Special offer – 600W audio amp module!

ELECTRONICS WORLD

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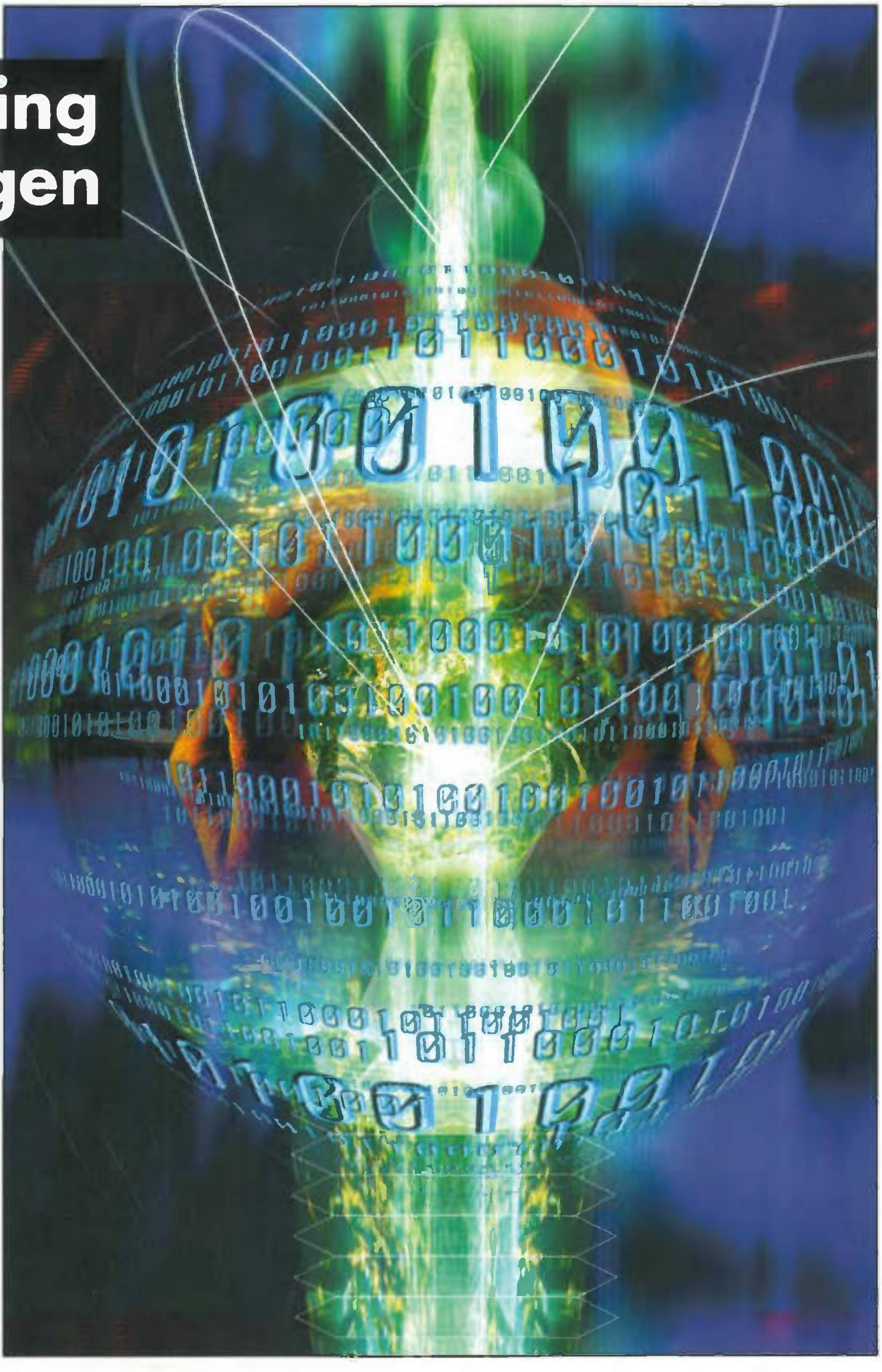


Phase-shifting digital sig gen

EMI tips and fixes

Pseudo random bits

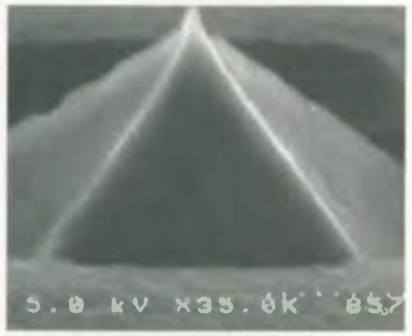
Colliding capacitors

Writing code for DSP

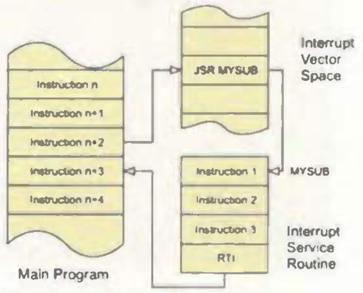
Circuit ideas:

Mains glitch detector
Phase shifter
Step motor PC interface
Sweep enhancer
Microcurrent amplifier
3-phase indicator
Serial keypad interface

CONTENTS

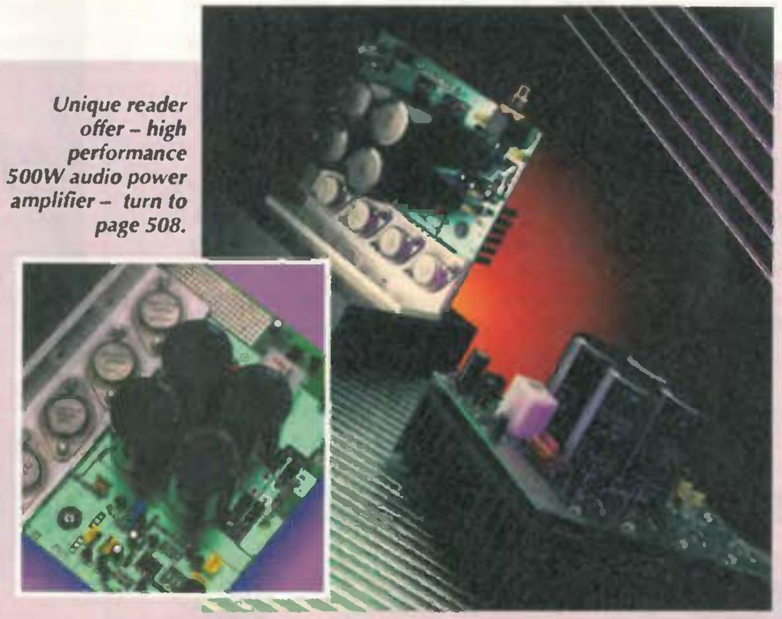


Solar cells using diamond films promise 50% efficiency and at a cost of just a dollar per square centimetre – this and much more news starting on page 492.



"Programming a DSP chip is no more difficult than programming a microcontroller," argues Patrick Gaydecki. Find out for yourself on page 522.

Unique reader offer – high performance 500W audio power amplifier – turn to page 508.



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A lesson in new technology

Many of you will be aware that the numbers of young people leaving school and considering a career in electronics and similar disciplines has been falling for years. Depending on who you listen to, this fall is either worrying or disastrous.

As someone who works with both schools and industry I am aware that many of my industrial colleagues are extremely concerned about future recruitment.

Here in Staffordshire, electronics companies are experiencing great difficulty in expanding due to the lack of quality electronics and other manufacturing engineers. As a result, electronics specialists are being appointed from all corners of the globe – but this source is not infinite.

Following the government hype regarding 'education, education, education', you could be led to believe that a new generation of high calibre students is in the pipeline about to burst out of our schools to revive the nation's manufacturing base. Don't hold your breath.

Unfortunately, governments have difficulty distinguishing between 'technology in schools' and 'design and technology' as defined in the National Curriculum.

