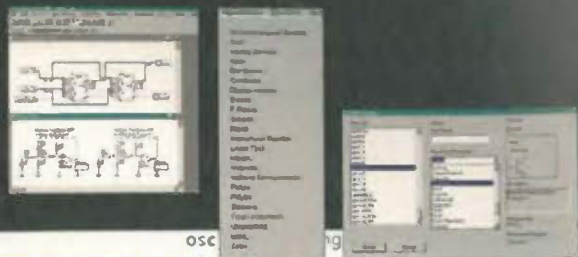
Capacitor guru **Cyril Bateman** has designed this 100kHz impedance meter with two ranges - 0-2Ω and 0-20Ω. Read about it on page 473.



A 10Mbit/s Ethernet card is a sophisticated interface, yet you can buy one for under a tenner thanks to mass production. On page 434, **Eddy Insam** describes how you can use them for real-world interfacing.

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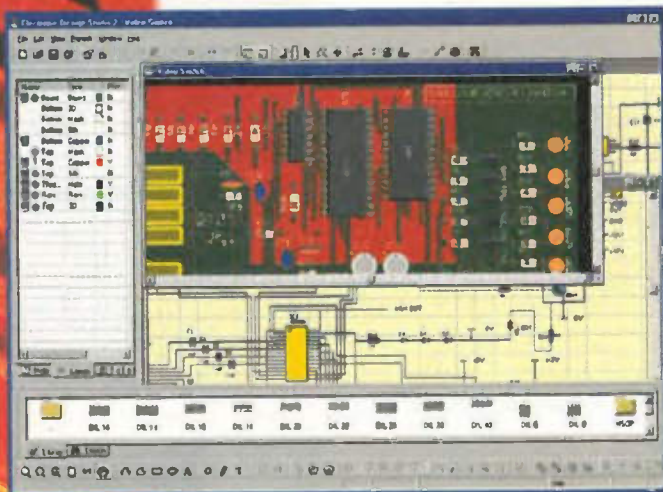
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Taxing times for the contractor

Very few tears are shed when IT contractors hit hard times. They are always the butt of jokes at work. There is often a simmering resentment between full-time employees and contractors who are paid twice as much and who have worked at the company for as many years. Their quality of work is generally low because of a lack of commitment – they are a pretty useless bunch really!

Managers take a different view though. Contractors can offer specialist services not immediately available within the company. They possess a flexibility not possible with full time employees. They require only minimal supervision and often produce more than regular employees.

They can also be hired for other purposes – use of contractors in Government Departments is a convenient way to keep the number of civil servants down while getting more work done. The down side is that a contractor can become very valuable and will try to hold the company to ransom.

If you are technically minded, contracting can be an attractive career because of the high wages, the freedom, the variety of work and the absence of politics in getting the job done. When companies employ contractors, the work is often better defined and more focused on results.

The down side is the need to keep one's skills up to date, ensuring that you have a good agent who pays you when he should, providing for sickness and pension and too much travelling. And there's the uncertainty, particularly in times of economic slowdown.

Unfortunately, the tax man has now added to that uncertainty. Rules were announced, in the 1999 Budget press release number IR35, to close a loophole in the national insurance rules for contractors.

The previous scenario ran like this: a contractor sets up a limited company of which he or she would become an employee. That company would then offer its services to another company and be paid a handsome daily rate to work at that company's premises.

The contractor and fellow directors – husband, wife, partner and the like – would draw a minimum salary on which national insurance was payable, but then take a big dividend every quarter, which was not subject to national insurance. Expenses could also be offset against company profits thus reducing the tax bill. Quite neat really.

IR35 identified these contractors as “disguised employees” and said that they would in future have to be taxed at source and expenses would be limited to 5% (www.inlandrevenue.gov.uk/ir35/).

In the present climate, where it is politically incorrect to increase direct taxation, the Inland Revenue has been told to squeeze everyone as tightly as possible and to apply the rules rigidly. These new rules have created considerable uncertainty for both contractors and companies.

The interpretation and advice from the Inland Revenue is so woolly that companies are afraid that they will face heavy fines. It has also created genuine hardship by the sudden imposition of IR35. For example, some contractors travel great distances to work or have to stop over locally – which is frighteningly expensive in London. These genuine expenses will no longer be offset against tax.

The Professional Contractors Association (www.pcggroup.org.uk) was formed in May 1999 to fight IR35. They recently took the Inland Revenue to the High Court in an attempt to prove that IR35 contravenes EC and human rights laws.

More down to earth, it was said to restrict free movement and unfairly help large companies that do contract work but employ people directly. The judge ruled that IR35 was “unfair and unworkable” but not illegal.

The Inland Revenue has been told to redraft the guidance manual used by officials to assess who falls under IR35. This is a small victory that will ease the pain but not end the “suffering”.

So what of the future for contracting? There have been the dire warnings that the UK will cease to be competitive in e-business, that contractors will move overseas etc, etc. In fact the impact of IR35 will probably be minimal.

Rates will increase – not immediately because of the economic slowdown – and some contractors will throw in the towel. If you are being taxed at source you may as well become an employee to enjoy the protection which that gives, and the lack of hassle from the tax man (and why is the tax system so complex?). The drop in salary could be compensated for by good company benefits such as a bonus scheme or share options.

There will continue to be a need for highly-skilled specialist contractors in the fast moving and risky high-tech world. Maybe the industry needed a shake out before Asian software competition really starts to bite.

Peter Marlow

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DERA breaks photon-level secure communications record

A team of UK scientists from the Defence Evaluation and Research Agency (DERA) is claiming a world record for transmitting quantum cryptography keys using a laser link. The keys are sent via free-space

quantum transmission.

"We transmitted successfully over 1.9km. That is 0.3km further than the previous best published result," said Professor John Rarity. The team working on the project are based at DERA's research site in Malvern.

Under 2km may not sound very far, but when you realise that "the demonstration experiment went down to 0.1 photon/pulse," according to Rarity, you get some idea of the achievement. Only one in ten pulses leaving the transmitter contains a photon.

Quantum cryptography key transmission relies on sending individual photons across a link with information encoded on their polarisation states.

Undetected eavesdropping is almost impossible because the photon disappears in the detection process.

Feedback by conventional means

DERA scientists in the lab. Pictured are (l to r): Phil Gorman, Professor John Rarity and Dr Paul Tapster. At the end of the telescope is the Wyche Inn – the pub that holds the transmission end of the new communication link.

tells the sender at what time a successfully detected photon arrived, then both sender and receiver can add this to a list of key bits that have completed the crossing.

The next step is, under the auspices of the European Quantum Cryptography Collaboration, to transmit over 20km between two mountains in the Alps.

The move to the Alps is to remove optical alignment problems in the experiment caused by turbulence in the air. "Turbulence wobbles the beam," said Rarity. "Britain is particularly bad, the Alps are better." Los Alamos, where the previous record was set, is in the desert and has one tenth of the turbulence typically found around Malvern, he said.

"The final aim, over the next four or five years, is to develop a key exchange between here and Los Alamos in the USA," said Rarity.

This future link will involve free-space transmission between Earth and satellites, and between satellites. It will be used for secure Government communication between the two countries.

Building to building communication links could be another development.

Steve Bush *Electronics Weekly*

UK company wins \$9m to develop a new generation of optical fibre

BlazePhotonics, a spin-off of the University of Bath, has won \$9m in first-round funding to develop photonic crystal fibre – also known as 'holey' fibre.

"Our fibres are going places," said Professor Philip Russell, who leads Bath's optoelectronics group. "Their properties give them the potential to out-perform and replace conventional fibres in a variety of applications. I am delighted that we now have the backing to build

BlazePhotonics."

The company will use the funding to establish a custom built facility and increase its R&D and operational teams. The operations will remain close to Bath and will maintain links with the University.

As part of the deal, the University will be a significant shareholder and will vest all of its existing and future intellectual property arising from research in photonic crystal fibres in BlazePhotonics.

Funding was led by Celtic House Investment Partners and included Quester Capital Management.

Pre-incorporation seed cash came from Sulis Seedcorn Fund.

Photonic Crystal Fibres are single-material all-silica optical fibres. Instead of the conventional core-cladding structure, light is trapped in the core by an array of tiny air holes running along the whole length of the fibre.

Steve Bush *Electronics Weekly*

New technology for home networking at 1.5Mbit/s over existing mains wiring

UK firm nSine has unveiled its powerline technology, providing 1.5Mbit/s network links through existing mains cables.

The company is aiming its nPlug technology at home and small office networking.

"The technology's been under development for three years now," says John Gilhooly, business development manager at nSine.

nPlug includes a digital baseband

controller, an analogue interface chip, and an AC coupler for connecting to the mains. The digital controller will be sold as a standalone chip or licensed as a core.

The system modulates data on to a carrier of between 10 and 30MHz.

"A low injected signal strength reduces the chances of that signal affecting other systems," said Gilhooly.

In order to ensure a robust system,

data is sent over multiple channels.

"You have various methods of ensuring robustness. We use multi-tone," said Gilhooly. "The early versions will send the same data down all four channels."

This is done because impulse noise on the mains can easily destroy the data on a channel.

Methods of avoiding this, such as forward error correction, are not used because of the cost implications.

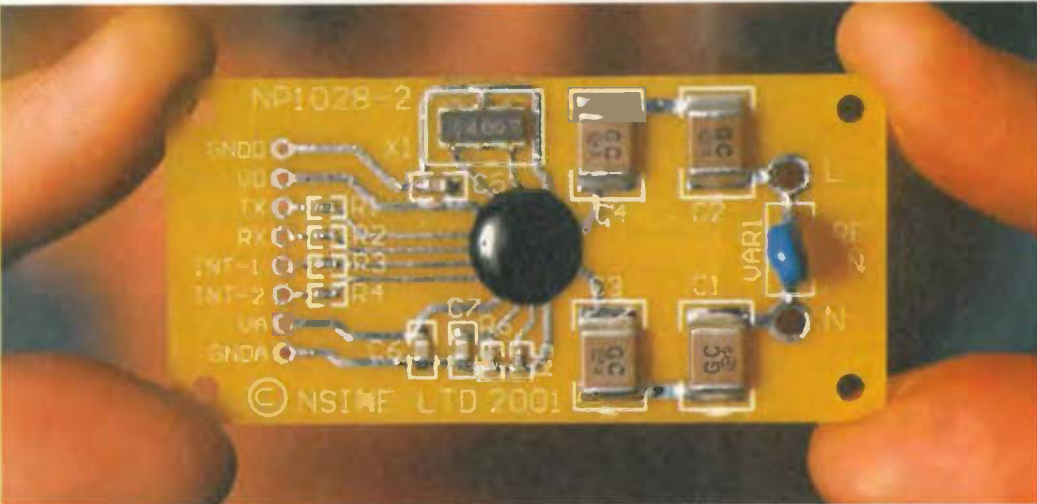
"The focus here is low cost," said Gilhooly.

nSine estimates that a complete network node – including digital baseband controller, analogue interface and main coupler – will cost \$5.

Devices that could attach to the network include PCs, printers, audio and video systems, white goods and Internet gateways.

Four reference designs are available. Two link PCs and peripherals to the network via either a USB or PCI-bus. Third is an ADSL gateway, while the last is a standard telephone connection.

Future versions of nPlug will offer increased performance of up to 40Mbit/s, the firm said.



Report highlights health concerns over Tetra mobile comms

Worries over the health effects of GSM mobile phones could be eclipsed by new concerns about the digital packet mobile communications system known as Tetra.

These new mobile phone health concerns have been highlighted in a report on the siting of mobile phone masts by the House of Commons Trade and Industry Committee.

While welcoming the review, the cross-party Committee said: "We are very disturbed that there seems to have been no examination of the evidence on potentially adverse health effects before bringing it (Tetra) into service."

Concern about the technology has led to a review currently being undertaken by DERA and the National Radiological Protection Board (NRPB).

The problem is Tetra terminals use TDMA which produces a burst of energy at a rate of 17.6 times per second.

The Stewart Report, commissioned last year to look at mobile phones and health, recorded studies that showed the calcium loss from brain tissue was greatest at 16Hz, close to the 17.6Hz present in Tetra transmissions. It suggested amplitude modulation around 16Hz should be avoided.

"We thought that that was an area that should be avoided if at all possible. Clearly it has not been avoided," said Sir William Stewart, chairman of the Independent Expert Group on Mobile Phones which wrote the report.

Stewart told the Committee he was sure this area would "become a major facet of any future investigation".

BT, whose Tetra 'Airwave' service is being rolled out for UK police forces, said it fully complies with guidelines on exposure to electromagnetic fields. And despite the references in the Stewart Report "no health risks were suggested in the report and none have been identified".

Dolphin Telecom, which launched a commercial Tetra network in August 1999, told *EW* that it supported additional research into wireless technologies and possible health issues. But said its handsets and masts were "well below the recommended exposure guidelines." Melanie Reynolds, *Electronics Weekly*

Website for R&D launched by DTI

The Department of Trade and Industry has set up a new web site, www.ukresearchanddevelopment.com. It styles itself as 'a one-stop-shop for information on science parks, universities, research and development organisations and national laboratories involving all aspects of R & D in the UK'. Patricia Hewitt, e-commerce minister, said: "Visitors to the new Web site will be able to immediately identify which UK centres have the technical expertise to help them develop their products."

UK electronics growth despite US slowdown

The electronics sector in the UK is still predicted to grow this year, according to a recent report, despite growing fears that a US slowdown could hit the industry hard.

The latest set of figures from the Engineering Employers' Federation (EEF) forecasts that the economic slowdown in the US, combined with the cooling of the global boom in IT and telecoms, will halve growth in the UK electronics industry from 17 to 9 per cent for 2001.

Electronics business is leading growth in the engineering sector as a whole in the UK, which continues to grow but is already showing early signs of impact from the US slowdown, according to the EEF report.

Growth in the UK engineering industry will outstrip the rest of the economy this year but growth in orders and output is down on the last quarter of 2000, said the latest RSM Robson Rhodes-sponsored quarterly survey of the sector.

The EEF's chief economist

Stephen Radley said that, despite signs of slower world growth, output and orders had generally held up well and companies were reasonably optimistic about the future.

"The overall picture for the sector remains optimistic," agreed John Nutton, head of RSM Robson Rhodes Engineering Group.

However, Nutton added one caveat: "The pressure on margins

persists as UK-based companies continue to struggle against unfavourable European exchange rates and lower cost economies."

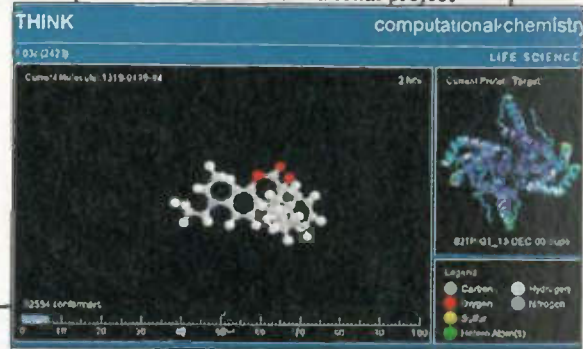
The report surveyed 1,396 companies in the UK during the period from February 12 to March 5, 2001.

It found that growth had slowed in the south east, south west and Scotland but identified growth in the north west and north east.

Helps fight cancer using your PC's idle time

Spare personal computer time is to be harnessed to help cure cancer in an international project

backed by the University of Oxford and Intel among others. Anyone can join in and their computer will run mathematical models to assess the fit of molecules to each other. Intel estimates 50Tflop/s of spare capacity will be made available. Check out www.intel.com/cure if you are interested.



Commercial 3D displays on the horizon

Three dimensional displays have taken a step closer to commercial reality with US firm Actuality Systems producing a 90 million voxel unit.

The display is a volumetric type system, with a flat, vertical screen rotating at 600rev/min. A series of images, between 4000 and 10 000 per second are flashed onto the translucent screen so the eyes and brain perceive a 3D image.

The downside of volumetric displays – compared to other types of 3D display – is the sheer amount of data that has to be stored and projected.

With an refresh rate of 20Hz, the 10000frames/s results in 500 unique frames per revolution, each being a 768 by 768 pixel image, pixels

having eight colours. Thus feeding the display requires a memory buffer containing up to 6Gbit of data.

Actuality has worked with Avnet and Xilinx on the design. Xilinx's Spartan II FPGAs are used in the raster engine. This converts 3D data from a PC via an ultra-SCSI link into the individual image slices.

Image data is fed to the 6Gbit of buffer memory, organised as 1Gbitx3x2 (three colours, double buffered). The images are projected using a Texas Instruments micromirror projector and some rather clever optics.

Three mirrors reflect the image onto the screen ensuring accurate focus and that there is no parallax errors.

Avnet's contribution to the system

is its team dedicated to start-ups, helping Actuality source hard to find components.

Richard Ball *Electronics Weekly*



Three types of display have shown themselves suitable for 3D viewing; stereoscopic goggles, parallax viewers and volumetric displays.

Actuality is an example of the volumetric. A series of image slices is projected onto a spinning screen. The more slices that are projected onto the screen rotates through 360°, the more lifelike the 3D image.

Parallax viewers are being looked at by firms such as Philips. A lenticular sheet, or other optics, in front of a display steers different images to each eye, causing a 3D effect.

Stereoscopic goggles are a simpler, but cumbersome, way of placing a different image at each eye.

New standard capacitor technique involves counting single electrons

In the Kingston-upon-Thames museum are standard yards from the past. Made of iron, they were the fundamental unit of length in their day.

The trouble is they differ in length, some by hundreds of millimetres, as they were based on the length of the Kings arm at the time.

If you had a short monarch, you got less cloth for your goat.

Times have changed and in an attempt to remove variability from

standards, they are being related to the invariant – as far as anyone knows – quantities discovered by quantum physicists.

First to fall was time with the invention of the atomic clock. Then in 1990 two more went to a Josephson junction-based voltage reference and the Hall-effect quantum resistance standard.

If work at the US National Institute of Standards and Technology (NIST) proves to be right, capacitance will be

next. So far, results compare favourably with an existing state-of-the-art standard capacitor – based on an accurately made physical arrangement of metal plates in a vacuum.

The Institute has tackled the capacitance problem from the formula $C=Q/DV$ using multiples of e , the invariant charge on an electron, as Q .

The proposed new standard is a cryogenic vacuum-gap capacitor of around 2pF.

The researchers measured it by counting 100 million individual electrons on to one plate then measuring the voltage change across the capacitor. The count takes around 40s and the voltage measurement takes 20s.

Key to the project is the electron pump. This took two years to develop and is a seven-stage single-electron transistor, or SET, pump – see photo.

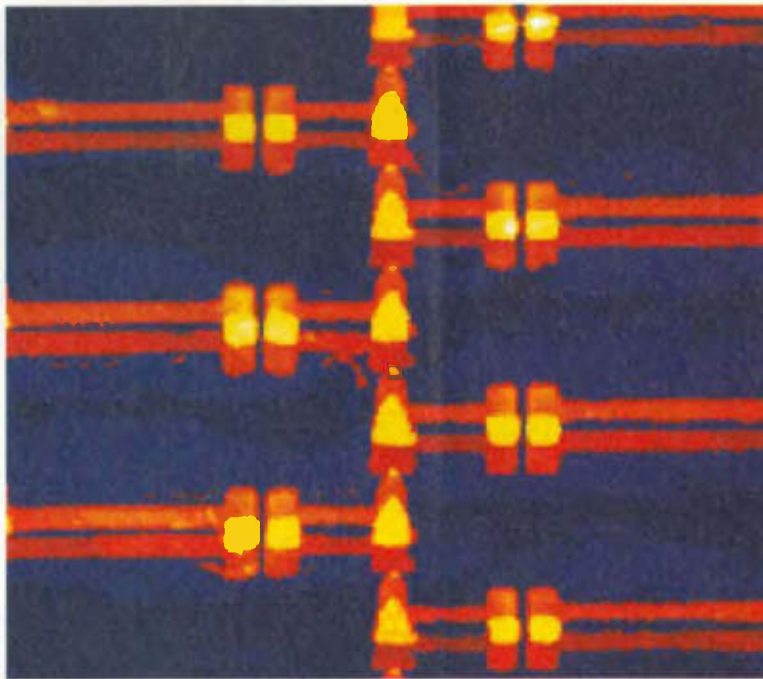
The pump is constructed from two layers of aluminium separated by aluminium oxide and shifted by 200nm. This forms 40 by 40nm tunnel junctions where the tip of an island overlaps its neighbour.

Sequential signals through gate capacitors – left and right – lower tunnel barriers and pump electrons between the metal islands.

The islands are small enough to hold only one electron when cooled to near absolute zero.

Steve Bush, *Electronics Weekly*

Scanning force microscope of a seven-junction electron pump. Its chain of tunnelling junctions move electrons with a certainty of 1 in 100 million. See text for more.



Quantum devices are the key

The upper circuit is the standard capacitor, C_{std} , and its measurement circuit.

The lower circuit is the standard capacitor in action, calibrating a room-temperature secondary standard C_{ref} .

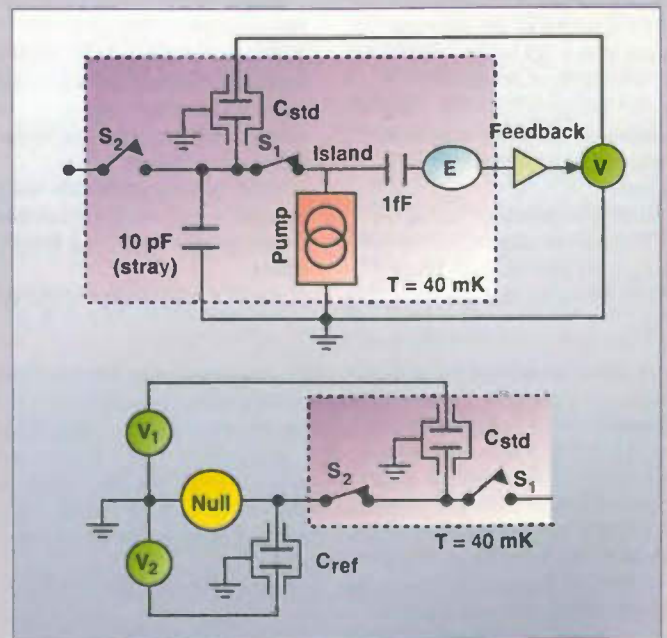
Switches S_1 and S_2 switch between modes and the dotted line is the limit of cryogenic isolation.

E and the 1fF capacitor form a two-SET electrometer that can measure charge down to $e^{-}/100$. These control the voltage source V to maintain the island as a virtual earth through C_{std} as N electrons are counted on to it.

The value of the standard capacitance is then $C_{std}=Ne^{-}/DV$.

Maintaining the virtual earth removes the effect of the 10pF stray capacitance and is necessary for the pump as it only works correctly with 0V across it.

To transfer the cryogenic standard value to a room temperature secondary standard, the two switches are toggled and a conventional AC bridge technique (V_1 , V_2 and the null indicator) is used to measure the ratio of C_{std} to C_{ref} . $C_{ref}=C_{std}V_1/V_2$



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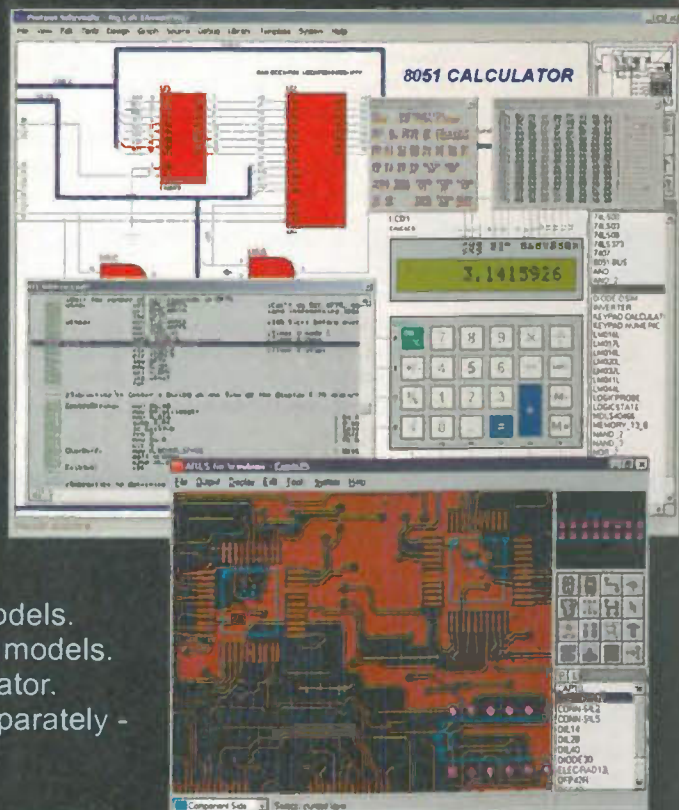
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E-fraud

If problems are opportunities in disguise, then mounting on-line credit card fraud should present fabulous opportunities – and not just for criminals. Making technology solutions work is harder than it seems though, as *Andrew Emmerson* reports.

Not all attainments are creditable and the recent revelation that the United Kingdom is the 'epicentre' of illegal credit card dealings in Europe is a decidedly dubious honour.

There's little consolation for neighbouring countries, however, since across the European Union card fraud rose by a staggering 50 per cent in the year 2000.

Facilitating this spate of swindling is the wired society; a large proportion of this massive increase in deception involves payments made by telephone or over the Internet, the so-called 'cardholder-not-present' fraud. Overall, illegal transactions amounted to an estimated 600 million euros (\$553 million) across the European Union.

The penalty – quite apart from the obvious cost to those defrauded – is a reduced potential for expansion of electronic commerce. While on-line trading has exceeded the most favourable estimates, it remains inhibited by lack of confidence in the privacy and security of payment transactions performed over the Internet.

The Grand Plan

An ambitious three-year plan aimed at tackling frauds across Europe is the European Commission's response to this rise in illegal transactions. The aim is to increase co-ordination with the industry, so as to introduce, "the highest economically viable level of security" for remote transactions by mid-2002, majoring in five main areas:

- Technological improvements
- Improved information exchange
- Better educational material and co-operation, new

Fears allayed?

Two thirds of private customers with access to the Internet lack the confidence to buy goods and services on-line. It was this research, carried out last year for Barclaycard, that led the company to launch a multi-million pound advertising campaign on television aimed at laying to rest shoppers' fears about credit card fraud on the Internet.

James Davison, managing director of e-commerce at Barclaycard, hoped the promotion would persuade consumers that they could shop as safely on the net as they can on the high street. He expected a huge surge in the number of on-line purchases to follow.

Private individuals concerned about items falsely debited to their credit cards on the Net can take comfort in the fact that under the Consumer Credit Act of 1974, the card issuer (and vendor) are liable for disputed personal purchases worth more than £100 – with some exceptions. This applies to cardholders in Britain.

Legislation is different in other countries. In the USA for instance card users can recover losses over \$50.

Doing the sums

- Between 20 and 40 per cent of on-line purchases are fraud attempts, according to US on-line fulfilment house Digital River.
- 'Cardholder not present' fraud in Britain alone cost an estimated £184 million last year.
- The cost of credit and debit card fraud in the USA this year is predicted to reach \$1221.87 billion.
- In the USA, Forrester Research estimates that e-commerce companies lost some \$2.8 billion last year on account of consumers' privacy concerns.
- While on-line credit card use accounts for only about 2% of Visa International's total credit card transactions, more than 50% of its discovered frauds and disputes relate to on-line use.

exposed

training programmes

- Specific fraud prevention measures
- Improved co-operation with non-EU countries.

In specific terms, proposals include:

- The introduction of a single phone number, operational in all EU member states – or at least a single phone number for all issuers based in each country – so as to make it easier for consumers to notify the loss or theft of their cards;
- The setting-up of an EU-wide fraud prevention web site with information on initiatives and links to all relevant organisations;
- Agreement on which key items of evidence are needed to investigate and prosecute fraud cases and on a common format for the exchange of relevant information.

While European in origin, these initiatives must be applied globally, the EC concludes. Action must be taken to prevent criminals from affecting the interests of the European Union by relocating their activities in third countries and only global co-operation will succeed in preventing fraud that's essentially international in scope.

Types of fraud

Most of the mechanisms built into credit card handling to prevent fraud fail when cardholders purchase using the Internet or telephone. So convenient and lucrative are these 'cardholder-not present transactions', however, that nobody would contemplate removing the facility, meaning that fraud proliferates seemingly unimpeded.

The types of frauds perpetrated on merchants revolve mainly around repudiation – when a customer refuses to pay for transactions already made – and impersonation – 'phantom' purchases made with stolen credit cards.

Users become the victim when they find their card statement indicates charges higher than agreed or repeat transactions that they never authorised. Both parties are the losers when thieves steal cardholder information from poorly protected web sites.

A victim-less crime?

Contrary to common belief, everyone pays for credit card fraud. Losses that card issuers cannot recover are simply spread over all users in the form of higher charges, but in general the banks aim not to be caught. The burden is thus transferred to the traders or merchants, also their customers, who may or may not be protected against losses at the hands of credit card thieves.

Whereas consumers' liabilities are generally restricted (see the panel entitled 'Fears allayed?'), the bulk of the burden of on-line fraud is borne by the traders themselves when credit card companies refuse to honour transactions.

A 'charge-back' is created when a cardholder refuses to accept a purchase on a monthly statement, a fertile source

being disputed visits to porno websites accessed using stolen credit cards.

For traders ('merchants' in card industry jargon) these chargebacks are very bad news, since not only do they receive no payment for the disputed transaction, they are also charged a penalty fee of around £10. Even worse, traders are under pressure from the card issuers to keep their charge-back rates to a minimum or face escalating 'review fee' penalties.

Smaller companies argue this policy could force them off the Web altogether, although it is felt that the real target of this policy is to 'punish' large merchants considered to have lax credit card authorisation policies.

A further charge on the profits of web site traders is a straightforward scam that exploits the all-too-simple way they handle purchases. Although their on-line shopping software correctly collects customers' orders, personal details and credit card data, the opportunity remains for buyers to alter the HTML code generated before clicking to confirm the purchase. Traders failing to examine that the sum paid matches the catalogue price of the good ordered on each transaction stand to lose large sums in aggregate.

Remedial measures

Techniques for counteracting these frauds rely on both technical and legal means. Credit cards were never designed to function on the Internet and the remedies to provide safer payment systems inevitably involve some redesign of methods in use.

Since April 2000, British card users have had to provide the three-digit identifier number printed on the signature

False security

Last year was a poor one for customer confidence in on-line trading.

A man logging on to Powergen's web site to pay his bill came across unencrypted credit card details, home address and payment records for himself and 7000 other customers.

Woolworths had to shut down its on-line store after customers' details and credit card numbers were spotted by another user.

In the USA, travellers with Northwest Airlines discovered a security breach in the company's web site that potentially exposed customers' credit card numbers and other personal information. Over there, American Express and Discover Financial Services had to replace the credit cards of all their

customers who shopped at CD Universe. This followed a security breach at the on-line music store that exposed some 350 000 customer credit card numbers.

In Wales, police arrested two teenagers on charges of hacking into Web sites and stealing more than 26 000 consumers' credit card numbers from web sites based in the United States, Canada, Thailand, Japan and Britain.

Back in the USA a federal judge ordered three Californians to pay \$37.5 million for billing 700 000 credit card holders for visits they never made to X-rated web sites. Some of the cardholders wrongfully charged did not even possess computers, according to the Federal Trade Commission.

Quotes of the year

"I would like to thank the nice people at *all* the sites I cracked for having left their entire sales database, readable & writable for any one who bothered to check their site out. Maybe one day people will set up their sites properly before they start trading because otherwise this won't be the last page I post to the NET. Also greetz to my friend Bill Gates. I think that any guy who sells Products Like SQL Server, with default world readable permissions can't be all BAD."

Credit card snaffler Curador.

"We're not blaming Microsoft, but that was the point of entry."
Chris Keller, founder of SalesGate, one of the sites that Curador cracked.

"We're still trying to make customers aware that there is a patch."

Microsoft spokesperson.

"Payment card fraud is the fastest-growing crime, bar none".
Independent fraud prevention specialist, John McVitie,

"Nine out of ten Internet frauds in the UK go unreported."
Richard Fiddis, Chief Operating Officer, Experian.

"Merchants who cannot control the flood of fraudulent purchase attempts will soon be out of business."

Alvin Cameron, Credit/Loss Prevention Manager, Digital River.

strip on the back of the card for cardholder-not-present transactions and also give their address. The companies have not introduced personal identification numbers (PINs), however, which many experts claim would be a highly effective means of preventing fraud.

The card companies are also working on the introduction of new chip technology, which they claim will put a stop to the counterfeiting of cards. Unfortunately, while a chip can guarantee that a card is genuine, it does not show that the person presenting the card is entitled to use it. Nor can it prevent 'phantom' purchases by telephone or Internet.

In the USA, American Express has announced a plan to give its US cardholders 'disposable', unique numbers for

each on-line purchase to avoid giving out their account number on the Web. By using the 'Private Payments' feature, the random number will expire even if a thief steals it from a merchant site or elsewhere.

MasterCard announced a similar plan replacing account data with customer numbers resembling PIN codes for use when shopping on-line or over cell phones.

Recourse to the law may provide a remedy, but it does not provide protection. It is debatable whether any sum recovered can compensate for the loss of status, reputation or competitive advantage following a security breach. Assuming that the police – or private investigators – will catch fraudsters is also unwise. Arrests under Britain's Computer Misuse Act of 1990 are by no means common and in most cases the culprits remain at liberty.

When the credit and business information company Experian UK questioned 800 firms about their attitudes to on-line fraud it transpired that just 57 per cent of them felt it worth reporting fraud cases to the police: of these a mere 9 per cent led to prosecution.

Furthermore, legal remedies lose effectiveness when incidents cross international borders. Even if culprits are caught, they are seldom in a position to compensate an organisation for extensive economic or other loss.

The cost of apathy

A recent survey of British companies involved in on-line trading indicated that most either did no checking or relied almost totally on manual fraud prevention measures. It's understandable, but equally unfortunate, that many companies implement effective information security technology only after a security breach has occurred, according to a report published by IDC Research.

Total suppression of fraud is impossible but Meridian Research calculates that investment in anti-fraud software would deduct \$2 billion from the cost of card fraud worldwide this year.

Virtually all damage is self-inflicted, either through inadequate discipline or by technical incompetence. When

Grand larceny - a practical example

The most sophisticated credit card fraud case ever seen in Britain came to light in February of this year when two Russian criminals were jailed for four years and served a deportation order for their part in a massive credit card fraud and pirate CD ring.

The Russian criminals had produced thousands of perfect-quality fake credit cards by 'stealing' and then copying the magnetic details from customers' cards in restaurants and hotels. Equipment seized by police included special duplicating plates to forge the hologram security marks of Visa and MasterCard.

The pirate music CDs were part of the same massive Russian organised crime operation, with couriers on false passports

smuggling supplies from St Petersburg to a secret counterfeiting factory in North London.

According to trade association IFPI, the case shows an increasingly prevalent form of credit card fraud. A network of accomplices working as waiters or shop assistants would secretly clone the credit details with special swipe machines. The information was then handed to the Russian syndicate and fed into a lap-top computer to create an exact copy of the card.

As well as selling on thousands of fake cards and defrauding Britain's banks of millions of pounds, the gang distributed their top-hit CDs to retail outlets across London and South-East England.



Northwest Airlines confirmed that a programming error had revealed customers' credit card details to third parties, the company blamed the problem on a programmer who forgot to turn the encryption software back on after doing maintenance on the page that yielded the stolen data.

Bug-ridden software is another means of gaining sensitive information; last year, American on-line brokerage company E*Trade acknowledged that its web site had allowed unauthorised access to customers' log-in cookies. Because these were encoded, not encrypted, a cracker could easily unscramble the cookie, yielding customers' user names and passwords. The task is simplified when miscreants use automated software tools to scan the Web for commerce sites with security holes open.

In a high-profile case last year, a cracker using the screen-name Curador claimed responsibility for at least eight Web site break-ins in four countries. He, or she, claimed to have taken advantage of a known bug in Microsoft software to read sites' commerce databases for downloading more than 23 000 credit card numbers. Although Microsoft released a patch for the security hole in mid-1998 and alerted software users, the task of downloading and installing it was clearly too much trouble for the companies whose web sites were breached.

Never-ending battle

If on-line traders are to outsmart their adversaries, it's clear they must get their act together. Implementing SET – secure electronic transaction – standards universally is an ideal but one that's considered unworkable and unaffordable by many observers.

Instead, merchants must put their own houses in order, closing the loopholes currently exploited by crackers. And the card issuers must apply the same kind of neural software used for detecting fraud patterns that mobile phone companies are now using so successfully to combat swindlers. Allowing the *status quo* to continue is unthinkable.

One rule for all

Firms handling credit cards on line are expected to exercise due diligence to protect their customers and themselves against fraud.

A further incentive to act responsibly is a new regulation that the EC intends to impose. It forces companies to comply with national law in every one of the 15 member states and exposes them to legal redress wherever they trade.

Compliance with national legislation in the trader's home country will no longer suffice; businesses will be liable in any EU territory where they do business.

The proposal, known as *Rome II*, would supersede the current principle that e-commerce should abide by the rules of its country of origin. Although this accords perfectly with natural justice, in that companies trading abroad must abide by the laws of that country, certain interests in the British e-commerce community allege that forcing companies to comply with national law in every one of the 15 member states would stifle e-commerce.



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It can be difficult finding a copy of Electronics World at your local newsagents. The number of magazines being published keeps increasing, which means that newsagents have less shelf space for the display of particular titles. Specialist magazines in particular get crowded out.

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- 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.
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- 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) on 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 and DOS 3.3 or higher.
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CIRCLE NO. 109 ON REPLY CARD

Pro audio-visual router

Available commercially as the AVRS8x4, Emil Vladkov's audio and video router system allows you to select one of eight video inputs and one of eight stereo audio channels using either a keypad or a PC. This first article looks at the multiplexer's hardware.

Audio specifications

Input	
Impedance	>20k Ω balanced (600 Ω option)
Level	+27dB μ maximum
Number of inputs	8 balanced stereo
Common mode rejection	>40dB, 20Hz to 20kHz
Output	
Impedance	600 Ω balanced
Level	+27dB μ maximum
Number of outputs	2 balanced stereo
Gain	1
Frequency response	< \pm 0.25dB, 20Hz-20kHz
Total harmonic distortion	0.03%, 20Hz to 20kHz
Intermodulation distortion	<0.05 % (SMPTE)
Crosstalk	> 60dB to 20kHz
Noise Floor	-75dB μ , 30kHz BW
Coupling	DC
DC on output	\pm 50mV maximum

This audio and video switching system is professional quality and suitable for studio use. It consists of two sections, one handling the audio, the other the video. Each section can be used independently, or connected in a stack to perform simultaneous switching of the video sources with their accompanying sound sources.

Called the AVRS-8x4 commercially, this audio-visual router has eight inputs and four outputs on the video module and eight stereo inputs and two stereo outputs on the audio module. Every output of the devices can be connected to every input, or be disabled. This equipment is very useful for TV studio production.