Several months ago, the government announced that a further £400m would be provided for 'Technology' in schools. Some took this as being the key to turning a new generation towards electronic design and manufacture.

In those schools where targeted investment has taken place, the experience known to many older readers as 'Craft' or CDT is now a cutting-edge subject, well matched to the 21st century. In the majority of schools 'hands-on' electronic design and manufacture teaching – if undertaken at all – takes place in design and technology lessons.

Unfortunately, I believe that little of this new funding will find its way into design and technology. If it doesn't, it will have little effect on the nurturing of the future engineers that industry and the country needs.

Although there has been considerable expenditure in general career development for teachers, we have only scratched the surface in updating the skills of specialist teachers in the new technologies – especially electronics. Unlike the wood and metal ages, these new technologies change rapidly, requiring regular career development for staff. Much of this practical career development was provided by local Education Authorities but sadly this has all but disappeared.

I doubt if the new funding will have any effect unless we see directives to equip the country's design, technology and science facilities with adequate levels of dedicated 'information and communications technology' hardware, ICT for short, suitable applications software and manufacturing facilities appropriate to 2001 rather than 1951.

Unfortunately, in many schools this new technology hardware will not be used for designing electronics systems, controlling electrical components via control interfaces, data logging, scientific analysis or allowing young people to design using some of the excellent CADCAM software now available.

Only a tiny percentage of schools will allow some of the funds to be spent on the purchase of dedicated design, technology and science computers, CADCAM equipment, PIC systems or ICT controlled manufacturing equipment. Instead the vast majority of this new funding will probably go on supporting ICT in English, business studies, leisure and tourism, geography – and every other subject area.

This of course is no *bad* thing. Our future engineers need a high quality general education in all subjects. But these areas of the curriculum have little to do with electronic design & manufacture.

The solution – or at least a move in the right direction – would be to set a standard benchmark of minimum ICT requirements for all schools, in design, technology and science departments. If £400m were ring fenced and only given to the country's high schools, design, technology and science departments, each would receive approximately £50 000 each.

You will appreciate that to purchase some small CADCAM manufacturing equipment, software and 20 computers will quickly swallow up this sum. This assumes that the environment is suitable for these modern technology activities – which is unlikely in many schools.

We also have the career development aspect, to teach electronics and integrate all the ICT requires considerable knowledge. This requires intensive long-term investment in training. All of this could cost in excess of £200 000 per school, minimum!

If science and technology is to be the base of this country's future economy, as stated by Lord Sainsbury and other government ministers over the past six months, we need some major dedicated funding for these subjects in the nations schools, primary and secondary. This is especially important in the secondary sector where the specialist ICT equipment, software and manufacturing equipment for design, technology and science is vital. ■

John Hindhaugh, Director Staffordshire SATRO

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UPDATE

Airship detects battlefield mines

Mineseeker, the British not-for-profit mine detecting airship, has proved that it can detect mines in a battlefield.

Tests were performed in Kosovo recently with special permission at the end of a six-week UN minefield photo-reconnaissance mission.

Post-flight analysis in the UK has shown that Mineseeker is very likely to be able to accurately delimit minefields.

Currently, whole fields and whole parts of settlements are fenced-off on suspicion after a mine has exploded. The entire area has to be laboriously probed before the land can be re-occupied. Delimiting allows the actual minefield to be fenced off which allows relatively normal life to resume around it and greatly reduces the land area that must be probed.

Made by The Lightship Group, or TLG for short, the airship carries an experimental radar developed by DERA, the Defence Evaluation & Research Agency in Malvern.

DERA's ultra-wideband synthetic aperture radar uses impulse radar techniques developed over a decade to provide what is believed to be the world's widest bandwidth, highest resolution radar of its type.

An exceptionally high resolution is achievable at a much lower mean frequency than conventional narrow band radar, allowing it to penetrate foliage and the ground.

The resolution, coupled with high digitisation rates allows detection of even small, plastic anti-personnel landmines.

Using an airship allows the radar to be swept predictably over the ground at a rate at which it can operate, without the vibration and cost associated with helicopters.

Although the radar has previously shown its ability to differentiate between types of mine in laboratory conditions and has operated from a boom on the back of a lorry over surface mines, it had not before been tried out over buried mines in a battlefield.

The test site was once a defensive position in the Kosovo war and includes surface laid, foliage-obscured and buried mines as well as unexploded ordnance – all rendered unlikely to explode – and is cratered and littered with post-conflict debris.

The site is normally used to evaluate munition-finding dogs.

Together TGL and DERA have formed the Lightship Foundation. www.mineseeker.org

Nanometre-scale tubes could be used to produce FETs

Nanotube transistors just a few nanometres across have been fabricated by a research team at IBM. The technique could lead to the bulk fabrication of nanotube FETs for the first time.

Details of the research came in a paper published in *Science*. IBM scientists Philip Collins, Michael Arnold and Phaedon Avouris showed how the oxidation of carbon nanotubes by injecting them with relatively high currents can change them from metallic to semiconductor in nature.

The properties of multi-walled nanotubes, or MWNTs, can switch from metallic to semiconductor depending on the numbers of atoms in the walls of the tubes – the number of shells it has.

Collins' group passed currents of up to 200μA though selected tubes causing the outer layer to oxidise. This is a controlled process, Collins said, because each layer sheds at a discrete current, after a few milliseconds.

As each shell is removed the characteristic of the whole tubes changes from metallic to semiconductor or *vice-versa*.

The technique also applies to bundles of single walled nanotubes, although the process is less straightforward. As the current is passed through the bundle of tubes, entire tubes are destroyed through oxidation, again changing the properties of the bundle.

IBM has shown that this technique, whether applied to multi or single walled tubes, can be done on a selective basis. Arrays of tubes can be left as a sea of metallic elements with a few semiconducting tubes making the FETs.

Class-D amplifier chip boasts 300W and 0.03% distortion

Digital reproduction is creeping up the audio chain and has reached the power stage of many low-cost amplifiers – particularly in those, like answer-phones and TVs, where quality is not top of the list.

Now Texas Instruments claims to have bought the well-known Class-D attribute of high-power efficiency to a real high-power hi-fi design.

It is to sell an all-digital amplifier subsystem chip that should deliver over 300W with THD+N below 0.03% (20Hz to 20kHz) and a 110dB dynamic range measured at the speaker – when the complete reference design becomes available later this year – claims the company.

It also says that a variable power rail design, which it is keeping up its sleeve, will allow a dynamic range of 130-140dB.

The TAS5015 digital modulator is the heart of the amplifier. It is a 24-bit multi-standard, multi-sample rate (32 to 192kHz) device that processes a digital pulse-code modulated input into a pulse-width modulated output ready for a power output stage.

TI has designed a suitable discrete power stage, which will only be available under licence, that converts the PWM signal up to the power level required at the speaker.

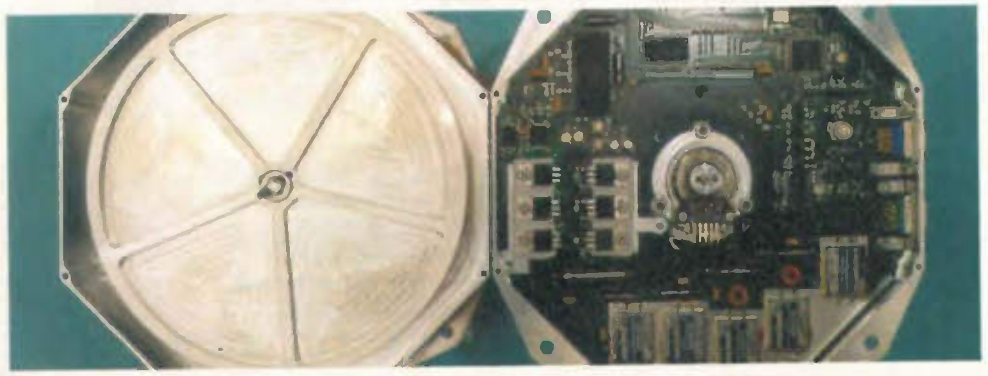
The company claims that the reference design will dissipate less than one-tenth the power of class-A and class-AB amplifiers.

The technology, called equibit, comes through TI's acquisition of Toccata Technology last year.

Samples of the chip have been distributed to licensees already. Price, to those who can afford a licence, will be around $25 in quantity.

Satellite's momentum wheel develops 5.2Nms from just 6W

Momentum wheel's end casing showing its interface circuitry.



Surrey Satellite Technology (SSTL) has made the momentum wheel which will fly on the Rosetta spacecraft, due to land on comet '46 P/Wirtanen' on 29th November 2011.

The wheel will be mounted inside the satellite so that the spin direction is inertially fixed, giving the satellite an inherent gyroscopic rigidity.

By changing the momentum wheel's speed over a limited range – less than 10% of the nominal speed of a few thousand revolutions a minute – the satellite can be rotated around the spin direction. In this way, rotational momentum can be exchanged between the wheel disc and the satellite's body.

This momentum wheel is unique, says SSTL, in that it provides 5.2Nms of momentum with a power dissipation of only 6W.

In flight, it will have to survive a nine year cruise in space before it is fired-up and used to manoeuvre the craft during the final 20 hours of approach and landing on the comet.

Dry lubricated bearings and low-friction operation are the keys to long life in a vacuum says SSTL. Details of the Rosetta Mission are available at,
http://spdext.estec.esa.nl/content/doc/e7/2279_.htm
and www.sstl.co.uk

New unique material focuses microwaves

A material with a negative refractive index at certain microwave frequencies has been created by physicists from the University of California in San Diego.

Future generations of the material could, the developers say, lead to highly directional aerials and perfect lenses that focus microwaves or light to a point.

"The experiments we report confirm earlier theoretical predictions that a new, unique, class of materials can cause electromagnetic waves, such as radar and microwaves, to bend in a direction opposite to the way they travel through all other known materials," said Sheldon Schultz, professor of physics at UCSD.

The work by Schultz' team proves an idea put forward by Professor John Pendry, head of physics at Imperial College, London.

He suggested that some composite materials with negative permeability and negative permittivity would have bizarre properties such as to make convex lenses behave as concave while flat sheets would focus light to a spot.

So-called metamaterials with negative permeability and negative permittivity lead to a negative refractive index, which means the material breaks Snell's Law.

Any EM radiation incident at a boundary will be refracted in an unexpected direction.

A microwave beam incident on Schultz' material refracts to the same side of the normal as the incident ray.

"If these effects turn out to be possible at optical frequencies, this material would have the crazy property that a small flashlight shining on a flat slab would produce a focus at a point on the other side," said Schultz.

LEDS for domestic lighting?

The domestic light bulb may be a step closer to the history books when Philips Research in the US has finished its latest development.

It is setting out to discover how leds can realistically be used for home lighting. Leds – even the latest which rival light bulbs in power efficiency – have to be used in bunches to get sufficient light for illumination.

Unfortunately chip-to-chip variation means that the output can be blotchy and vary from assembly to assembly.

The company is developing ways to optically mix the light coming out of the leds, as well as control algorithms to allow the effective colour of the assemblies to be varied to suit mood and tastes within the home.

One aim is to generate control algorithms that carefully balance the many variations which occur in leds, said the company.

Algorithms so far developed also include compensation for device to device light output variations, output drop with age, change with temperature and wavelength variation with current.

At the moment, some form of initial calibration is required. Eventually an in-room sensor may feed back enough information to allow 'plug-and-play' swapping of led-based lamps.

Backlighting of LCDs may be the first application for the new algorithms, which have been written for Philips' microcontrollers.

MPEG4 edges closer

Audio and video compression system MPEG4 has taken another step closer to full industry acceptance as three key organisations agree on a standard.

The International Telecommunications Union (ITU), 3G Partnership Project (3GPP) and Internet Engineering Task Force have pledged their support for a format proposed by five Japanese companies.

Their support for any MPEG4 format is said to be critical for delivery of MPEG4 over the Internet.

The five developers of the format are Matsushita Electric, NEC, NTT, Oki Electric and Toshiba.

Rather than using TCP/IP, the conventional Internet protocol, they have based their standard upon the real-time transport protocol (RTP).

Audio and video are a problem for TCP/IP because of their inherent real-time character. When data is lost, TCP/IP retransmits, which can cause annoying delays to the service or even out of order sequencing.

On the other hand RTP does not retransmit and adds time stamps to the packet headers in order to keep audio and video synchronised. Sequence numbers make sure that packet losses don't cause major timing problems.

Designed for lossy media such as mobile phones, MPEG4 is expected to become the main compression standard for telecoms, local-area networks and the Internet.

Video over wired or wireless links can have data rates of between 5 and 64kbit/s, while standard TV is between 2 and 8Mbit/s.

The data, whether audio or video, is split into a series of objects, for example a background, table and a human body. If only the human face moves from one scene to another, only the relevant data is transmitted, reducing bandwidth requirements.

High-performance Bluetooth

A Bluetooth radio chip has been developed by STMicroelectronics using intellectual property from Parthus Technologies, the Irish design house.

Aimed at battery operated products such as mobile phones and PDAs, the radio uses a low intermediate frequency design.

ST will use its BiCMOS silicon germanium process to manufacture the chip. Current consumption is claimed to be 30mA in receive mode and 30mA when transmitting at maximum power.

"Exploiting ST's silicon germanium technology has enabled us to create the highest performing Bluetooth radio on the market," said Kevin Fielding, president of Parthus.

On its own the chip can transmit at Bluetooth's Class 2 and 3 levels, good for up to 10m. With the addition of a extra power amplifier, it can manage Class 1 operation, which takes its range up to 100m.

In order to bring Bluetooth products to market quickly, the specification has very relaxed requirements. Bit error rate, for example, is only required to be 10^{-3} at –70dBm. The ST/Parthus chip has a claimed BER of 10^{-6} at –75dBm.

Optical fibre survives oil-well drill tip

New Mexican research centre Sandia Labs has shown that optical fibre is tough enough to survive in an operating oil-well drill without additional armour.

Previously it was thought that the fibre would be too fragile to last long enough to do any useful work within the hollow drill, which also carries abrasive liquid 'lubricant'.

The findings come out of a research project to increase the data rate of live in-well sensing during drilling operations.

Fibre, protected only by a thin protective plastic coating, survived the few hours needed to make measurements in a 1000m test drill, transmitting information about temperatures, pressure, chemistry, and rock formation at 1Mbit/s. Researchers predict the scheme should be good for wells 6000m deep.

After use the fibre is ground up by the drill bit.

Why it survives is not yet proven, but it is suspected that it 'gives' when brushed by particles in the lubricant and also tends to cling to the inside wall of the drill.

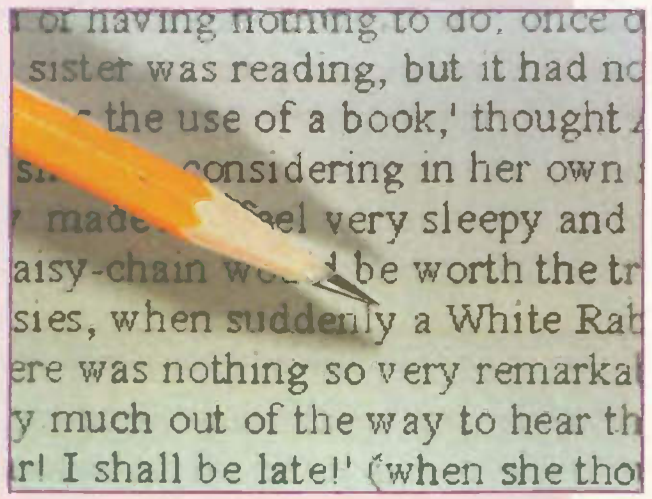
LCD read-out with no power

E Ink has demonstrated a 12-inch active matrix 'electronic ink' display that retains its image on power-down.