Circuit details – video section

The schematic diagram for the VRS-8x4 video switcher is given on Fig. 1. The heart of the circuit is IC_1 , a Maxim MAX459 cross-point switch with integrated buffers.

The eight inputs are supplied through eight BNC connectors J_1 to J_8 . Each input presents a standard 75 Ω load via terminating resistors R_1 to R_8 . The signals can be AC or DC coupled to the switching matrix IC_1 . Capacitors C_1 to C_8 provide the AC coupling. Correct biasing of the inputs of IC_1 is performed with resistors R_9 to R_{16} .

An industry standard 80C51 microcontroller, IC_2 , is used. Code for the controller is stored in EPROM IC_4 and IC_3 serves the necessary address latching in a typical 8051 system.

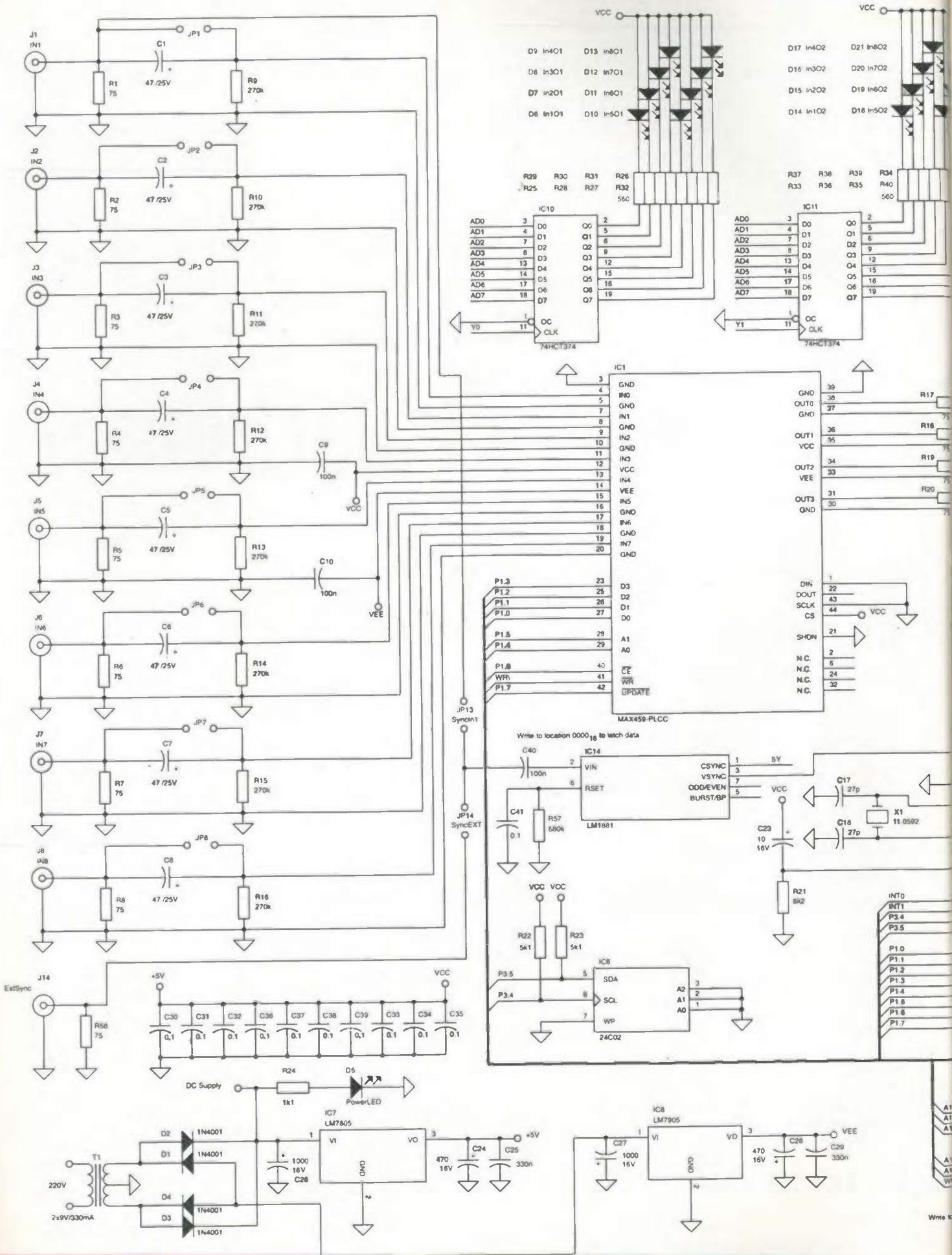
A quartz oscillator is used, and the frequency of 11.0592MHz proved to be more than suitable for the switching purposes of the equipment.

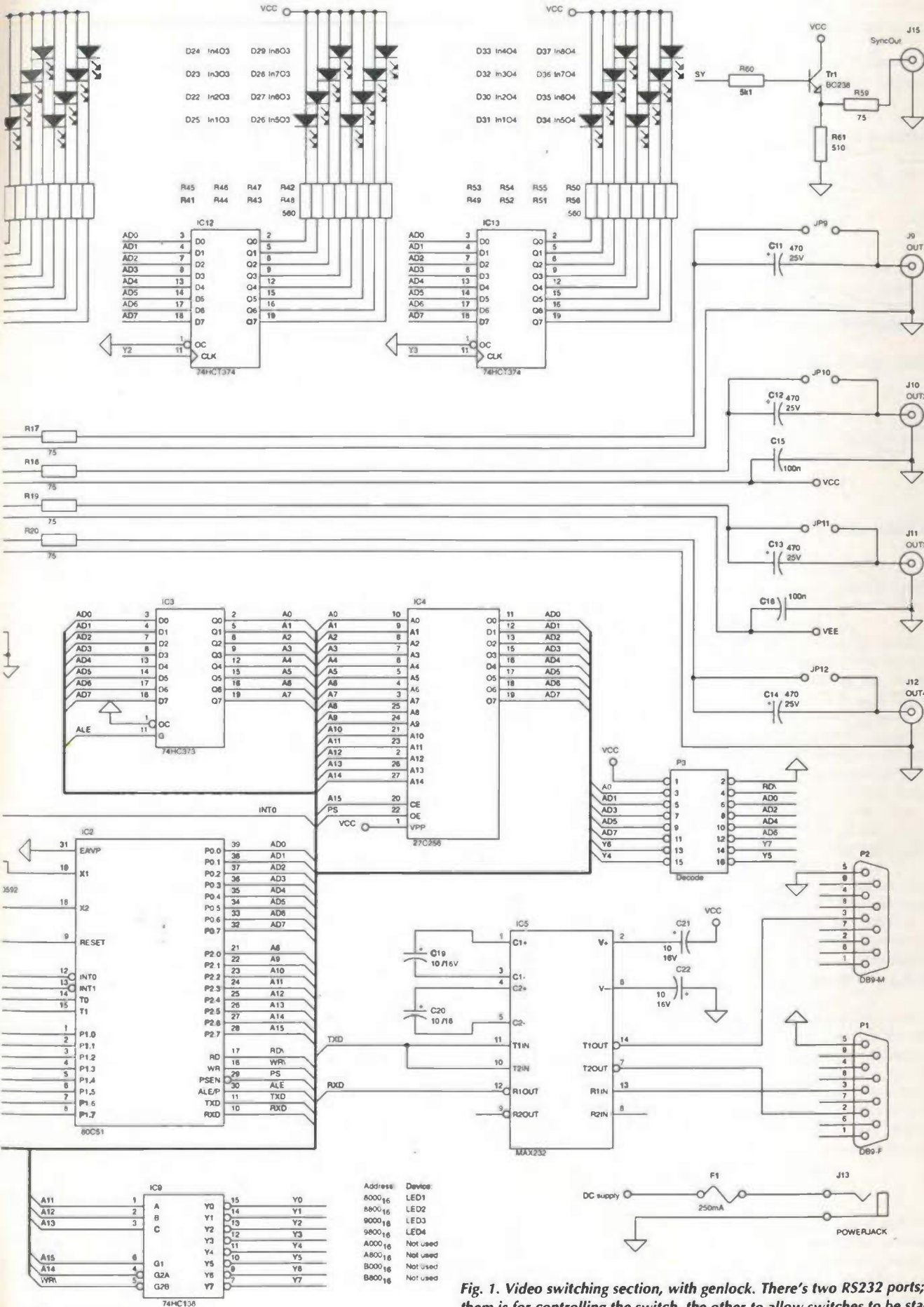
Configuration data is loaded into the switch matrix IC_1 by connecting the signals of Port 1 of the microcontroller to the appropriate digital controls of the matrix. The transfer is performed in two passes because of the internal double latched architecture of the Maxim IC.

In the first pass, the input-select data, presented on lines D0-D3 (P1.0-P1.3), is latched in the input registers of IC_1 . The writing operation is performed by asserting the CE(P1.6) line low and applying a low-going write pulse WR. This signal is generated automatically by the microcontroller when accessing external data memory.

In the second pass, data from the input registers is transferred to the switch registers of the outputs. At this point, the actual change in the switch configuration affecting the outputs takes place. This is done by asserting the UPDATE line (P1.7) low.

In the first pass, the output is selected by the address lines A0-A1

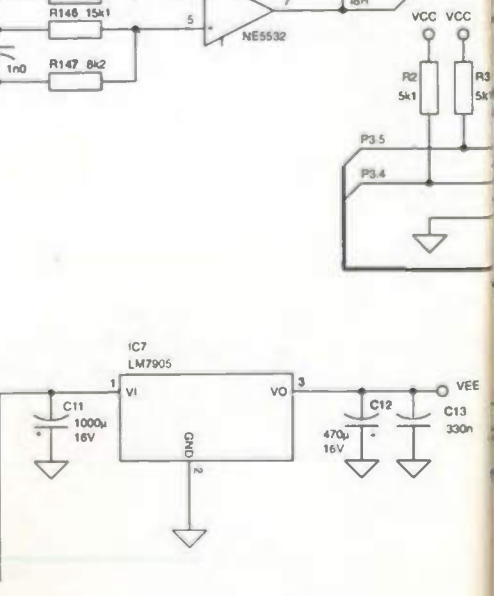
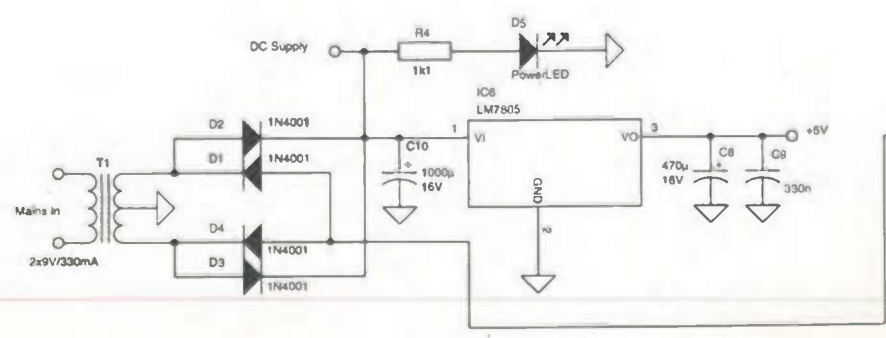
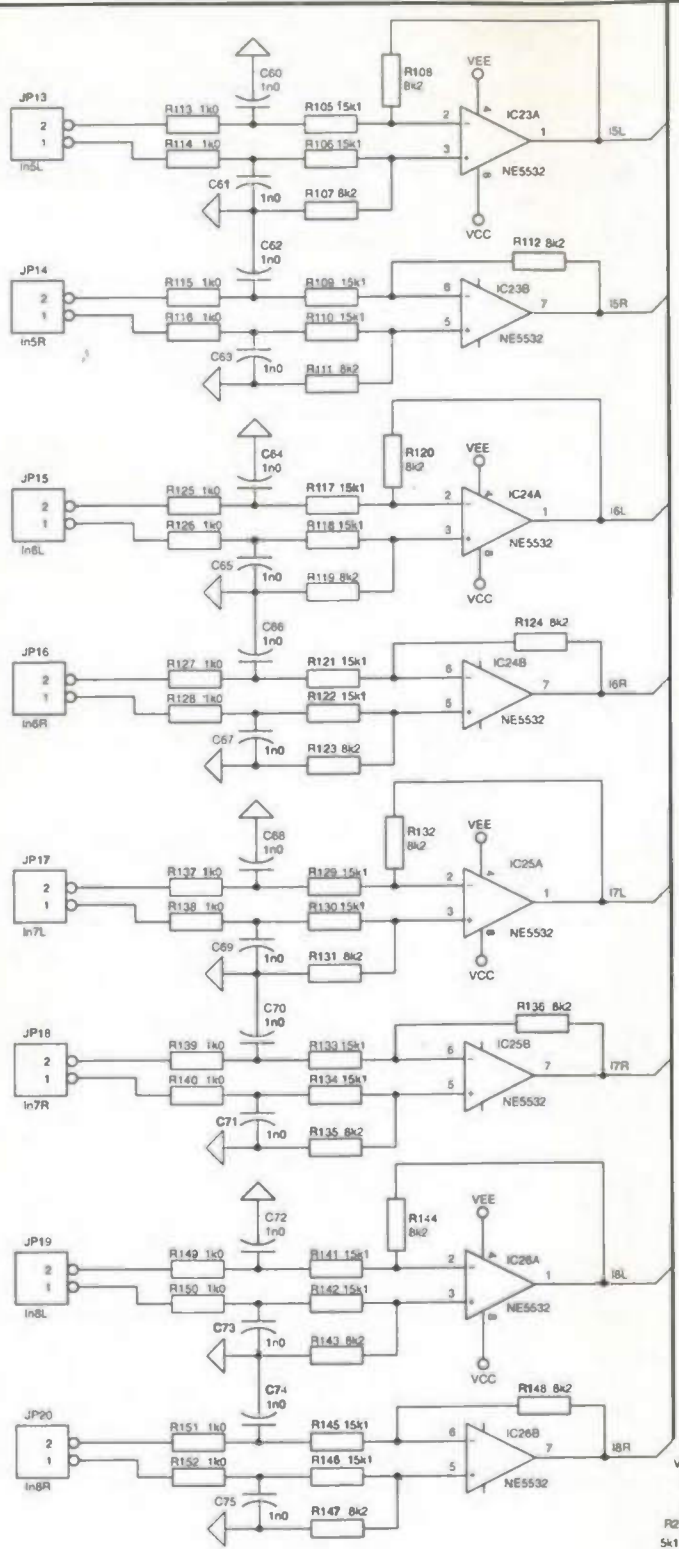
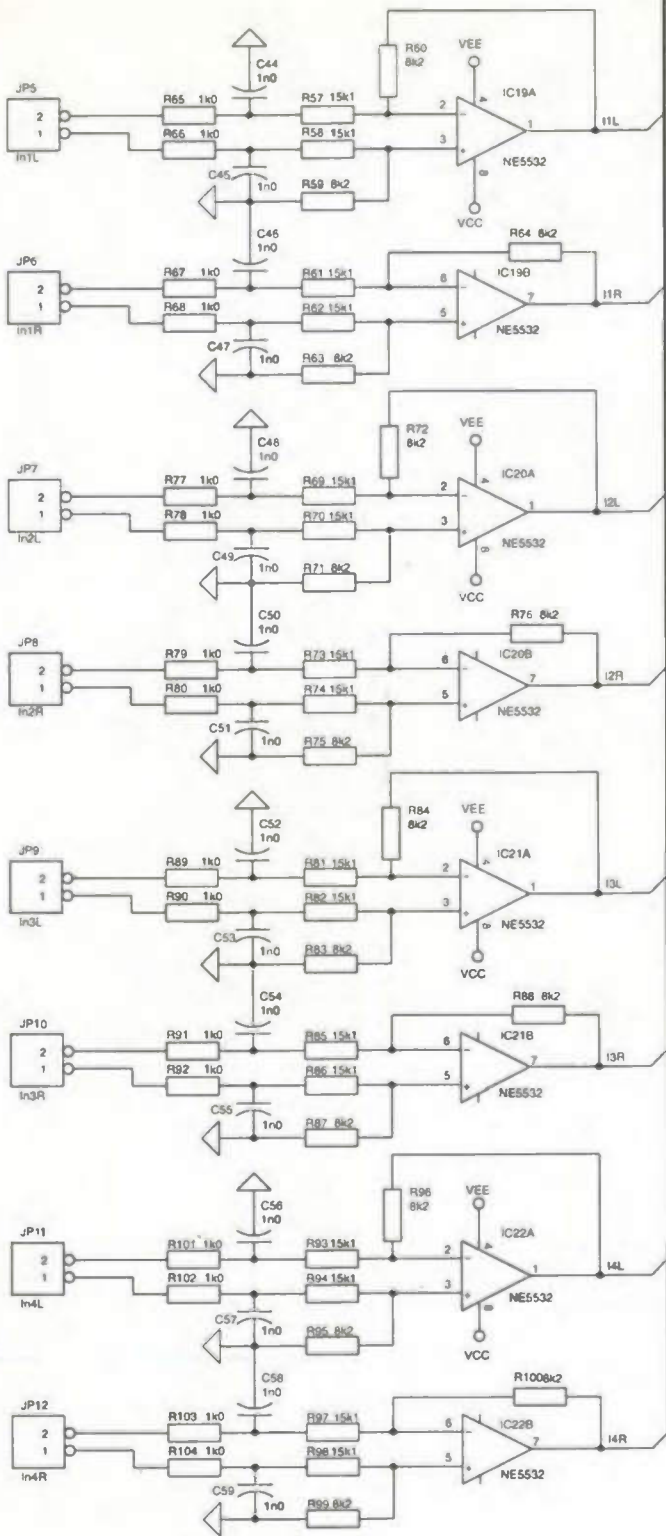




Address	Device
8000 ₁₆	LED1
8800 ₁₆	LED2
9000 ₁₆	LED3
9800 ₁₆	LED4
A000 ₁₆	Not used
A800 ₁₆	Not used
B000 ₁₆	Not used
B800 ₁₆	Not used

Fig. 1. Video switching section, with genlock. There's two RS232 ports; one of them is for controlling the switch, the other to allow switches to be stacked.

Write to these address locations to enable specific LEDs



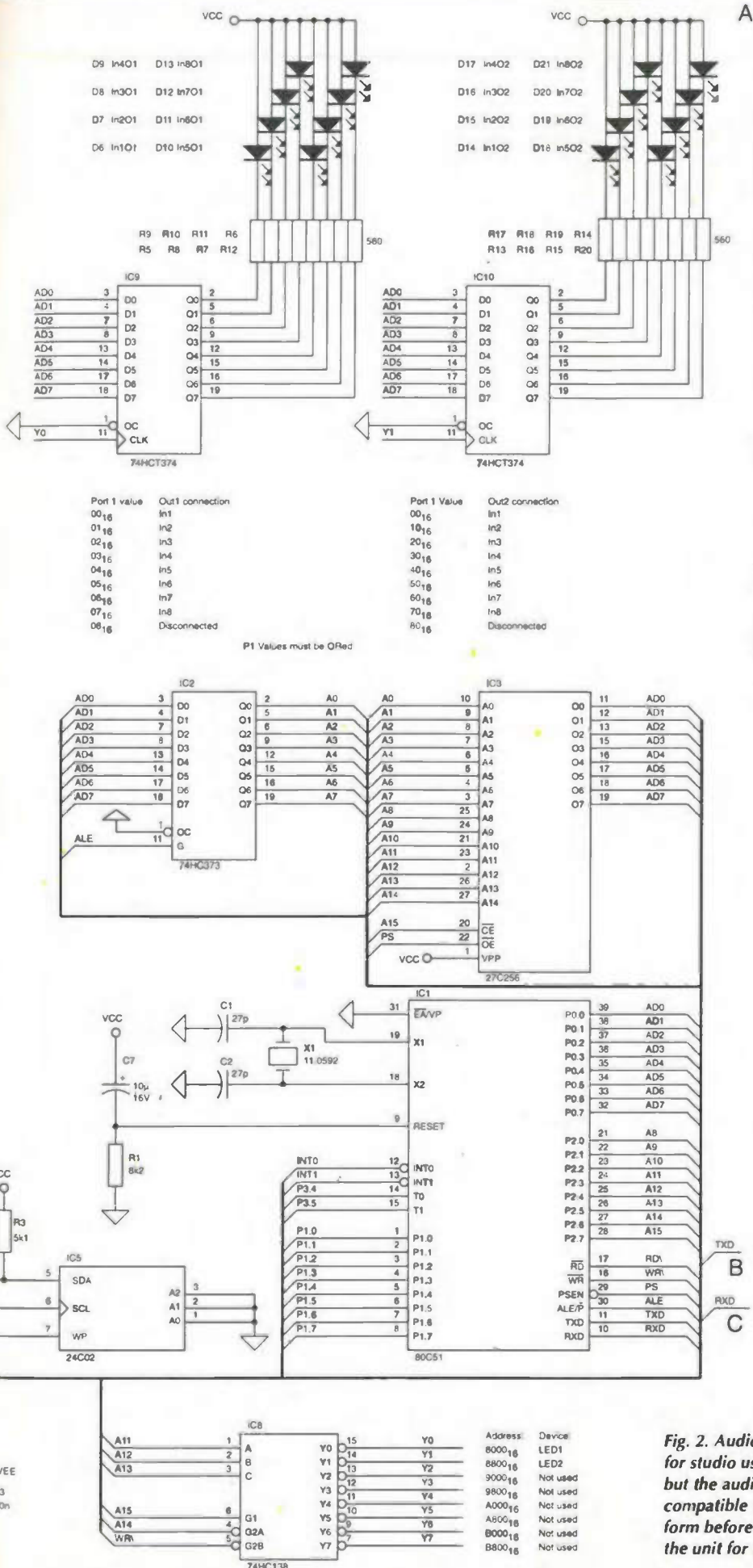


Table 1. Address lines A0 and A1 on the MAX459 determine which output is active.

A1	A0	O/p amp selected
L	L	0
L	H	1
H	L	2
H	H	3

Table 2. Three data inputs on the MAX456 video switch IC are used to select one of the eight input channels. A fourth line, D3, provides an active-high disable input.

D3	D2	D1	D0	I/p channel selected
L	L	L	L	0
L	L	L	H	1
L	L	H	L	2
L	L	H	H	3
L	H	L	L	4
L	H	L	H	5
L	H	H	L	6
L	H	H	H	7
H	x	x	x	Disable output amplifier

(P1.4-P1.5). The amplifier selection and the input selection data is presented in Tables 1 and 2.

The input configuration data, Table 2, is written to each of the four output registers by asserting the corresponding address, Table 1. So the first pass consists of four write cycles. In the second pass, the update is performed only once.

The video switcher is programmed to load the new configuration at the vertical sync. pulse, so a genlock-feature is provided. The source of the sync. signal can be selected through jumpers JP₁₃ and JP₁₄. It can be either input 1 or the external sync. input, J₁₄.

Vertical sync. pulses are separated by IC₁₄, a National Semiconductor LM1881 sync. separator. The VSYNC-output of this IC is applied as interrupt-source 0 (INT0) to the microcontroller, so that the firmware can decide when to switch the matrix.

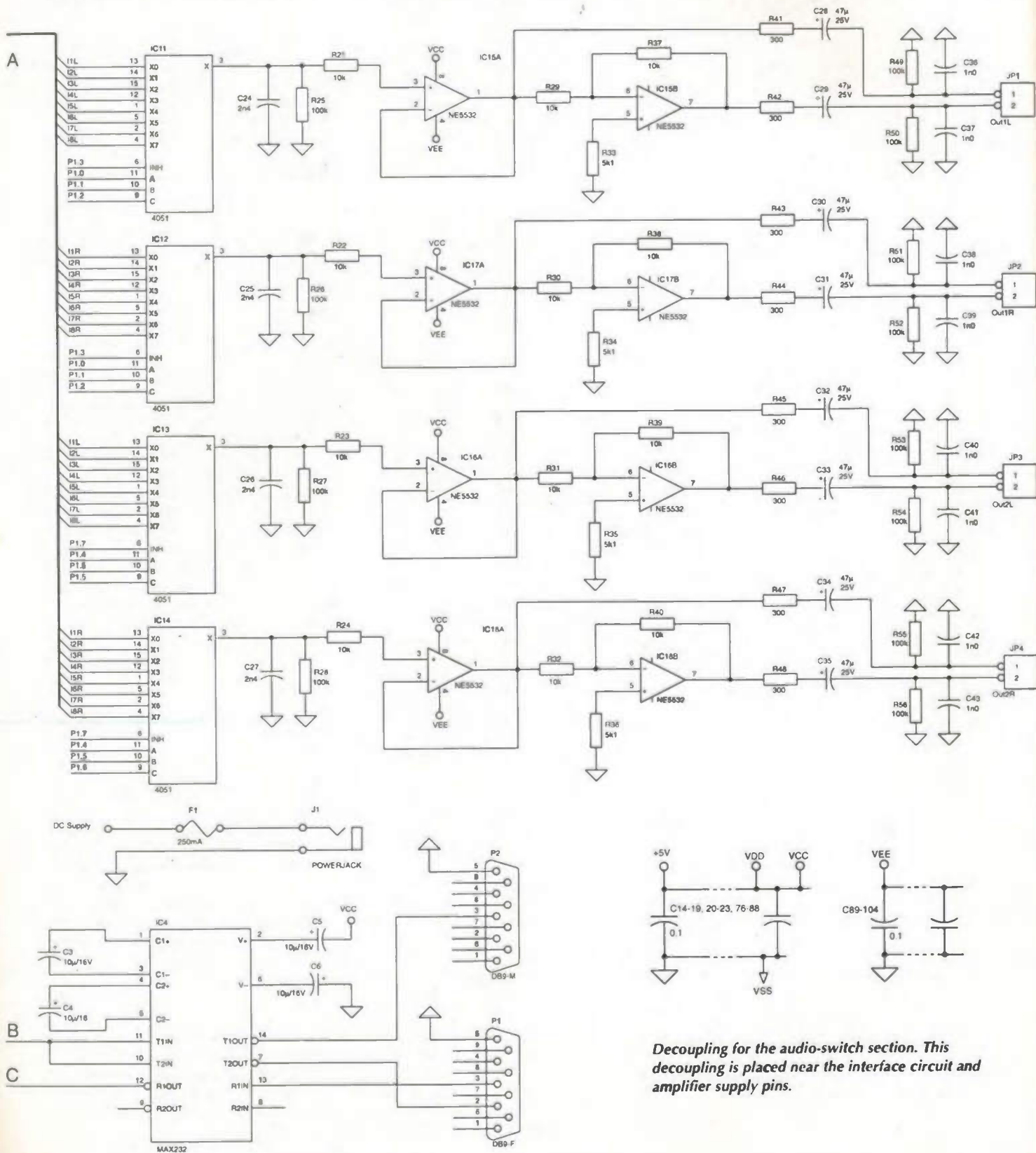
Actually the switching is performed in interrupt driven mode. If the sync. signal fails, because there is no source at input 1 or external sync input, the microcontroller waits for about two frame intervals and performs a timeout switching event. So locking of the video router is avoided, which can be disastrous in critical live studio applications.

Fig. 2. Audio switch section. As this router was designed for studio use, its audio inputs and outputs are balanced, but the audio is converted to unbalanced form to make it compatible with the 4051 switch IC, then back to balanced form before output. This should make it very easy to adapt the unit for unbalanced use.

The outputs of the video matrix are standard 75Ω, which is assured by resistors R₁₇₋₂₀. Again the outputs can be AC-coupled, through C₁₁ to C₁₄, or DC-coupled through the jumpers JP₉₋₁₂. Output connectors J₉₋₁₂ are BNC types on the commercial design.

A composite sync. output from the IC₁₄ sync separator is provided at J₁₅ (SyncOut). The signal is buffered by T₁. This composite sync. signal can be used for supplying additional equipment with a clock signal. Visual indication of the active configuration

is provided by four rows of eight red LEDs, D₆₋₃₇, with their current limiting resistors R₂₅₋₅₆. The LEDs are driven by IC₁₀₋₁₃ 74HC374 latches. These are loaded with data by the firmware of the microcontroller. Each of the four latches is visible as a loca-



Decoupling for the audio-switch section. This decoupling is placed near the interface circuit and amplifier supply pins.

tion in the external memory address space of the microcontroller. The decoding of the address, applied in a standard write MOVX operation of the micro, is performed by IC₉. To light the corresponding LED, the micro has to write to the corresponding latch a word with logical 0 at the LED's location.

The microcontroller subsystem includes a non-volatile EEPROM memory IC₆, which is a 24C02 or similar. This memory stores the data for the start-up configuration to be loaded into the matrix immediately after power is applied. Control signals for the serial EEPROM are emulated by the lines P3.4 (serial clock) and P3.5 (serial data).

The video router uses a linear power supply. It comprises T₁, which is 2x9V, 330mA transformer, rectifying diodes D₁₋₄ and voltage regulators IC₇ and IC₈ for the +5V and -5V rails respectively. In my design, the transformer is mounted on the PCB. Supply voltages of +5V and -5V are needed for the Maxim crosspoint switch.

Capacitors C₃₀₋₃₉ are power supply bypassing capacitors. The raw power supply after the rectifier can be used to supply an external keyboard, which can be used to control the switcher with a simple button interface. Such a keyboard would connect to the video switcher through the P₁ serial RS-232 port and the J₁₃ power jack.

In the actual implemented configuration, I use only a small banana plug to apply the DC voltage supply to the keyboard. Ground is provided via the 'D' connector. Fuse F₁, rated at 250mA, protects the power supply in the case the keyboard cable is damaged, resulting in possible short circuit.

A serial communication channel is used to control the switcher and to enhance its functions. This channel is incorporated in the 80C51 microcontroller. Voltage-level converter IC₅ is necessary to supply the RS-232 standard voltage levels.

There are two serial 9-pin D-type connectors at the back of each switcher module – one male and one female. The female connector P₁ is the master control input, where the external keyboard or the host controlling computer is connected. This port provides full duplex communication, which means that the switcher can not only receive configuration data through it but also return configuration status data.

Male port P₂ is the slave port, where the next device in the stack (discussed later) can be connected. This port provides no return receive channel. Only configuration data to the next stacked video router – or audio switcher – can be transmitted through it. Status information can not be received.

Power-on LED D₅ provides the user with a visual indication that there is no power supply failure and the system is on.

Circuit of the audio router

The schematic diagram of the stereo audio switcher is presented on Fig. 2. The routing

Video specifications of the 8x4 audio-visual multiplexer

Input	
Video type	Composite DC or AC coupled, backporch at ground
Impedance	75Ω, internally terminated
Number of inputs	8
Return loss	>40dB to 10MHz
Level	1V p-p nominal +1.5V to 0.5V
Sync input	1V p-p composite video or 0.4V p-p composite sync
Synchronisation	From input 1 or sync input jumper selectable
Output	
Number of outputs	4
Impedance	75ΩDC (or AC) coupled
Return loss	>40dB to 10MHz
Frequency response	<±0.5dB to 5MHz 3dB bandwidth 90MHz
Gain	1 ±0.05dB
Differential phase	<0.140
Differential gain	<0.13%
Tilt	<0.1% (DC coupled mode)
Sync output	2.5Vp-p (loaded 75Ω) from sync Input or input 1 depending on jumper position
Crosstalk	>55dB worst case @ 4.43MHz >65dB typ. @ 4.43MHz
S-to-n, inc. hum	>70dB unweighted wideband >76dB unweighted to 10MHz
Switch	Vertical interval (external sync or from master input signal)
Electrical	
Power consumption	Approximately 8W

Note: some specifications will depend on PCB layout. Figures given are for author's design.

matrix has eight balanced stereo inputs and two balanced stereo outputs.

The balanced input/output feature makes the device suitable for work in noisy environment. Connections JP₁ and JP₂ are for the left and right channels of output 1 and JP₃ and JP₄ are for output 2.

The differential inputs are provided by NE5532 operational amplifiers IC_{19,26}, wired as differential amplifiers. The differential nature of the processing involved means that resistors with tolerance <1% have to be used for R₅₇₋₁₅₂ to achieve a good CMRR value. An alternative method is to select matched resistor pairs.

The internal processing of the signals is unipolar, minimising the number of channels for the audio multiplexers needed. The unipolar output signals of the multiplexers have to be converted back to differential signals. This is done by op-amps IC_{15,18}, which are wired as followers and inverting amplifiers. The output is a standard 600Ω type, which is assured by resistors R₄₁₋₄₈. The signal is AC-coupled through capacitors C₂₈₋₃₅.

My PCB is designed so that additional 600Ω resistors can be placed at the inputs to terminate properly the audio path if necessary. This is recommended only if you are sure that the source audio equipment can drive the loads. If

Router control system

Local control	
Display:	4 (2) rows of 8LEDs
Configuration:	Indication only
Keyboard control:	PVRS-1 device, connected to the serial port of the first device in the stack with the power plug inserted
PC control	
Type:	GUI-application, running under Windows95
Configuration:	Full control, stackable
Interface:	Serial RS-232, 9600bit/s

it cannot, as in the case of my prototypes, these resistors should be omitted.

Four 4051 8-to-1 multiplexers, IC₁₁₋₁₄, switch the audio signals. Multiplexer IC_{11,13} switch the left channel while IC_{12,14} handle the right-channel signals. The multiplexers are controlled through the P₁ port of the 80C51 microcontroller IC₁, clocked at 11.0592MHz (x1).

When writing the appropriate values to the port pins P_{1,0}, P_{1,1} and P_{1,2}, the micro selects the source for the left output channel. Using line P_{1,3}, the microcontroller can inhibit the work of the multiplexer, so that no input is connected to the corresponding output. This is the so called 'disabled' state of the output.

Selection of the right-hand channel is controlled through lines P_{1,4}, P_{1,5} and P_{1,6}, while the right-hand 'disable' function is controlled by P_{1,7}. The values, which should be written (after the OR-operation) to the P₁ port, are given in Fig. 2.

Code for the micro is stored in the EPROM IC₃. The start-up configuration for the switcher is stored in EEPROM IC₅. Address decoding is performed by IC₂. This is necessary for the microcontroller subsystem.

A serial RS-232 transceiver is incorporated in the microcontroller and is supported by interface circuit IC₄. Serial port connections P₁ and P₂ are built in a similar manner to those of the video switcher section, assuring the possibility of stacking mode of operation.

Visual indication of the actual connections is given via LEDs D₆₋₂₁, driven by latches IC_{9,10}. Resistors R₅₋₂₀ are current limiting resistors for the LEDs.

To select the appropriate LED latch, decoder IC₈ is necessary. The two latches are at their corresponding locations in the address space of the microcontroller.

Linear power supply for the audio switcher consists of T₁ transformer, rectifiers D_{1,4}, voltage regulators IC_{6,7} and the accompanying capacitors. Because of the great number of interface circuits and amplifiers, special attention has to be paid to power supply filtering. A great number of bypass capacitors are needed - C₁₄₋₂₃ and C₇₆₋₁₀₄.

Because the audio switcher can be used without the video switcher, the power supply option for an external keyboard is provided by J₁ power jack and the F₁ fuse.

The keyboard

The audio-visual routing system can be controlled in two ways - through a computer, or via an external keyboard.

Despite having a powerful and versatile PC-based GUI interface, I found that some studio staff would prefer to work with simple buttons. So I developed this 8x2 keyboard.

This keyboard was published as a circuit idea in the March 2000 issue of *Electronics World*, pp. 259-260. I will not discuss it in detail here, but for the case of completeness the circuit diagram is given in Fig. 3.

The keyboard is build around the 89C2051

microcontroller. It has simple push buttons with LED user feedback. When pressed once, the corresponding connection is established. Pressing the button a second time releases the connection.

Communication with the rest of the system is performed through RS-232 (P₁) and the power is supplied to the power jack J₁ from the host system. Alternatively, you could use one of the unused D-connector pins to supply power, so that the keyboard is connected to the router with only one cable.

I have hard-wired the serial cable to the key-

board, and at other end (the AVRS-side) I have branched the cable into two connectors - one a standard male D-type connector and one a power-supply jack. So the serial port of the switcher remains standard (it does not carry power) and at the same time the power is transmitted through the dedicated single cable to the keyboard.

The keyboard has a very useful feature, namely a lock switch. When activated, it prevents its operator from disturbing the whole production in the studio by accidentally sitting on the keyboard.

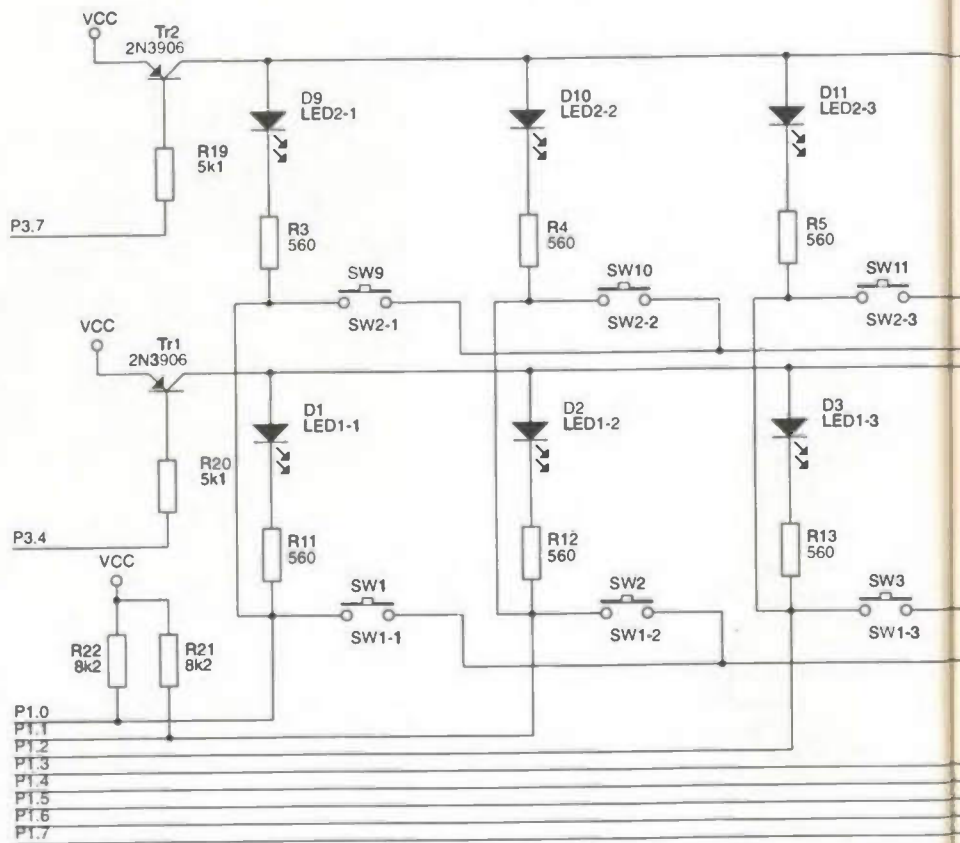
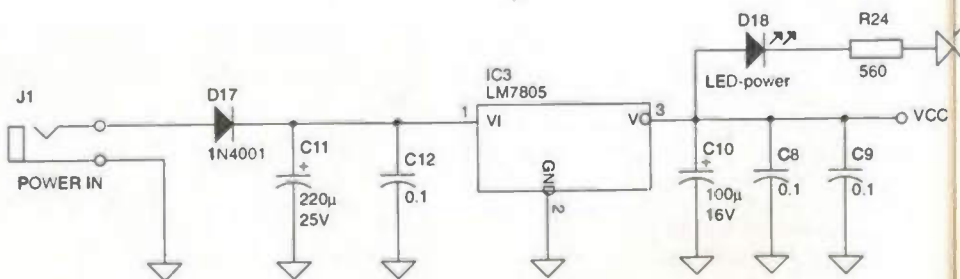


Fig. 3. If you don't need the versatility of a PC-controlled AV router, you can add this keypad to the audio and video switches and control them manually.



Putting it together

I strongly advise that you use sockets for expensive components like the MAX459 crosspoint switch and the microcontrollers. The same is valid for the serial EEPROMs and the firmware EPROM.

For the transformers I, used ANG parts. These are readily available here in Bulgaria – they are manufactured here. They are sealed and have excellent reliability. If you can not find the same components, use substitutes with comparable reliability.

In my system, the video and audio switchers are in separate single RU 19in metal cases. In

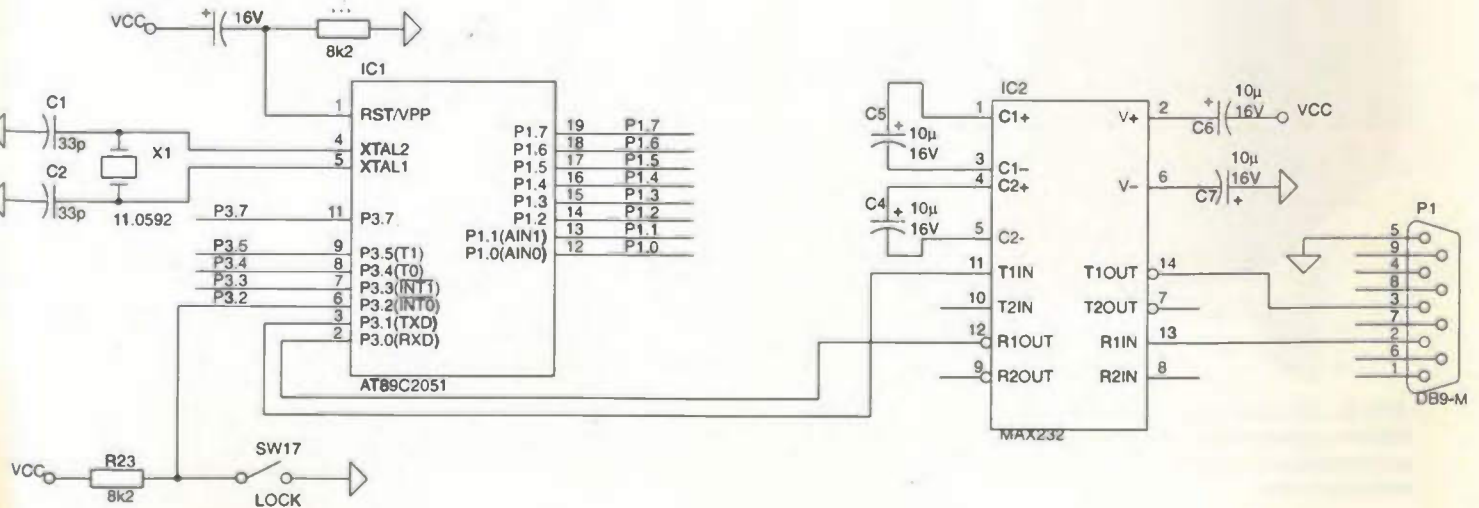
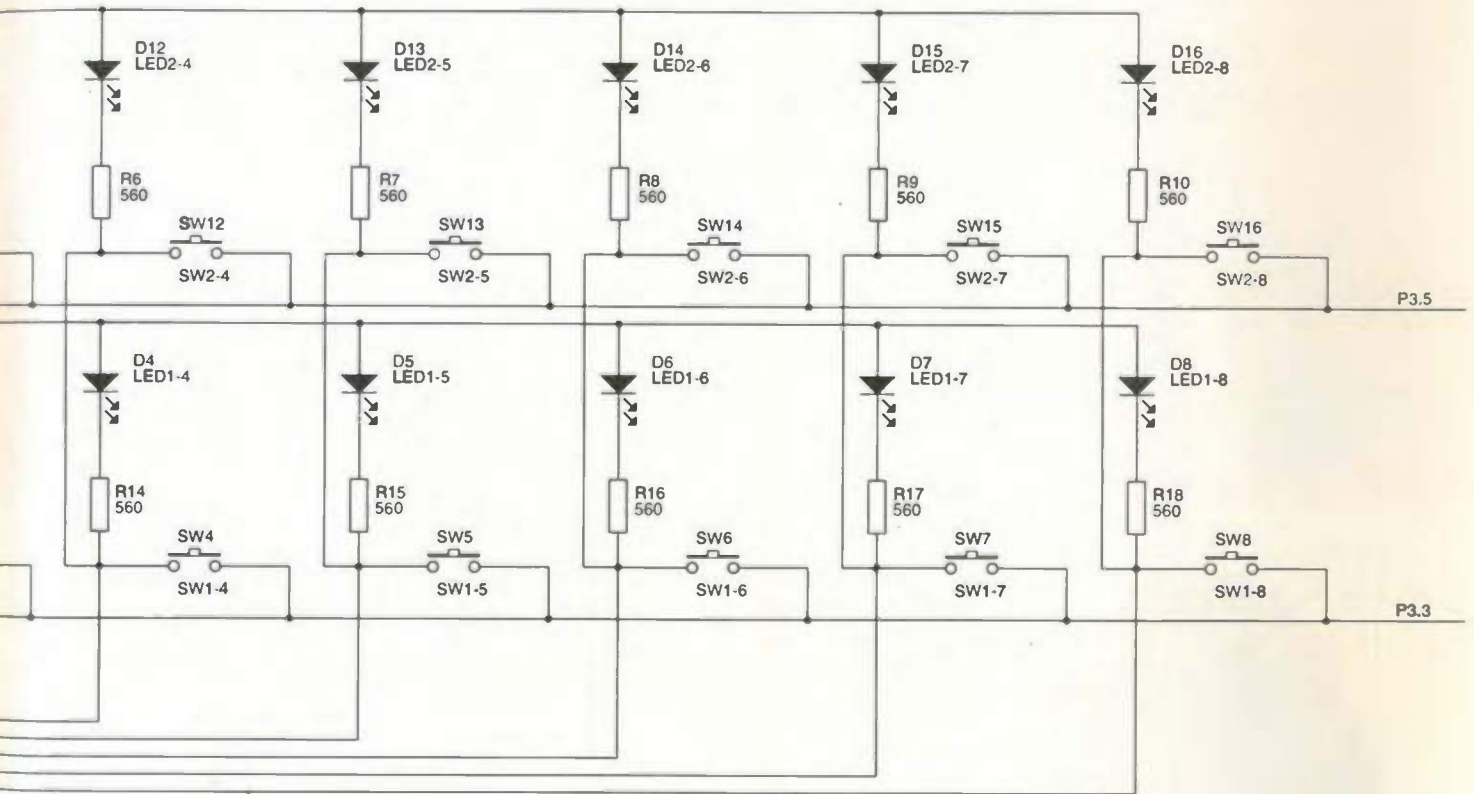
the commercial system, only the active connection LEDs and the power supply green LED are mounted on the front panel of the video router. The front panel of the audio switcher differs from the video panel only in the number of LEDs and their colour, because of the fewer output channels available.

The rear panels of the systems carry all the connectors. A mains switch, the AC-power connector and the fuse socket are also on the rear panel.

Master and slave connectors for the RS232 interface are standard 9-pin female and male D-type connectors. The inputs and outputs of

the video switcher use standard BNC-type connectors. The balanced inputs and outputs of the audio switcher are wired to two rows of screw type connectors that allow audio connecting cords to be connected directly. ■

Next month, Emil discusses the software needed for the audio and video section microcontrollers, and for the PC GUI. Details on how to obtain the software will also be revealed.



Technology

After decades of developing speech recognition technology there are a few commercial applications emerging like turning on the car radio but nothing dynamic that will fulfil its promise. David Manners* talks technology

*David is Components Editor with *Electronics Weekly*.

The question is always asked: 'When will speech recognition technology fulfil its promise?' The answer from the experts these past twenty years has always been: 'In the next three to five years'.

The difficulties have always been under-estimated. Back in the 1980s, the UK government's 'Alvey Programme' tried to crack speech recognition and concluded that the necessary hardware processing power didn't exist. It still doesn't.

Nonetheless, Alvey produced some valuable

software that was widely licensed to, among others, Microsoft, AT&T, MIT, HP, Sharp and Logica.

Today, in Sweden, if you phone the directory enquiry service of network operator Telia, you will be asked to speak a postcode and street name and the phone number is automatically spoken back. That system was supplied to Telia by the UK company Vocalis which was spun out of Logica in 1993. The technology being used by Telia derives from the Alvey technology.

There are other examples of practical voice recognition technology. If you're in Los Angeles and dial 411 on a telephone, you'll get an automated voice asking for the name of the party you wish to call. You speak the name into the handset. The automated voice replies with the telephone number.

Pop into an up-market car salesroom and you'll find

that the top-of-the-range Jaguars, BMWs and Mercedes will be featuring speech recognition capabilities for the settings on their radios and for telephone dialling.

Last October, NTT Data Corporation of Japan started selling a speech recognition telephone system to network operators and corporations which costs around \$20 000 for four lines aimed at providing information services for tourism, medical help, educational assistance and corporate guides.

At the end of last year, AOL announced a service called AOLbyPhone by which users could telephone voice commands and receive information in return – including having their e-mails read to them. It attracted 200 000 subscribers in its first three months of service.

So, after decades of development, we are seeing a few commercial applications emerging. But they are only a few, and they're not making anyone rich.

After all, Lernout & Hauspie (L&H), one of the world's great practitioners of speech recognition technology, is bankrupt. It went into Chapter 11 in the US last year and, in January, went into bankruptcy in Belgium. This is despite big investments. Intel invested \$30m in the company; Microsoft invested \$45m.

With L&H looking as though it can no longer remain independent, the main speech recognition developers are IBM, Philips and Microsoft. All three are said to be considering buying L&H, because everyone regards the technology as potentially valuable. That's because it has always been thought that good speech recognition technology would be the key to a new electronics boom.

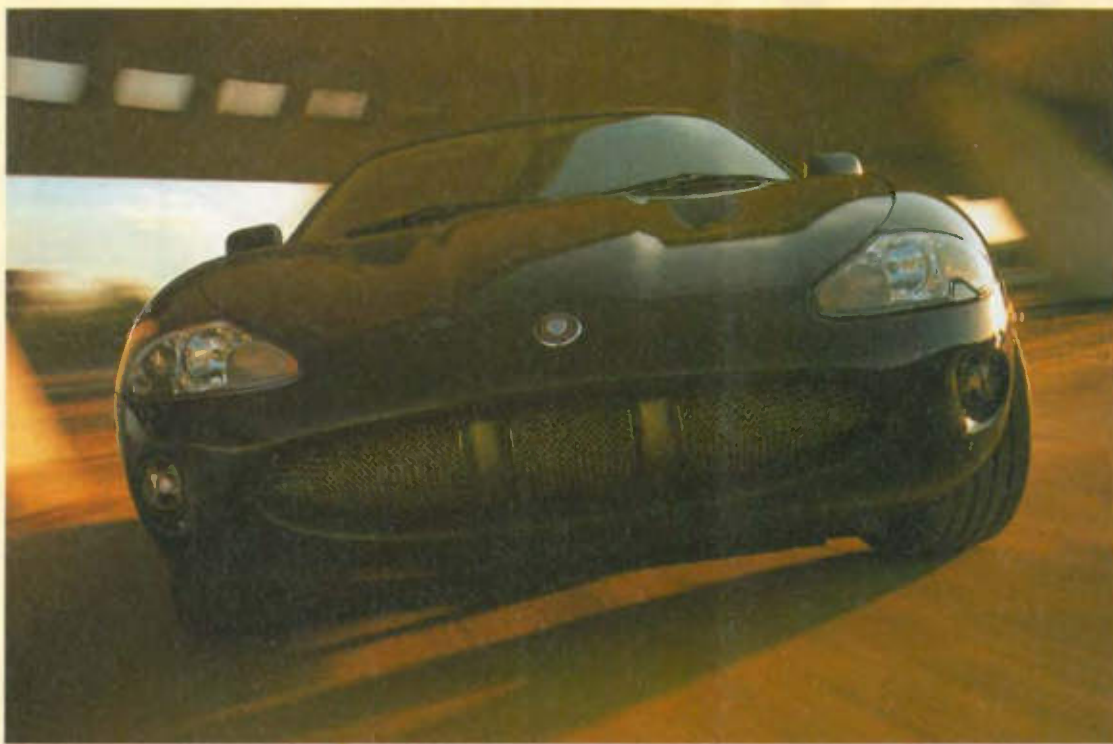
For instance take China where the language has over eight thousand characters making keyboards a problem. Or take the 50 per cent of the Western world who don't own a PC because it takes effort and training to use them and requires you to use your eyes, brain and fingers.

A voice-controlled PC would not need such effort. It would be more of an electronic personal servant rather than a quirky, complicated machine seemingly intent on baffling its owner.



NEC has developed a personal robot called PaPeRo (partner-type personal robot) that can recognise 650 spoken phrases and speak more than 3000.

that's all talk



Some Jaguars, BMWs and Mercedes feature speech recognition capabilities for the settings on their radios and for telephone dialling.

For 3G mobile phones, speech recognition is thought to be valuable if their full potential is to be realised, and many companies are working on it. One is the Israeli company Advanced Recognition Technologies (ART). "From the time we set up the company we designed our algorithms to fit into a limited space", says ART's v-p Yakov Schulman. Accordingly ART's basic speech recognition programme fits into 60kbyte.

Thereafter, the storage requirement depends on the number of commands the user wants recognised. For 20 commands, the storage required is 50kbytes. For recognising 50 names, each taking one second to say, needs 30Kbytes.

Philips has developed special chips for adding speech commands to any product. Called SPHERIC, the speech recogniser chips require half a megabyte to store up to 120 words. In a one-user, one-language telephone application requiring 20 words of user-dependent vocabulary, the SPHERIC chip would need 47kbyte of storage of which 27kbyte would be user-dependent vocabulary. Other applications for SPHERIC are remote controls, telephones, answering machines, personal digital assistants, car navigation and audio systems and toys.

IBM produced the technology for adding speech

recognition to the PalmPilot. According to Tom Houy, of IBM's embedded speech operation, the add-on hardware device which is attached to the PalmPilot contains 400kbyte of memory and is capable of storing 500 words.

3G phones are one thing, but there's a better thing – a Holy Grail of a product which successful development of speech recognition will make possible – the portable, real-time translation machine.

Companies estimate, that this application will require 3Gips microprocessors, which are expected in 2003. Two years after that should come versions cool enough, and low power enough, to fit into portable products.

If the cost of processing power in terms of Mips/\$ continues to decline at the current rate – i.e. by two orders of magnitude per decade, from \$100 to \$1 in the '90s – then, in 2004/5, a microprocessor Mips should cost ten cents. This would price a 3Gips microprocessor at \$300, and a portable translator at an affordable \$1000.

The software people think they can do it. "In three to four years you will be able to say something in one language and the computer will say it in, for example, Japanese," said Jo Lernout at an L&H demo last year.

So it's three to four years away – so what's new? ■

PC via

interfacing Ethernet II

Connecting external devices to your PC via a local-area network card is not as difficult as it may seem. In this second article on the topic, Eddy Insam describes a development device that can get you going in no time at all.

In last month's article, I covered some of the theory behind designing embedded microprocessors and local networks. In this second part, I describe a practical implementation using a preprogrammed PIC 16F877 controller. Together with a cheap, standard Ethernet PC card, this PIC can be used for demonstrations or as the basis for a prototype interface project.

The circuit

Figure 1 shows the main interface circuit diagram. To avoid using difficult-to-obtain parts such as Ethernet line transformers, I opted for a ready made cheap ISA PC LAN

card; these are readily available for less than £10.

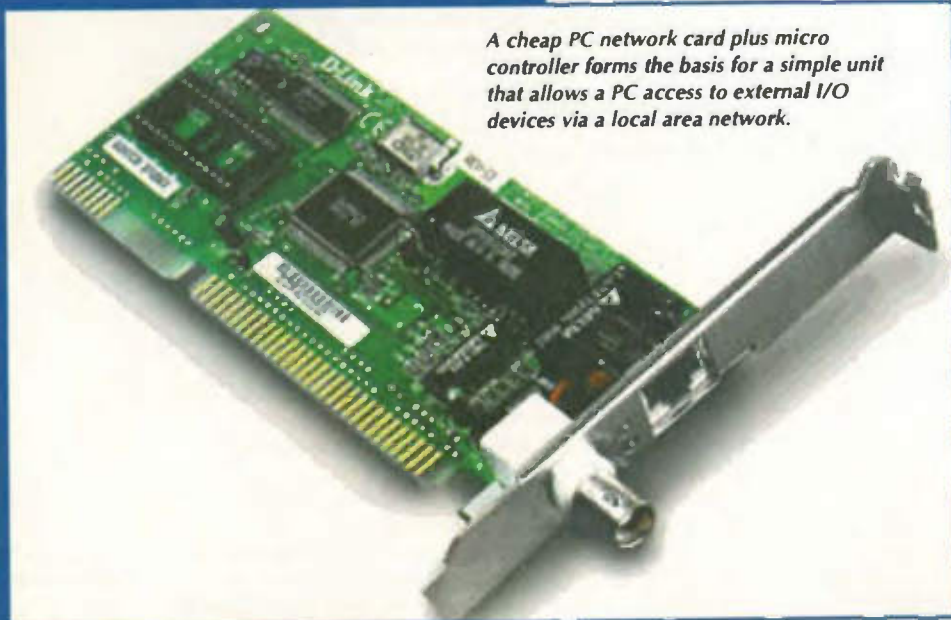
The 16F877 uses quite a few of its available I/O ports to drive the PC card: port D is used for bi-directional 8 bit data transfers, five bits of port B are used to address the sixteen I/O mapped registers, and the three available bits of port E are used to drive the read/write strobes. Remaining ports are available for user interfacing, including serial communications, reading the DIL switch, and for general purpose digital and analogue I/O.

As mentioned, only five address lines are used with the PC card. The Ethernet chip communicates via 32 I/O-mapped registers, so five bits are all that is needed. The other address lines are just hard wired to map to the nominal address range 0x300 to 0x30F, the usual default for this type of card.

You will notice that ISA PC cards have two edge connectors, and that I am using only one here. This is because these cards were designed as plug-compatible 16-bit enhancements of the original PC XT bus, which only work in 8-bit mode. The second edge connector carries the remaining 8 data bits plus a few more control lines. The electrical design of the card works in such a way that leaving the second connector floating or unconnected will automatically configure the card for XT compatible 8-bit transfers.

The only other IC device in the diagram is an RS232 converter. This is only needed if the PIC is to communicate with another controller, PC or microprocessor via its serial

A cheap PC network card plus micro controller forms the basis for a simple unit that allows a PC access to external I/O devices via a local area network.



interface. Of course, this is only relevant if you want to develop your own application and drive the PIC via the serial port.

You may decide to dispense with the converter chip and provide the output at TTL levels. After all, there is not much point in converting your data to RS232 levels only to be converted back to TTL levels at the other end. The serial interface operates at either 57600bit/s or 2400bit/s, depending on DIL settings.

I added a DIL switch to read start-up conditions during reset, such as port serial speed. The PIC also operates a 'clean-up' reset to force

the contents of the internal EEPROM to absolute default conditions. This is done by forcing port C6 down while resetting the chip, hence the two DIL positions connected to these pins.

Other DIL positions are connected in series with the LEDs to allow the same ports to be used as inputs and avoid being loaded by the resistance of the lights.

The demonstration software mentioned later uses these ports to flash the LEDs. No project is worth anything unless it can flash some lights, so this is a good reason for including them here. Of course, you can use these ports for anything you like, input or output.

The various spare I/O port data lines are brought out to connector J3. The assumption here is that there would be another PC board containing the 'user' section and other custom interfaces.

The 16F877 has a built-in 10-bit a-to-d converter, and can read analogue data from any of the 5 lines in port A. This can be quite useful for reading sensors and other devices. The analogue outputs must be conditioned to supply an output voltage range between 0 and 5 volts, as this is the range of the a-to-d inputs.

Figure 2 shows some simple examples of sensors that may be used

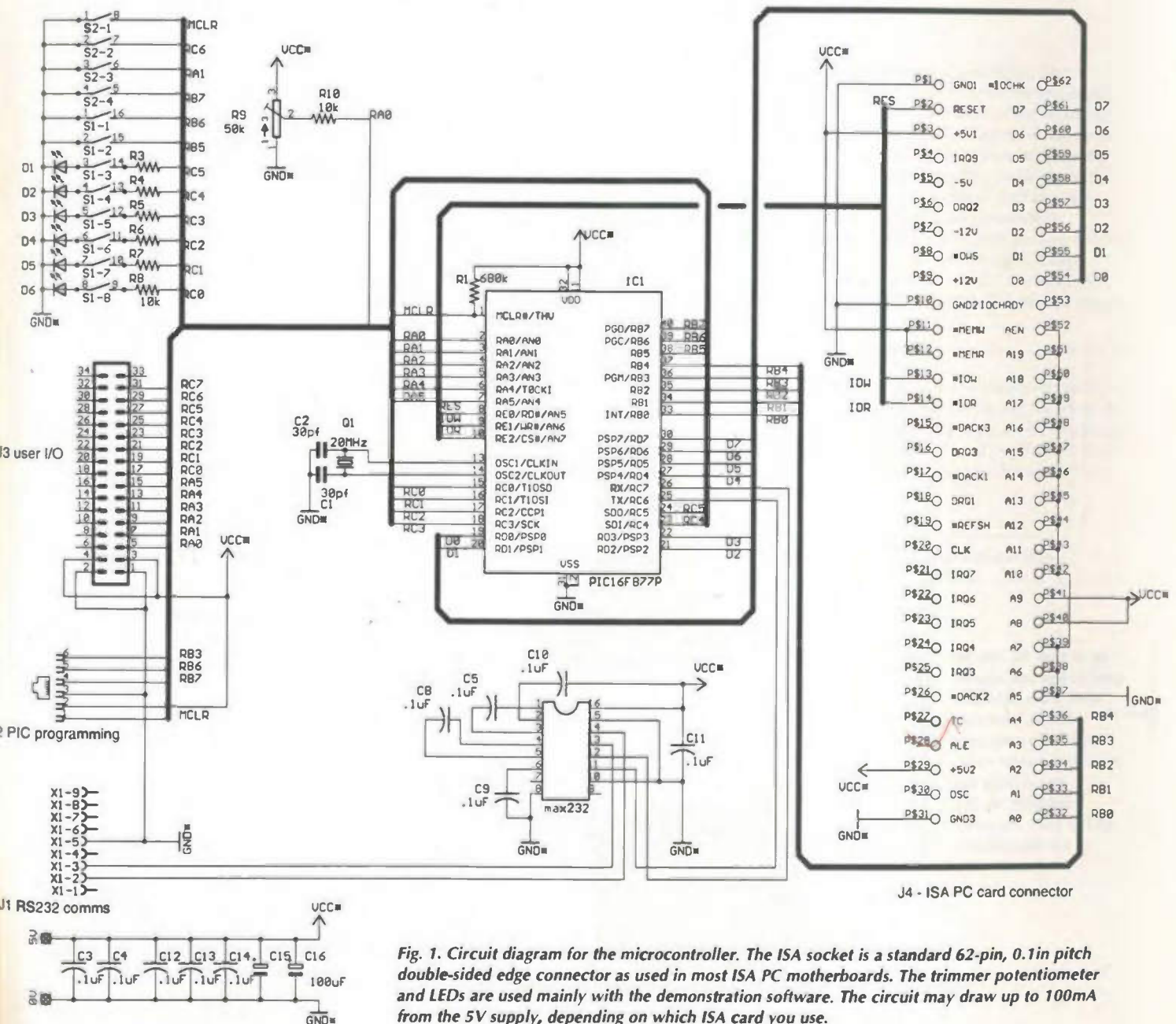
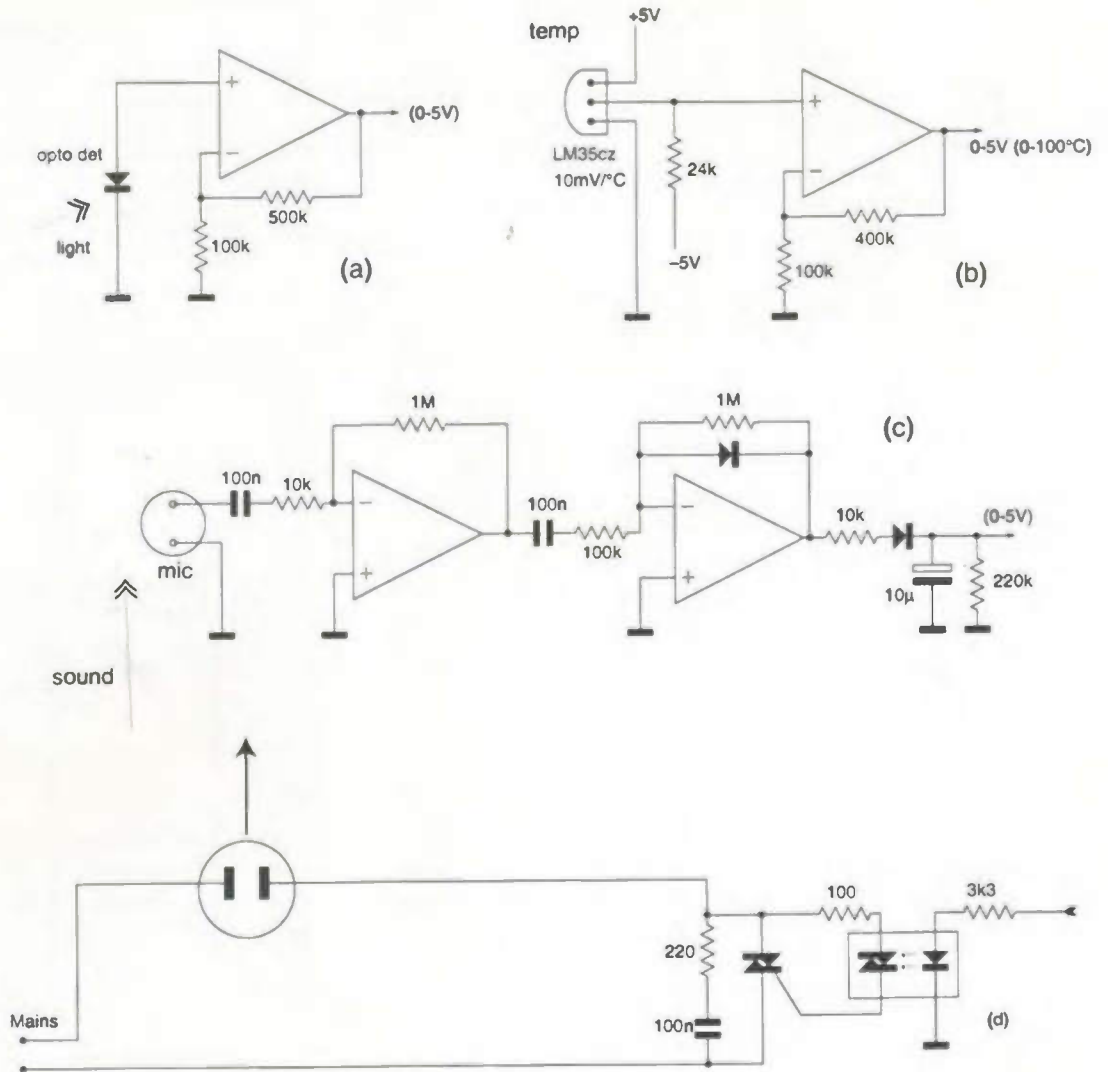


Fig. 1. Circuit diagram for the microcontroller. The ISA socket is a standard 62-pin, 0.1in pitch double-sided edge connector as used in most ISA PC motherboards. The trimmer potentiometer and LEDs are used mainly with the demonstration software. The circuit may draw up to 100mA from the 5V supply, depending on which ISA card you use.

Fig. 2. Ethernet capuccino maker for the person who has everything. Various sensors for temperature, sound and light intensity can be used to feed back data to the PC on request. Data from the PC can be used to drive digital outputs such as the mains switch and timers.



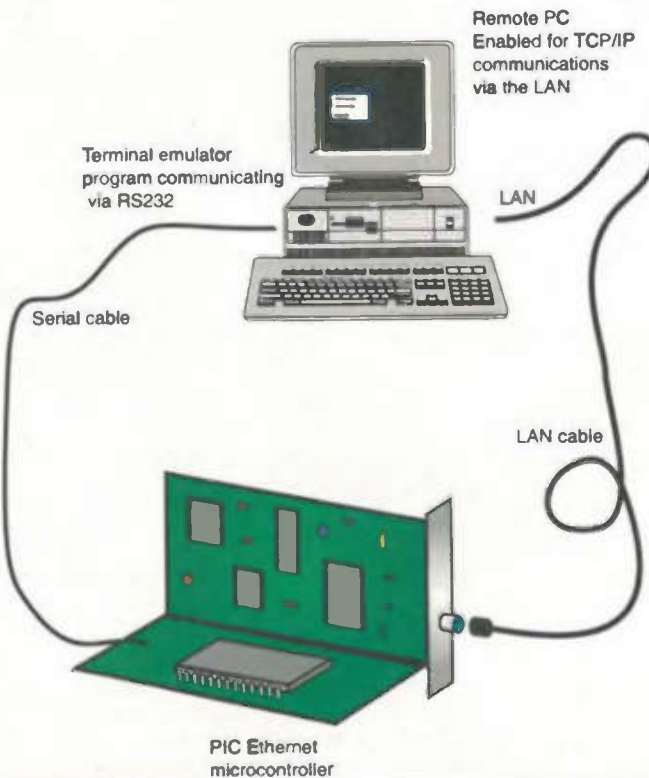
in a hypothetical Ethernet controlled coffee maker. In Fig. 2a), a standard opto detector diode is connected in voltage mode, making it respond to the logarithm of light intensity. The two resistors provide a gain of 6-8 to bring the output to within the 0-5 volt range.

Similarly, in Fig. 2b), a temperature sensor, namely a LM35CZ device with a sensitivity of 10mV/° output, is used with an op-amp with a gain of 5 to bring up the voltage output to cover a 0-100°C span within the required 0-5V range. The LM35 is a pretty accurate device, so 1% resistors should be used if you want to read temperatures to the nearest tenth of a degree or so.

In Fig. 2c), the ambient noise level sensor uses a basic electret microphone with a simple peak level detector to give an output roughly proportional to overall ambient noise levels. Gain and time constants can be adjusted to suit your requirements.

Figure 2d) shows how one of the

Fig. 3. One PC can be used to test and also to drive the controller. The PC uses two Windows programs open at the same time, one driving the controller via the RS232 port, the other via the Ethernet network.



outputs can be used to switch the kettle on and off via an opto isolated triac circuit.

Those of you who are digitally inclined could use devices such as the Dallas one-wire bus temperature probes. These sensors do not rely on external analogue components for accuracy, and deliver their outputs as serial bit encoded data.

Outputs from these can be fed to a separate small PIC, a 16F84 for example, suitably programmed to generate a serial stream at either 57600 or 2400bit/s, which is then feed to the 16F877 via its serial port.

Putting it all together

You now need a network card. Ensure the card is an ISA card, not PCI, and uses the Realtek 8019AS chip-set. The 'AS' version of the 8019 chip includes an internal 16Kb buffer – the non-AS version does not. Many NE2000 generic cards are compatible with the 8019AS, but not all.

There may be a problem if you want to use an existing or obsolete card from your old computer. The 'Plug-and-Play' system in your PC may have initialised it to a different base address value from the default of 0x0300. If this is the case, the micro controller may not be able to access the registers.

Try reconfiguring the card by

Useful references:
 It is impossible to give a thorough list of the products available, as they appear – and disappear – faster than you can write them down on paper. This is just a very short selection. You may also like to try searching on the web under the keywords "embedded and TCP".
www.Siteplayer.com
www.rabbitsemiconductor.com
www.beck-ipc.com/products/catalog/a/chip.asp?status=0_EN_products_a
www.j-works.com
www.lantronix.com/products/embedded/coboxmicro/
www.compulab.co.il
www.ibutton.com

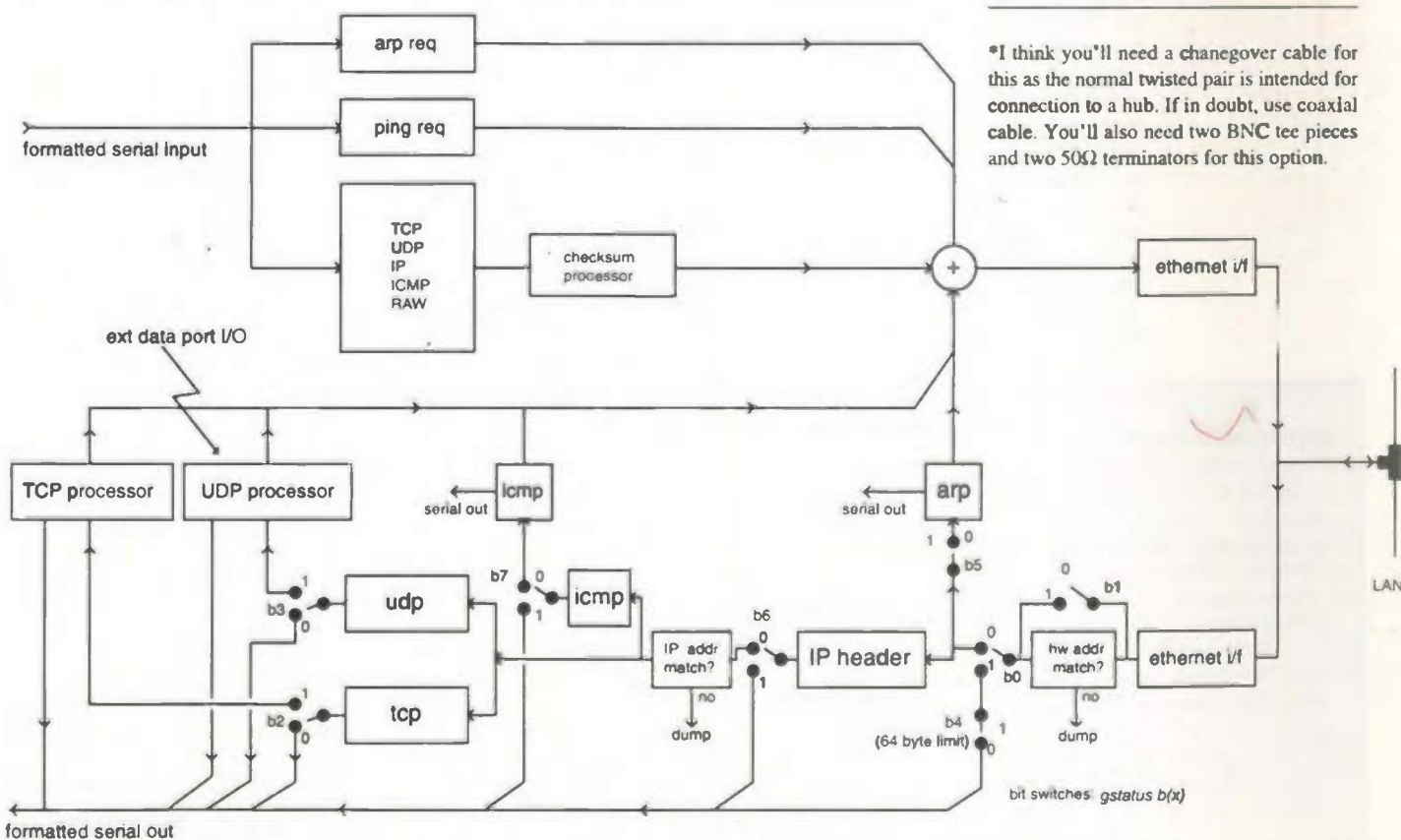
placing it back in your PC. Using the Control Panel in Windows 95 or 98, reconfigure the base address to 0x300. There is no need to change the interrupt settings as they are not used. It may be also possible to disable plug-and-play on the chip by hard wiring one of the pins on the 8019AS chip to V_{CC}. Refer to the chip's data sheet to find out the exact pin number, which will depend on the device package used. You will need a very small soldering iron.

After the PCB has been wired up and connected together, it is time for some tests. Plug the PC Ethernet card into the socket and power the board with 5V. The LED should start flashing at about once a second, indicating the PIC is working and addressing the Ethernet card.

Connect the unit to your PC LAN via the BNC or twisted cable*.

Ensure DIL switches B5-B7 are in the 'off' position. This will default the chip to 57600bit/s, and also enable demonstration mode. It is not essential to use the serial port at this

Fig. 4. Internal signal flow path for the controller. The switches correspond to bit flags in an internal register byte, and can be programmed to modify the path of the incoming signal. This seemingly unconventional architecture allows for various pre-processing levels for the incoming and outgoing data blocks. An external processor handling serial character streams is then presented with 'pre-cooked' data blocks, ranging from Ethernet raw packets to processed TCP segments. The controller also adds check sum and size entries to outgoing packets to simplify the user software. Using simple external processing power, the device can form the basis for a network analyser, embedded mail system or simple web server.



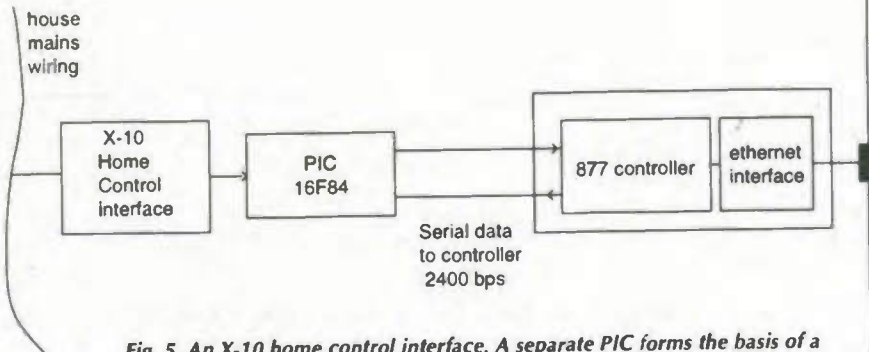


Fig. 5. An X-10 home control interface. A separate PIC forms the basis of a simple interface between X-10 home automation and the local area network. No interface standards exist at present for this kind of interactivity.

stage, but it could be very useful for performing initial tests and for generally seeing what's going on. Connect the PIC serial port interface to a spare COM port in your computer.