The company started off with an active matrix donated by IBM and added the ink, which works through the electrostatic attraction of coloured particles.

The black ink on white background avoids the need for backlighting, claims the company, while brightness is up to six times better than LCD, it said.

E Ink and IBM will deliver a joint paper describing the display at the Society for Information Display conference in San Jose in June.



Solar cell promises 50% efficiency

Scientists at Vanderbilt University in Tennessee are promising a solar cell that converts sunlight at 50 per cent efficiency. Moreover the design will cost just $1 per square centimetre, the researchers claim.

Professor Timothy Fisher says his polycrystalline diamond films would be well suited to space applications as they can withstand the high level of cosmic radiation.

Efficiency of conventional solar cells can be reduced by a half after ten years in space.

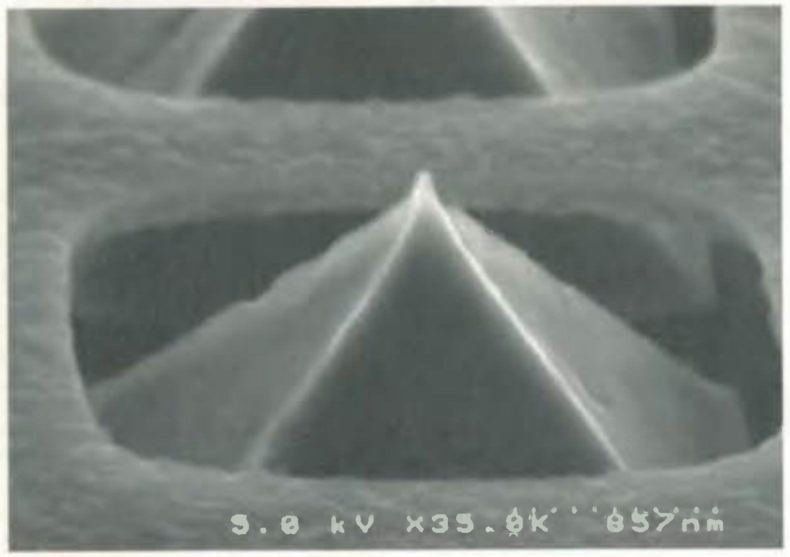
Fisher's diamond-film structure includes pyramids, about ten million per square centimetre, separated from the cell's anode by a small gap. The cathode is attached to the back of the film.

Sunlight hits the cathode and heats it to around 1000°C, causing streams of electrons to flow from the pyramid tips to the anode.

The resulting large current and small voltage are passed though a DC-to-DC converter at around 90 per cent efficiency, says Fisher.

Costs are kept down as the diamond film can be deposited from methane using chemical vapour deposition.

The goal of current research at the University is to produce a prototype 1cm square cell turning out 10W of power at 1000°C.



Steerable solid-state antenna for mobile phones

Cambridge start-up Antenova has garnered £3.4m in venture capital funding to develop a steerable solid state antenna.

Launched late last year, Antenova is developing directional and steerable antennas aimed at mobile phones.

By directing the RF, mobile phone systems can physically split cells and radiation to the user can be reduced.

The steerable side of the design ensures a phone stays in contact with the base station as the user changes position.

By avoiding 360° transmission, power consumption is also reduced, the firm says. And because the antenna is directional, signal processing needs are reduced.

The antenna is formed from three strip-lines on the mobile phone's PCB. This gives it directional control with a 60° resolution.

First working prototypes are due by the end of this year.

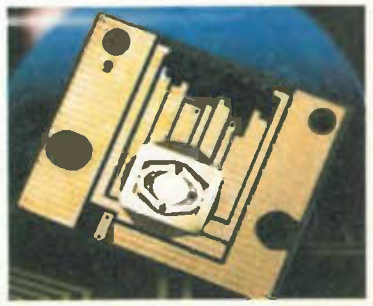
A spin-off from radio frequency location company ActiveRF, Antenova holds licences for technology developed at Sheffield University and Griffith University in Australia. Funding has come from the Cambridge Gateway fund.

Steerable mirrors at heart of Texas Instruments' optical comms systems

Texas Instruments has announced an optical communication system based on steerable mirrors.

It is a point-to-point system which will send 100Mbit/s Ethernet hundreds of metres or up to 50 metres with a class 1 laser.

Optical terminals made from TI parts could be tennis-ball size and will be handled much like a Webcam. Users will plug into a computer and point them at a similar terminal to establish a link. No precision alignment is needed.



Central to operation is a two-axis analogue steerable mirror.

Far bigger than its micromirrors, but small at only 3.2 by 3.6mm, it is controlled by a DSP to establish alignment and compensate for vibration.

TI is making an evaluation kit, including a DSP, analogue chip, packaged mirror and software, in prototype quantities, production is probably a year away. ■

WAP addresses reduced to a number... Typing in WAP addresses using only the 12 keys of a phone keypad is an irritation that may go away with a development from AirClic. In a trial in Sweden the company has replaced WAP addresses with a key-pad-friendly number. The user either types in the number, or scans a special mini bar code into the phone. Apparently, it only takes minutes to register any phone, PC or PDA with the company to use the system.

Elastic capacitors

Electronic engineers understand capacitors but they tend to view colliding bodies as mysterious – a strange mindset since they are both expressions of the same formulae, as Andrew Robertson explains.

What do colliding bodies and capacitors have in common? Not a lot one might think, apart of course from both being described by the ubiquitous $a=bc$ and its integral $a={}^1\!/_2bc^2$, i.e.,

momentum$=mv$
kinetic energy$={}^1\!/_2mv^2$

and,

charge$=CV$
charge energy$={}^1\!/_2CV^2$

Beyond this, few textbooks make more than passing comment of correspondence, but one must bear in mind that formulae only describe – they don't define.

Non-elastic interactions

If a body travelling with a velocity of v collides *non-elastically* full face with a stationary body of identical mass, the two bodies unite and move in the same direction as the original moving body but with a velocity of ${}^1\!/_2v$. Momentum, being inviolate, prevails.

The energy embarrassment to satisfy ${}^1\!/_2mv^2$ is liberated ultimately as heat. Likewise, but with a change of terms, this dissipation of energy occurs in the classical example of Fig. 1.

With identical capacitors and initially only C_1 charged to V, closing the switch results in a voltage of ${}^1\!/_2V$ in each capacitor. Charge conservation, also inviolate, substitutes for momentum, capacitance for mass and voltage for velocity.

Regardless of the impedances of the physical components, be they large or diminishingly small, energy is conveniently disposed of in heating, electromagnetic radiation and even sound. In purely theoretical conditions – which would evoke Maxwell's equations – the 'spare' energy will be still be liberated from the system. We have therefore in a sense 'inelastic' capacitors.

Momentum and charge of the whole universe is by all accounts absolute, and probably zero. In all known interactions, momentum and charge will always be preserved at the expense of system energy. So, regardless of relative masses or capacitances where the second body is initially stationary (or uncharged), it follows that, for inelastic collisions, combined velocity is,

$$\frac{mass_1 \times velocity_1}{mass_1 + mass_2}$$

or, for (inelastic) capacitors, combined voltage is,

$$\frac{C_1 \times V}{C_1 + C_2}$$

Elastic interactions

Consider now the *elastic* collision of two identical masses, initially one moving and the other stationary. For simplicity, this is restricted to a full face collision.

The first body will transfer all its momentum *and* energy to the second body. This absolute condition is only realised at atomic level but can be approached in the 'real world'.