You can use any terminal emulator software if you have one in your PC. Alternatively, use the supplied program "877driver.exe". This program has facilities for transmitting pre-programmed ASCII sequences making any setting up easier. The program communicates at 57600bit/s, so make sure that bit 5 of the DIL b5 is 'off', i.e. PORTB-5 or pin 38 of the PIC is not connected to ground, Fig. 3.

With '877driver.exe' running, resetting the controller will cause the string '<Pp>' to be sent to the PC, and be shown on the screen. This is an indication that the card is working. The character pair "Pp" is specific to Realtek 8019 chips. You may get a different character pair if your card uses other chip sets.

If the string is not displayed at all, the Ethernet card is not being

addressed by the PIC properly. In this case, refer to the previous note on plug-and-play addressing.

Once you get a response, try querying the controller for its IP address. You do this by entering the single letter Q in the transmit Edit control, or alternatively by clicking the 'Query variables' tick box, followed by a click on the 'xmit' button to send the command to the PIC.

The PIC will return a string of bytes corresponding to the current MAC and IP allocations for the card. To display these properly, you must ensure the screen display is set to receive in hexadecimal notation. Just click on the appropriate 'hex display' box.

If this is the first time that you are using the device, you may get the 'factory default' values. You can change these if you like, using the various options provided in the Windows software.

The default IP address is '192.68.0.15'. Some people prefer to use numbers ranging from '10.0.0.1'. The exact numbers used will not matter as long as each workstation in your domain has a different address.

If your PC is part of a larger system, contact your network manager, who will suggest a non conflicting address that you can use.

You will also need to know the IP address of your PC workstation. This can be found in Windows' Control Panel under the Network applet. If your PC is part of a larger network, your IP address may be dynamically allocated, so it may change from session to session. Consult your network manager, who will be able to allocate your workstation with a fixed IP address.

The address values are retained in

the PIC's EEPROM for future reference, so you will only need to enter them once. If you know your PC's allocated IP address, enter it under 'remote IP'.

Open an MS DOS console on your PC and use the *ping* and *arp* commands to check that the PC is talking to the controller and that all is working properly. If there is no response, ensure the controller address matches the requested ping address.

Also make sure that the network card green LED is flashing. This shows that packets are being fired across. Disk files supplied with the chip cover the above instructions in more detail, including examples and lists of commands available.

And now for some excitement...

With '877driver.exe' running, enter the two digits '03' in the 'set status flag' box. Then set the tick next to it, and click on 'xmit' to send this command to the PIC.

Now watch the PC screen. Maybe nothing will happen. If this is so, get your PC to access other workstations on the network, e.g. via Network Neighbourhood in Explorer, or start a network printer. The screen display will suddenly start filling with packet information.

You are now watching your network connection and all packets that float past it. The command you have just entered has set the PIC to act in a 'receive all packets' mode.

Figure 4 shows how the internal flow in the PIC depends on this internal global flag. The data will quickly fill your screen, so press the 'clear screen' button every so often. You may notice that even when all the workstations are idle, there will be the odd packet being displayed. Soon you will easily differentiate between ARP, ICMP, DHCP and TCP messages, and with a bit of practice – and help from the literature – deduce where they come from and their purpose.

Can it flash lights?

On the 'set status flag' box, enter the two digits '08', set the tick next to it, and click on 'xmit'. This sets the PIC to demonstration mode. At this point, you don't need '877driver.exe' any more.

Now run the program '877demo.exe'. This program is a Windows version of the listing shown last month. It sends a UDP packet to the PIC via the network, retrieves a response, and then displays the results.

The GUI front end lets you tick

Technical support

A set of parts is available for this project for £35 including VAT and p&p in mainland UK. The kit consists of a preprogrammed PIC, PC demonstration software, application notes and an instruction manual.

More information on devices for the other applications mentioned is also available. Please contact Eddy for more details: e-mail edinsam@eix.co.uk or write to Eddy Insam, BelleVue, London Road, Harrow on the Hill, Middlesex HA1 3NH with an SAE.

You can obtain the hex code for the controller from Eddy's web site at:

<http://www.eix.co.uk/Ethernet>

boxes to set any LEDs you want to turn on, and will also display the returning data as digital or analogue values. The program is very simple. It operates on the five least-significant bits of port C (digital output), and on return, reads port A bit 0 as an analogue voltage, which is displayed as a voltage between 0 and 5 volts.

With '877demo.exe' running, ensure the IP addresses shown in the top two windows correspond to that of your PC, and that of the PIC controller. Click 'send request', and wait for a response. If nothing happens, check again that all the IP addresses are correct. Do not forget to ensure the PIC board is set to demo mode as described. Reset the PIC with all the DIL switches in the off position to guarantee this.

Once you start getting a response, tick some of the LED selector boxes and click on 'send request' again. The LEDs on the PIC board should start turning on or off according to your settings. Try adjusting the trimmer on the PIC board, and watch how the progress bar follows it on the PC screen every time you transmit a request.

Full C++ source code is provided so that you can customise this program to your needs. The software was developed in Microsoft Visual C version 6. It should not be too difficult to port this to other environments such as Borland or Basic.

In practice you will only need to deal with one function or subroutine, which is fully documented to make it as easy to understand as possible. The rest of the program is just the standard template functions used by the Visual C environments to provide visual support for dialogue boxes, user buttons and others.

How fast can it work?

You should be asking this question, but a more relevant one is, "how fast can data be moved across the network?" I have used the controller to deliver a stream of 8-bit speech from a microphone at 8kbit/s, and I am sure it could easily run faster than that. The controller itself is purely limited by its ability to read or write the 8019 buffers via its I/O ports.

However, there are many other factors to consider as well. In practice, the main limit is the network itself, i.e. how busy it is. There is a major difference in performance whether you want to transmit short bursts of high speed data, or a continuous stream of bytes at a guaranteed rate. Starting a network printer or running a file transfer

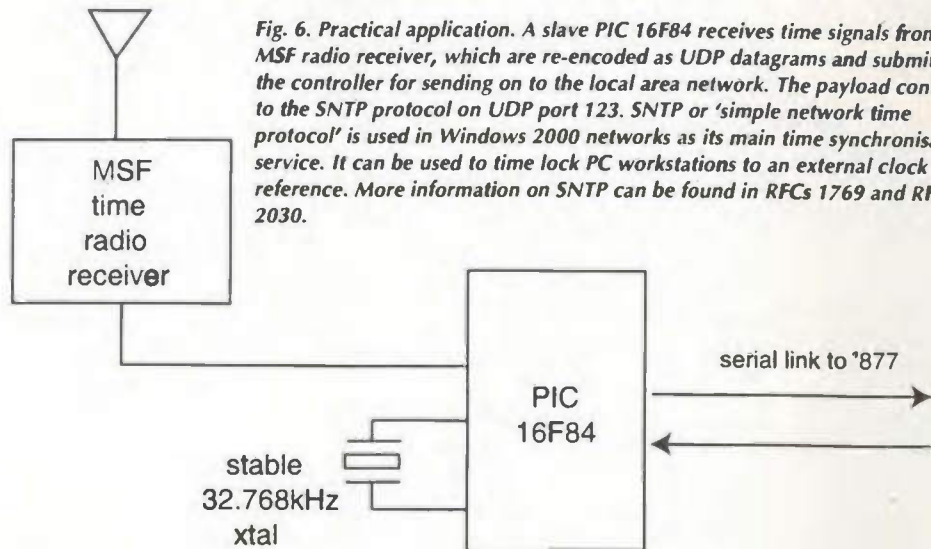


Fig. 6. Practical application. A slave PIC 16F84 receives time signals from the MSF radio receiver, which are re-encoded as UDP datagrams and submitted to the controller for sending on to the local area network. The payload conforms to the SNTP protocol on UDP port 123. SNTP or 'simple network time protocol' is used in Windows 2000 networks as its main time synchronisation service. It can be used to time lock PC workstations to an external clock reference. More information on SNTP can be found in RFCs 1769 and RFC 2030.

between two PCs causes a noticeable flow disruption. This makes it impossible to guarantee a particular data transfer rate.

What else can it do?

The controller provides various levels of functionality. At the lowest level it can deliver raw formatted Ethernet IP or TCP packets via its serial port. The packets are enclosed in a simple header trailer arrangement, and simple software can be written to decode them. This is useful if you want to build a simple network 'sniffer', want to develop your own protocols, or just want to see what's going on in your network.

At the highest level, it can provide a point-to-point data stream with all necessary ARP and ICMP support handled internally and transparently. A single flag variable controls the overall data flow, Fig. 4.

On transmission you just supply the data component of the packet, the PIC will add all check sum and size fields, thus simplifying their generation.

A powerful combination is obtained when using the controller with another device, e.g. another PIC. The serial port is an effective method of communicating between the two and for passing both commands and data.

In a typical context, a separate PIC is used to accept data signals from the environment, perform the simple conversions, and to control the 16F877 PIC controller using plain byte and text commands. This is a very powerful option as you can develop your own software using your own device, avoiding the complexities of network protocols.

The example in Fig. 5, duplicated

here from last month's article, shows how an interface to a home X10 network could be wired up. The 16F84 shown has a simple job; it only needs to convert serial data sent from the PIC controller – at 2400bit/s in this case – to X-10 type commands.

Another example of a two-chip application is shown in Fig. 6. This forms the basis of a simple PC network-wide atomic clock reference. The circuit shows a 16F84 used to decode bit data stream signals from a standard Rugby MSF receiver. The '84 performs some simple conversion on the time received, and generates a simple pre-formatted UDP data packet containing the clock information which is reformatted as a 32-bit binary word.

The exact protocol and format is known as SNTP and is fully described in RFC 1769 and 2030. SNTP is the main method used in Windows 2000 and other networks to align all workstations to an external precision clock. SNTP is enabled in Windows by using the NET TIME command. More information can be found in the <http://msdn.microsoft.com> technical information web site by searching under 'SNTP', 'Q216734' or 'Q224799'.

What you can't do

This Ethernet interface system's main limitation is the very small amount of RAM space available within the PIC chip. Input data is streamed, i.e. not stored, so incoming packets can be of any size. However output datagrams are buffered within the chip.

The buffer is dynamically allocated

to whatever remaining RAM is available at the time, usually between 80 and 200 bytes. This is enough for quite a few applications, but not enough for everything. I tried using the internal RAM buffer of the 8019AS for this purpose, but it appears it cannot be used in full duplex mode, i.e. for doing DMA read and write transfers at the same time.

This buffer ceiling places a limitation on any applications that needs to deal with large amounts of data. This means you won't be able to build sophisticated applications such as a web server that require large amounts of data. For this, you'd be better off buying ready-made cards and devices with sufficient spare RAM or ROM.

Network security

The controller was originally developed as a custom tool for validating security and throughput aspects of routers and fire walls in both IPv4 and IPv6. Recent concerns about overloading systems using low level SYN attacks, incomplete IP fragments and other difficult to trace methods have made this a very relevant issue.

Note that using the controller in a shared network can cause real havoc if used irresponsibly. It can be distressing to think that such a small device can be used to bring an apparently secure network to a grinding halt, and without leaving any traces.

So if you want to use the controller in a shared office, lab or a college, make sure that you know what you are doing. Otherwise disconnect the cable to the rest of the network.

New Standards. Do we need them?

Nowadays, print servers are using TCP and SNMP for their configuration screens. The user just clicks open a standard internet browser to communicate with the print driver, or with the printer itself by means of a standard web page.

This has great advantages for the

printer manufacturer. It means that special configuration software programs for the PC are unnecessary. Everything is done via web page commands and using a normal web browser.

Of course, the Ethernet connected printer or appliance has to act like a fully-fledged web server. But as shown in this article, this is not a difficult task, if you know how and have enough computer power.

For point-to-point and time-critical applications, some of these protocols are not efficient ways of providing communications. Some concepts such as safety, feedback paths and constant throughput have barely been considered in the present standards, which are mainly aimed at user oriented systems. So there is a possibility that new, or variations on existing standards may be developed specifically for certain areas of appliance and embedded device control and communications.

Printers? In a nutshell, IP provides for basic routing facilities, so any new standard will reap their benefits if developed on top of this. Protocols based on UDP are fast and easy to implement, but as UDP contains no in-built error management, facilities would have to be provided by the higher layers. This could result in performance degradation if not done properly.

TCP on the other hand, provides a reliable flow, but it requires disproportionate overheads both in terms of extra data and time delays. Many embedded devices may require a reliable connection with quick response, but not necessarily a time consistent circuit, so TCP could easily degrade efficiency. There is plenty of scope for experimentation and new ideas.

Of course, this may not happen at all. The Hayes AT modem command set is still in common use for computer driven applications, complete with its idiosyncrasies and computer unfriendly command set.

What next?

I hope that these two articles will have made you a bit more knowledgeable about using the Ethernet in your applications. My main purpose was to bring the concept of network aware devices from a 'bottom up' point of view, and help to dispel some of the black magic associated with ready-made boxed solutions.

If there is enough interest, I will produce more articles, perhaps on more specific projects, and using some of the latest devices available. ■

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The author

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If you are considering a serious application for this device, Eddy will endeavour to answer your queries via e-mail.

Designing with DSP

Using a design example incorporating audio data converters, **Patrick Gaydecki** reveals that the DSP hardware needed for communicating with a PC, memory addressing and codec interfacing is not too difficult to implement.

In the first article in this set, I introduced the general concept of real-time DSP. I also looked at some of the internal design features that allow DSP chips, such as the DSP56002, to perform arithmetic operations so quickly and efficiently.

Typical hardware requirements are covered in this second article, together with a simple design example incorporating the DSP56002 as the core processor.

Implementing DSP hardware

Figure 1 shows the pin assignments of the DSP56002; pins are grouped according to function. These are described in detail in the DSP56002 technical data document supplied by Motorola.¹

Although quite straightforward in concept, the design is suitable for advanced stereophonic signal processing of audio bandwidth signals, as is shown in Fig. 2. This design involves a dual-channel audio codec and 32K-words of external memory for holding both code and data. It also has facilities for communicating directly with the serial interface of a PC.

Because the design has been kept simple, it does not service interrupts generated by external hardware. However, this is not an issue here since the system is intended as a stand-alone processing environment.

It can of course respond to interrupts generated both by software and internal hardware systems: there's more on this later.

As Fig. 2 also indicates, a number of the control signals

are tied to the power rail or ground by 15k Ω resistors. This is because of the operating modes that have been selected for this system. For more detailed information, you should investigate the relevant Motorola literature.^{2,3}

Resetting and clocking

The DSP56002 is reset by a low-going pulse on its input. A suitable reset circuit is easy to construct, involving a capacitor, resistor and a Schmitt trigger such as a 74HC14.

As Fig. 1 shows, the DSP56002 may be clocked by connecting a crystal, a resistor and two capacitors to the pins labelled XTAL and EXTAL. It can also be clocked more simply by connecting a clock oscillator module to EXTAL. If you are using an oscillator module, XTAL should be left unconnected.

In common with many modern microprocessors, the DSP56002 incorporates a phase-locked loop (PLL) that can be programmed to multiply the externally applied clock frequency, generating a higher internal clock frequency. This enables a relatively low-frequency clock oscillator to be included in the design, minimising interference and EM radiation.

The PLL can be programmed not only with a multiplication factor, but also simultaneously with a division factor, thus generating non-integer multiples of the externally applied clock frequency.

If the PLL is used, a capacitor must be connected between the PCAP input and V_{cc} . Its value is given by $540\text{pF} \times MF$, where MF is the multiplication factor chosen. Hence if the external clock is 10MHz, and an internal clock of 60MHz is required, the capacitor must be 3.24nF; 3.3nF is suitable.

Initially, the PLL should be disabled by taking the PINIT

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input low. This condition is changed via programming after reset, as discussed in a subsequent article.

Communicating with other hardware

As I mentioned in the previous article, port C is a triple-function port that can act as a general-purpose I/O interface, as a three-pin serial communications interface (SCI) port or as a six-pin synchronous serial interface (SSI) port.

Since port C comprises nine pins, SCI and SSI modes can be made available together. To configure port C in a given mode or combination of modes, you load the associated internal control registers with an appropriate word, i.e. a bit-pattern; more on this in a later article.

For communicating with a PC, the SCI port is used. This is a full duplex interface; in asynchronous mode, used here, it can communicate at a maximum of 625kbit/s assuming a 40MHz clock. This port can also operate in synchronous mode, allowing a maximum communication rate of 5Mbit/s, again with a 40MHz clock.

Since the serial interface of the DSP56002 employs TTL, the voltage levels must be converted to those expected by the RS232 protocol, involving $\pm 12V$ inverted logic, prior to interfacing with the PC's serial port. This is achieved with a MAX232CPE chip, as in Fig. 2.

Note that this device uses a number of capacitors that are not shown in the diagram for purposes of clarity. These capacitors act as reservoirs for the charge pumps. Refer to manufacturer's data sheets for more information on these.

Under normal circumstances, the SCI can be made to operate like a standard minimum-mode connection serial

interface, requiring TX, RX and a common ground.

However, in bootstrap mode, an external clock signal must be supplied to the SCLK input. This should be a square-wave signal that is 16 times the bit rate of the interface. Hence if used at 19200bit/s, the clock signal frequency should be 307.2kHz. After reset booting, communication with the SCI no longer needs this clock.

As I have already shown, the DSP56002 can operate in one of several modes; the system described here operates under 'Mode 1', the special bootstrap mode. In this mode, following a hardware reset, the DSP downloads a program either from the SCI, as used here, from external boot memory, or from the host interface (HI). It then stores the code in its internal memory, and starts executing it.

Programs may be written on any text editor using DSP56000 assembly code and assembled into an object file. It is also possible to simulate the software's operation with a PC-based simulator. Both the assembler and the simulator are available free of charge from the Motorola DSP web site (<http://www1.motorola-dsp.com/>).

The assembled object file has a specific format known as a CLD file, which has some header information. In order to download this directly to the memory space of the DSP56002, this header information must be stripped off to leave a pure binary file. This file can now be directly transmitted to the DSP56002 from a PC via the serial interface.

The serial interface bootstrap mode is selected by setting appropriate logic levels on the control pins termed MODA, MODB and MODC during reset (0, 1, 1 respectively, as shown in Fig. 2).

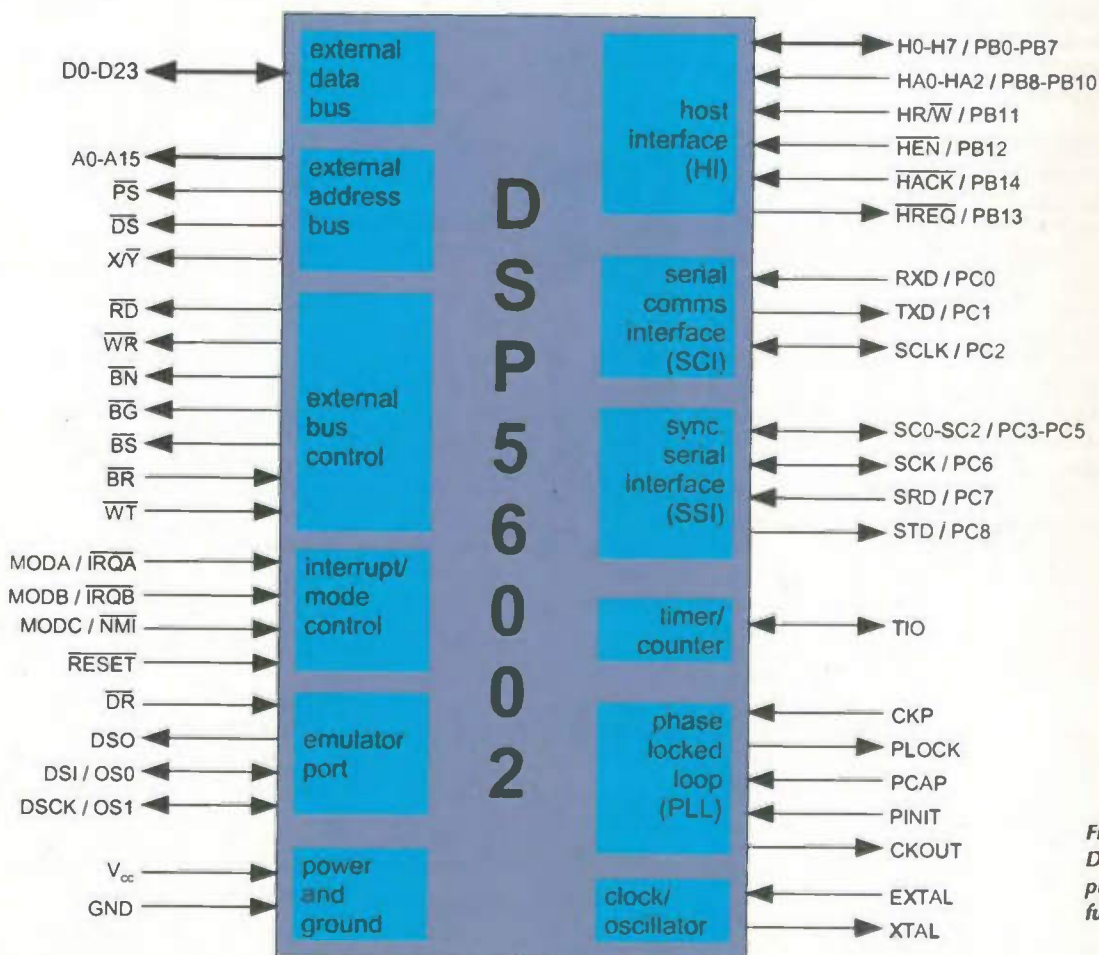


Fig. 1. Pin-outs of the DSP56002 digital signal processor, grouped by function.

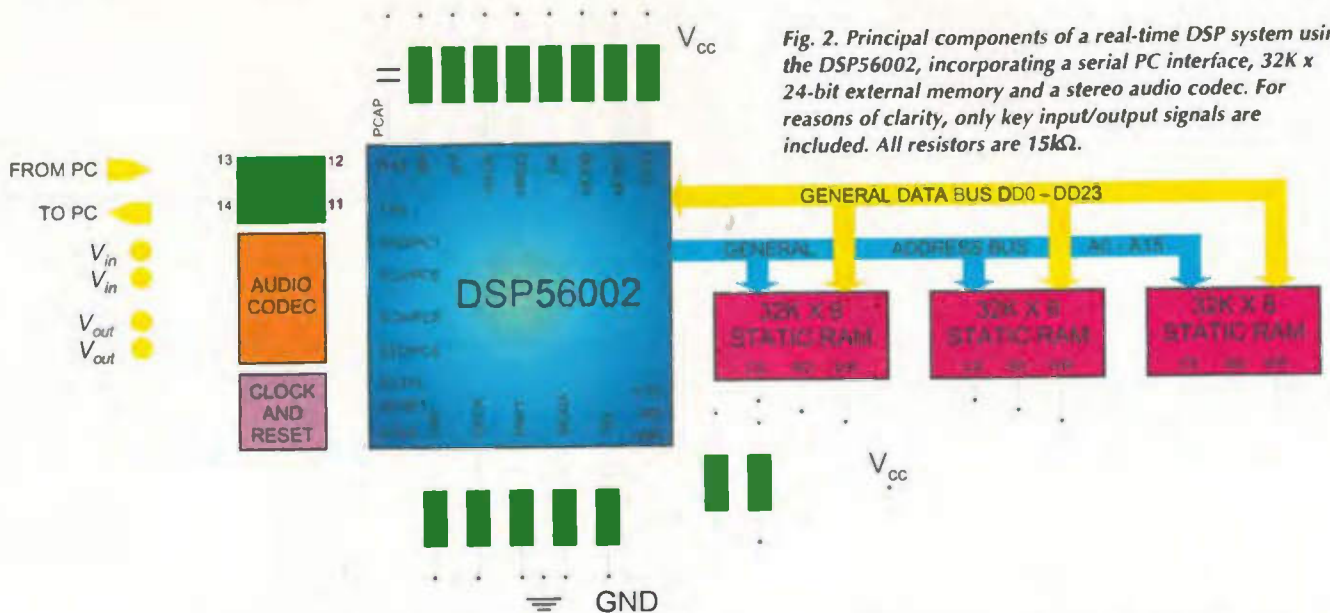


Fig. 2. Principal components of a real-time DSP system using the DSP56002, incorporating a serial PC interface, 32K x 24-bit external memory and a stereo audio codec. For reasons of clarity, only key input/output signals are included. All resistors are 15k Ω .

Bootstrap code must be preceded by three bytes that specify the number of program words and three bytes that specify the program start address. Following this, the code proper is downloaded, comprising three bytes for each program word loaded.

It is essential that the least significant byte is received first and the most significant last. After the final byte has been loaded, the code immediately commences execution at the start address specified.

It goes without saying that this facility is enormously useful. Effectively, it means that a simple emulator or evaluation module can be constructed for less than £50, and the results of any program changes can be seen within seconds.

External memory system

If program and data requirements are small, the DSP56002 can operate using just its internal memory. For more adventurous designs, external memory can be added.

The DSP56002 has 16 address lines and can therefore address 64K words of a particular memory type, each word being 24 bits in width. This requires a little explanation.

As mentioned in the first article, the DSP56002 exploits super Harvard architecture. Code and data memory space is segregated, and the data memory space is further subdivided between X data and Y data memory. Internally, all three memory spaces have their own address and data buses.

The three address and data buses are multiplexed into single bus structures externally. Harvard design is maintained by using control signals that determine the type of the external memory being accessed. Thus in total, the DSP56002 can address up to 64K words of code, 64K words of X data and 64K words of Y data memory.

This design incorporates 32K words of mixed-mode external memory; in other words, the design makes no distinction between the different memory types and the external memory can be used to hold code, data or both, in any combination. Although this does not compromise the speed of the system – since the external buses are multiplexed – it does mean the programmer must be aware of where code and data reside.

Figure 2 shows that three 32K by 8-bit static RAM chips are combined to hold the 24-bit words. The first chip holds the low order byte and is connected to data lines DD0-DD7; the second holds the mid-order byte, and is

connected to data lines DD8-DD15; the final chip holds the high-order byte, and is connected to data lines DD16-DD23.

All address lines are connected to A0-A14. The chips are enabled, i.e. selected, by connecting the A15 line to the inputs of each memory device. Only if this line is low during a memory cycle, i.e. an address is generated equal to or below 32K, will the external memory be activated.

At this point it is worth mentioning that the internal program memory resides at addresses between 0 - 511 (0-1FF₁₆), and the X data and Y data memory reside at addresses between 0 - 255 (0-FF₁₆).

If code or data are read from or written to any of these locations, the DSP56002 will not activate the external memory, i.e. it will keep line A15 high. This ensures that internal memory is used, which has a clear advantage in terms of access time.

Timing requirements

In order to select the appropriate external memory devices for the system, it is essential to consider the speed of the various bus signals and the timing relationships that exist between them. A handful of simple equations is all that is required for this procedure. It is important though to navigate the method with due diligence, since a mistake will lead to a faulty design which will simply not work.

Standard static RAM suitable for general-purpose microprocessors and microcontrollers is no use here; fast static RAM must be employed, with access times of less than 30ns.

Specifications necessary to determine the access time of the memory to be used are the external clock period, t_{ce} , the internal clock period, t_c , the internal clock low period, t_l , the PLL division factor f_d and the PLL multiplication factor f_m . These are discussed in detail in various Motorola technical documents.⁴

For a read cycle, these specifications allow you to calculate the time between the address becoming valid and the data appearing on the data bus. This is termed t_{aa} . For a write cycle, they allow you to calculate the time between the data becoming valid on the data bus and the de-assertion of the signal. This is termed t_{sd} . The relationships are shown in Fig. 3. The two parameters t_{aa} and t_{sd} are obtained using the following expressions,

$$t_{aa} = t_c + t_l - 7ns \quad (1)$$

and,

$$t_{sd} = t_l - 0.4ns$$

where,

$$t_c = t_{ce} \frac{f_d}{f_m}$$

and,

$$t_l = 0.467t_{ce} \frac{f_d}{f_m}$$

Any memory used in the system should have values for t_{aa} and t_{sd} less than or equal to those calculated for the DSP56002, for given external and internal clock rates. For example, if the DSP56002 is clocked externally by a 10MHz oscillator with a PLL multiplication factor of 6, this yields a value for t_{aa} of 17.5ns and a value for t_{sd} of 7.4ns.

Motorola and other semiconductor manufacturers supply fast static RAM suitable for real-time DSP systems. In this case, a Cypress CY7C199-10 (32K x 8-bit fast static RAM) would be ideal. Figure 3 shows the timing involved in memory accessing.

The audio coder/decoder

Most DSP chips are designed to work with a wide range of different a-to-d and d-to-a converters, produced by different manufacturers, all of which have slightly different data formats.

Furthermore, in order to minimise the physical number of pins of VLSI devices, many a-to-d and d-to-a converters and DSP chips make use of serial communication for digitised data streams.

Digital processors, and the DSP56002 is no exception, can be configured for these different formats by loading appropriate bit patterns in the relevant control registers. If a-to-d and d-to-a converters are combined into a single package, it is generally referred to as a codec.

Figure 2 shows a simple interconnection strategy for interfacing a codec – or separate a-to-d and d-to-a converters – to the SSI of the DSP56002. Because the device is synchronous, for any one device it requires a minimum of three signal lines.

Considering the a-to-d converter requirement first, it requires a framing signal that delimits the start and end points of a datum word sent by the a-to-d converter. This connects to pin SC2 of the SSI – also known as pin PC5, since it can be configured as a general purpose I/O pin of port C.

Next, it requires a serial bit clock whose negative transitions indicate to the DSP56002 when to sample the individual data bits. This connects to pin SCK/PC6. Finally, it requires a signal line carrying the data in serial form, connected to pin SRD/PC7.

For a d-to-a converter, an identical arrangement is used except that the serial data are transmitted by the DSP56002. Thus instead of SRD/PC7, the pin labelled STD/PC8 is connected to the data input of the converter.

If the a-to-d and d-to-a converters are combined as a codec, then the framing signal and serial bit clock are common to the input and output sections of the device; thus a codec will require a minimum of four signal lines.

The SSI can be configured to operate in slave or master mode; if in slave mode, the framing signal and serial bit clock are provided by the codec; if in master mode, they are generated by the SSI.

Figure 4 illustrates the various timing waveforms associated with the analogue voltage input/output process operating in normal mode. More complex arrangements

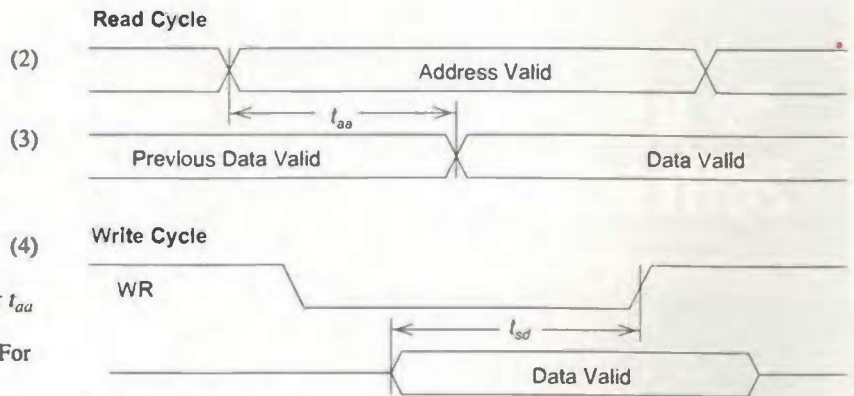


Fig. 3. Critical timing parameters for calculating access times of static RAM used with the DSP56002.

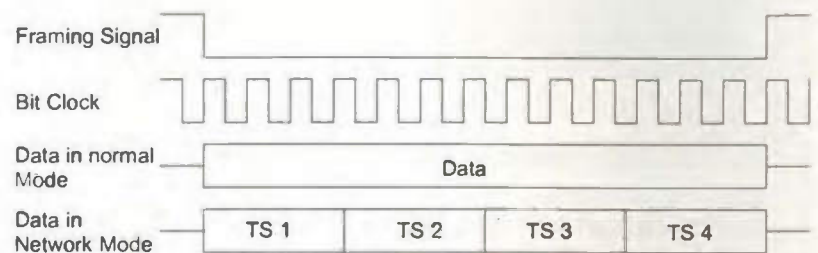


Fig. 4. Synchronisation and data signals for transfer of information between a codec and the SSI.

are also possible, and further information is provided in an article that has appeared previously in this magazine.⁵

An important property of the SSI interface is that the number of signal lines required does not increase if a multi-channel codec is used – for example a stereo or quadraphonic device.

With a multi-channel system, the SSI is set up to operate in what is termed network mode, also shown in Fig. 4 and is used in this design. In this protocol, the framing signal delimits the start and end points of all the datum words within one sample period.

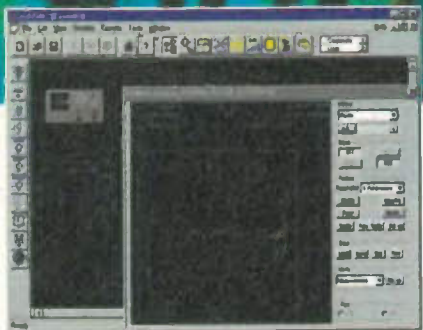
The bit clock remains the same, and the serial data are multiplexed on the lines connected to SRD/PC7 and STD/PC8 in packets called time slots. Clearly, when the SSI is configured, information must be supplied that details the resolution of the a-to-d/d-to-a converter or codec system. Information is also needed as to whether the device must operate in normal or network mode, and the length of the time slots if the latter mode is used. How the SSI registers are configured for codec operation is discussed in a fourth article.

In the third and next article, I will introduce the DSP56K assembly language set, and explain why it facilitates powerful, efficient and high-speed real-time digital signal processing. ■

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5. Gaydecki P, 'Interfacing Digital Audio', *Electronics World*, December 1999, pp. 1026-1030.

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"Liesegang diaty" automatic slide viewer with built in high quality colour TV camera. It has a composite video output to a phono plug (SCART & BNC adaptors are available). They are in very good condition with few signs of use. For further details see www.dfatv.co.uk

£91.91+ vat = £108.00

Board cameras all with 512x582 pixels 8.5mm 1/3 Inch sensor and composite video out. All need to be housed in your own enclosure and have fragile exposed surface mount parts. They all require a power supply of between 10 and 12v DC 150mA.

47MIR size 60x36x27mm with 6 infra red LEDs (gives the same illumination as a small torch but is not visible to the human eye).....£37.00 + vat = £43.48

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40MC size 39x38x27mm camera for 'C' mount lens these give a much sharper image than with the smaller lenses.....£32.00 + vat = £37.60

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1206 surface mount resistors E12 values 10 ohm to 1M ohm 100 of 1 value £1.00 + vat 1000 of 1 value £5.00 + vat

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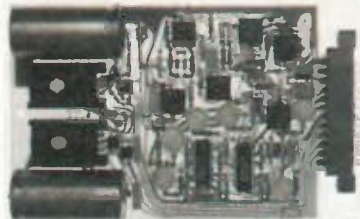
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NEW PRODUCTS

New Products are edited by
Electronics Weekly

Power supplies suit network 1U uses

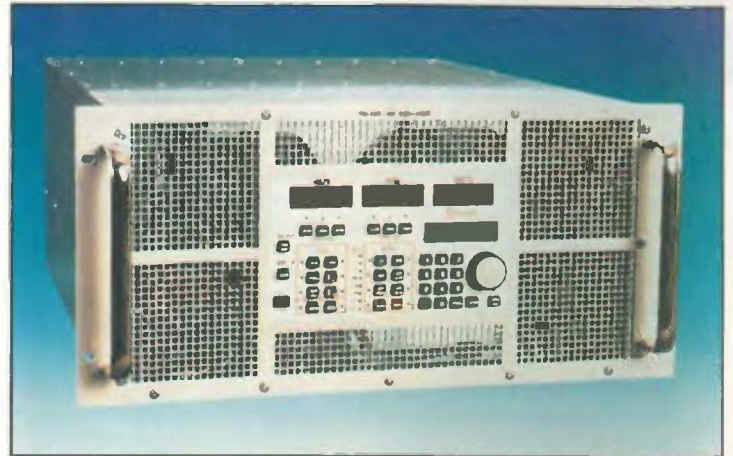
Lite-On Power supplies are for networks and communications in 1U applications. They deliver 12, 24 and 48V DC with combinations available. Active PFC complies with IEC1000-3-2, having a typical power factor of 0.99. Size is 17.8 by 12.7cm by 1U for 350W units and 19 by 10.2cm by 1U for 200W models. Efficiency is 85 per cent at full

load. With an input of 85 to 264V AC, they are also available with active current share and hot plug ability, while airflow requirements are 11.9m³/hr.

Lite-On Power
Tel: 0191 427 0255
www.liteon-power.com

Electronic loads rate at 1kA to test cells

The Dynaload RBL single-channel electronic loads have ratings of 1kA and are for testing batteries, power supplies, fuel cells and related DC power equipment. From Thurlby Thandar, the loads have a constant power feature to aid battery testing and analysis. The internal pulse generator provides variable slew rate up to zero to full scale in 10µs. They can perform below 0.5V at up to 1kA making them suitable for fuel-cell testing. They operate in five modes – constant-current, constant-resistance, constant-power, constant-voltage and



pulse. Multiple units can be synchronised as master and slave units that are controlled simultaneously using analogue control or the GPIB interface on the RBL488 model. Voltage is from 0 to 1kV and there are three power ranges: 0.8, 2 and 4kW.

Thurlby Thandar Instruments
Tel: 01480 412451
www.ttinst.co.uk

Flexible encapsulant in transparent layer

Dymax 9-20558 thick film coating from Intertronics provides a transparent layer to protect circuits and components from moisture, dust and contaminants. It cures in



seconds on exposure to UV and visible light. The flexible encapsulant adheres to and forms a coating over glass-filled epoxy, metal, ceramic and Kapton. It has a secondary heat-curing capability and flow characteristics for quick coverage without running. Cure time is 20 to 30s, even under 200mW/cm² longwave light. Cure depths of 0.635cm are possible.

Intertronics
Tel: 01865 842842
www.intertronics.co.uk

Electrolytic capacitor can handle heated situations

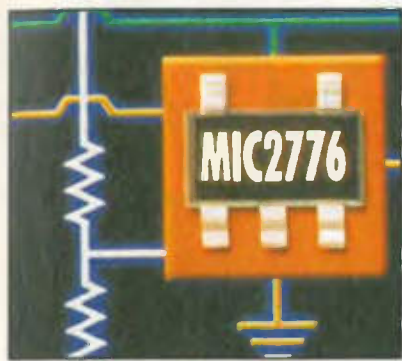
BC Components has announced an SMD electrolytic capacitor, the 140 CLH. Applications include mobile telephone network infrastructure, engine management, ABS systems and ovens. It can be used in environments at temperatures up to 150°C for short periods; useful life is up to 1500 hours at 125°C. It can also be used for filtering, buffering or smoothing voltages in applications where life times over ten years at 60°C are required.

BC Components
Tel: 00 31 40 259 0724
www.bccomponents.com



NEW PRODUCTS

Please quote *Electronics World* when seeking further information



Power supervisor monitors supplies

An adjustable supervisor IC for monitoring power supplies down to 300mV has been developed by Micrel. Available in the firm's Ittybitty SOT23-5 packaging, the MIC2776 has a choice of output options. A manual reset input is provided for a reset switch or for daisy-chaining it onto existing power supervision circuitry. It is made using a BiCMOS process. The three versions are the MIC2776H-BM5 with an active-high output, the MIC2776L-BM5 with an active-low output and the MIC2776N-BM5 with an open-drain, active-low output. *Micrel Semiconductor*
Tel: 01635 524455
www.micrel.com

Line interface uses ADSL card chip set

Advanced Power Components has launched the APCC78120 line interface hybrid component for ADSL line cards. It uses Centillium's multimode ADSL chip set. The hybrid contains the necessary transmit and receive filtering components, line drivers and matching components between the chip

set and the line interface transformers, occupying less than 3.9cm² per channel. It replaces discrete components with one module.

APC
Tel: 01634 290588

Embedded workbench includes promo

The latest release of IAR Systems' embedded workbench for Mitsubishi M16C microcontrollers includes a promotional version of the Segger Embos real-time operating system. *IAR Systems*
Tel: 00 465 18 167800
www.iar.com

Quad amplifier aids signal processing

Zetex has launched an amplifier for video and high-speed signal processing applications. The ZXFV201 has a bandwidth of 300MHz, slew rate of 400V/ μ s and output drive of 40mA. The device is stable with up to 400pF capacitance load. Differential gain is 0.01 per cent and differential phase 0.01°. Power supply requirements are $\pm 5V$ at a current of 7mA per amplifier. The device is for -40 to +85°C operation. Applications are in video amplification and buffering, RGB driving, video distribution, high frequency instrumentation and medical imaging. *Zetex*
Tel: 0161 622 4422
www.zetex.com

Bluetooth tester checks true RF

Anritsu has introduced the MT8850A Bluetooth test set for



design proving and production test. The instrument can test Bluetooth modules or any product with a Bluetooth wireless interface, such as mobile phones, PDAs, PCs and other office or consumer devices. It lets developers conduct true RF measurements rather than simple functional testing. Test time is 10s for typical production requirements, including measurements of power, frequency, modulation and receiver bit error rate. Preprogrammed or user-defined test scripts are initiated by one press of the run key. The reference Bluetooth transceiver incorporated has a better than 1kHz frequency accuracy at the start of any packet and complies with dirty transmitter requirements.

Anritsu
Tel: 01582 433200
www.anritsu.co.uk

Dual output buffer adds to DMM support

Fairchild has introduced a dual output registered buffer to add to its DDR Dimm support for 184-pin, 2.5V PC200 and PC266, 72-bit wide registered SDRAM modules. They comply with the Jedec DDR SDRAM Dimm design. The SSTV 16859 buffer is a 13-bit register for address and control buffering with 184-pin and 232 DDR-I



stacked memory modules. The device has a differential input clock, SSTL-2 compatible data inputs and an LVC MOS compatible reset input. It has been fabricated on a sub-micrometre CMOS process. *Fairchild Semiconductor*
Tel: 001 888 522 5372
www.fairchildsemi.com

Space mapping of Trillium Lynxos

Trillium Digital Systems and Lynxosworks have announced a certified application-space mapping of the Trillium SSI system services interface to Lynxos. Trillium compliant certification of Lynxos for the Trillium Tapa advanced portability architecture means developers can choose Lynxos for the Trillium SSI that can run Trillium protocol stacks in the application space. Tapa is a set of architectural and coding standards that is independent from the target system's compiler, processor, operating system and architecture. Telecoms equipment makers can use the Trillium software for developing IP telephony, session initiation protocol and next-generation converged network infrastructure. With Trillium's SSI layer, Lynxos will let companies migrate tens or hundreds of thousands of lines of protocol stack code out of the kernel into the application space. This reduces kernel software maintenance and improves system reliability because any protocol code failure that may occur is less catastrophic in application space than in the kernel. *Lynxosworks*
Tel: 00 33 1 30 850600
www.lynxosworks.com

Power switch for chargers

Fairchild's latest power switch is a voltage mode PWM controller and SenseFET

Tester handles GSM

Rohde & Schwarz has introduced the CMU300 mobile base-station tester for all GSM bands and 8PSK Edge modulation. It can be upgraded for 3G mobile radio. It is available for GSM 400, 850, 900, 1800 and 1900 and provides various RF layer-one measurements on the transmitter and various receiver bit-error-rate measurements. Measurements are based on real-time coding and decoding of channels for GMSK and 8PSK modulation and circuit and packet switched modes.

Rohde & Schwarz
Tel: 01252 811377
www.rohde-schwarz.com



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(LDMOS) power switching element integrated into a single chip in an 8-lead DIPH. The FSDH0165 consumes 0.3W, it is a self-biased and protected linear control current-to-duty cycle converter with an open drain output.

Fairchild Semiconductor
Tel: 001 888 522 5372
www.fairchildsemi.com

SM-compatible connectors suit PIHIR

Surface-mount compatible connectors for use with high-temperature production



processes, such as pin-in-hole intrusive reflow (PIHIR) soldering, are available from Townsend Coates. Made by Harting, they can withstand the temperatures associated with reflow soldering -210 to 225°C. They have controlled lead lengths for compatibility with the PIHIR process. The range includes straight and angled DIN41612 male and females, headers to DIN41651, D-sub and 1.27mm pitch SCSI connectors. Among the products in SM format is the recently introduced Har-Bus 64 with 160 contacts for VME 64 extensions bus applications.

Townsend Coates
Tel: 0116 276 9191
www.townsendcoates.co.uk

Models added to signal generators

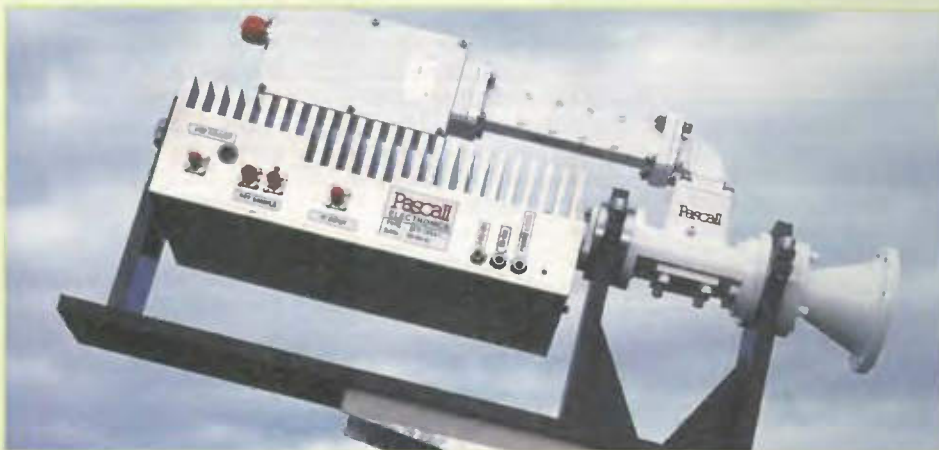
IFR has announced two frequency models for its 2026 multi-source generator. The 2026B 2.51GHz model is for the 2.4GHz ISM frequency band, used for wireless LAN applications, including Bluetooth. The 2026A is a 2.05GHz model with two or

Transceiver to hit satellites

Later this year, Pascall plans to start shipping commercially the STR-1812 solid-state 18GHz transceiver for use in private satellite networks. It transmits at 17.3 to 18.4GHz and receives at 10.7 to 12.75GHz. Data rates of up to 2Mbit/s from a 1.2m antenna and up to 4Mbit/s from a 1.8m antenna are supported, making it suitable for V-Sat type networks

where data such as intranet information needs to be transmitted almost instantaneously around the globe. The L-band interface eliminates the up-conversion stage between interface and transceiver.

Pascall Microwaves
Tel: 01621 816655
www.paradise.co.uk



three sources. For research, development and manufacturing where there is a need for two or three combined sources for conducting tests such as intermodulation and selectivity performance of components and receiver assemblies, the signal generators use a built-in combiner, switches and cables to reduce measurement uncertainties introduced by connecting separate signal generators. Adding Option 116 to either extends the functionality to test amplifiers in GSM base stations and, when used with a radio test set, for providing the interference signals for GSM receivers and intermodulation tests.

IFR Systems
Tel: 01438 772087
www.ifrsys.com

SM crystals follow dual stocking deal

Flint has expanded its range of leaded and surface mount crystals following a dual stocking agreement with Acal to distribute standard and custom quartz devices from Advanced Crystal Technology. The agreement covers SMD metal can quartz crystals, surface mount oscillators, watch



crystals, through-hole microprocessor crystals, through hole oscillators, ceramic resonators and crystal oscillators. HC49 can microprocessor crystals also form part of the range with frequencies from 1.8432 to 160MHz. The HC49 metal can SMX-4 and SMX-6 SMD quartz crystals have frequencies from 3.2 to 75MHz, with frequency tolerance options between 10 and 50ppm. They have four and six pads respectively. A 2.5mm high version is also available. The ZTA, ZTB and ZTT through hole ceramic resonators are stable from -20 to +80°C. The ZTA and ZTT are available over a 1.8 to 50MHz frequency range, with the ZTB covering 190 to 1250kHz.

Flint
Tel: 01530 510333
www.flint.co.uk

Development kits based on SBC

Arcom has announced development kits for Microsoft Windows NT Embedded 4.0 and Windows CE 3.0 for embedded systems designers. Both kits are based on the firm's SBC-Geode-GXm single board computer providing EBX compatibility with multimedia functions. The embedded NT kit contains a set of components and comes in several preconfigured builds. The options support workstation products or server systems. They are licensed and ready to run out of the box. A CD includes pre-built implementations of the operating system. The component definitions for the SBC let developers use Microsoft Target Designer and Component Builder to generate customised implementations of the operating system. The Windows kit contains the Windows CE 3.0 kernel preloaded into an onboard flash drive on the SBC. It ships with a copy of Microsoft's Embedded Visual Tools. Both kits provide software drivers, cables and documentation. Also included are a PS/2 mouse, power supply

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and cables for interfacing the SBC to a VGA-CRT display. An optional NEC 16.5cm TFT colour LCD module is available.

Arcom Control Systems
Tel: 0800 411300
www.paradise.co.uk

SD connectors keep memory cards secure

Molex has introduced connectors for use with secure digital (SD) memory cards for products such as PDAs, portable audio players, digital cameras and cellular phones. The connectors can handle SD cards with a capacity of up to 64Mbyte and incorporate card detect and write protect switch functions for secure downloading of digital data such as audio or image files. Connectors come in top and bottom mount styles and are 3.1mm high. Metal grounding covers that provide EMI and RFI protection are incorporated. A push-push eject system is used for card insertion and extraction and secure card retention. The 2.5mm pitch connectors are rated at 0.5A and supplied on embossed tape for automatic board mounting.

Molex
Tel: 01252 720720
www.molex.co.uk

Low-noise stereo volume controller

Profusion has introduced the WM8816 low-distortion low-noise stereo volume controller for audio applications. Total harmonic distortion plus noise is 0.001 per cent at 100dB. It combines digitally controlled volume adjustment and peak level detection on one chip. This makes it suitable for audio

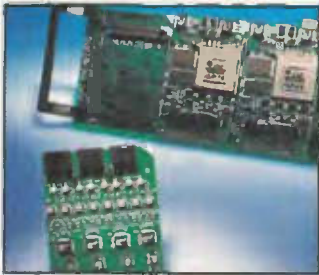


amplifiers and preamplifiers, home entertainment systems, mixing desks and audio recording equipment. Maximum input signal level is $\pm 18V$ and positive gain up to 15.5dB. The device is based on a resistor chain design with external op-amps, so users can tailor the system. It is controlled via a three-wire digital interface with daisy-chain capability for multi-channel systems. The gain of each channel can be individually adjusted from -111.5 to $+15.5dB$ in 0.5dB steps. On-chip peak detection circuitry can be used to implement volume monitoring and automatic gain control systems without the need to digitise the audio signal. It can be configured in double differential mode to reduce hum and other audible noise.

Profusion
Tel: 01702 543500
www.profusionplc.com

Eight-channel d-to-a 14-bit converter

Hunt Engineering has announced that the HEGD14 14-bit DAC is shipping. Part of the Heron range, the digital-to-analogue converter is for high output rate applications, such as wireless, especially digital radio



transmission. It can also be used in test, high-speed control and instrumentation. The device provides eight channels of 14-bit d-to-a converter at up to 33MS/s. This is a mirror of the HEGD9 12-bit a-to-d converter.

Hunt Engineering
Tel: 01278 760188

Multiband transceiver for unlicensed use

The nRF903 is a single chip multichannel UHF transceiver for the unlicensed 433, 868 and 915MHz low power radio device frequency bands. Available from Clere, the

Nordic VLSI product has GMSK and GFSK modulation and demodulation capability at an effective bit rate of 76.8kbit/s for 153.6kHz bandwidths. With transmission power adjustable to a maximum of +10dBm and a differential antenna interface, it is suitable for PCB antenna applications. Configuration data is programmed by a 14-bit word architecture via a serial peripheral interface. Operating from a +3V DC supply, it has power down and standby modes, and complies with EN300 220-1 and CFR47.

Clere
Tel: 01635 298574
www.clere.com

Single board supports Celeron and Pentium

From Advanced Modular Computers is the AMC-3782EVS Socket 370 architecture board. The CPU card supports both Celeron and Pentium III processors, up to 133MHz front side bus, up to 512Mbyte SDRAM and dual Ultra DMA66 IDE interface. The Intel i810E chip set uses a shared system architecture

allowing the on-chip VGA and audio to access system memory. Onboard utilities include dual Intel 82559 10 or 100Mbit/s Ethernet controllers, Ultra 160 SCSI chip and one Compactflash socket. Because the VGA and audio is onboard, slot space is freed up.

AMC
Tel: 01753 580660
www.amcuk.com

Solid-state relays in 6 and 8-pin models

Crydom has introduced six and eight-pin DIP solid-state relays. The G2 devices are available in contact configurations, including form A, dual form A, form B, dual form B and one form A one form B. All can be wired for AC operation or one of three DC configurations. They can switch AC or DC voltages from the microvolt and nanoamp range up to 400V and 400mA, and can be wired in series to achieve switching in the kilovolt range. Typical input voltage and current are 1.25V DC at 5mA. Dielectric strength input to output is 3750V and capacitance is 0.8pF for single-relay and 1.2pF for dual-relay devices. Power dissipation is



Button has split screen

Omron has launched a rectangular lighted pushbutton switch with split screen options for process and motor control. The A3PJ measures 18.9 by 26.1mm and comprises a switch unit and indicator and pushbutton unit, which are joined with a snap action. The indicator unit is available in solid colour as well as two, three and four split screen options in horizontal and vertical orientations. Colours include white, red, green and orange and can be specified to 5, 12 or 24V DC. There are three types of indicator unit – LED-lit with built-in LED chip, LED ball-lit with removable LED and incandescent lamp-lit.

Omron
Tel: 0208 450 4646
www.omron.co.uk

Please quote *Electronics World* when seeking further information

500 or 600mW respectively. Relays are available in through-hole and surface-mount variants, including a package 2.2mm high, making it suitable for PCMCIA applications.

Crydom
Tel: 01444 473555
www.crydom.com

1.2kW supplies will support hot swapping

Artesyn has launched two 1.2kW front-end AC/DC power supplies with configuration and control facilities for system integration and power management. For data processing, datacoms and telecoms applications that use distributed power architectures, the AFE1200 supplies have current sharing capabilities and support hot-swapping. Rack-mountable power shelves are available for users wishing to configure multiple power



sources. There is a choice of 48 or 24V DC outputs. The 48V model can deliver up to 25A and the 24V model up to 50A. Both provide an auxiliary 12V, 1A DC output for powering items such as external housekeeping circuitry. The auxiliary output is isolated from the main output, and both outputs are floating, letting users adopt whatever rail reference configuration they require. They have an 85 to 264V AC universal input range and accommodate mains frequencies from 47 to 63Hz. The supplies incorporate power factor correction, providing a minimum power factor of 0.99.

Artesyn
Tel: 00 353 24 25572
www.crydom.com

Single board supports Celeron and Pentium

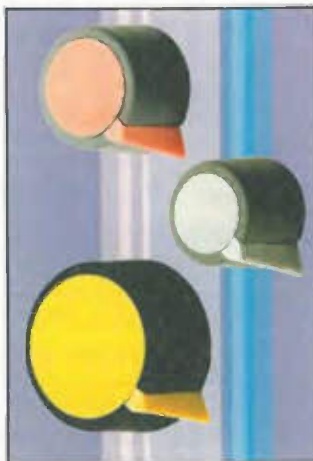
Azzurri is shipping the ZX4500 Openarchitect from Znyx Networks. The open-source Linux Ethernet switch provides a line-rate hardware switching environment that lets OEMs deploy standard and customised IP switching applications within a 6U CompactPCI blade. It handles packet manipulation within the switch silicon to achieve line-rate speeds of 6.6 million packet/s. Layers two to seven filtering is accomplished by downloading and running existing Linux networking applications. Media conversions are possible because it allows direct access to its switching fabric from its open PPMC slot. It has 12 or 24 10 or 100Mbit/s Ethernet ports, two Gigabit Ethernet ports, embedded Motorola 8240 PowerPC processor and a PMC or PPMC slot for an expansion processor or additional I/O. The design is hot swap compliant, with all Ethernet ports implemented as rear-panel I/O. There are 64Mbyte of buffer RAM and a non-blocking switch fabric. Up to 30 switches may be stacked via one or both Gigabit ports, putting up to 720 10 and 100Mbit/s ports under the same management as the rest of the equipment in the chassis. It can be combined with Znyx's Rainlink technology to provide switch-to-switch or port-to-port failover and link aggregation and load-balancing services. Applications include VoIP gateway, network load balancing, web load balancing, Firewall IVPN, multimedia multicasting and SS7 signalling.

Azzurri Technology
Tel: 0118 944 5900
www.azzurri.com



Ergonomic tuning knobs in six sizes

OKW has introduced the Com-Knobs series of ergonomic tuning knobs. They come in six sizes from 16 to 50mm in diameter. The three-part construction comprises cover, knob and marking element. Individual marking elements, which slide into the side of the



knob moulding, are available in five standard colours – coral, beach, lagoon, sky and mineral. They can be attached to control spindles of 4, 6 or 6.35mm diameter using a collet system in the top of the knob moulding. The knobs are suitable for the fine control and adjustment of machines and devices and for electronic measurement applications.

OKW
Tel: 01489 583858
www.okw.co.uk

Open-frame switcher has output choices

From Ultimate Renaissance is a 60W AC/DC power supply that provides power densities above 0.43W/cm³ in a 50.8 by 101.6 by 20.3mm package. The VLT60 open-frame switcher comes with single or multiple outputs, at various voltages, and has efficiencies above 85 per cent. It has universal inputs



from 90 to 264V AC. Operating between 0 and 50°C, the supply has overcurrent protection and convection cooling. Weighing 0.17kg, it can be used by OEMs in POS, data networking, computer peripherals, touch-screen displays, telecoms, medical equipment and industrial products. Maximum leakage current is 500µA and it qualifies as a class B FCC and CISPR EMI filter.

Ultimate Renaissance
Tel: 01793 439310
www.ur-home.com

The Headphone Amplifier Box

Balanced or unbalanced microphone or line input to headphone output

Professional portable units operating from an internal PP3 battery or external mains adaptor



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- * Bridged headphones output drive
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- * Loop through facility
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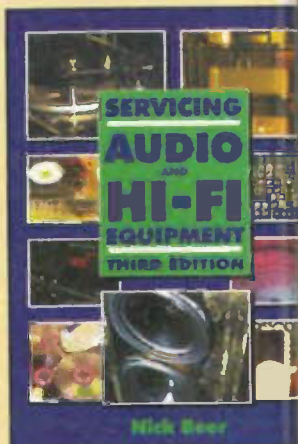
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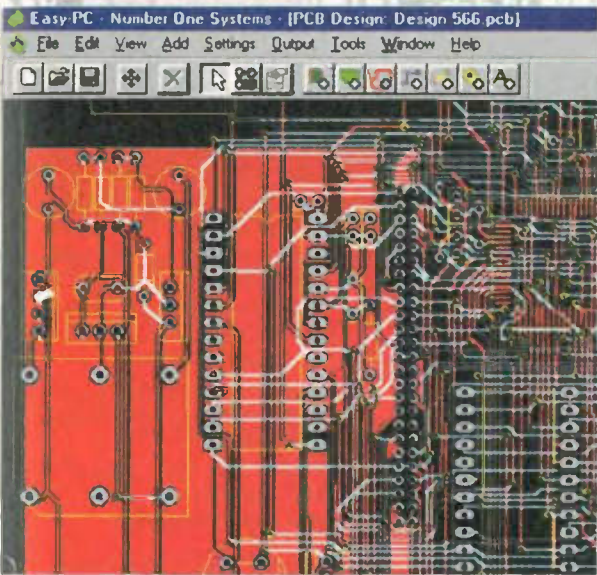
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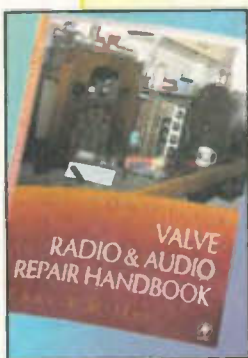
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Condor has announced a 125W multi-output switching power supply in an 89 by 152 by 38mm package. Called the GPFM125, it is approved to UL2601, CSA 22.2NO.601.1 and EN60601-1 for medical systems applications. Power density is $0.24\text{W}/\text{cm}^3$ and

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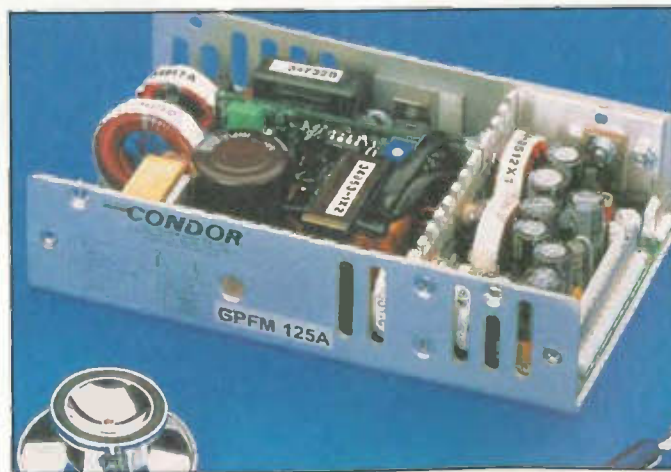
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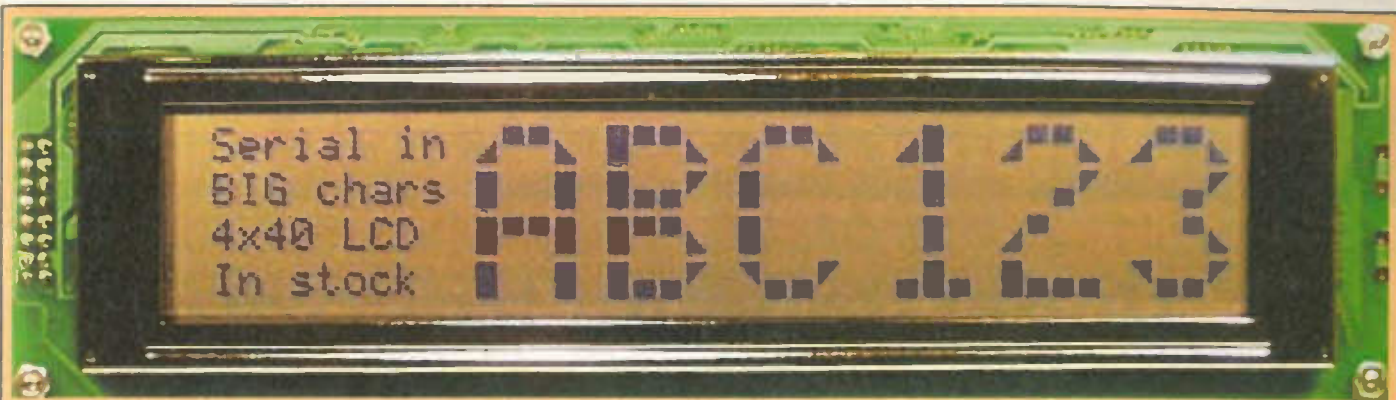
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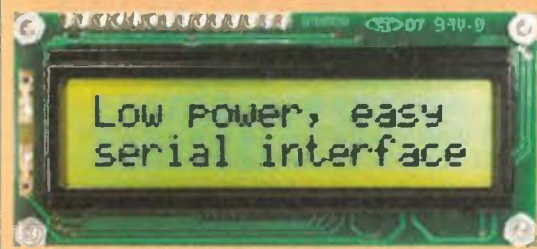


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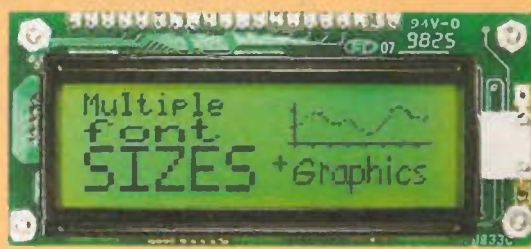
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Understanding transformers

As **Ian Hickman** explains, electricity and magnetism have many similarities – but they also have many differences. It is important to understand them before you can fully appreciate the main topic of this set of two articles – transformers.

I promised in the January issue to write about transformer equivalent circuits and believe me, I will. But in considering how to start, where to start seemed the more important question. So in time honoured fashion, I'll begin at the beginning, in this case with basic electromagnetism.

This is a topic most of you will have studied at some stage in the past. But you will probably find that you are not as comfortable with it now as you are with Ohm's Law and the like.

I will develop the topic of electromagnetism by analogy with electric circuits.

Rivers and ring roads

Some phenomena are circular, like London's M25 ring road; the traffic goes round and round. Other phenomena are open ended, such as a river. But most phenomena are ultimately circular; everything thing forms part of a circuit.

Water that a river empties into the ocean evaporates, falls as rain on the land and finds its way back into the river. Similarly, electric currents can only flow in a circuit – even if intermittently like lightning – and the same applies to magnetic flux. Whether this also applies to electric flux is a point I hope to be able to expand on in a future article.

The electric circuit

Figure 1 shows a source of EMF (electromotive force), a battery of voltage E volts say, capable of supplying current to a load consisting of a resistor of value R ohms. A potential difference of E volts exists between points A and B , due to the EMF of the battery.

If the switch is closed, the same potential difference exists between points A' and B' , and a current of I amps flows through the resistor, where $I=E/R$. This is Ohm's Law; most of you will have been familiar with it for

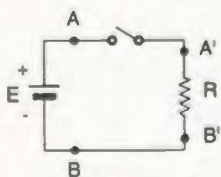


Fig. 1. An electromotive force causes current to flow in a circuit on closure of the switch.

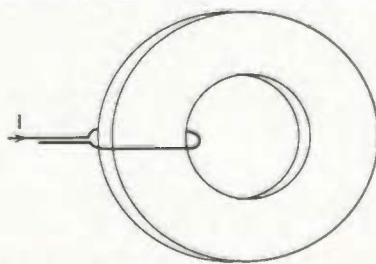


Fig. 2. A magnetomotive force causes flux to flow in a magnetic circuit.

longer than you care to remember.

Forgive me for stating the obvious, but if the switch is open, then no current flows out of the battery. This is an important difference between the electric circuit and a magnetic circuit.

It is generally assumed that the resistance of the connecting wires is zero, or at least negligible compared with R . It is also assumed that the resistance of any insulators – in the switch and elsewhere – is infinite. If this were not the case, if air had a resistance which, although high, were finite, then some small current would flow even if the circuit via R was broken. This would be analogous to the magnetic circuit.

Considering resistor R in more detail, assume that it's a wire-wound type, the length of whose resistance wire is l metres. Now the current equals E/R , where $R=\rho l/A$, A being the cross-sectional area of the wire in square metres. So,

$$I = \frac{E}{R} = \frac{EA}{\rho l}$$

where ρ – the Greek lower-case letter 'rho' – is called the resistivity, in ohm.metres, of the material that the resistance wire is made from.

In the case of a high-wattage resistor, this would typically be Nichrome, a nickel-chromium alloy with a reasonably low temperature coefficient of resistance, and good resistance to oxidation.

Rearranging,

$$I = \frac{EA}{\rho l}$$

a little gives,

$$\frac{I}{A} = \frac{E}{l} + \rho, \text{ or } J = \frac{\epsilon}{\rho}$$

where J is the current density in amps per square metre of cross-section in the wire, and ϵ is the electric field strength gradient along the wire, in volts per metre. These terms will all prove to have analogous equivalents in a magnetic circuit.

The magnetic circuit

Figure 2 shows a source of magnetomotive force, or MMF for short, consisting of a single turn of current-carrying wire. In this instance, it is wound on a core of ferromagnetic material.

Imagine that this core is toroidal with a mean path length or effective circumference of l metres, and cross-sectional area $A\text{m}^2$. If the current in the wire is I amps, the MMF is I amp.turns. The fashion nowadays is to

drop the 'turns', and say simply that the MMF is I amps, or NI amps if there are N turns.

My venerable textbook¹ defines the unit of MMF as the "ampere turn". But certainly in the example quoted, the MMF would still be I amps (or amp turns), if the current were one amp flowing in N turns, instead of I amps in 1 turn, if $N=I$.

You may find it easier, like me, to think in terms of amp turns, but as the number of turns N is a pure dimensionless number, it comes to the same thing.

The symbol for the units of MMF is 'A' (for amps). It is convenient though to also have a symbol to use in equations, to denote an MMF as a variable. I use 'F'. Thus in general, $F=NI$; the MMF equals the number of amperes times the number of turns.

Note the convention that symbols denoting variable quantities in equations – such as N turns, I amps – are in *italics*, while constants and symbols denoting units, such as A for amps, V for volts, etc., are not.

Just as the EMF E volts in Fig. 1 establishes a voltage gradient on the wire of the resistor, forcing a current of I amps to flow through it, the MMF in Fig. 2 does something very similar. Where the voltage gradient $\epsilon=E/l$ volts/metre, the MMF gradient is denoted by H , where $H=F/l$ amps/metre, (ampere turns per metre).

Note the difference between the variable H with units $A.m^{-1}$, and the H that denotes the unit of inductance, the henry. Unfortunately, there are only 26 letters in the Roman alphabet, so inevitably some get used to denote both a variable and a unit.

The value of H is called the magnetic field strength. In Fig. 2, it is virtually constant at all points around the magnetic circuit of the toroidal core. Likewise, ϵ would be constant around the circuit in Fig. 1, if the resistor, switch and copper wires were all replaced by a single length of resistance wire of constant cross section.

In Fig. 1, ϵ causes a current to flow, of value related to the length, cross-section and resistivity of the wire. Something similar happens in Fig. 2. The magnetic field strength $H=F/l$ causes a flux of Φ Wb, where Wb denotes webers, to flow in the core, where,

$$\frac{\Phi}{A} = \frac{F}{l} \times \mu_0 \mu_r$$

The flux density $\Phi/A=BT$, where T denotes teslas, is analogous to $I/A=J$ in an electric circuit. So $B=H \times \mu_0 \mu_r$, just like $J=\epsilon \times A$. A run-down of the equivalent terms is given in the Table.

Rearranging one of the earlier equations gives,

$$I = E \times \frac{A}{l} \times \frac{1}{\rho}$$

where $1/\rho$ – the reciprocal of resistivity – is the

conductivity, σ , of the conductor. Conductivity is a measure of how easy it is to make current flow. The unit of conductivity is the siemen, and its abbreviation is S. Formerly, an inverted Ω sign was used, and the units were called 'mhos'.

A similar rearrangement gives,

$$\Phi = F \times \frac{A}{l} \times \mu_0 \mu_r$$

where $\mu_0 \mu_r$ is a measure of how easy it is to make flux flow, and is called the permeability of the material. Parameter μ_0 is the permeability of free space, while μ_r is the permeability of the core material relative to μ_0 . Permeability μ_0 has a value $4\pi 10^{-7}$, specified in teslas per amp per metre, or T.m/A. On the other hand, μ_r may be anything from two or three for the highest frequency grades of ferrite, to 300000 or more for certain types of transformer laminations in mu-metal.

By contrast, the conductivity of free space is zero, and virtually the same for air and all common insulators. Thus in Fig. 2, if the core were removed, there would still be some magnetic flux produced by the magnetic field. In Fig. 1 though, if the wire of the resistor were replaced by air, no current would flow.

Rearranging the last equation yet again, gives,

$$\frac{F}{\Phi} = \frac{l}{A \times \mu_0 \mu_r}$$

The term,

$$\frac{l}{A \times \mu_0 \mu_r}$$

is called the reluctance, S , of the magnetic path, and is specified in amperes per weber.

Comparing with,

$$\frac{E}{I} = \frac{l\rho}{A}$$

reluctance is to a magnetic circuit what resistance is to an electric circuit. And by analogy with resistivity,

$$\frac{1}{\mu_0 \mu_r}$$

might be called 'reluctivity' – though I have never heard of the term – while permeability $\mu_0 \mu_r$ is analogous to $1/\rho$ or conductivity σ . So just like,

$$I = \frac{E}{R}, \quad \Phi = \frac{F}{S}$$

The crucial difference

Despite the similarities in the formulae involving EMF and MMF, the crucial difference concerns energy. In Fig. 1,

Table. Run-down of commonly used electrical quantities and their magnetic counterparts.

Name	Symbol in equations	Unit name	Abbreviation
Electromotive force (EMF)	E	volt	V
Magnetomotive force (MMF)	F	amp	A
Electric current	I	amp	A
Magnetic flux	Φ	weber	Wb
Potential gradient	ϵ	volts per metre	V/m
Magnetic field	H	amps per metre	A/m
Current density	J	amps per square metre	A/m^2
Flux density	B	tesla (webers per square metre)	T (Wb/m ²)
Conductivity (1/resistivity)	$\sigma(1/\rho)$	1/(ohm.metre)	(Ωm) ⁻¹
Permeability	$\mu_0 \mu_r$	webers per amp.metre	Wb/Am

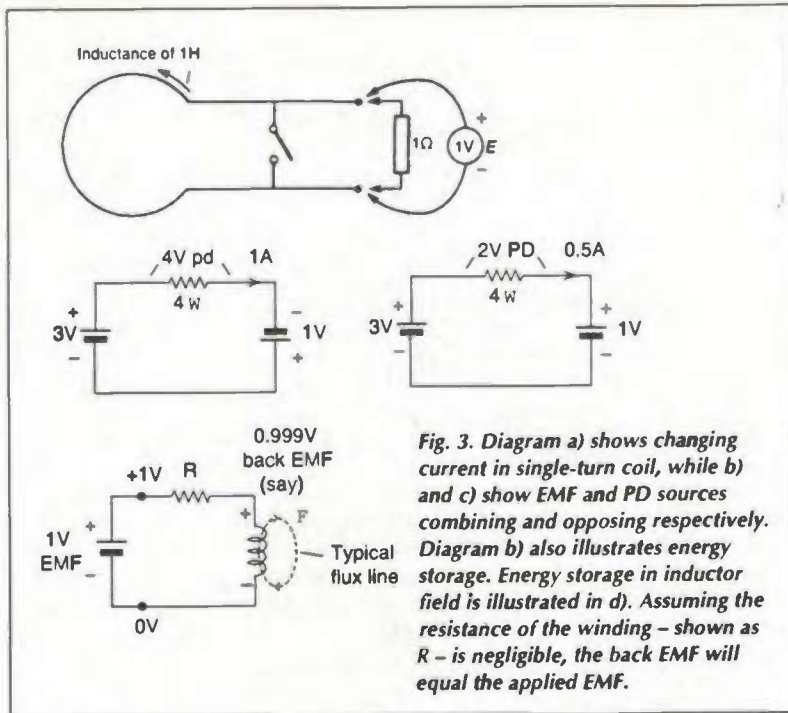


Fig. 3. Diagram a) shows changing current in single-turn coil, while b) and c) show EMF and PD sources combining and opposing respectively. Diagram b) also illustrates energy storage. Energy storage in inductor field is illustrated in d). Assuming the resistance of the winding - shown as R - is negligible, the back EMF will equal the applied EMF.

the current through the resistance is responsible for the dissipation of electrical energy, measured in joules, abbreviated to J, into heat. The rate of dissipation is measured in joules per second, where $J/s=W$, where W is the abbreviation for watts. Dissipation W is I^2R .

By contrast, no energy is dissipated in Fig. 2, assuming that the turn of wire producing the MMF has zero resistance. But in the very act of the current starting to flow and creating a magnetic field, energy was stored in the resultant magnetic flux. In the steady state shown in Fig. 2, that energy just sits there, waiting for something to happen. The flux is conventionally indicated by lines of flux, forming closed loops; in Fig. 2 there will be some in the air and many more in the core.

Just as the individual resistances around an electric circuit can be added up when working out the total EMF E needed to cause a current I to flow, so in a non-homogeneous magnetic circuit - for example a core of magnetic material with a permeability μ_r , having an air gap - the reluctances of the core and gap can be added up to find the total MMF needed to cause a given total flux.

So far, I have looked at the field produced by a constant current of I amperes, but what happens when the current changes? Indeed, how can the current get to flow in the first place? Figure 2 rather begs the question by assuming that the current is already flowing.

Consider what happens when you connect an EMF of one volt to a very large single-turn coil, as in Fig. 3a). Assume that the coil has negligible resistance. Nothing can change instantaneously, so the moment after connecting the supply the current must be the same as the moment before, i.e. zero. Clearly, you can expect the current to increase thereafter, but how fast?

Assume that the current increases at one ampere per second, so that after one second the MMF F is just one ampere turn, and assume that the reluctance S is $1A/Wb$ (or, one amp per weber), so that the resulting flux Φ is one weber.

In fact, for this to be so, the coil would have to be very large indeed, or immersed in a magnetic medium with a huge relative permeability. But that is a minor practical

point that does not affect the principle of the thing.

Having assumed the coil to have negligible resistance, the current will ultimately become very large; so why isn't it already huge after just one second? The reason is that the steadily increasing flux linking the coil induces an EMF in it, in opposition to the applied EMF: this is known as Lenz's Law.