So what of a capacitor equivalence? Conventional wisdom conjures up the 'inelastic' situation, but there is a way. In Fig. 2, where only C_1 is initially charged and $C_2=C_1$, with idealised components the entire charge *and* energy are transferred when the switch is closed. As with collisions, losses cannot be avoided but can be minimised with a high-Q inductor and synchronous switch replacing the diode and switch.

At first glance, the inductor's action may appear strangely obliging but the arrangement is simply a tank circuit, with the condition at the instant of switch closure being the voltage crest of an oscillatory cycle. After 180°, the voltage across the inductor will be reversed, current will have fallen to zero and further oscillation and charge exchange is blocked by the diode.

The system energy will be again solely capacitive, having been transferred by way of the inductor intermediary from C_1 to C_2. Of course for the oscillatory cycle, the two capacitors are in effect only one in series combination, charged to V_3. This is true for both the initial and final conditions since our voltage reference is their junction. In this simplest case it is zero, as initially is the voltage of the second capacitor. Ergo that capacitor equates to an initially stationary body in our voltage reference frame.

Notably, the voltage V_3 reversing between the top ends of the capacitors is analogous to a primary law of elastic collision, namely, ***the relative velocity of two particles in a full face elastic collision is unchanged in magnitude but reversed in direction***. We have now 'elastic' capacitors.

Dissimilar capacitances

The special case of equal masses is familiar and almost intuitive, but what of the resultant velocities of dissimilar masses in an elastic collision? Although second nature to the atomic physicist, this may not be so for the electronics engineer. The formulae are not difficult to derive by manipulation of momentum and energy equations, but can also be derived from basic electronic principles. Consider Fig. 2 with dissimilar (or identical) capacitors. Series capacitance is,

$$\frac{C_1 \times C_2}{C_1 + C_2}$$

In the elastic interaction – i.e. 180° of an undamped oscillation – the voltage across the capacitors is reversed, which is to say that voltage *change* across the capacitors is twice, V_{3i}. Note that the subscripts i and f denote the initial and final states before and after switch closure. This means that charge flow is,

$$C_{series} \times 2 \times V_{3i}$$

Charge in C_2 was initially zero so charge flow is equal to charge transferred from C_1 to C_2. But since V_{2i} is zero, $V_{3i}=V_{1i}$, and since,

$$V=\frac{Q}{C}$$

resultant voltage V_{2f} across C_2 is,

$$\frac{Q_{transferred}}{C_2}=\frac{C_1 \times C_2}{C_1 + C_2} \times \frac{2 \times V_{1i}}{C_2}$$

$$=\frac{C_1 \times 2 \times V_{1i}}{C_1 + C_2} \qquad (1)$$

resultant voltage V_{1f} across C_1 is,

$$\frac{Q_{initial} - Q_{transferred}}{C_1}$$

$$=\frac{(C_1 \times V_{1i}) - \left(\frac{C_1 \times C_2}{C_1 + C_2} \times 2 \times V_{1i}\right)}{C_1}$$

$$=V_{1i}\left(1 - \frac{2 \times C_2}{C_1 + C_2}\right)$$

$$=V_{1i}\frac{C_1 - C_2}{C_1 + C_2} \qquad (2)$$

It should come as no surprise that with a change of terms, equations 1 and 2 are the same as the equations for elastic mass interactions.

The quantitative world

Simple though the formulae are, it is quite pleasing to have arrived at them from knowledge of a different discipline. It could be inferred from this method of calculation that two masses in an elastic collision act arithmetically like two capacitors in series. In other words, the product divided by the sum. Intuitively it is difficult to appreciate why this should be so, although an atomic physicist might feel otherwise.

A point of caution, apparent from equation 2 – beware of inductors feeding capacitance in power feeds or signal paths. Under favourable conditions, as exploited in applications such as thyristor commutation, the transient voltage across capacitance can reverse or aspire to reach twice the applied voltage.

With extreme values of C_1 and C_2, if C_1 is much less than C_2, C_2 will remain at almost zero volts, while the voltage on C_1 will be almost unaltered in magnitude but reversed in polarity. Where C_1 is much greater than C_2, C_2 will charge to nearly $2V_{1i}$, while the voltage on C_1 will remain almost unaltered.

The voltage reversing and doubling ties up nicely with the mass analogy. If a super bouncy ball is thrown at a wall it comes back at the same speed – almost. Change the reference and it is the same as throwing a wall at a bouncy ball, whereby the resulting velocity of the ball will be nearly twice that of the wall.

The derived formulae relating to inelastic and elastic collisions between one moving and one stationary object can be augmented to describe interactions between bodies or capacitors with any initial starting condition. Interaction between two moving bodies equates to both capacitors having an initial charge but the analysis still holds true. One can even incorporate coefficients of restitution and resistance as imperfect components in the mechanical and electrical systems.

There are limits though. In more complex states, such as where the collision is not full face or asymmetry of the bodies impart spin, the elegance of the equivalence becomes strained. Also, the assertion of charge conservation is appropriate but not altogether accurate – capacitor charge is perhaps better viewed as charge differential.

When a body – a capacitor would do – is thrown into the air, the world moves in the opposite direction. The charge of that capacitor – the complementary effect of momentum – is not quite so absolute.

If, for instance, two identical capacitors are charged in series, then in their charged state reconnected in parallel, the voltage halves but the capacitance quadruples. By manipulation of each capacitor's zero reference, there is a doubling of charge. In the world of momentum, contrary to what, say, advocates of gyroscopic levitation would appear to expound, there is no likely analogue.

So is there benefit from viewing capacitor charging as elastic and inelastic collisions, or *vice versa*? Certainly, if one's inclination is to subjectively 'feel' the physical world. The two environments are merely different expressions of underlying fundamentals, not formulae coincidence. It is akin to describing electricity in terms of water flow.

The appreciation is not Grand Unification Theory, but such comparison can aid insight and hence possibly improve circuit design.

Finally, thanks to Glenn O'Dell for his help in preparing this article. ■

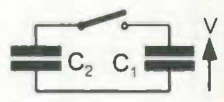


Fig. 1. Capacitor C_2 is uncharged, C_1 is charged at voltage V. Closing the switch results in both capacitors being at the same voltage, ${}^1\!/_2V$.

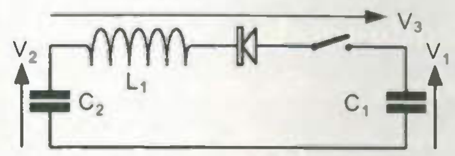


Fig. 2. In this circuit, with idealised components, the entire charge and energy are transferred from the charged capacitor when the switch is closed.

Pseudo-random bits

Pseudo-random bit sequences have uses in communications, testing, audio and many other areas. Here, Ian Hickman sets out to demystify the topic.

Pseudo-random bit sequences, or PRBSs, are readily generated, either in software, or in hardware using linear-feedback shift registers.

If the output of the last stage of an n-stage shift register is fed back to the input of the first stage, the contents of the register will cycle around indefinitely as the device is clocked. The length of the output sequence will be n bits per repetition. But if the output of the last stage and the content of an earlier stage or stages are combined to give the feedback signal, the length of the repeating output cycle can take various values up to 2^{n-1}, depending on how the feedback is arranged.

Modulo arithmetic

In modulo arithmetic, the result can never exceed some arbitrary maximum number. If the result exceeds that, the number, called the modulus, is repeatedly subtracted until the modulus is not exceeded.

Ordinary decimal arithmetic uses the ten digits 0 to 9 inclusive, and is capable of representing any number however large. But in modulo-10 arithmetic, the result can never exceed nine. So, for instance, nine plus seven would equal six, not sixteen, and 25+98=3.

Table 1

In GF(2), addition of two numbers X and Y follows these rules.