If the flux Φ increases by a small amount $d\Phi$ in a fraction of a second dt , so that the rate of increase is $d\Phi/dt$, then the back EMF E_B induced in the single-turn coil is $E_B = -d\Phi/dt$. But $\Phi = MMF / \text{reluctance} = F/S$, i.e. $\Phi = NI/S$. As the current and flux are both increasing, for this to remain true their rates of change must also be equal, i.e.,

$$\frac{d\Phi}{dt} = \frac{N}{S} \times \frac{dI}{dt}$$

This rate of flux increase will induce a voltage $-E_B = d\Phi/dt$ in series with each and every one of the N turns of the coil, so in all,

$$E_{B(\text{total})} = -N \frac{d\Phi}{dt} = -N \frac{N}{S} \times \frac{dI}{dt} = -\frac{N^2}{S} \times \frac{dI}{dt}$$

The term N^2/S , which determines the induced voltage resulting from a unit rate of change of current, is called the inductance L . It is measured in henries: that is, $L = N^2/S$ henries.

You must keep the difference between an electromotive force (EMF) and a potential drop or difference (PD) very clearly in mind, to understand the minus sign in $E_B = -d\Phi/dt$. To illustrate this, consider two secondary - i.e. rechargeable - batteries and a resistor as in Fig. 3b).

The total EMF round the circuit, counting clockwise, is $3+1$ volts. This is balanced by the PD of IR volts across the resistor. The batteries supply a total of $4W$ of power, all of which is dissipated in the resistor.

If now you reverse the polarity of the $1V$ battery, as in Fig. 3c), the total EMF acting is $3-1V$, so the current is $0.5A$. The $3V$ battery is now supplying $3 \times 0.5 = 1.5W$, but the dissipation in the resistor I^2R is only $1W$. The other $0.5W$ is disappearing into the $1V$ battery; but it is not being dissipated, it is being stored as chemical energy.

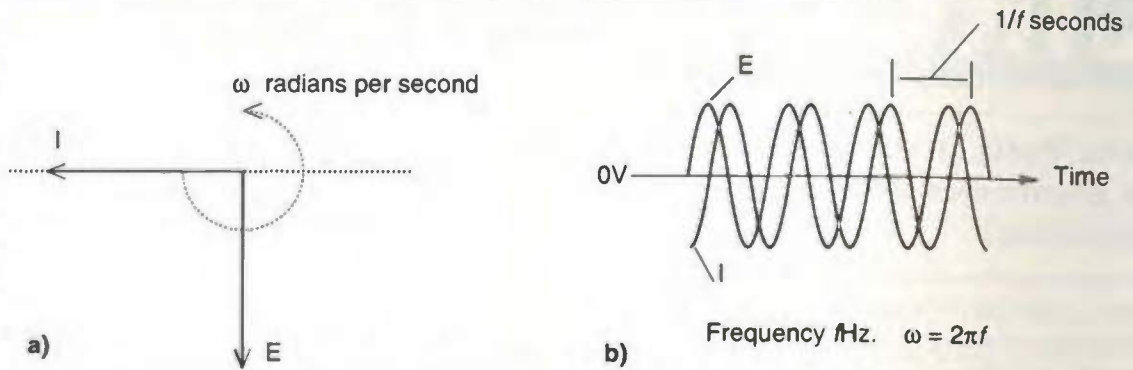
In Fig. 3d), the situation is just the same; the applied EMF of the battery is opposed by the back EMF of the inductor - which in turn is determined by the inductance and the rate of increase of the current - while energy from the battery is being stored in the steadily increasing magnetic flux. If the internal resistance of the battery and the resistance R of the inductor are vanishingly small, the current will continue to increase indefinitely; if not the current will reach a limit set by the applied EMF and the total resistance in the circuit.

Returning now to Fig. 3a), if the switch is closed one second after connecting the battery, at which time the current has risen to $1A$, then there is no voltage across the ends of the coil. No back EMF means that $d\Phi/dt$ must be zero, so dI/dt is also zero. Hence the current now circulates indefinitely its value frozen at $1A$ - provided that the coil really has zero resistance.

In the meantime, imagine the battery disconnected and replaced with a 1Ω resistor; you will see why in a moment. Thus energy stored in a magnetic field is preserved by a short-circuit, just as the energy stored in a capacitor is preserved by an open-circuit.

Now open the switch in Fig. 3a), thus substituting the 1Ω resistor in place of the short-circuit. At the moment the switch opens the current of $1A$ continues to flow in the resistor; it cannot change its value instantaneously. So the energy stored in the magnetic flux is being dissipated and

Fig. 4a) Vector diagram showing current lagging voltage by 90° in a perfect inductor. In 4b), you can see the associated voltage and current waveforms.



therefore the current must fall. This will establish a 1V potential difference across the resistor, of opposite polarity to the now disconnected battery. In other words, the top end of the resistor will be negative with respect to the lower end, because dI/dt is now negative.

The coil is now acting as a generator feeding its stored energy into the resistor, initially at a rate of 1 joule per second, i.e. 1W. How much energy is there stored in the field, and how long before it is all dissipated as heat in the resistor?

Initially the current must be falling at 1A per second, since we have 1V across the resistor, and $E = -LdI/dt$, where the inductance is unity in this case. Of course dI/dt is itself now negative, i.e. current is decreasing, as the polarity reversal witnesses.

After a fraction of a second, the current being now less than one ampere, the voltage across the resistor will have fallen likewise; so the rate of decrease of current will also be lower.

How much energy?

To find the energy stored in the field, you could integrate the product $E_B \times I$ from $t=0$ – i.e. the point at which the switch was opened – to infinity. But there is an easier way.

Suppose you use a variable resistor and increase its resistance, keeping the value inversely proportional to the current. Now IR will be constant at 1V and the current will fall linearly to zero in 1 second.

Since the induced voltage across the resistor has, by this dodge, been maintained constant at 1V, the energy dissipated in it is easily calculated.

On opening the switch, the dissipation is $1V \times 1A$, and this falls linearly to zero over one second. So the average power is 0.5W maintained for one second giving a stored energy of 0.5J.

If the inductance had been 2H and the current 1A when the switch was opened, the initial rate of fall would have been 0.5A per second and the discharge would have lasted 2s, dissipating 1J in the resistor. This assumes that its value was adjusted to maintain 1V across it as before. Thus the stored energy is proportional to the inductance L .

On the other hand, if the current was 2A when the switch was opened, the voltage across the 1Ω resistor would have been 2V, so the rate of fall would need to be 2A/s, assuming 1H inductance. Thus the initial dissipation would have been 4W, falling to zero over 1s, giving a stored energy of 2J, or four times as much as with 1A.

So the stored energy is proportional to the square of the current. In fact, the stored energy is given by,

$$J = \frac{1}{2} L \times I^2$$

Uses of inductors

An inductor can be, and often is, used as an energy store in

power supplies, but there are certain limitations to the inductor's energy storing ability. In particular, the energy stored in the magnetic field of a short-circuited inductor is rapidly lost due to dissipation in the resistance of its windings.

For a high-quality capacitor, the shunt loss resistance R is almost infinite, whereas it is not possible to make the series loss r of an inductor negligible. At very low temperatures though, the electrical resistivity of certain alloys and compounds vanishes entirely – a phenomenon known as superconductivity. Under these conditions an inductor can store energy indefinitely in its magnetic field, as none is dissipated in the conductor.

In addition to use as energy storage devices, inductors have several other applications. For example, inductors with cores of magnetic material are used to pass the direct current output of a rectifier to later circuitry while attenuating the alternating hum components.

Air or ferrite-cored inductors – RF chokes – are used to supply power to radio-frequency amplifier stages while preventing RF power leaking from one stage to another via the power supply leads. This application, and others, makes use of the AC behaviour of an inductor.

Since the back EMF, E_B , is given by,

$$E_B = \frac{-Nd\Phi}{dt} = -L \frac{dI}{dt}$$

it follows that the higher the frequency, the smaller the alternating current required to give a back EMF balancing the applied alternating EMF. In fact, the reactance X_L of an inductor is given by $X_L = 2\pi fL = \omega L$, where f is the frequency in hertz, ω is the angular velocity in radians per second, and L is the inductance in henrys.

This may be represented vectorially as in Fig. 4a), where the projection of the voltage E and current I vectors on the horizontal gives the instantaneous value of those variables. Thus the position of the vectors in Fig. 4a) corresponds to the start of the waveforms shown in b).

You can see that when the voltage is at its positive peak, the current is zero but increasing. In fact, the current is increasing – or becoming less negative – all the time that the applied voltage is positive and *vice versa*. The net energy flow is zero.

In calculations, the 90° phase shift between the voltage and lagging current can be accommodated by using the j operator and writing $X_L = j\omega L$.

The foregoing provides all the background necessary to start unravelling the intricacies of transformers. I plan to cover those in another article. ■

Reference

1. Fundamentals of Electrical Engineering in MKS Units, E. Hughes, Longmans, Green and Company, 1954.

CIRCUIT IDEAS

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Send your ideas to: Jackie Lowe, Room L514, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Colour-bar generator

This is a colour bar generator intended to soak test colour video monitors. It had to be low cost and small, therefore it had to have a low parts count. Some cunning tricks were employed to achieve this.

The heart of the colour bar generator is a 74LS90 decade counter. Its outputs are used as in the table.

If IC₃ is used as a straightforward

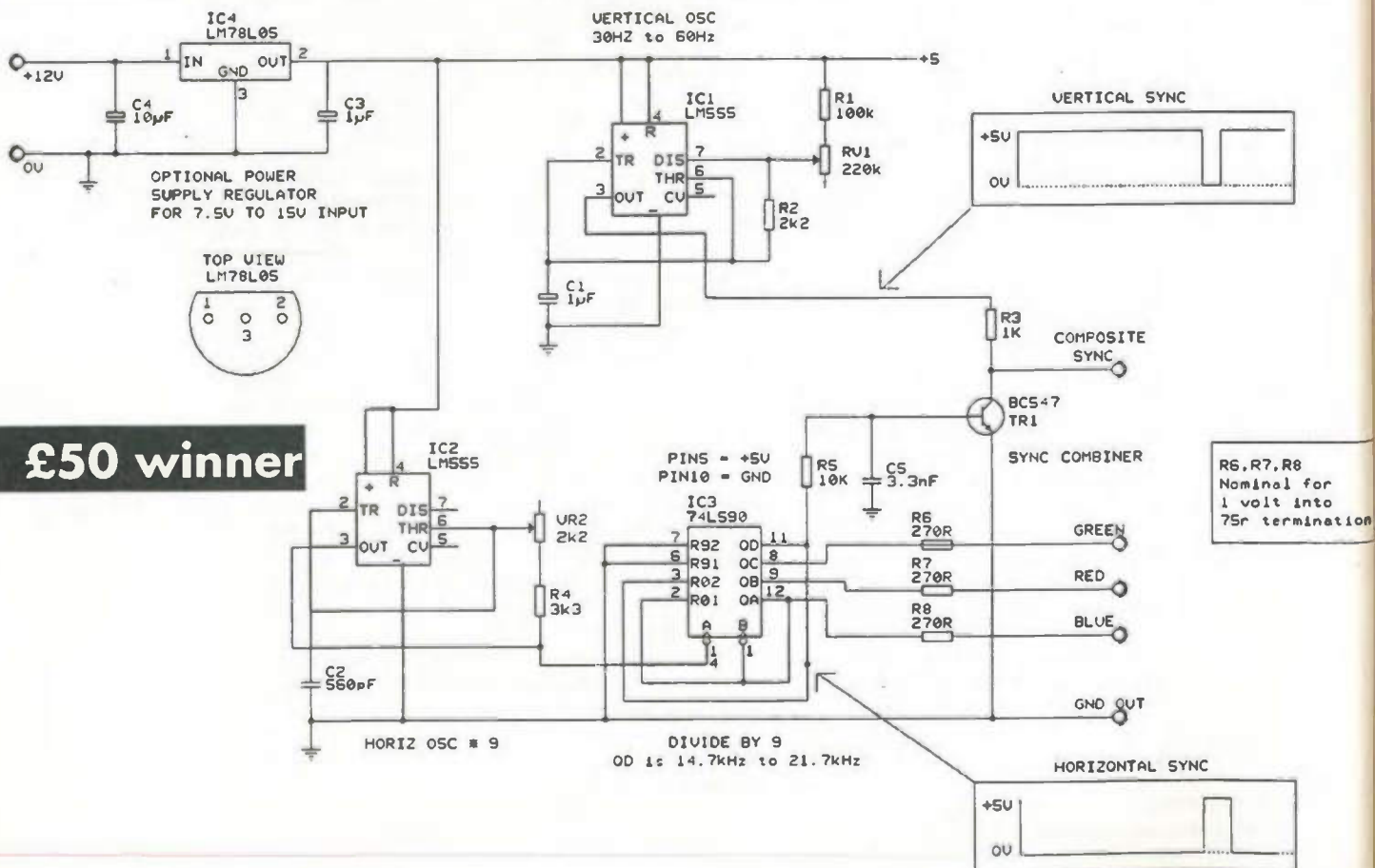
divide-by-10, an error occurs in some monitors. Technically correct monitors use black level clamping on the back porch.

The back porch is defined as the period after the sync pulse but before the active video starts. If the counter were allowed to count to 10 there would be a blue bar in this time period. This would be clamped by the

Table. Outputs of the 7490 counter have the following functions.

Count	Blue A	Red B	Green C	Sync D	Colour displayed
0	0	0	0	0	Black
1	1	0	0	0	Blue
2	0	1	0	0	Red
3	1	1	0	0	Magenta
4	0	1	0	Green	
5	1	0	1	0	Cyan
6	0	1	1	0	Yellow
7	1	1	1	0	White
8	0	0	0	1	Sync. (Black)
9	1	0	0	1	

To eliminate blue bar, reset to count 0.



monitor and so it would display no blue; to avoid this the counter is reset to zero at the ninth count, see above.

Timer IC_2 is the horizontal oscillator, running at 9 times the line frequency. It is used in the schmitt trigger mode to reduce component count – it saves a resistor. Output QD is inverted horizontal sync, settable in the range 14.7kHz to 21.7kHz.

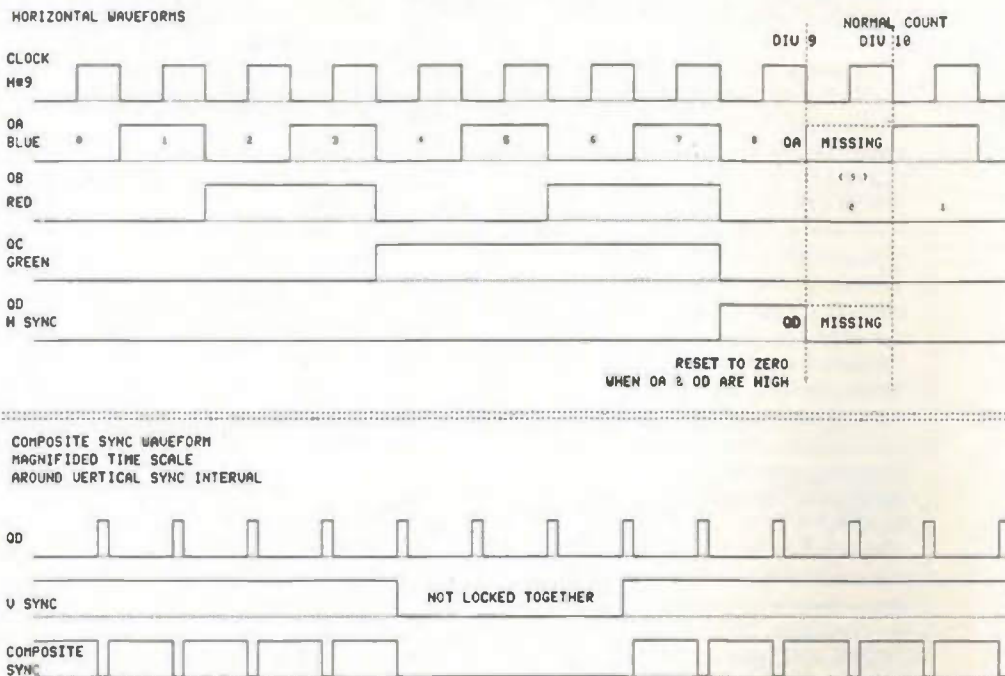
The QD output of IC_3 is inverted by Tr_1 to give correct polarity horizontal sync. Timer IC_1 is the free running vertical oscillator, settable in the range 30Hz to 60Hz. Resistor R_2 sets the vertical sync time, while RV_1 & R_2 set the vertical frequency.

Vertical sync output is normally high and goes low at sync time. This is used as collector supply to Tr_1 and so mixes horizontal and vertical syncs together.

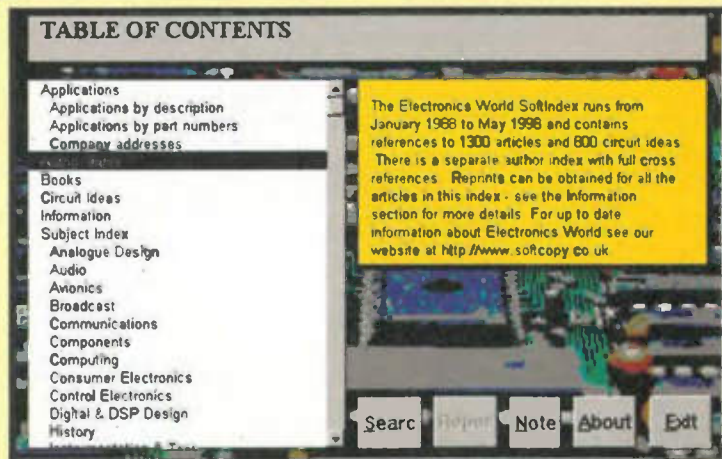
Outputs are at TTL levels but can be attenuated to 1V by inserting 270Ω series resistors and feeding into nominal 75Ω terminations.

Colin Pye
Coventry

COLOUR BAR GENERATOR WAVEFORMS



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Measure revs-per-second on a PC

This design uses the BIOS INT1Ch of a PC, together with a few inexpensive components, as a high-speed rev/min meter. The meter can measure speed in the range of 1 to 65500rev/s, which equates to around 1 to 3932160rev/min.

A dozen ICs are used, in conjunction with the computer's BIOS INT1Ch, without affecting its normal ISR. Printer port LPT1's data and status ports are also used.

Figure 1 shows the circuit diagram, which interfaces with LPT₁. The design was originally developed for remotely measuring the speed of turbo molecular vacuum pumps, but it can be used for any application where the speed range is up to 65500rev/s.

Input pulses are generated by a Hall element or slotted disk with infra-red photo diode, at a calibrated rate of 1 pulse per revolution. These pulses are fed via a 74LS08 AND gate to a 16 bit up/down counter.

Output from the counters and the flip-flops' Q outputs are routed to the status port (at 0x379h) of LPT₁ through

74LS244 buffers. A 74LS138 3-to-8 line decoder, controlled from data port at 0x378₁₆, provides various control signals as dictated by the software.

A simple Turbo C program controls the speed measurement. The BIOS INT1Ch occurs 18.2 times a second. It is used only for keeping time of day. In this design, it is exploited for generating a timing window of 1 second by incrementing the TIMERINTR variable, which is set to zero initially.

At the start, input pulses are inhibited by disabling the AND gate. All the counters are loaded with 0x00₁₆ and the flip flops are cleared. When the INT1Ch interrupt occurs, it immediately enables the AND gate and the counters count up.

When each counter's output crosses from 0x0F₁₆ to 0x00₁₆, its carry output sets the 74LS74 D-type flip-flop connected to it to logic 1. When the TIMERINTR reaches 18, the pulses are again inhibited by disabling the AND gate and flip-flop outputs Q₀, Q₁, Q₂ - initially set to 000 - are read through the status port. The timing window is actually 0.989

seconds, but the measurement is subsequently corrected to a one second measurement by multiplying by 18.2/18.

The binary word Q₀, Q₁, Q₂ indicates progress in counting, the counters involved and the maximum speed expected as given in the Table. This relieves the PC from reading all the counters. Instead it reads only the counters that actually counted, thereby reducing the number of read operations and subsequent manipulations.

The SPEEDREAD routine reads the counter output, a nibble at a time, starting from counter 1 (nib0), depending on the binary word, up to counter 4 (nib3).

Nibbles are manipulated and the speed in revolutions per second and revolutions per minutes are displayed.

At the end of one measurement cycle, the TIMERINTR is reset to zero and the counters and the flip flops are cleared, leading to another speed measurement cycle.

K Suresh
Kalpakkam
India

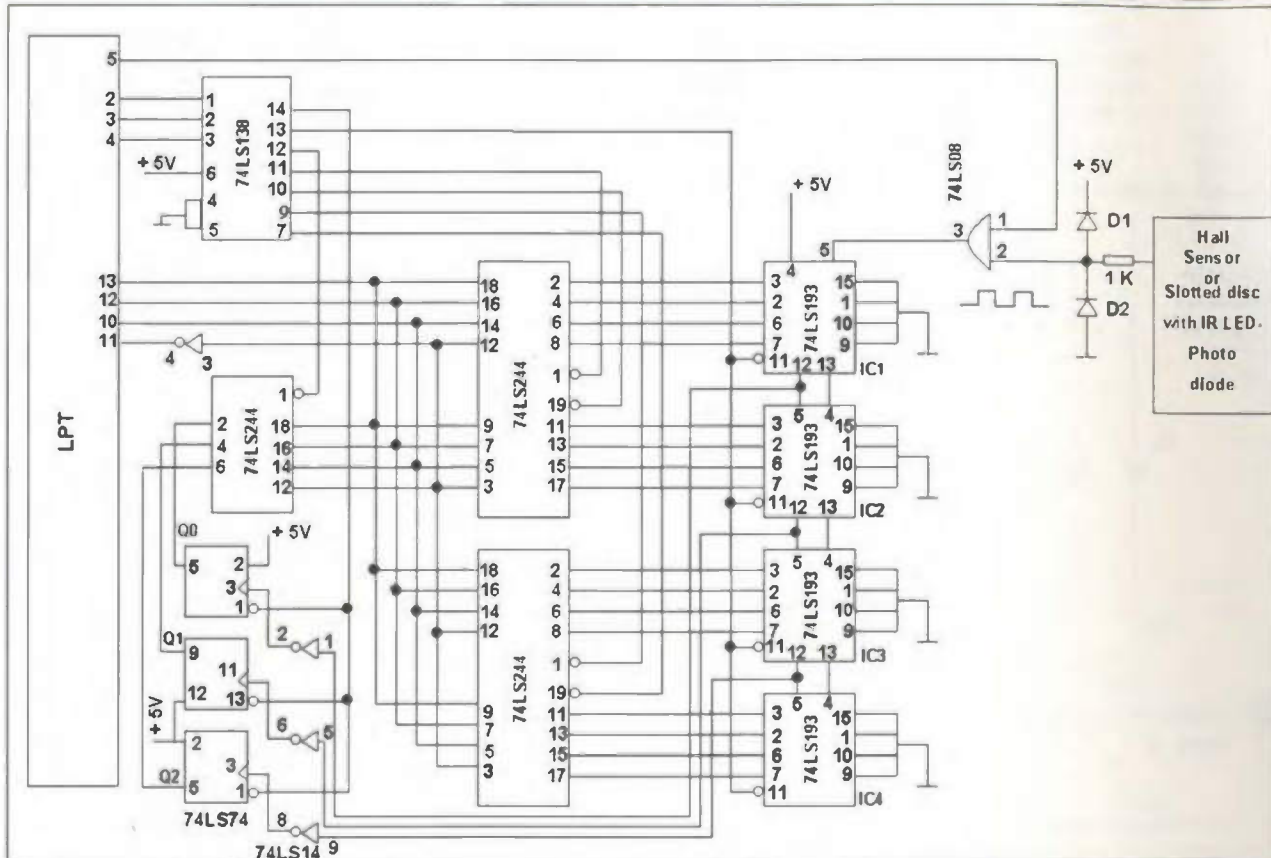
Software for the PC-based high-speed rev counter.

Turbo C code for "BIOS INT1Ch turns LPT of a PC into a high speed measuring device"

```
#include<stdio.h>
#include<dos.h>
#include<conio.h>
#include<process.h>
#include<math.h>
#define INT1C 0x1C /*INT 1Ch*/
int TIMERINTR=0; /* Time Base Variable*/
float SPEED=0.0; /* Measured SPEED*/
int CWORD=0,MCR,DP,SP;
float y1 = 0.0,y2=0.0;
void interrupt (*oldvect)(); /* INT1Ch pointer*/
void interrupt TIMEBASE(); /*Routine for Time
window of 1 sec*/
unsigned int SPEEDREAD(); /*Routine declaration
for reading the*/
/* counter output*/
void interrupt TIMEBASE() /*Our ISR for INT1Ch*/
{
    disable();
    outp(DP, 0x08); /*Enable AND gate and feed
input pulses*/
    TIMERINTR++; /*Increment continues till
TIMERINTR becomes 18*/
    oldvect();
    enable();
}
unsigned int SPEEDREAD()
{
    unsigned char nib0,nib1,nib2,nib3,byte1,byte2;
    int temp,i,a;
    y1=0.0;
    y2=0.0;
    byte1=0;
    byte2=0;
    outp(DP,0x03);
    a=((inp(SP)>>4) & 0x07); /*Read the Flipflop
output Q0Q1Q2*/
    switch(a)
    {
        case 0: /*Counter 1 alone counted, read nibble
0*/
            outp(DP,0x04);
            nib0=inp(SP); /*Read nibble 0*/
            nib0=nib0>>4;
            byte1=nib0|0x00;
            byte2=0;
            break;
        case 1: /*Counter 1 & 2 involved in counting*/
            outp(DP,0x04);
            nib0=inp(SP)>>4; /*Read nibble 0*/
            outp(DP,0x05);
            nib1=inp(SP)&0xF0; /*Read nibble 1*/
            byte1=nib1|nib0;
            byte2=0;
            break;
        case 3: /*counters1,2 & 3 involved in counting*/
            outp(DP,0x04);
            nib0=inp(SP)>>4; /*Read nibble 0*/
            outp(DP,0x05);
            nib1=inp(SP)&0xF0; /*Read nibble 1*/
            outp(DP,0x06);
            nib2=inp(SP)>>4; /*Read nibble 2*/
            byte1=nib1|nib0;
            byte2=nib2&0x0F;
            break;
        case 7: /*all counters involved in
counting*/
            outp(DP,0x04);
            nib0=inp(SP)>>4; /*Read nibble 0*/
            outp(DP,0x05);
            nib1=inp(SP)&0xF0; /*Read nibble 1*/
            outp(DP,0x06);
            nib2=inp(SP)>>4; /*Read nibble 2*/
            outp(DP,0x07);
            nib3=inp(SP)&0xF0; /*Read nibble 3*/
            byte1 = nib1|nib0;
            byte2 = nib3|nib2;
            break;
        default:
            printf("\n Error ! ...Check the flip flops");
            exit(1);
    }
    /* Convert binary words byte1,byte2 to decimal*/
    for(i=8; i<16;i++)
    {
```

Table. ICs involved for the four speed ranges of the PC-based rev counter.

Flip-flop output word	Count advanced to	Counters to be read	Maximum expected speed (rev/min)
Q2 Q1 Q0			
0 0 0	IC ₁	IC ₁	16°60
0 0 1	IC ₂	IC ₁ , IC ₂	256°60
0 1 1	IC ₃	IC ₁ , IC ₂ , IC ₃	4096°60
1 1 1	IC ₄	IC ₁ , IC ₂ , IC ₃ , IC ₄	65536°60



```

temp=byte2;
byte2=byte2 &(0x01);
y1=y1+byte2*pow(2,i);
byte2=temp;
byte2=byte2>>1;
}
for(i=0; i<8;i++)
temp=byte1;
byte1=byte1 &(0x01);
y2=y2+byte1*pow(2,i);
byte1=temp;
byte1=byte1>>1;
)
CUWORD=(y1+y2);
return(CUWORD); /*Return decimal equivalent*/
}
int main() /*Main Program starts here*/
{
float CF=18.2/18;
clrscr();
printf("BIOS INT1Ch turns LPT into a high speed
measuring);
printf("\n\t\t\t device by");
printf("\n\t\t\t K.Suresh,MSD,IGCAR,Kalpakkam,Tamil
Nadu,India");
DP = peek(0x40,8); /*check up for availability of
printer port */
if (DP==0)
{
printf("\n\n LPT NOT AVAILABLE!
EXITING");
exit(1);
}
}
)
printf("\n\n LPT address = 0x%X",DP);/* Address of
DATAPORT*/
SP = DP-1; /*Address of STATUS PORT*/
while(!kbhit())
{
oldvect=getvect(0x1C); /*store vector address
of INT1Ch*/
outp(DP,0x00); /*Disable AND and inhibit
pulses to counter*/
outp(DP,0x02); /* Load the counters
with 0x00h*/
outp(DP,0x01); /*Clear all Flip
Flops*/
while(TIMERINTR<=18)/*Wait for time base
generation of 1 sec*/
{
setvect(0x1C,TIMEBASE);
}
outp(DP,0x00); /*Disable AND*/
setvect(0x1C,oldvect);
SPEEDREAD(); /*Read speed routine */
SPEED= CUWORD*CF;
gotoxy(10,10);
printf("\nSPEED in RPS=%.1f", SPEED);/*Display speed in
RPS*/
printf("\nSPEED in RPM=%.1f",SPEED*60); /*Display
speed in RPM*/
TIMERINTR=0;
}/*Speed measurement continues till a key is
pressed*/
setvect(0x1C,oldvect);
}

```

Negative high-voltage regulator

When a regulated negative supply is required, the LM337T may be used, provided the input never exceeds -40V. For higher voltages, the circuit shown in Fig. 1 is simple but effective.

The p-channel MOSFET Tr_1 acts as the pass element, and can withstand up to 100V. It is regulated by means of the op-amp N_1 and voltage-

reference Z_2 , with feedback from the output provided by the potential divider consisting of R_5 and R_6 .

Output from N_1 is coupled to the gate of Tr_1 by the common-base transistor Tr_2 . If the voltage at the output terminals tries to exceed 24V, resistor R_4 moves upwards and current is drawn through Tr_2 and hence R_1 , tending to turn Tr_1 off by

reducing its gate voltage.

Zener diode Z_1 protects the gate against excessive voltage - which might occur when V_{in} is first applied. With the components shown, the circuit is stable with C_3 omitted.

The current that may safely be drawn depends primarily on the heat sinking of Tr_1 . Sense-resistor R_2 begins to turn the clamp Tr_3 on at approximately 250mA, thus robbing Tr_1 of more gate drive and limiting any further increase in output current.

Tests with the input varying from -65V down to -30V yielded less than 1mV shift in the -24V output; in fact the only observable change was from the warm-up of Z_2 , roughly 200ppm/°C. This could be improved by using a better device, such as 1N821 (100ppm/°C).

With the load current varied from zero to 200mA, output voltage fell by only 0.25mV, which implies an output resistance of less than 1.5mΩ. Using a raw supply that gave ripple of 4V pk-pk at 100Hz, the measured output ripple and noise was 0.2mV pk-pk, which is a stabilisation ratio of 20000:1.

The minimum input/output difference for regulation is less than 6V at 250mA loading. For inputs higher than -85V, the value of R_1 should be increased to reduce its dissipation.

If V_{in} is likely to exceed 100V, Tr_1 should be updated, for example to an IRF9630 for 200V, or an MTP2PSOE for 500V. Output voltages other than the -24V shown in Fig. 1 can readily be obtained by scaling R_6 accordingly, but for more than -36V it is advisable to restrict the voltage supplied to N_1 ; this is easily done by adding R_8 and Z_3 , and re-arranging the feedback as illustrated in Fig. 2.

C J D Catto
Cambridge

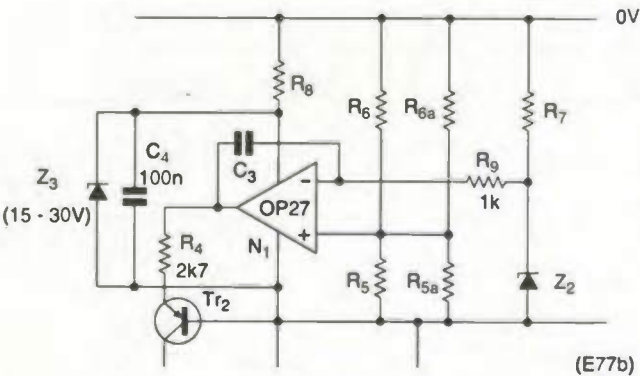
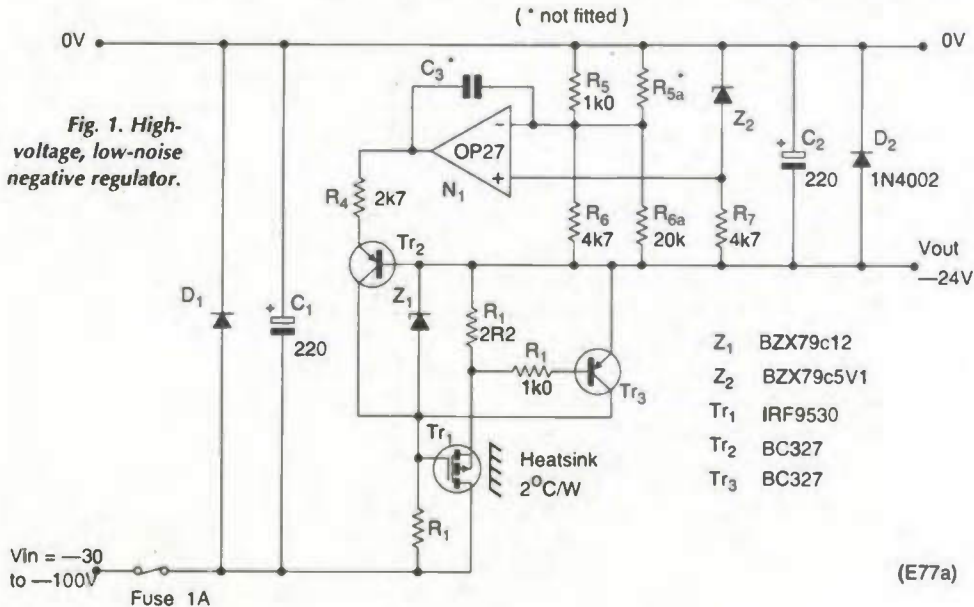


Fig. 2. Rearranging the circuit around the op-amp in this way allows the circuit to give outputs greater than 36V.

High-gain amplifier uses medium-power MOSFET

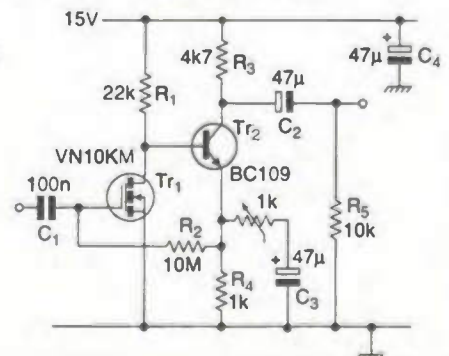
This is a very good little general-purpose pre-amplifier, which was designed as part of a bigger project. It is rather unorthodox in that it uses a medium power MOSFET as the input device, but this is what gives it the edge over other designs.

It has a medium to high gain, (300 to 4000 adjustable or -50 to 72dB), a high input impedance of 10MΩ and a bandwidth of 10Hz to 100kHz making it suitable for many audio applications. It is self-biasing, which takes care of the wide spread of threshold voltages for the FET, and it is stable with a

wide variety of input and output devices.

Output voltage can be as high as 8V pk-pk and so it can drive into most high power amplifiers directly. The VN10KM may be replaced with a 2N7000, which is probably more readily available.

Jack Paterson
Livingston
Scotland
E82



Radio works off 2V

Originally, the low-power radio circuit in Fig. 1a) appeared in circuit ideas in the April 2000 issue. The circuit was so puzzling that I just had to see if it worked. The circuit reappeared as the winning entry in the June 2000 edition.

What is even more intriguing is that with a few changes, the sensitivity of the radio can be drastically increased, and at the same time made to consume even less power.

The original circuit is a negative resistance oscillator. Bases of the transistors are tied to each other collectors, making the oscillation swing very low in amplitude, and resulting in very low power consumption. The low slope of negative

resistance means that the oscillator can be very responsive and lock easily to external disturbances such as a radio signal.

By rearranging the circuit slightly, and by adding an extra 10uH coil, the circuit can be turned into a super-regenerative receiver Fig. 1b).

RF oscillations are blocked by the quenching action of the RC network in the emitter circuit. The radio only consumes power during oscillations, and sensitivity is sharply increased due to the regenerative action. Consumption is less than 12µA with a 2V supply.

Eddy Insam
Via e-mail
F71

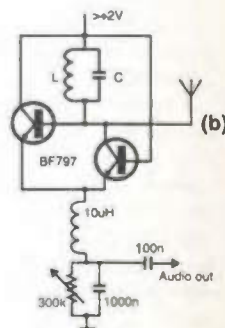
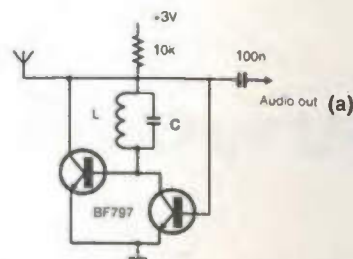


Fig. 1. In a) is the low-power radio receiver, as published in the April 2000 issue. Fig. 1b) shows how the original circuit can be modified slightly to produce a super-regenerative receiver.

Constant-current ripple reducer

Often, it is necessary to reduce the ripple on a power supply line. Usually, the solution is just to increase the filtering capacitance.

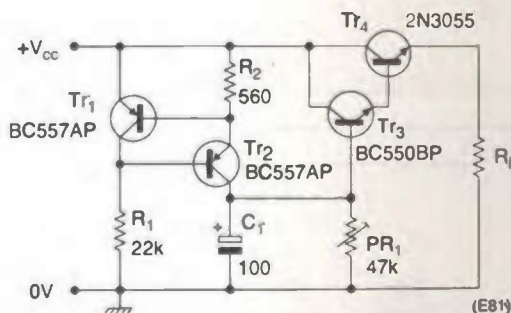
There is however an alternative possibility, using a capacitance multiplier circuit. The circuit shown was specifically designed as part of a Class-A amplifier power supply. It reduced 2V pk-pk ripple to less than 100µV while delivering 2A into the load. The only disadvantage is the 2.5V headroom required by the circuit.

Transistors Tr_1 and Tr_2 , together with R_1 and R_2 , form a constant current source providing 1mA into

PR_1 . Transistors Tr_3 and Tr_4 form an output Darlington to provide current to the load. The high level of ripple reduction is due to the high impedance at Tr_2 's collector, which forms an AC voltage divider with the low impedance of C_1 . Potentiometer PR_1 allows the output voltage to be adjusted from zero to about 45V given a suitable power supply.

If the load current is less than an amp, Tr_3 and Tr_4 can be replaced with a TIP120 device.

Jeff Macaulay
Chichester
West Sussex
E81



Low-dissipation Class A amplifier

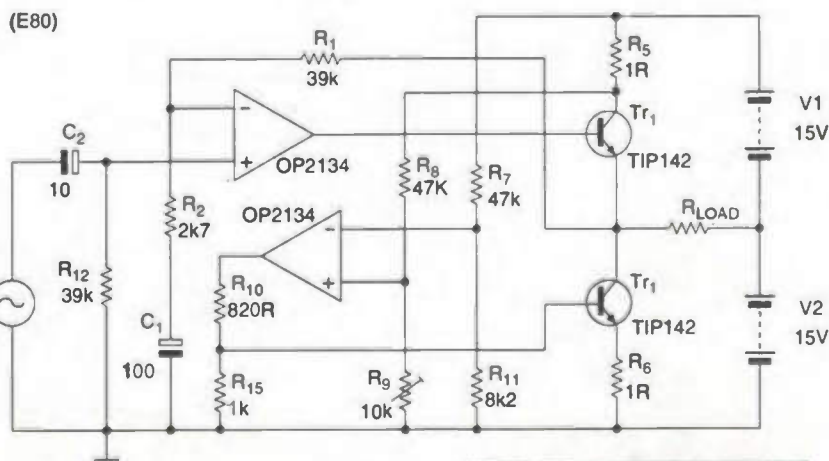
This amplifier keeps the output transistor Tr_1 operating in Class A regardless of signal potential but without the large standing current that would normally be required. The circuit is initially set up with the preset R_9 so that 50mA flows in the output stage.

Op-amp A_2 continuously monitors the current flow through Tr_1 via the voltage drop across R_5 . When Tr_1 's base is driven positive it turns on in the normal manner delivering power to the load. As it does so the non inverting input of A_2 goes positive and its output goes negative cutting off Tr_2 .

On negative going signal excursions A_2 biases Tr_2 in such a way as to maintain the 50mA flow through R_5 . Transistor Tr_2 thus supplies the load current, plus 50mA. Hence Tr_1 is always conducting and operating as a single-ended class-A stage.

Signal voltage supplied to the load is determined by A_1 and the associated negative feedback loop R_1, R_2 and C_1 . By using the low noise, low distortion 2134 IC, distortion levels are below 0.01% and the bandwidth extends from 2Hz to around 100kHz.

Jeff Macaulay
Chichester
West Sussex
E80



£75 winner

Opto couplers detect three-phase faults

In many applications using three-phase AC mains, such as motors, high-current DC power supplies, lighting systems, etc., presence of all the three phases is essential. Failure of even one phase can lead to a major failure of the systems being powered.

A simple and inexpensive opto-coupler based circuit shown in the diagram continuously monitors the presence of all three phases, and in the event that one or more phase fails, it immediately cuts off the three-phase AC input to the system and lights a lamp to indicate the failure. In the circuit, each phase input is attenuated and converted to a dc voltage to drive the LED of an

opto coupler, OC, by R_1 , D , C and R_2 . The diodes D_1 to D_3 and the transistor Q form a NOR gate.

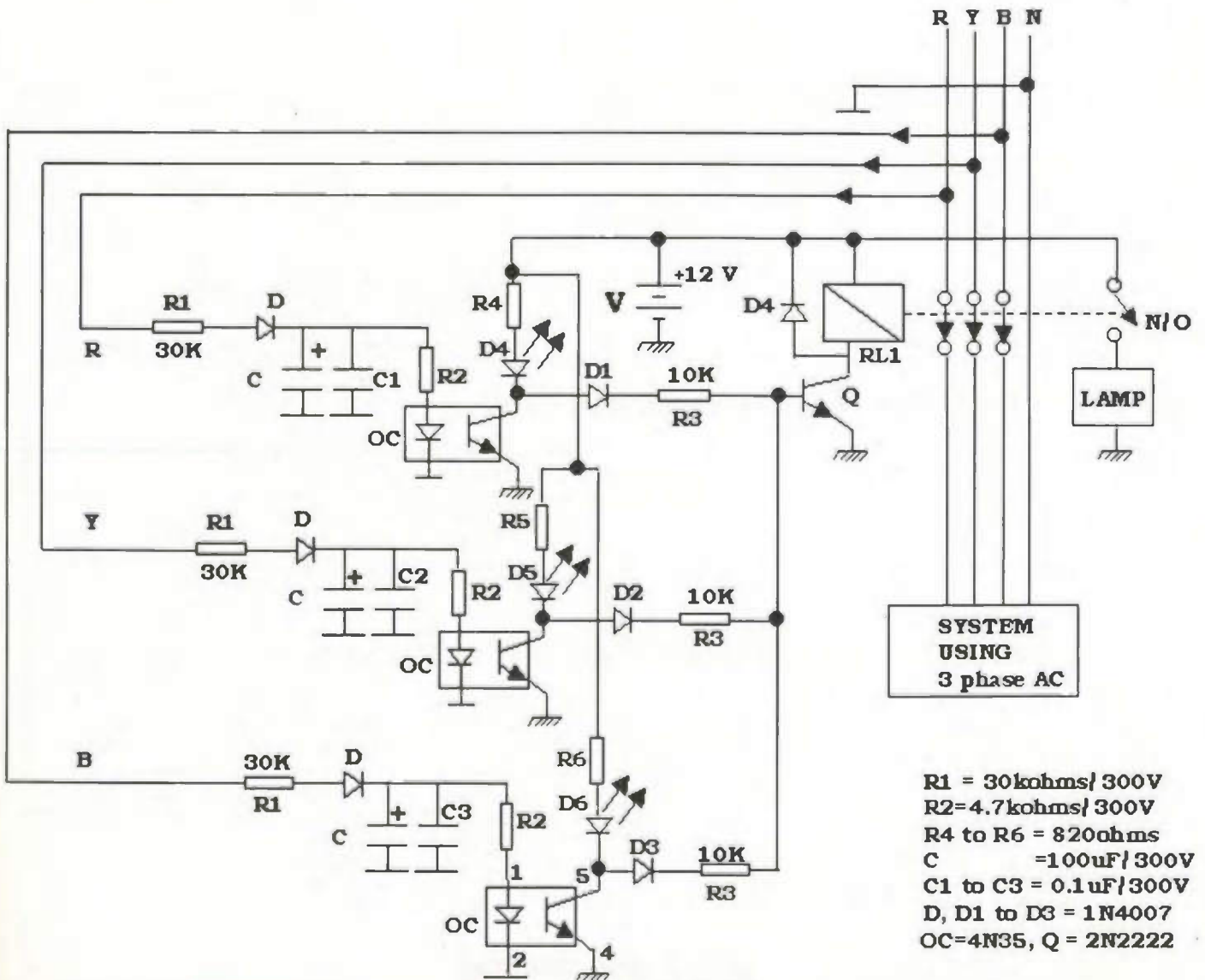
An independent 12V battery powers the photo transistors of the opto coupler and the driver transistor Q to energise the relay RL_1 . In normal conditions, where all three phases are present, the photo transistors of all the opto couplers are on and LEDs $D_{4,6}$ are lit, indicating the presence of all the three phases, and the NOR gate does not energise the transistor Q . Consequently RL_1 is in the off condition and the three phase AC input is applied to the system through its N/C contacts.

When one or more phases fail, the opto couplers in the failed phases are

off. Hence the NOR gate turns on transistor Q to energise the relay RL_1 . This cuts off the three phase supply to the protected system while switching on the lamp through one of its normally-open contacts.

By being off or on, LEDs $D_{4,6}$ indicate which phases have failed or are OK. If the supply to the protected area also supplies its lighting, the lamp can be arranged to light the area during a fault, facilitating fault-finding and repairs. For this reason, the lamp is powered by the 12V battery rather than the mains.

K Suresh
Kalpakkam
India
E89



- R1 = 30kohms/ 300V
- R2 = 4.7kohms/ 300V
- R4 to R6 = 820ohms
- C = 100uF/ 300V
- C1 to C3 = 0.1uF/300V
- D, D1 to D3 = 1N4007
- OC = 4N35, Q = 2N2222

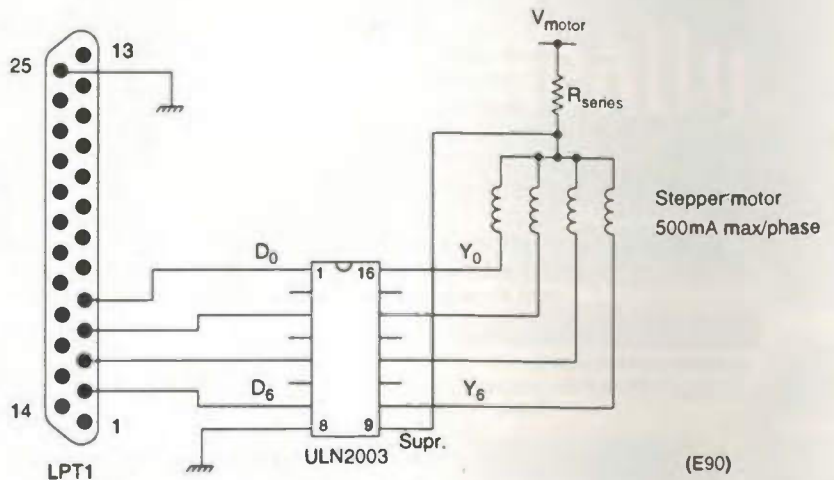
Stepper motor driver using PC printer port LPT1

Small stepper motors are often used for motion control. This simple and inexpensive stepper motor driver is for small stepper motors taking up to 500mA. It uses only one external component, a ULN2003, and the printer port LPT1 of a PC. The stepping sequence is generated with PC parallel port and the phase current is amplified with ULN2003. This IC has internal diodes to suppress the inductive transients of the motor coils, and can sink up to 500mA per phase. This is a resistive-inductive unipolar type of drive. The value of the series resistance is calculated as follows,

$$R_{series} = (V_{supply} - I_{motor} \times R_{coil}) \div I_{motor}$$

The program generates the sequence and outputs it through LPT1. The program given in the listing implements only a few basic features like half step, full step, number of steps and direction control. The software can be readily modified to include more features, such as acceleration, deceleration and variable speed.

Vikas W Meshram
Indore



(E90)

Listing for the LPT controlled stepper motor driver. This software only implements a few basic features like half step, full step, number of steps and direction control but it can be modified to include features such as acceleration, deceleration and variable speed.

```
#include <stdio.h>
#include <dos.h>
#include <ctype.h>
#include <conio.h>
#define LPT1 0x378
#define R 1
#define F 0
void FullStep(int steps);
void HalfStep(int steps);
void park(void);
struct FSData { unsigned char data;
                }DataFS[ ] = {10,9,5,6};
struct HSData { unsigned char data;
                }DataHS[ ] = {10,8,9,1,5,4,6,2};
int i;
char StepMode,dir;
void main (void)
{
    int NoOfSteps;
    park ();
    printf ("\nFull/Half Step(H/F):");
    StepMode = getche ();
    printf ("\nNo of Steps:");
    scanf ("%d",&NoOfSteps);
    printf ("\ndir: (F=1,R=0)");
    scanf ("%d",&dir);
    switch (toupper (StepMode) ) {
        case 'F' :
            FullStep (NoOfSteps) ;
            park () ;
            break ;
        case 'H' :
            HalfStep (NoOfSteps);
            park () ;
            break;
        default:
            FullStep (NoOfSteps);
            park () ;
            break;
    }
}
void FullStep (int steps)
{
    char BufStart ;
    char Full=4;
    if (dir) {
        BufStart=0;
        for (i=0; i<=steps;i++) {
            output (LPT1,
                    DataFS[BufStart].data);
            delay (30);
            BufStart++;
            if (BufStart==Full)BufStart = 0;
        }
    }
    else{
        BufStart=4;
        for (i=0; i<=steps;i++){
            output (LPT1,
                    DataFS[BufStart].data) ;
            delay (30);
            BufStart--;
            if (BufStart==0)
                BufStart = 4 ;
        }
    }
}
void HalfStep (int steps)
{
    char BufStart;
    char Full=8;
    if (dir){
        BufStart=0;
        for (i=0; i<=steps; i++) {
            output (LPT1,DataHS[BufStart].data);
            delay (30);
            BufStart++;
            if (BufStart==Full)
                BufStart = 0;
        }
    }
    else{
        BufStart=8;
        for (i=0; i<=steps; i++) {
            output (LPT1,DataHS[BufStart].data);
            delay (30);
            BufStart --;
            if (BufStart==0)
                BufStart = 8;
        }
    }
}
void park (void)
{
    output (LPT1,0);
}
```

Single IC oscillator/doubler

Analog Devices' AD725 RGB-to-PAL/NTSC encoder integrated circuit needs a $4F_{SC}$ clock signal of 17.734475MHz ($4 \times F_{SC}$ PAL frequency). This is usually not widely available as a crystal frequency.

The chip's application note suggests a design for a low-cost crystal oscillator with a 17.734475MHz crystal¹ and I have used the idea to build a clock with a

commonly available 8.867238MHz crystal. The generated frequency thus needs to be doubled.

The crystal oscillator uses one inverter in a 74HC04 IC, and a second as a buffer. The frequency doubler has been designed using the other four inverters in the same IC. The result is a frequency doubler using six inverters, two diodes and a few passive components.

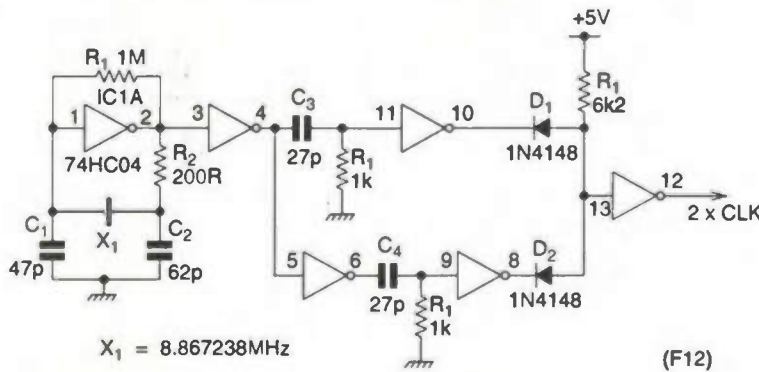
Crystal oscillator U_{1A} and

associated components is the design suggested in the application note, but operating at half the required frequency. Inverter U_{1B} buffers the signal, to avoid loading the oscillator.

The two edge-detector circuits, C_3 - R_3 and C_4 - R_4 , differentiate complementary squarewaves. Positive spikes are amplified and buffered by U_{1D} , U_{1E} , and combined by the diode negative OR gate D_1 , D_2 . Gate output is buffered and negated by the inverter U_{1F} .

The circuit works well and can be used with various crystal frequencies if the appropriate modifications in the crystal oscillator component values are made.

Emil Vladkov
Sofia
Bulgaria



Reference

1. Low Cost RGB to NTSC/PAL encoder with Luma Trap Port., AD725 Data Sheet, Rev. 0, Analog Devices Inc., 1997, pp. 12-13.

Low-cost load-sensing and switching circuit

This arrangement allows the operation of an ac mains electrical load of 250W or less to be sensed, and for power to be switched to a slave circuit. This is particularly useful where the load is remote and the wiring difficult of access.

Where the load also has a local switch, such as a time-switch, thermostat or PIR, additional wiring to carry a control signal is eliminated.

Applications could include:

- Synchronisation of a ventilation fan with other equipment
- Basic remote monitoring of a load, e.g., positive indication of load status in a home automation circuit, or security system
- Slave indicator of PIR operation, or synchronisation of several floodlights with one PIR detector
- Slave operation of several domes-

tic lamps through switching on one lamp.

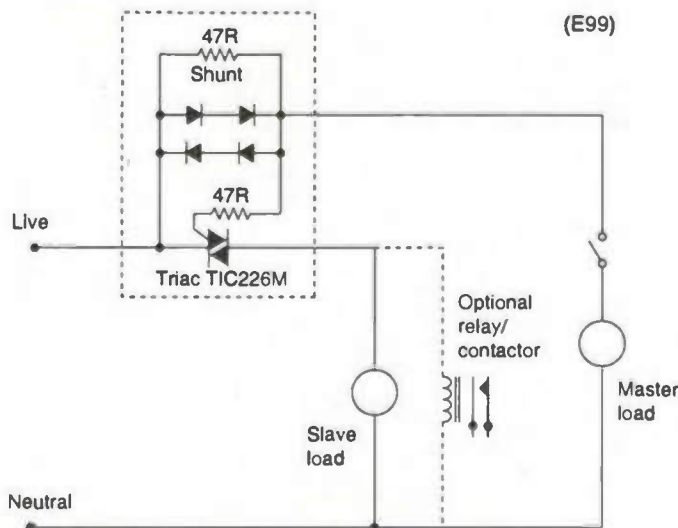
When the master load is switched into the circuit, alternating current flowing through the diodes develops a positive and negative voltage, equivalent to double the diode forward voltage drop. This triggers the triac via the 47Ω resistor within a few degrees of zero crossover.

Although RF levels in the power circuit caused by the triac switching are relatively low, there may be some configurations where additional filtration may be required.

If the load being sensed includes a residual current drain, such as a time-switch or sensor circuit, discrimination against a load of up to 5W can be achieved by including a 0.25W 47Ω shunt resistor in the position shown.

Thermal considerations caused by current flow in the diodes and triac affect the master and slave circuit ratings, and heat sinking or relays should be deployed if necessary. The circuit as illustrated is adequate for 250W master and slave loads without additional heat sinking.

R J Phipps
Bedford
E99



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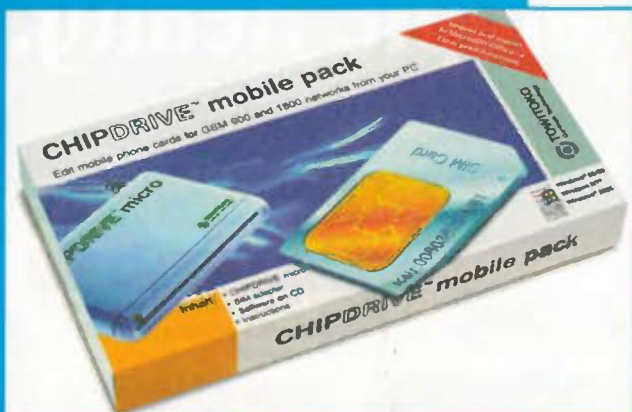
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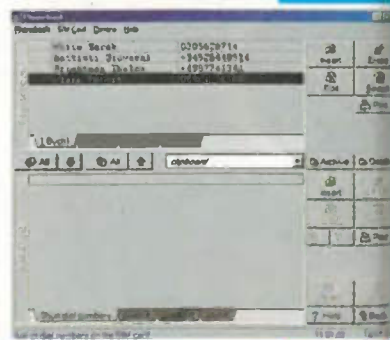
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Impedance meter

Although designed for checking electrolytic capacitor health, **Cyril Bateman's** impedance meter has a variety of uses. Operating at a test frequency of 100kHz, the meter has two ranges, 0-2 Ω and 0-20 Ω .

Combining the charged capacitor protection circuits and four terminal input stage designs from my tan δ meter¹ with the impedance measuring stages from my high-frequency impedance meter², I have designed a self contained, hand held, battery powered 100kHz impedance meter. Using a 150mV maximum test voltage, it provides two measurement ranges – 0-2 Ω and 0-20 Ω . The result is displayed on a low-cost 3.5-digit PM128 liquid-crystal panel meter.

While designed for measuring the impedance of capacitors, this meter can be used to measure any impedance within its ranges, whether inductive, capacitive or resistive. Indeed 1% metal film resistors were used to confirm measurement accuracy.

This 100kHz meter is intended to aid circuit design by accurately measuring the impedance of stock capacitors inserted into a four terminal test jig. Suitable jigs can be assembled using commercial Kelvin contact four-terminal test sockets, to accept axial or radial leaded devices.

Attached to 'panel mount plugs',

these four-terminal jigs plug into the 100kHz impedance meter front panel. This facilitates the measurement of axial or radial lead components. My prototype used four-terminal test sockets³ manufactured by Textool, Fig. 1.

At higher frequencies the 'soldered into stripline' test jig approach of my high frequency |Z| meter² becomes mandatory, but at 100kHz, using plug-in Kelvin contact test jigs is feasible.

Fig. 1. Measuring a 220 μ F OSCON capacitor that has been inserted into the radial-lead jig. This Textool test socket is available from RS Components.



Table 1. Wayne Kerr 6425/HP16047A measurements of a Philips 1000µF 25V type 135 capacitor.

Frequency	Capacitance	Tanδ	ESR (mΩ)	Z (mΩ)	Phase angle (°)
100Hz	980.2µF	0.06405	103.95	1627	-86.34
300Hz	964.7µF	0.1457	80.15	555.6	-81.72
1kHz	940.4µF	0.4175	70.595	183.4	-67.3
3kHz	898.9µF	1.1228	66.335	88.64	-41.7
10kHz	836.2µF	3.3024	62.83	65.78	-16.82
30kHz	820µF	9.496	61.5	61.6	-6.0
100kHz	0.7nH	115	59.4	59.5	+0.4
300kHz	5.4nH	5.37	57.6	69.6	+28.0

using a co-located sense resistor, two voltage measurements suffice, Fig. 3.

Four-wire test leads and jigs

At DC and low frequencies, designing four wire test leads/jigs is straightforward, provided the current leads used have low resistance compared to the generator's output impedance. As test frequency increases – and especially when measuring impedances⁴ less than 1Ω – a large test signal current flows through the current leads. This current couples an error voltage into the voltage sensing leads, resulting in incorrect readings.

At 100kHz and above, test lead inductance combined with mutual inductive coupling provides the main source of measurement errors. These can only be minimised by using the 'four-terminal pair' configuration.⁴ There's more on this in the panel entitled, 'Four terminal pairs'.

The classic four-terminal pair configuration uses four coaxial test cables. All four cable outer screens are bonded together only at the end of the cables where the device is being tested. This provides four fully shielded paths from the test meter to the four-terminal test jig. Two coaxial cables supply the test current and two monitor the voltage drop across the device being tested.

When measuring capacitors mounted on a PCB, a four terminal test jig with these essential earth paths is not practicable. Two test prods must be used. Routing a four-terminal pair system into two test

Despite using Kelvin contacts, mutual coupling between their voltage and current contacts introduces minor measurement errors. This can be corrected by offsetting a voltage measurement.

Alternatively, these plug-in jigs can be replaced by four-wire test leads fitted with test probes. The meter's 150mV maximum test signal then allows it to be used for accurate in-circuit capacitor measurements, Fig. 2.

While the general circuit arrangement looks quite different, this meter uses exactly the same 'method 3' used in my high frequency |Z| meter² but with one essential difference. The current-sense resistor for that design was co-located with the device under test. The current sense resistors for this 100kHz portable meter are located within the test meter's body.

To ignore the voltage drop along the return current lead of the device being tested now requires the measurement of three voltages. When



Fig. 2. Prototype meter performing an in-circuit measurement using my 'best' four-wire test-lead set.

Fig. 3. Two versions of my 'Method 3' impedance measurement. Using the stripline test jig and co-located sense resistor my High Frequency |Z| meter needed two voltages only. With the sense resistors mounted in the meter body, the voltage drop along the current return wire necessitates an extra measurement V3, the voltage at the sense resistor.

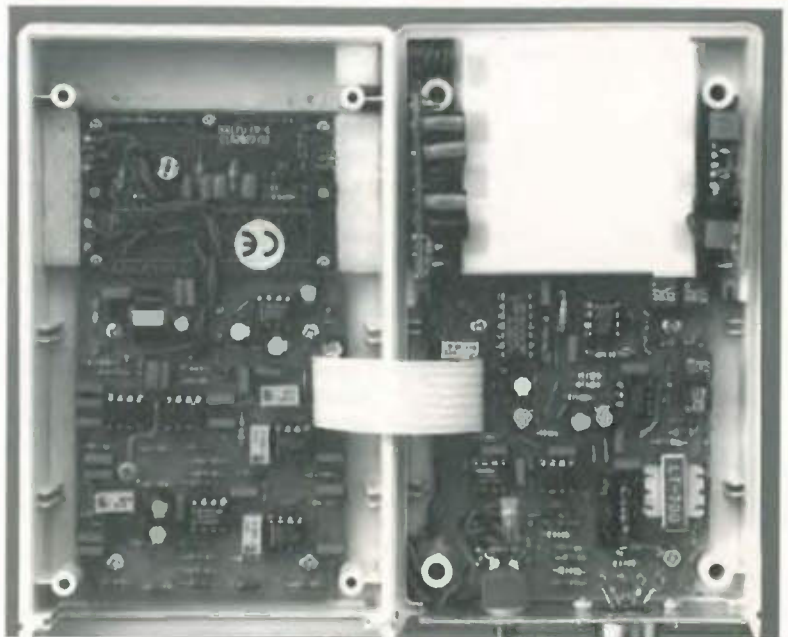
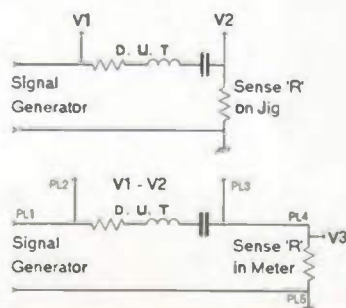


Fig. 4. Prototype meter opened to show the general arrangement. The PM128 display and the 'top' board can be seen on the left. The +5V power supply and auto-off boards are located either side of the battery box top right. The top and

prods while maintaining adequate earth paths and minimising mutual coupling, proved extremely difficult. More on this later.

Why not simply use two test leads?

A conventional test-lead pair exhibits some 100mΩ of DC resistance and around 2μH of self inductance. Even worse, this inductance is not a finite value; it increases significantly with separation of the leads and test prods.

Every capacitor also exhibits some self inductance. This is why most aluminium electrolytics become series self resonant between 10kHz and 1MHz, depending on capacitance value⁵. See the panel entitled 'Capacitor impedance versus ESR' for details.

For example, with just 5nH self inductance, a typical 1000μF radial-lead capacitor resonates close to 100kHz. Above resonance, the capacitor exhibits an inductive reactance, its measured impedance increases rapidly with frequency. For accuracy it is essential that any test leads used do not influence the capacitor's series self resonant frequency, Table 1.

To illustrate this test lead effect, I measured a 47μF 50V Panasonic radial-lead electrolytic capacitor using a Wayne Kerr 6425 precision component analyser, equipped with the Hewlett Packard 16047A four terminal test fixture. This expensive test fixture is usable up to 13MHz. At 300kHz, which is the analyser's highest test frequency, this 47μF, 50V capacitor measured as -1° phase and 171.6mΩ impedance. This confirms that the capacitor's self resonance frequency was above 300kHz.

Inserting a pair of test leads caused this 47μF electrolytic capacitor to resonate between 15 and 20kHz. This is substantially less than the desired 100kHz measuring frequency, resulting in gross impedance measurement errors when the device is re-measured at 100kHz.

Using a Wayne Kerr analyser at 100kHz with the capacitor mounted directly in the Hewlett Packard test fixture, the impedance of this capacitor measured 171.6mΩ. Replacing the test fixture with a pair of standard two wire test leads, the capacitor/test lead combination measured 1.26Ω.

As capacitance value increases, a two-wire test leads/capacitor combination will resonate at ever lower frequency, further increasing measured impedance errors. These errors can only be eliminated by

using the 'four-wire pair' test lead configuration. There's more on this in the panel entitled, 'Four-terminal pairs'.

To ensure a professional finished appearance, I used the same OKW case as used for my tanδ meter design¹. The PCB is again split into two halves, one mounted in the

bottom of the case adjacent to the batteries, the other in the top half adjacent to the display module, Fig. 4.

Bottom board

This board houses the test signal generator, the range switching relay and both the voltage and current

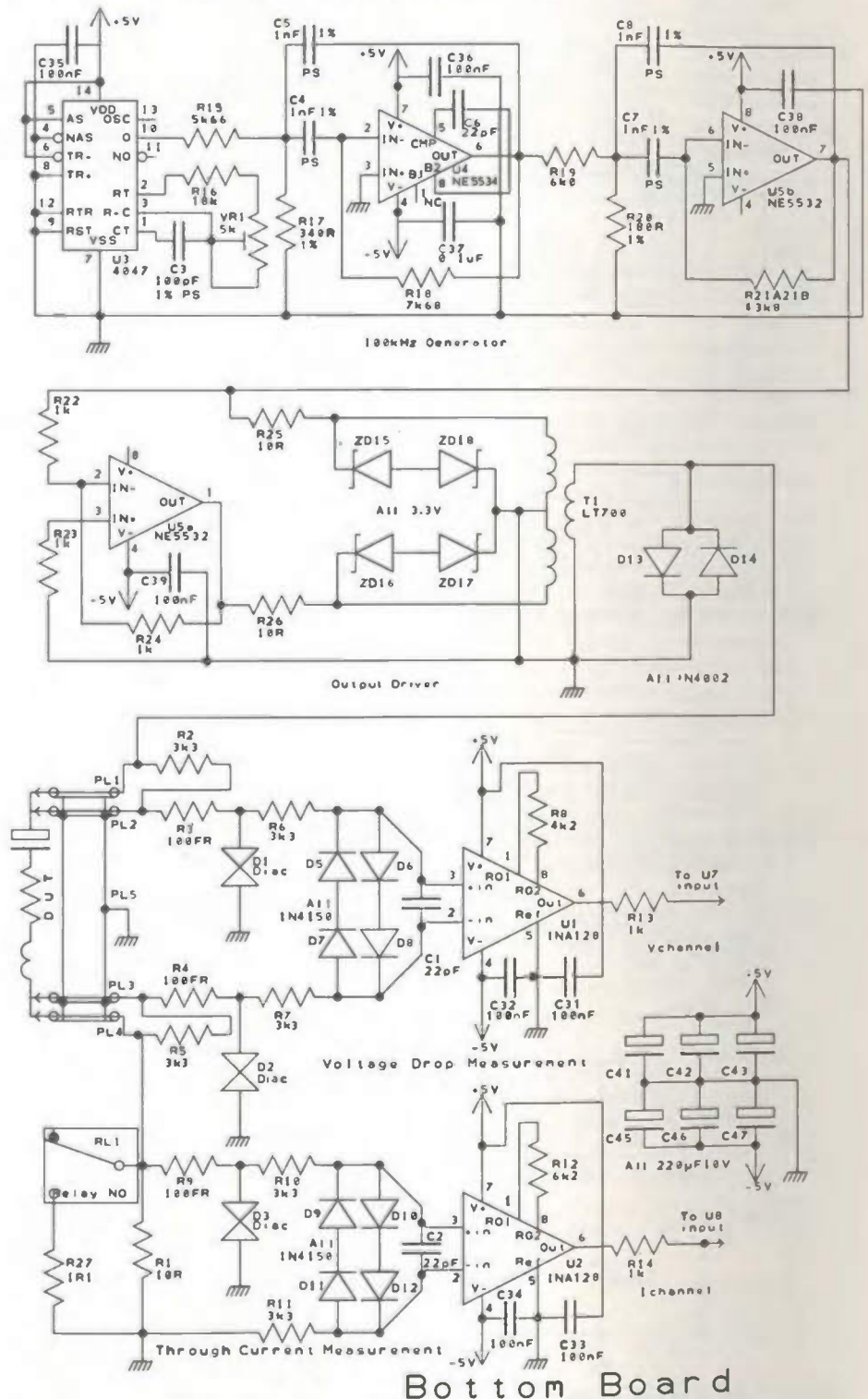


Fig. 5. Full schematic for the 'Bottom' or right hand board of Fig. 4, including all decoupling used. The 100kHz signal source is provided by the upper circuits, the voltage and current channel front ends comprise the lower half.

measuring differential input amplifiers together with their charged capacitor protection circuits, Fig. 5.

Isolating the four-terminal current and voltage sensing paths within the test meter PCB proved relatively easy. Adequate separation of components together with Faraday

shielding earth tracks was found to suffice for the original prototype single range design.

Current-sensing resistors. The relay switching needed to provide two measurement ranges complicates the design because of the inductance

introduced by the relay contacts. The physical size of a relay also increases the PCB path lengths. At 100kHz these added inductances become important, so they must be compensated for by reducing the actual sense resistor value used for the 2Ω range.

While it preferable to use non-inductive current-sense resistors, at 100kHz normal metal-film resistor inductance is small compared to the relay and PCB inductances. However,

Setting up

With a frequency meter connected between PL_1 and ground, adjust VR_1 to read 100kHz. Alternatively, connect a DVM between the V_{out} test-pin, adjacent to U_{10} , and ground. Adjust VR_1 to maximise this voltage.

With nothing connected between pins PL_1 through PL_4 , connect a DVM between R_{73} and earth at the end nearest to U_8 . Adjust VR_{S2} to read zero volts.

Solder a good short circuit directly to pins PL_2 and PL_3 , taking care to make contact only to these pins. Connect a DVM between R_{53} and earth at the end nearest to U_7 . Adjust VR_{S1} to read zero volts.

Replace the short circuit with a known 10Ω resistor connected to PL_2 and PL_3 , link PL_1 to PL_2 also PL_3 to PL_4 and set the 100kHz |Z| meter to its 10Ω range.

Connect a temporary short circuit across R_{84} . Adjust VR_{S3} to read 10Ω on the PM128 display meter.

Replace the 10Ω resistor with a known 1Ω resistor. Set the 100kHz meter to its 1Ω range. If necessary, adjust the value of R_{27} until the display reads 1Ω.

On the prototype meter a 6.8Ω resistor in parallel with R_{27} was needed to compensate for the relay contact resistance.

Remove the resistor and links from the four test pins.

With the axial-lead four-terminal test jig fitted to the meter, insert a good short circuit into this test jig. My usual short circuit is a 7mm wide 25mm long piece of 1mm thick, polished copper sheet.

Remove the short from R_{84} and adjust VR_{S4} until the PM128 display reads zero.

Removing the short circuit from the test jig completes the setting up.

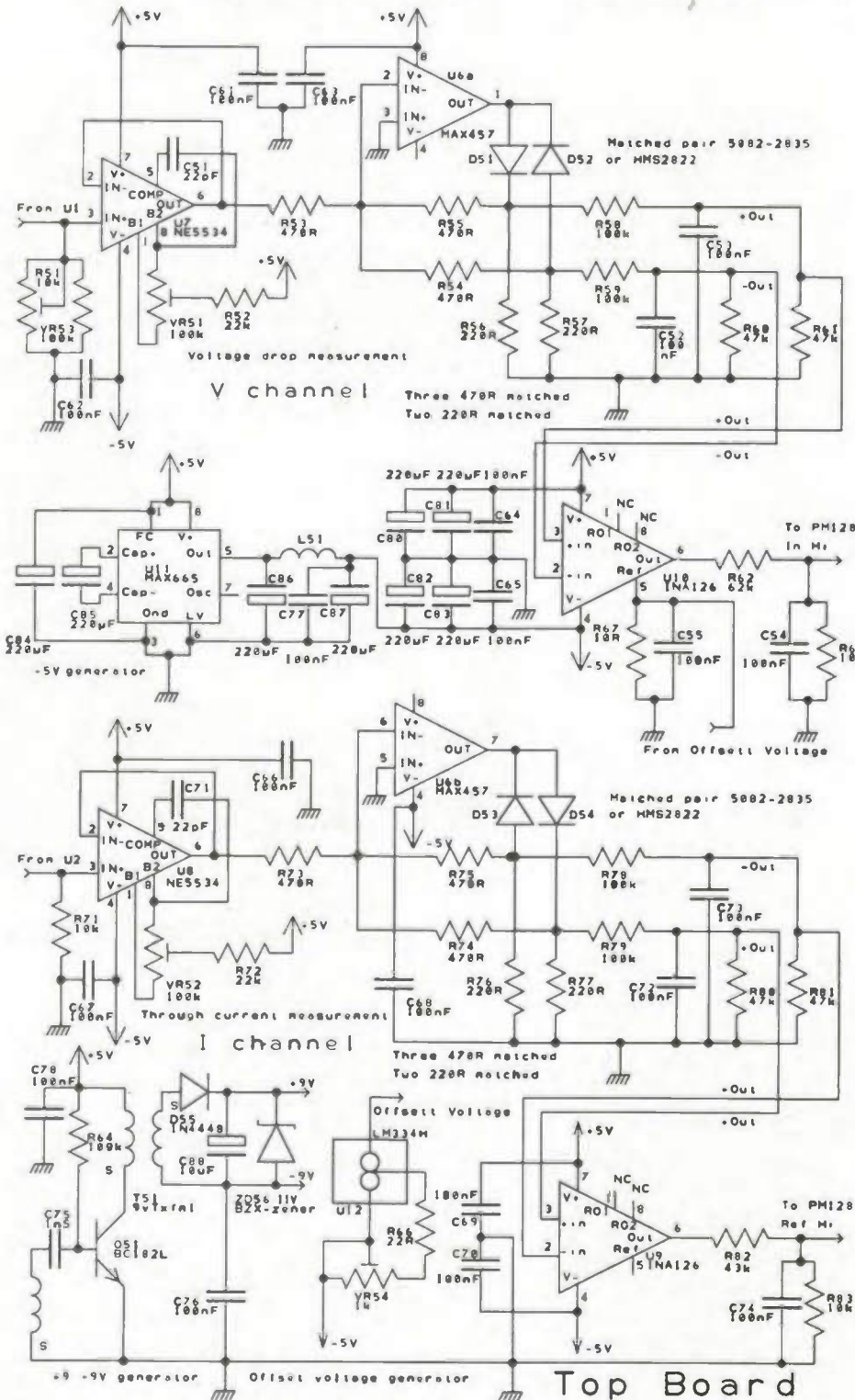


Fig. 6. The full schematic for the 'Top' or left hand board of Fig. 4. This board houses both Voltage and Current channel precision rectifiers, together with the INA126 output amplifiers. The 9V supply for the PM128 and the offset generator can be seen bottom left, the -5V generator is shown centre left of the circuits.

I designed my PCB to accommodate either low-inductance surface-mount current-sense resistors or conventional metal film types.

Signal generator. The test signal generator for my tan δ meter used a low cost LT700 audio transformer to ensure a low source impedance. I could not find a suitable high-frequency transformer for this new meter. Eventually I decided to see if the LT700 could be used at 100kHz. I was pleased to find that its reduced output could be accommodated simply by increasing drive levels.

I evaluated a number of ICs for the



Fig. 7. This photo shows the test probe ends of the three test lead sets I assembled. The 'best' set, with its external braid link, can be seen at the top of the picture. The easiest to assemble lead was made using the black four-way screened flat strip cable at the bottom.

Test-lead assembly

My 'easy' test-lead set was assembled using a low-cost four-way screened audio cable, available in cut lengths from Maplin as part number XS95D. Comprising four individually lap screened 10/0.12mm wire cores, co-jointed into a flexible flat strip cable, Fig. 8d).

All four wire screens were connected together inside the meter plug and isolated inside the test probes to produce effectively a five-terminal test arrangement. Using this test lead, resistance values of 0.47 Ω and above could be measured.

I decided I needed to use much lower resistance current leads, but four screened or coaxial cables with thicker inner conductors were much too large to fit into most four-pin plugs.

I decided to tightly pre-twist and double screen the voltage wire pair for as far as possible, and use heavier non-screened current wires to ensure an acceptable cable diameter.

My remaining two lead sets used identical voltage lead pairs but with different current leads and test prods.