X	Y	Sum
0	0	0
0	1	1
1	0	1
1	1	0

A scheme of arithmetic operating with a modulus is called a Gallois field. Of particular interest is the modulus-bounded arithmetic of just the two digits nought and one, known in the trade as the field GF(2). In GF(2), addition of two numbers X and Y follows the rules shown in **Table 1**.

This is the same as the truth table of an exclusive-or gate, so such a gate performs modulo 2 addition.

Polynomials

As the name suggests, these are expressions having several or many terms, involving a variable, usually denoted by x. The general form of an nth degree polynomial is,

$$a_n x^n + a_{n-1}x^{n-1} + a_{n-2}x^{n-2} \mathrm{K} \ + a_1 x + 1$$

In the field GF2, both the variable x and the coefficients a are restricted to taking the value 0 or 1. To the electronics engineer, the interest in this apparently arcane branch of mathematics is that polynomials provide the key to designing PRBS generators using linear-feedback shift registers, or LFSRs. Usually, of particular interest are LFSRs capable of providing a maximal length sequence 2^{n-1} bits long.

Published tables are available, providing details of such polynomials. One of the earliest and best known references is Appendix C of reference 1. As in some other references, the polynomials there are listed in a 'shorthand' consisting of a listing of the coefficients a, in octal notation. Octal uses the digits 0 to 7 to indicate three binary digits or bits. The best way to see how to use the listings is by an example.

One of the polynomials listed under degree nine is 1131. Translating these four octal digits into binary gives 001 001 011 001 as the coefficients a of the polynomial. The two leading zeros are an artefact of the octal-to-decimal conversion, and can promptly be ignored. The final 1 is the 1 at the end of the polynomial and the other digits are the coefficients a_9 to a_1, reading from left to right. So this particular polynomial reads,

$$1.x^9 + 0.x^8 + 0.x^7 + 1.x^6 + 0.x^5 + 1.x^4 + 1.x^3 + 0.x^2 + 0.x^1 + 1$$

Deleting the redundant terms with a zero coefficient a leaves the polynomial,

$$x^9 + x^6 + x^4 + x^3 + 1$$

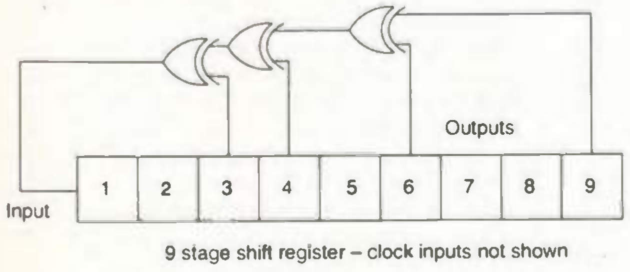


Fig. 1. Showing a nine-stage shift register plus exclusive-or gating, providing a 511 bit long pseudo random bit sequence.

Ignoring the constant 1 at the end of the expression, the powers of x give the stages of a 9-stage shift register from which modulo-2 feedback should be applied to give a sequence of noughts and ones, as detailed below.

An LFSR giving a maximal-length sequence

As '1131' is taken from a listing of 'irreducible' ninth-degree polynomials, the feedback arrangement mentioned will provide a maximal length sequence of $(2^9-1)=511$ bits, consisting of 256 ones and 255 noughts, arranged in a particular way.

If the input to the nine-stage shift register is obtained by sequentially EXORing the output of the ninth stage with the outputs of the 6th, 4th and 3rd stages, as in **Fig. 1**, the desired maximal length sequence will be obtained. With a nine-stage shift register, the possible number of arrangements of nine binary digits is 2^9 or 512, including all 0s, all 1s and all other possibilities.

From the first line of Table 1, if the two inputs to an EXOR gate are both noughts, then the output is also a nought. So in Fig. 1, if all the stages of the shift register come up as 0s at switch-on, the sequence will consist of an indefinite string of 0s. But if any other combination of 0s and 1s appears, even a single 1 in any stage, then the shift register will hold, at various times, all possible combinations of nine binary digits (except nine zeros), and the output will be the maximal length sequence of 511 bits, continually repeating. Thus nine noughts is the only non-allowed state; any other combination of nine digits, 0 or 1, must by definition be one of the $(2^9-1)=511$ valid states.

In some designs, it may be necessary to include arrangements to force a 1 in a stage of the register at switch-on, to prevent any possibility of the degenerate all 0s case occurring. Note that you could alternatively use exclusive-nor gates, but in that case, the non-allowed degenerate combination is all 1s.

At any one time, nine bits of the circulating sequence of 511 bits will appear in the shift register stages. The sequence may be represented by a sequential 'state diagram' of 'bubbles', Fig. 2. Each state or bubble represents a sequence of nine bits, of which the last eight are the first eight of the previous state. The first bit is the result of the exclusive-or operation on the previous state.

As the sequence repeats indefinitely, it is academic which state bubble 1 refers to. But you could define it as whatever arrangement of 0s and 1s, usually indeterminate, comes up initially in the nine shift register stages, just before the first shift when the first clock pulse arrives.

If the arrangement of exclusive-or gates does not correspond to an irreducible polynomial, the repeating sequence produced will be less than 2^{n-1} bits long. If, at switch-on, the shift register stages come up in one of the states of this shorter sequence, then the state diagram will be similar to Fig. 2, but with fewer than 2^{n-1} bubbles in the ring.

By definition, there must be other states that do not feature in the sequence. I believe that if, at switch-on, the shift register stages come up in one of these, the states will proceed via a 'side arm', as in **Fig. 3**, until the non-maximal length repeating sequence is entered at some point.

It is possible that a feedback arrangement exists that can generate two different non-maximal length sequences. Which one is generated would be determined by the initial state of the shift register stages at switch-on. I am still investigating this point. No doubt one of you will already have the answer. If so, please write in.

For LFSRs with more than two stages, there is more than one maximal length sequence, each having a particular series of 0s and 1s. For an LFSR with three stages, there are only two maximal-length sequences, of seven bits. One is the time-reverse of the other; the same bit sequence but in back-to-front order.

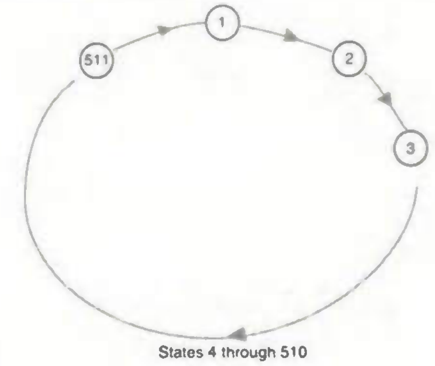


Fig. 2. The unique state diagram relating to Fig. 1.

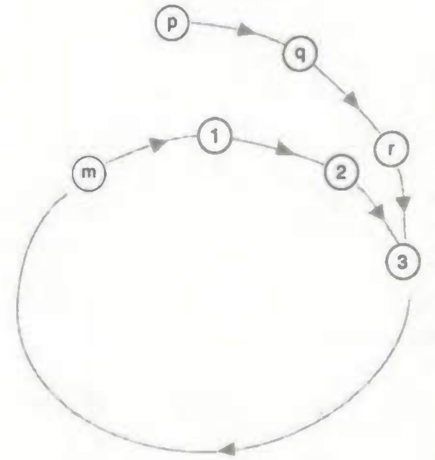


Fig. 3. A non-maximal length sequence may be entered directly at switch-on, or from a 'side chain' starting with a state that does not occur in the non-maximal length sequence itself.

Many different sequences

As the number of shift-register stages increases, the number of different maximal length sequences escalates dramatically. For example, Appendix C of reference 1 lists 28 primitive polynomials of degree 9. Some require more exclusive-or gates – i.e. some have more non-zero coefficients of powers of x – while some require fewer than the example "1131" given.

For each of these, there is another arrangement that provides the time-reverse of the sequence, making 56 maximal length sequences in all. For degree sixteen, there are six and a half densely-packed pages of them, while for higher degrees up to 34, there's just a small subset for each[1].