The voltage pair leads were twisted and double screened to minimise mutual coupling from the current leads. I used PTFE insulated 7/0.2mm wire with some screening braid removed from RG174 coax.

At 20cm from one end of two 1 metre lengths, I carefully opened the braid wires to form exit holes. Both braids were then expanded, sliding one onto 5mm the other onto 3.5mm knitting needles.

Fig. 8. Schematic drawing of the three possible four-wire test lead configurations. The traditional four-terminal arrangement (a) is restricted to low frequencies. The five-terminal lead set (b) offers improved performance. The shielded four-terminal (c) extends both usable frequency and impedance measurement range. Bottom line, the three leads I assembled. The 'easy' lead is shown in (d), the 'intermediate' lead in (e) and the 'best' lead is (f).

The smaller diameter was drawn inside the larger, to emerge via that exit hole.

A double braid, emerging as two short, separate, single braids, produced a 'Y' jointed cable. Using the knitting needles I carefully aligned both exit holes, ensuring an easy wire path from each single braid into and through the double braid section.

Two 1 metre lengths of the PTFE coated wires were twisted together for a distance of 70cm, leaving the remainder separate. Two pull wires were inserted, one into each single braid to pass into and through the doubled braid section, and used to draw the PTFE wires into position.

This 'Y' jointed assembly was covered in yellow heat-shrink sleeving. It is clearly visible in the photograph.

Both braid screens were connected to ground in the connecting plug.

For the intermediate test leads, two lengths of 0.5mm² extra flex and the double screened length were drawn into clear heat-shrink sleeving using a slow twist to aid flexibility of the completed test lead. Emerging from the 'Y' joint, one current wire and one screened wire were passed through each FH32K probe body, then soldered together into the

brass probe tip, Fig. 8e).

Using this intermediate test lead resistance values of 0.22 Ω and above could be measured.

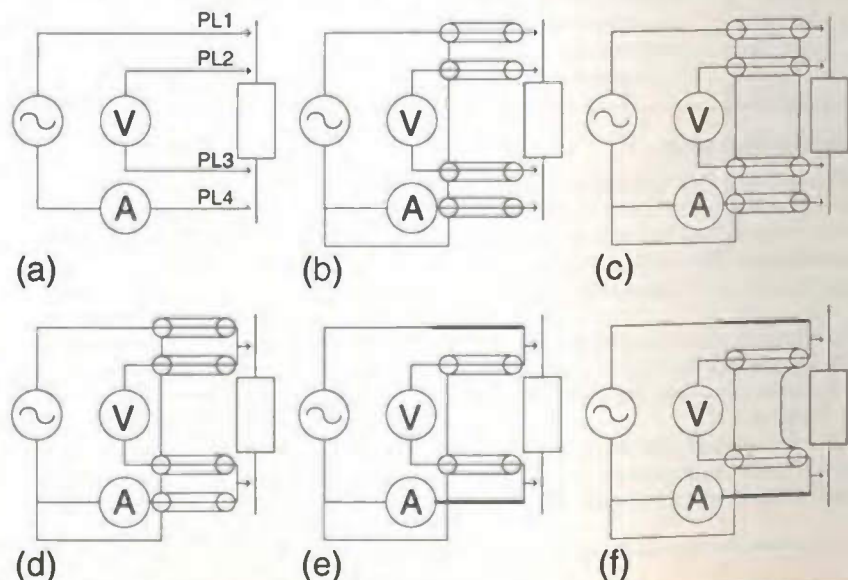
For the best accuracy test lead, a pair of Maplin HF33L pre-assembled test leads replaced the 0.5mm² wire and FK32K test prods. Otherwise, assembly generally followed that described for the intermediate lead set except for the screen braid ends at the test prods.

For this lead set the PTFE voltage wires were soldered to the test probes 1cm from the pointed ends. Both screening braids were then interconnected using another 10cm length of braid as in the photograph, Fig. 8f).

Using these best practical test leads, board mounted resistance values of 0.1 Ω and above can be measured in-circuit.

Using each of the above test leads to measure a good 100 μ F electrolytic I found much smaller differences, ranging from a measured impedance of 0.142 Ω using the best leads to 0.28 Ω for the worst.

For in-circuit diagnosis, all three test leads were able to identify a failed capacitor.



filter and output stages. Using NE5534 and NE5532 ICs with 1% polystyrene capacitors, the resistor values used in the filter stages were adjusted to provide a bandpass response. The resulting strange resistor values were obtained by parallel combinations of 1% metal films, mounted on Vero pins.

At 100kHz, variations in internal resistance in the HEF4047BP oscillator IC could result in a 7% frequency variation between ICs. This is compensated by the 5k Ω pre-set resistor VR₁.

While it is preferable to adjust this oscillator frequency to exactly 100kHz, if a frequency meter is not available, the oscillator can be adjusted to maximise the output signal. Using 1% components, the filter centre frequency for maximum output will usually be between 99 and 101kHz. This frequency accuracy should suffice.

Four-terminal inputs. The diac, fusible resistor and diode protection methods used in my tan δ meter design¹ have been closely followed except for two resistor values R₂ and R₅ which are much increased. These resistors are only needed to maintain earth continuity should the test jigs/test leads be removed. So as not to affect the voltage measurement, their value must be much larger than the impedance of any test leads used at 100kHz.

To monitor the voltage drop at the test capacitor and across the current sense resistor, two instrumentation op-amps are used. For accurate measurements, better than 40dB of common-mode rejection at 100kHz is needed.

The INA118 in-amps used in tan δ meter have been replaced by INA128 instrumentation amps. These offer the same degree of input self protection and improved high frequency performance.

To optimise the input levels to the

rectifier stages, the voltage channel INA128 IC has been set to a gain of 12.9 while the current channel INA128 IC gain is set to 9.1 times. These gain differences are compensated after rectification.

Top board

This board contains the rectifier stages needed for the V and I measurement channels, together with a floating 9V supply for the PM128 meter and the -5V converter needed to power the op-amps, Fig. 6.

Rectifier stages. I originally hoped to be able to drive the rectifier stages direct from the INA128 instrumentation amps, but these could not supply sufficient drive to the 470 Ω input resistors. So a buffer stage was needed. This could be provided using a dual amplifier, but for accuracy I also needed to minimise input offset voltages. Two single channel NE5534 ICs with offset nulling were used.

Both rectifier stages are evolved from those proven in my RF millivoltmeter and the high-frequency IZL meter. Designed to work with a 1V nominal maximum input signal, their output attenuators have been modified. This allows two INA126 instrumentation amps to be used to convert the floating outputs to single ended for the PM128 display. These have a minimum gain of five.

My earlier designs used two AD712 ICs to perform this task in a circuit with a minimum gain of two. While I was building the prototypes though, AD712s were in short supply.

With U₆ mounted in a turned-pin socket, various ICs, including the NE5532, TL2062, and TLE2072, were evaluated for use in these rectifiers. I had hoped that at 100kHz, lower cost devices would prove suitable. However as with my other designs, the Maxim MAX457 provides by far the best dynamic range and measurement accuracy.

Unfortunately it also consumes significant power – some 30mA at 5V.

Used only for in-circuit measurement, these other ICs could be used, almost doubling battery life but with the penalty of some reduction in low impedance measurement accuracy. For in-circuit diagnosis, this level of accuracy should suffice.

Measurement accuracy

After calibration, this measurement method ensures that a short circuit connected directly across the PCB terminals PL₁ through PL₄ reads as zero ohms. However with the mutual inductive coupling along the length of the four terminal contacts in the Textool jigs, the same short circuit mounted in these test jigs measures as a small resistance.

The mutual inductance produces a small, near-constant error voltage. This voltage can be overcome by introducing a similar error offset voltage into the 'Reference' input of U₁₀, pin 5. Usually, this pin connects directly to earth but I found that by connecting a 10 Ω resistor to earth, the resulting few millivolts provided the necessary error correction.

This correction voltage is arranged using an LM334 IC to generate a constant current into R₆₇. This current can be varied from near zero to several milliamps by adjusting VR₅₄, to ensure accurate measurement.

If the meter is used only with test leads for in-circuit diagnosis using the 20 Ω range, R₆₇ can be shorted and the LM334 circuitry omitted.

Power supplies

The floating 9V supply is based on Siemens' low-cost B62152A4X30 two-hole ferrite 'balun' beads. This design has proved repeatable and reliable, consuming less than 2mA at 5V to power the PM128 meter.

The ferrite is conductive, so to avoid accidental stripping of the

Four-terminal pairs

At DC and very low frequencies, the conventional four-wire test cable with Kelvin contacts can provide accurate measurements, but mutual inductance generated errors increase with frequency, Fig. 8a).

The most accurate four-wire measurement – especially for high frequency or low impedance – uses the 'four-terminal pair' method.⁴

The five-terminal and shielded four-terminal cable arrangements provide a capability between these two extremes,

Figs. 8b) & 8c)

The four-terminal pairs or '4TP' with Kelvin contacts are the best solution for any wide-range impedance measurements. From my own tests, they can be used at 100kHz with 1 metre leads without needing error correction.

This 4TP method relies on using four coaxial test cables with all four cable screening braids connected together by a low impedance path, immediately adjacent to the device under test.

Two coaxial cables supply the measurement current to and return from the

device under test. The remaining pair monitor the voltage drop across the device being tested. The inner cores of each cable should be connected separately to the component being tested to form Kelvin contacts.

The four earth braids must be directly interconnected, using typically a 1.6mm thick, 30mm by 50mm aluminium plate, or equivalent low-impedance connection, immediately adjacent to the device under test.

Unfortunately this 4TP method is not practical for in-circuit measurements.

winding wire enamel, any sharp edges around the holes should first be deburred then the cores varnished before winding.

Negative 5V supply is derived using MAX665 or TC660CPA switching inverters with low ESR electrolytics. This design provides a low noise, reliable, near -5V supply when supplied from +5V.

Display meter

Measured impedance is output as two voltages, one representing the voltage drop across the device under test, the second representing the DUT through current.

From Ohms law, these two voltages divided produce the unknown impedance. As with my $\tan\delta$ meter design, this division is most easily arranged using the PM128 digital panel meter, modified to ratio mode.

This modification is easily made by removing two resistors, R_2 and R_3 , from the PM128 module. The 100kHz $|Z|$ meter 'Ref-Hi' output from U_9 connects to the 'Ref-Hi' pad which originally linked both resistors. The 'In-Hi' output from U_{10} connects to the PM128 normal 'In-Hi' terminals and the 100kHz $|Z|$ meter ground to the PM128 input ground.

Four-wire test leads

Considerable time and experimentation was needed to devise usable test leads that could measure in-circuit impedances with acceptable accuracy. I evaluated nine different possible configurations using three differing wire types. In each case my 100kHz impedance meter was calibrated to read 0 Ω when measuring a short circuit mounted in the axial four wire test jig.

Carrying 100mA DC, the unscreened four-wire test leads from my $\tan\delta$ meter measure 35m Ω per lead. At 100kHz using these leads, a good short circuit measured 0.19 Ω to 0.4 Ω depending on lead orientation. While excellent working at 100Hz, this lead was found to be suitable only when measuring impedances greater than 1 Ω at 100kHz.

Similarly, a two-wire standard test lead that measured 62m Ω at DC per lead, at 100kHz the same lead measured this short circuit as varying between 0.74 Ω and 1.04 Ω , depending on lead orientation. This two-wire test lead was not usable at 100kHz either.

By measuring the same test capacitors and resistors inserted directly into my four-terminal test jigs and then using test leads, I was able to ascertain the measurement errors caused by my differing test

Capacitor impedance versus ESR

When an alternating current is passed through an ideal, or perfect, capacitor, having neither inductance nor resistance, the voltage waveform lags that of the current by 90°. The capacitor produces an impedance with a phase angle of -90°.

At any one frequency, a practical capacitor can be represented by a series combination of inductance, capacitance and resistance. In a practical capacitor the inductive component may be relatively constant, but not so the resistive and capacitive components. Both are strongly frequency dependent. These combine to produce an impedance $|Z|$ with a much reduced phase angle. Depending on frequency, the measured phase angle can be either positive or negative. Fig. 9.

Using the standard mathematical conversion from polar to rectangular notation, we find,

$$|Z|\angle\theta = R \pm jX$$

where R is the capacitor equivalent series resistance at the measured frequency and X

is the capacitor's net reactance at that frequency.

But as you probably know, this X comprises two elements⁴ – one a capacitive reactance X_C , the other an inductive reactance X_L .

$$|Z| = \sqrt{ESR^2 + (X_C^2 - X_L^2)}$$

From the above, you can see that ESR is not a finite value. It varies widely with frequency. Furthermore, with aluminium electrolytic capacitors it is also extremely temperature sensitive.

At one frequency only the capacitive reactance X_C and inductive reactance X_L , being equal and opposite, cancel. At all other frequencies $|Z|$ is greater than ESR, but at this one frequency only, then $|Z|$ is exactly equal to ESR.

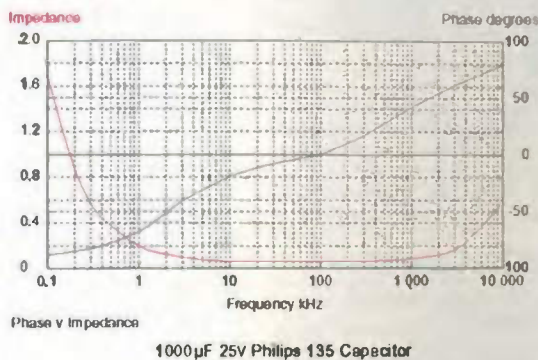


Fig. 9. Impedance and Phase Angle curves of another low ESR 100 μ F 25V Philips type 135 capacitor, measured using my High-frequency impedance meter².

lead designs exactly.

Ultimately, three test lead sets were assembled for final testing. One offered ease of assembly and use, one gave best accuracy for low impedance measurements but less ease of assembly or use, and the third represents an intermediate or compromise set, Fig. 7.

Using the 'easy-build' test leads, my short circuit measured 0.23 Ω , but resistance values of 0.47 Ω and above could be measured in-circuit, Fig. 8d). With the 'intermediate' test leads, my short circuit measured 0.075 Ω , so resistance values of 0.22 Ω and above could be measured, Fig. 8e). My best test-lead set measured this short circuit as 0.015 Ω . Board-mounted resistance values of 0.1 Ω and above, could then be measured in-circuit, Fig. 8f).

Performance

Using each of the above test leads to measure a good 100 μ F electrolytic in-circuit, I found much smaller differences. They ranged from a measured impedance of 0.142 Ω with the 'best' leads to 0.28 Ω for the 'easy build' or worst. In practice, for in-circuit diagnosis measurements, all

three assembled test leads were able to identify a failed capacitor.

Using either the axial or radial four terminal test jigs, with the meter set to its 2 Ω range, impedances of a few milliohms to 1.999 Ω were accurately measured, extending to 19.99 Ω using the meter high range.

I plan to offer professionally made PCBs for this meter together with a self adhesive front panel label, and a set of drawings and assembly notes. ■

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3. Axial or Radial Lead Sockets. (Textool) 3M Electronic Products Division, In-line socket (radial lead) RS Components part No 160-7090.
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5. Bateman C., 'Understanding Capacitors', *Electronics World*. June 1998.

Technical support

For PCB details please send a stamped, self addressed, A4 sized envelope to Cyril at the address below. Overseas readers should send an International Reply Coupon and self-addressed envelope. Cyril Bateman, Nimrod, New Road, Acle, Norfolk NR13 3BD, UK

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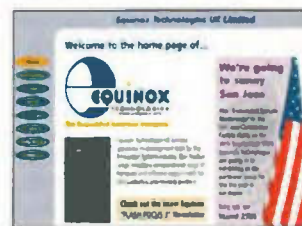
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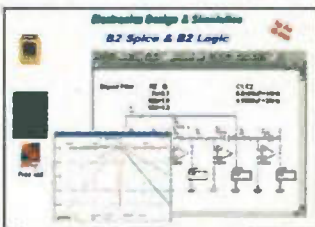
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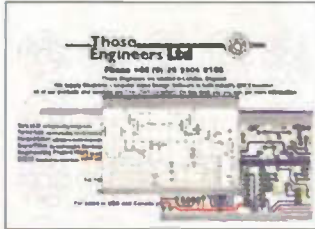
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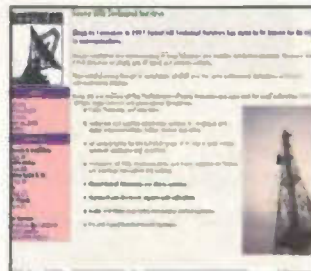
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LC resonant tanks

Joe Carr takes a look at how LC tank circuits work and describes how to get the best from them.

When you use an inductor, L , and a capacitor, C , together in the same circuit, the combination forms an LC resonant circuit. This is also sometimes called a tank circuit or resonant tank circuit. These circuits are used to select one frequency, while rejecting all others – as in to tune a radio receiver.

There are two basic forms of LC resonant tank circuit: series, Fig. 1a) and parallel Fig. 1b). These circuits have much in common, and much that makes them fundamentally different from each other.

The condition of resonance occurs when the capacitive reactance X_C and inductive reactance X_L are equal in magnitude ($+X_L = -X_C$). As a result, the resonant tank circuit shows up as purely resistive at the resonant frequency, Fig. 1c), and as a complex impedance at other frequencies.

LC resonant tank circuits operate by an oscillatory exchange of energy between the magnetic field of the inductor, and the electrostatic field of the capacitor. A current between them carries the charge.

Because the two reactances are both frequency dependent, and because they are inverse to each other, the resonance occurs at only one frequency f_r . You can calculate the standard resonance frequency by setting the two reactances equal to each other and solving for f . The result is:

$$f = \frac{1}{2\pi\sqrt{LC}} \quad (1)$$

Series-resonant circuits

Like other series circuits, the series-resonant circuit of Fig. 1a) is arranged so that the terminal current I from the source V flows in both components equally. The vector diagrams of Fig. 2a) and Fig. 2b) show the situation under three different conditions.

In Fig. 2a), the inductive reactance is larger than the capacitive reactance, so the excitation frequency is greater than f_r . Note that the voltage drop across the inductor is greater than that across the capacitor, so the total circuit looks like it contains a small inductive reactance.

In Fig. 2b), the situation is reversed: the excitation frequency is less than the resonant frequency, so the circuit looks slightly capacitive to the outside world. Finally, in Fig. 1c) the excitation frequency is at the resonant frequency, so $X_C = X_L$ and the voltage drops across the two components are equal but of opposite phase.

In a circuit that contains a resistance, inductive reactance and a capacitive reactance, there are three vectors to consider Fig. 3, plus a resultant vector. As in the other cir-

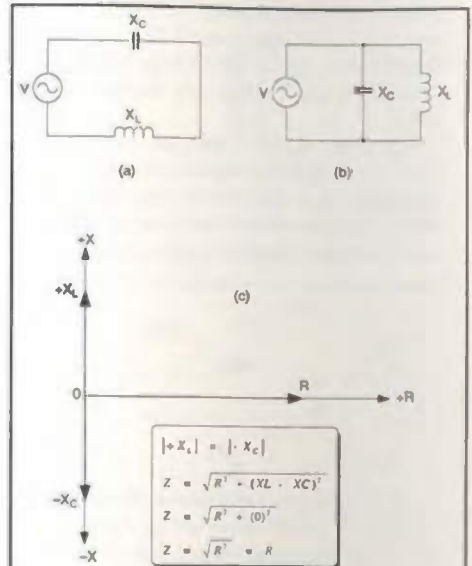


Fig. 1. There are two forms of LC tank – series, a), and parallel, b). At the resonant frequency, the tank is purely resistive, c).

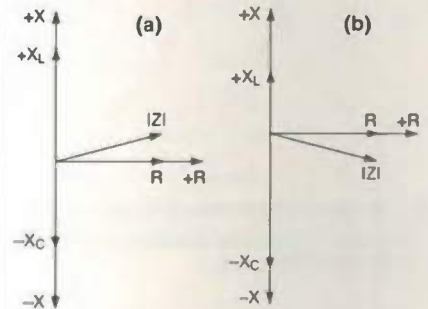


Fig. 2. In a), the inductive reactance is larger than the capacitive reactance while in b), the situation is reversed.

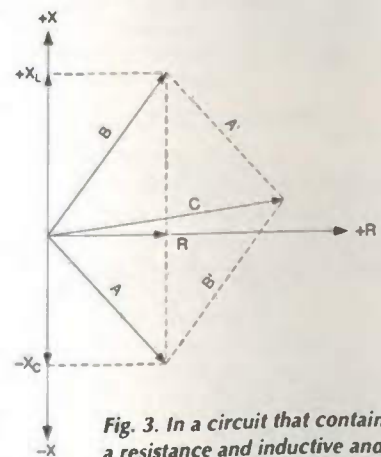


Fig. 3. In a circuit that contains a resistance and inductive and capacitive reactances, there are three vectors to consider.

Sadly, Joe passed away late last year. I don't know enough about Joe's background to produce a proper tribute, but I do know that he was extremely popular and an unbelievably prolific writer. I met Joe twice, but only briefly. He was a warm and friendly man and a great conversationalist. His passing is a great loss.

cuit, the 'north' direction represents X_L , the 'south' direction represents X_C , and the 'east' direction represents R .

Using the parallelogram method, first construct a resultant for the R and X_C , which is shown as vector 'A'. Next, construct the same kind of vector 'B' for R and X_C . The resultant, 'C' is made using the parallelogram method on 'A' and 'B'. Vector 'C' represents the impedance of the circuit; the magnitude is represented by the length, and the phase angle by the angle between 'C' and R .

Figure 4a) shows a series-resonant LC tank circuit, and Fig 4b) the current and impedance as a function of frequency. The series-resonant circuit has a low impedance at its resonant frequency, and a high

impedance at all other frequencies. As a result, the line current I from the source is maximum at the resonant frequency and the voltage across the source is minimum.

Parallel-resonant circuits

The parallel-resonant tank circuit Fig. 5a) is the inverse of the series resonant circuit. Line current (I) from the source splits and flows in inductor and capacitor separately.

The parallel-resonant circuit has its highest impedance at the resonant frequency, and a low impedance at all other frequencies Fig. 5b). Thus, the line current from the source is minimum at the resonant frequency, Fig. 5c), and the voltage across the LC tank circuit is maximum. This fact is important in radio tuning circuits, as you will see in due course.

Tuned RF/IF transformers

Many of the resonant circuits used in RF circuits – and especially radio receivers – are actually transformers that couple signal from one stage to another. Figure 6 shows several popular forms of tuned, or coupled, RF/IF tank circuits. In Fig. 6a), one winding is tuned while the other is untuned.

In the configurations shown, the untuned winding is the secondary of the transformer. This type of circuit is often used in transistor and other solid-state circuits, or when the transformer has to drive either a crystal or mechanical bandpass filter circuit. In the reverse configuration, where L_1 is the output and L_2 the input, the same circuit is used for the antenna coupling network, or as the inter-stage transformer between RF amplifiers in TRF radios.

The circuit in Fig. 6b) is a parallel-resonant LC tank circuit that is equipped with a low impedance tap on the inductor. This type of circuit is often used to drive a crystal detector or other low impedance load.

Another circuit for driving a low impedance load is shown in Fig. 6c). This circuit splits the capacitance that resonates the coil into two series capacitors. As a result, we have a capacitive voltage divider. The circuit in Fig. 6d) uses a tapped inductor for matching low impedance sources – antenna circuits for example – and a tapped capacitive voltage divider for low impedance loads.

Finally, the circuit in Fig. 6e) uses a tapped primary and tapped secondary winding in order to match two low impedance loads while retaining the sharp bandpass characteristics of the tank circuit.

Construction of RF/IF transformers

The tuned RF/IF transformers built for radio receivers are typically wound on a common cylindrical form. They are surrounded by a metal shield can that prevents interaction of the fields of coils that are in close proximity to each other.

Figure 7a) shows the schematic for a typical RF/IF transformer, while the sectioned view Fig. 7b) shows one form of construc-

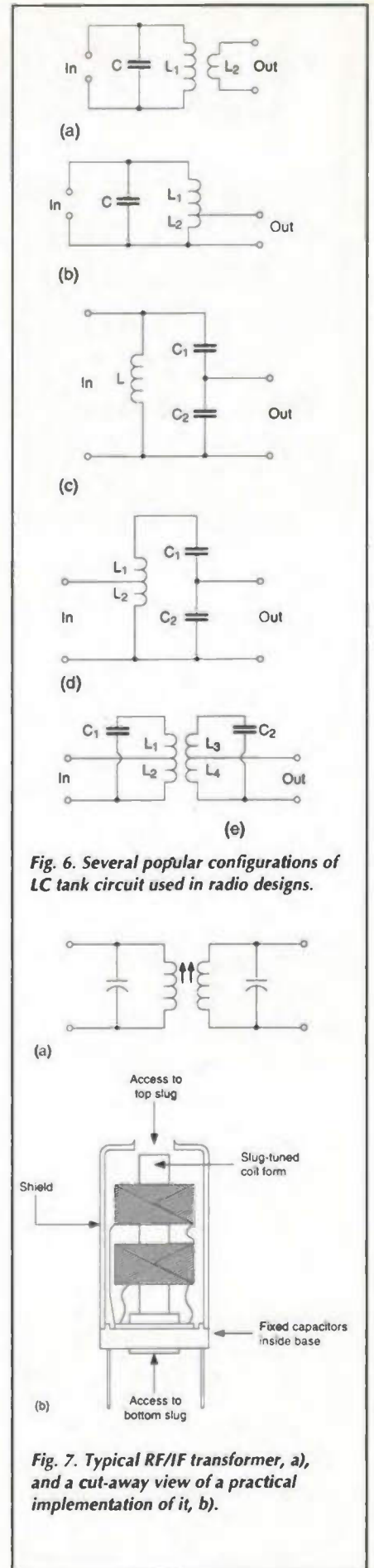


Fig. 6. Several popular configurations of LC tank circuit used in radio designs.

Fig. 7. Typical RF/IF transformer, a), and a cut-away view of a practical implementation of it, b).

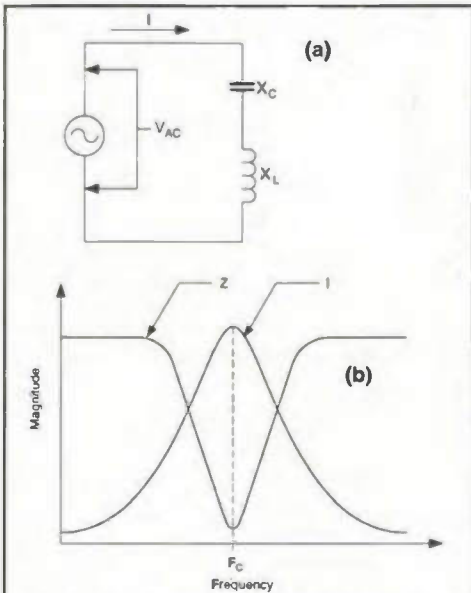


Fig. 4. In a) is a series-resonant-tank circuit and in b), its current and impedance plotted as a function of frequency.

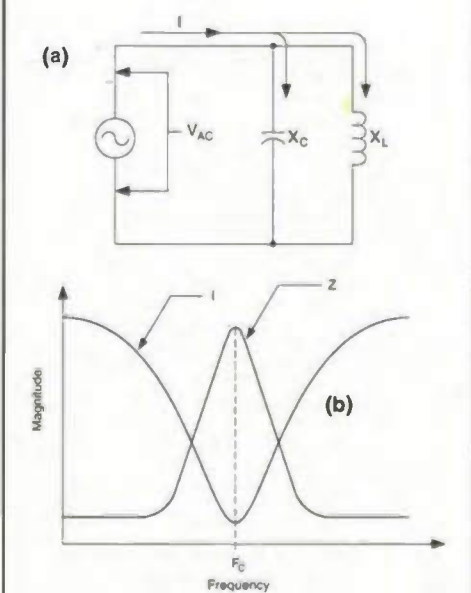


Fig. 5. Parallel-resonant tank circuit, a), and its current and frequency characteristics, b).

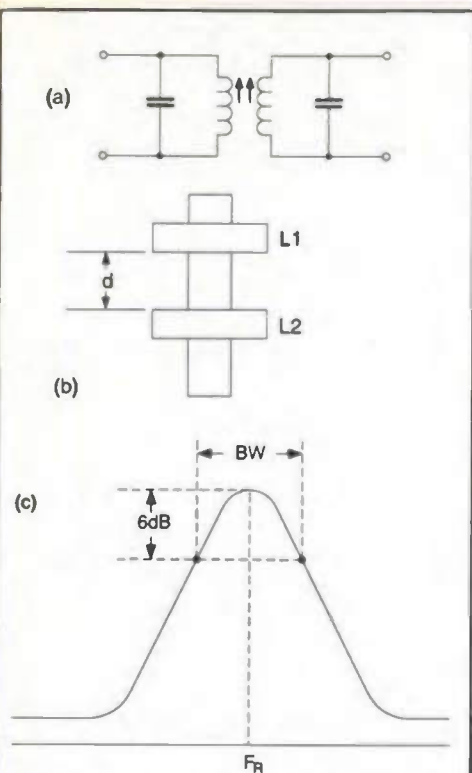


Fig. 8. Parallel-resonant RF/IF transformer, a), the two coils of which are normally wound as in b). Bandwidth of such a transformer is the range of frequency between the -6dB points on the bandpass curve, c). The shape of this response curve represents critical coupling.

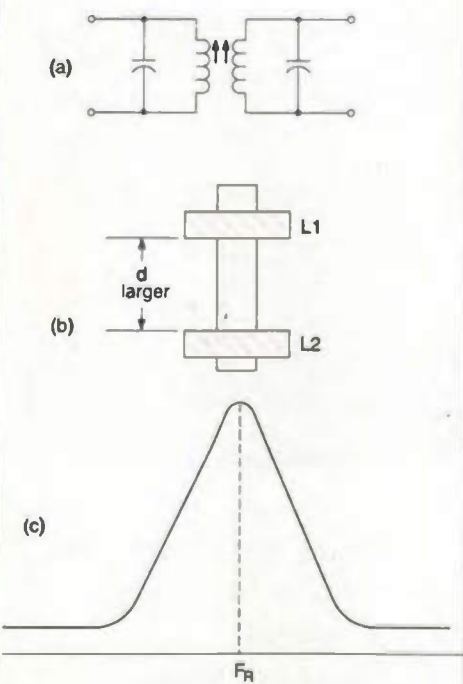


Fig. 9. In the undercoupled RF/IF transformer, a), the gap between the windings is greater than it is with critical coupling, b). This makes the bandwidth narrower, c).

tion. This method of building the transformers was common at the beginning of World War II, and continued into the early transistor era.

The methods of construction shown in Figs. 7c) and 7d) were popular prior to World War II. The capacitors in Fig. 7b) were built into the base of the transformer, while the tuning slugs were accessed from holes in the top and bottom of the assembly. In general, expect to find the secondary at the bottom hole, and the primary at the top hole.

The term 'universal wound' refers to a cross-winding system that minimises the interwinding capacitance of the inductor, and therefore raises the self-resonant frequency of the inductor – a good thing.

Bandwidth of RF/IF transformers

Figure 8a) shows a parallel resonant RF/IF transformer, while Fig. 8b) shows the usual construction in which the two coils L₁ and L₂ are wound at distance d apart on a common cylindrical form.

Bandwidth of the RF/IF transformer is the frequency difference between the frequencies where the signal voltage across the output winding falls off -6dB from the value at the resonant frequency f_r, as shown in Fig. 8c).

If F₁ and F₂ are -6dB – also called the -3dB point when signal power is measured instead of voltage – frequencies, then the bandwidth BW is F₂-F₁. The shape of the frequency response curve in Fig. 8c) is said to represent critical coupling.

An example of a subcritical or undercoupled RF/IF transformer is shown in Fig. 9. As shown in Figs. 9a) and 9b), the windings are farther apart than in the critically coupled case, so the bandwidth, Fig. 9c) is much narrower than in the critically coupled case. The subcritically coupled RF/IF transformer is often used in shortwave or communications receivers in order to allow the narrower bandwidth to discriminate against adjacent channel stations.

The overcritically-coupled RF/IF transformer is shown in Fig. 10. Note that in Figs 10a) and 10b) that the windings are closer together, so the bandwidth, Fig. 10c), is much broader.

In some radio schematics and service manuals – not to mention early textbooks – this form of coupling was sometimes called 'high fidelity' coupling because it allowed more of the sidebands of the signal – which carry the audio modulation – to pass with less distortion of frequency response.

The bandwidth of the resonant-tank circuit, or the RF/IF transformer, can be summarised in a figure of merit called Q. The Q of the circuit is the ratio of the bandwidth to the resonant frequency

$$Q = \frac{BW}{f_r} \tag{2}$$

An overcritically-coupled circuit has a low Q, while a narrow bandwidth subcritically

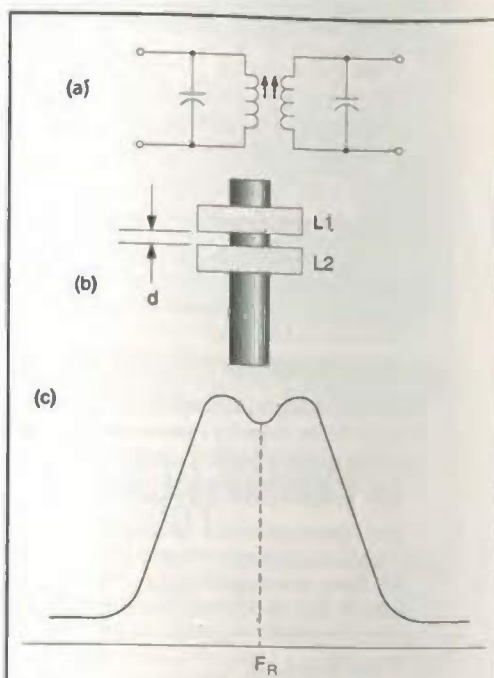


Fig. 10. In the overcritically coupled RF/IF transformer, a), the gap between the windings is relatively small, b), and the bandwidth much broader, c).

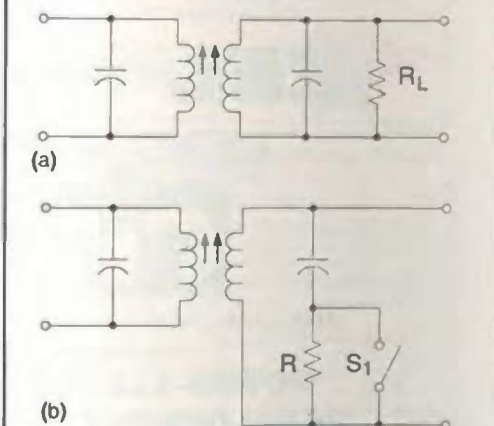


Fig. 11. Introducing resistance into the tank circuit causes its Q to go down, a). A switched resistor is sometimes used in radios to allow the bandwidth to be tweaked. Such a switch might be labelled 'fidelity' or 'tone'.

coupled circuit has a high Q.

A resistance in the LC tank circuit will cause it to broaden, that is to lower its Q. The resistor is sometimes called a 'de-Qing resistor'. The 'loaded Q' – that is, Q when a resistance is present – as in Fig. 11a) is always less than the unloaded Q. In some radios, a switched resistor, Fig. 11b) is used to allow the user to broaden or narrow the bandwidth. This switch might be labelled 'fidelity' or 'tone' or something similar.

In a second article on this topic, I will be looking at how LC tank circuits are implemented and the practical problems associated with them in radio designs.

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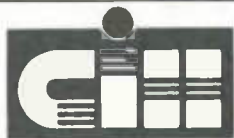
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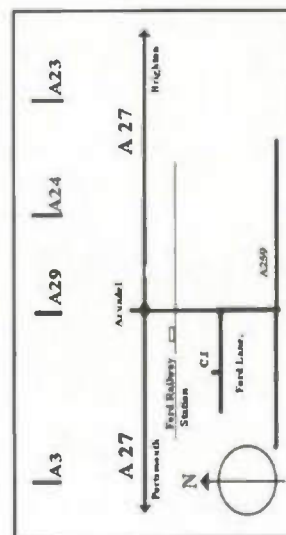
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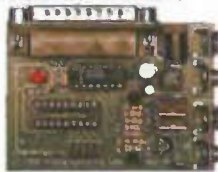
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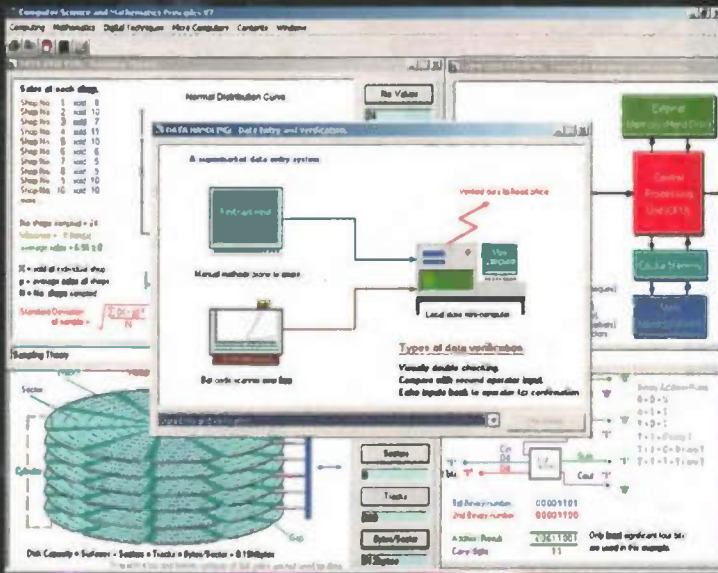
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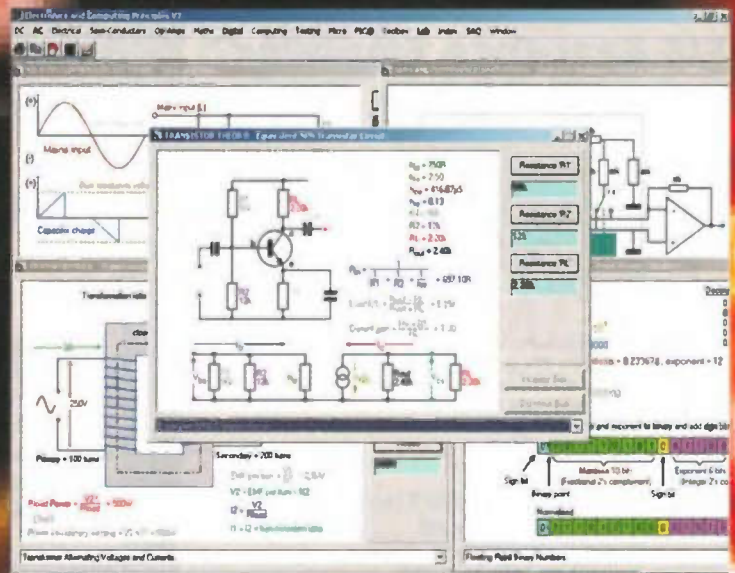
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