The ability to produce a pair of sequences, one the time reverse of the other, is important in some applications. An example is military communications requiring high security. This application might employ direct-sequence spread spectrum (DSS) communications. Here, the digital data stream of 0s and 1s to be transmitted is "chipped", before being modulated onto the final RF carrier.

Thus for example, a 100kbit/second data stream might be EXORed with a maximal length PRBS of 63 bits, produced by a six-stage LFSR clocked at 6.3MHz. In this way, the final signal out of the transmit mixer occupies 63 times the bandwidth which would have been occupied by the unchipped signal. So for a given transmitter power, the power spectral density is 63 times or 18dB lower.

The wider occupied bandwidth makes the signal more difficult to jam, while the lower level means that the signal might escape an enemy's attention entirely.

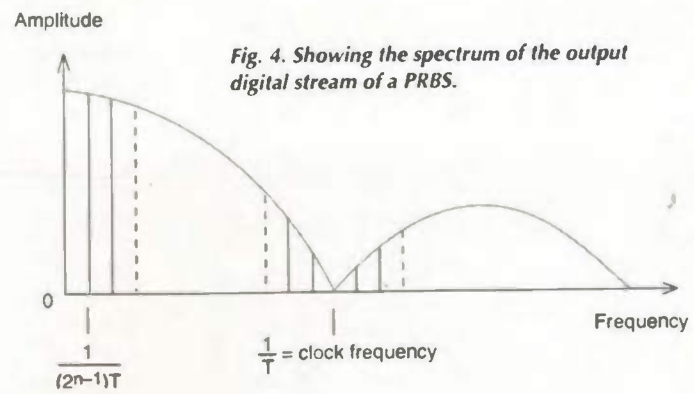


Fig. 4. Showing the spectrum of the output digital stream of a PRBS.

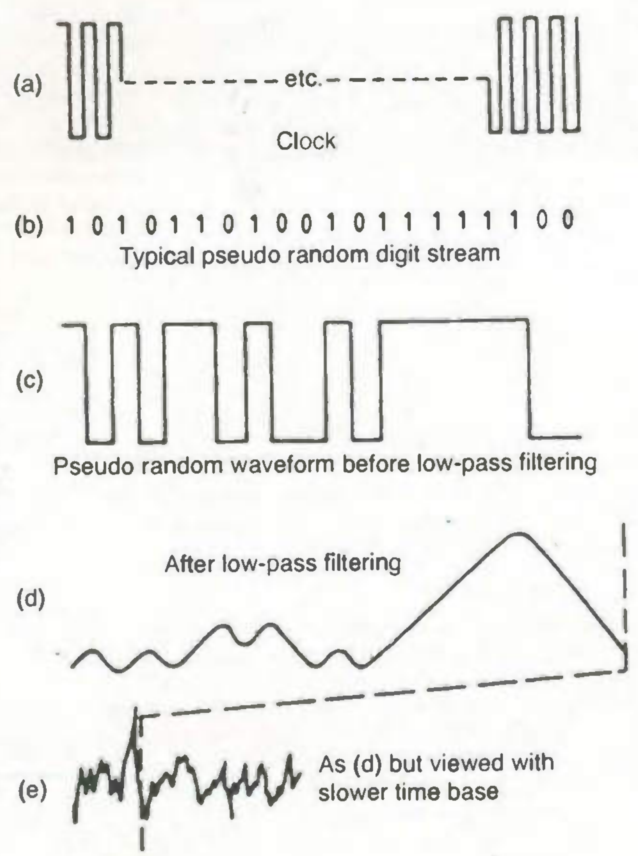


Fig. 5. Showing how low-pass filtering the sequence provides a random noise-like waveform. This illustration is reproduced from reference 3.

Time-reversed sequences

The arrangement to produce the time reverse of a given sequence can be derived from the polynomial of that sequence, by deriving the 'reciprocal polynomial'. Given an irreducible polynomial f(x) – one that gives a maximal length sequence – the reciprocal polynomial is $x^n[F(1 \div x)]$

Taking the ninth degree example of Fig. 1, where the generating polynomial is $x^9+x^6+x^4+x^3+1$, the reciprocal polynomial is,

$$x^9\left[\left(\frac{1}{x}\right)^9+\left(\frac{1}{x}\right)^6+\left(\frac{1}{x}\right)^4+\left(\frac{1}{x}\right)^3+1\right]$$

Multiplied out, this gives,

$$1+x^3+x^5+x^6+x^9 \text{ or } +x^9+x^6+x^5+x^3+1$$

So an LFSR with exclusive-or feedback from the ninth, sixth, fifth and third stages will provide the reverse time sequence of one with EXOR feedback from the ninth, sixth, fourth and third stages. Stated more simply, every term x_r is replaced by a term x_{n-r}. Here, r is less than n.

Convolvers for spread-spectrum systems

With a chipping rate of 6.3Mb/s, the spread-spectrum signal needs to be 'collapsed' at the receiver, back to the original unspread 100kb/s data. This is usually done at a convenient intermediate frequency, the receiver being up to this point a conventional superheterodyne type.

In one arrangement, the IF signal is fed into one transducer port of a convolver consisting of a SAW (surface acoustic wave) device using, typically, lithium-niobate material. The signal proceeds across the device as a surface acoustic wave, passing underneath a 'collector' electrode, forming the device's output. Another input, physically at the opposite end, is fed with a signal at the IF centre frequency, consisting of a CW (continuous wave) signal that has been chipped with the time reverse sequence.

Under the collector, the two surface waves, travelling in opposite directions, interact due to the piezo-electric non-linearity of the material, to provide an output at the original 100kb/s data rate. Thus the spread data is collapsed back to the original data rate. On the other hand, any interference, such as an attempt at jamming with a CW signal, is spread out by the local reverse sequence signal, to a low level. With the chipping rate at 63 times the data rate, the device provides – ideally – 18dB of 'processing gain' to the wanted signal, hopefully raising it above the level of the now spread interference.

Creating a noise source

Since the maximal length sequence for an LFSR with n stages is 2^n-1, there is a component in the output at a frequency of $f_{min}=f_c \div (2^n-1)$, where f_c is the clock frequency. There are components also at all harmonics of the minimum frequency f_{min}, up to the clock rate.

The envelope of this spectrum is the well known $\sin x \div x$ function, and the first zero of this occurs at the clock rate, as in Fig. 4. Only the first of the higher order loops of the spectrum is shown. This spectrum is caused by the distribution of different length sequences of 0s and 1s in the overall sequence, which is as shown in Table 2. This table is taken from reference 2, which contains much further useful information.

As the longest runs in the overall sequence are n ones and (n–1) noughts, if the sequence is low-pass filtered with a cut-off frequency lower than f_c/n, different length sequences of ones and noughts will result in different length positive- or negative-going ramps, all of the same slope.

Table 2

Number of runs of consecutive 0s and consecutive 1s, up to n ones and n–1 noughts in the maximal-length output sequence of an n stage LFSR using exclusive-or gating. In this case, n noughts is the forbidden degenerate state.

Run length	Consecutive 1s	Consecutive 0s
n	1	0
n–1	0	1
n–2	1	1
n–3	2	2
n–4	4	4
...		
2	2n–4	2n–4
1	2n–3	2n–3

The result is a random noise-like waveform, as illustrated in Fig. 5. This is taken from reference 3, which describes an instrument that did not work exactly as planned. It was supposed to produce audio-frequency noise from a maximal-length sequence, but the article stated that start-up was like Fig. 3, whereas a maximal length sequence must start up immediately, as in Fig. 2.

The error was promptly pointed out by a reader, it being always open season for shooting down authors – a healthy tradition of this journal. The noise source was however used in a more recent article[4] and the operation of the instrument was investigated.

This investigation cleared the original fault, and uncovered an interesting and unexpected fact: a low-pass filtered PRBS does not necessarily provide a Gaussian amplitude distribution. A redesign was therefore incorporated to provide the desired distribution, but that must be a subject of a later article.

Meanwhile, if you want to experiment with LFSR PRBS generators of different lengths, the necessary taps for shift registers with up to 32 stages long are given in reference 5. For most register lengths, there is a 'trinomial', –a polynomial with just the three terms x^n, x to some lower power, and unity. These can be implemented with a single EXOR gate.

Reference 6 gives taps for lengths up to 45 stages for those degrees that have a trinomial. This list of trinomials is extended, for degrees up to 1000 (!), in reference 7.

Reference 8 points out that a pseudo-random binary sequence generator can be made from just a CD4006 18-stage shift register and a CD4070 quad exclusive-or gate – and of course a clock source. The length of the sequence is 2^{18}–1 or 262 143 bits long. ■

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EMI tips and fixes

Leading off with an EMI backgrounder, Joe Carr presents tips and fixes for minimising interference.

Electronic pollution is all around us. Radio and noise waves impinge on us all of the time. Never has electromagnetic interference (EMI) been so great as it is today.

One source defines EMI thus: "... electromagnetic interference is a degradation in performance of an electronic system caused by an electromagnetic disturbance." At worst, EMI can cause a loss of human life, as when it interferes with an aircraft or automobile electronic system. At best, it will pass unnoticed or will interfere with the electronic system on a sub-audible basis.

The European Community has issued regulations pertaining to EMI in all manner of electrical and electronic equipment. Any electrical and electronic product sold in Europe must exhibit that it neither emits nor is affected by radiation and conduction of EMI. In other words, it must be electromagnetically compatible (EMC).

Means of EMI transmission

EMI is transmitted from the source to the victim system in two basic ways: conduction and radiation. The difference is that the EMI travels along a wire in conduction, and travels by air in radiation. In general (but not always), radiation **Fig. 1a)**, occurs at high frequencies above 30MHz, and conduction Fig. **1b)**, occurs at frequencies less than 30MHz.

In some cases, both radiation and conduction can occur. In those cases, either radiation occurs first, and then the wave is conducted into the equipment on a line, Fig. **1c)**, or the radiation occurs after conduction Fig. **1d)**.

In general, the existence of EMI can occur only if:

- there is a source of energy,
- there is a receiver of that energy, and
- there is a transmission path between the two.

If any of the three does not exist, EMI cannot occur.

What we do about EMI depends on the situation. For example, in the case of some noise sources we can turn it off or otherwise suppress it. In other cases, we might have to live with the effects of EMI as best we can.

Electronic noise

Noise comes in two different types: continuous and transient. By definition, noise has been standardised as 'transient' if it lasts less than a sixtieth of a second, i.e. 16.67ms, and continuous if it lasts longer.

Continuous noise. Low-frequency noise sources include fluorescent lights, electric motors, and switching-mode DC power supplies. High-frequency noise is mostly radio-frequency interference, or RFI. It can originate in either radio transmitters, computer clocks or other sources.

In the typical RFI environment, signals levels can be between a few microvolts/metre and 300V/metre. While the latter field strengths are only found close to transmitting antennas for high-power radio and radar stations, anything in excess of 1V/m can cause damage to unprotected circuits.

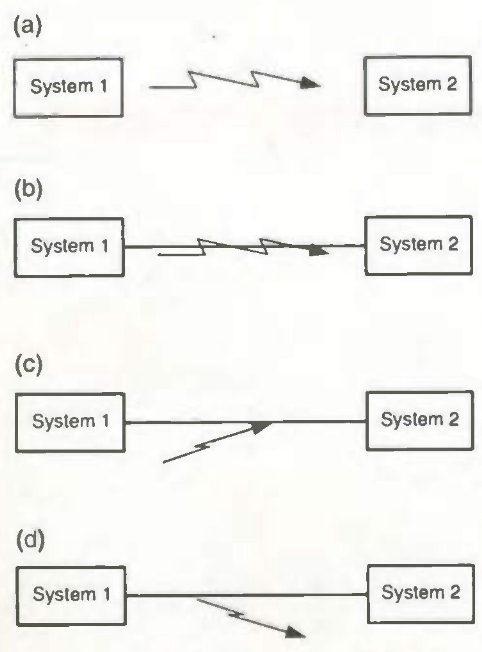

Fig. 1. There are two basic ways in which EMI is transmitted; radiation, a) and conduction, b). Conduction from radiation entering the system is represented in c), while d) shows radiation resulting from conduction.

Test specifications for commercial systems may call for protection to 10V/m, while automotive and medical environments can call for up to 200V/m and military environments up to 300-400V/m.

Analogue circuitry tends to be more influenced by RFI than digital circuitry because it can be interfered with by lower voltage fields.

Transient noise. A 'transient' is any temporary over voltage or over power condition that lasts for less than 16.6ms. Transients are either repeatable or random in nature. An example of the repeatable type of transient is the discharge of an inductor or capacitor. Examples of random type transients include electrostatic discharge (ESD), lightning, and the nuclear electromagnetic pulse.

In the case of lightning, **Fig. 2** shows the exposure of the US electrical power system to lightning strikes. Clearly, high-voltage lines are struck more frequently than low voltage lines. Unfortunately, that exposure transmits along the lines to your home or business to disrupt electronic circuits.

In fact, the lightning doesn't have to actually strike the line. Coupling via inductance, it can cause disruptive currents to flow in the power system by striking something close!

Counters to EMI

There are two effective ways to counter EMI: shielding and filtering, **Fig. 3**. Shielding is used to guard against radiation interference, whilst the filter is used to guard against conduction interference. The filters have the advantage of being bi-directional, so they also prevent interference from flowing out of the system as well as prevent it from flowing in.

Shielding. Shielding is used to attenuate the interfering RF signal before it reaches the protected circuitry. Very frequently, the hidden difference between a higher-priced appliance and a lower-priced one is in the internal shielding that one gets. Consider computer monitors for example. The principal difference between a high-priced model and a cheap one is in the shielding that is provided.

Unfortunately, all shielded enclosures for electronic projects are not created equal. Some enclosures are butt fitted, and have dimples or notches to hold the half shells together. As far as I am concerned, the minimum requirement for low-frequency radiation noise is shown in **Fig. 4a)**. Note that the flange of the lower half shell overlaps the upper half shell by at least 4 to 6mm.

Four screws, of which only two are shown, hold the assembly together. For high frequencies, even this box is insufficient, but it can be made suitable by adding of more screws, Fig. **4b)**. The rules are that the screws should be not more than one half wavelength apart, and one-eighth wavelength is better.

Filtering. Filtering can take on different meanings for different situations. In general, most EMI filters are low-pass filters, although high-pass and bandpass filters exist. In some cases of a particular frequency being the cause of interference, a notch filter may be used.

In general, a perfect, ideal single component filter – either a capacitor or inductor – has a theoretical roll-off or gain of –20dB/decade, with a practical maximum of something between –60 and –120dB. In fact, real components do not achieve that theoretical goal. Capacitors are more useful in high-impedance circuits, whereas inductors are more useful in lower impedance circuits.

Perhaps the simplest form of single component filter is the feedthrough capacitor, also known as an 'EMI filter'. When combined with good shielding, such a capacitor can be quite sufficient.

Figure 5 shows two methods of passing a feedthrough capacitor through a shielded panel. Figure **5a)** shows the screw-in variety. The threaded nut is cinched tight against the chassis or panel. Figure **5b)** shows the installation of a solder-in type of feedthrough capacitor. A small fillet of solder is used to hold the capacitor against the chassis or panel. This type of capacitor assumes a solderable chassis or panel, thus it eliminates the use of aluminium.

Where greater suppression is needed, a combination of *L* and *C* elements is needed. A two-component L-section filter is shown in **Fig.**

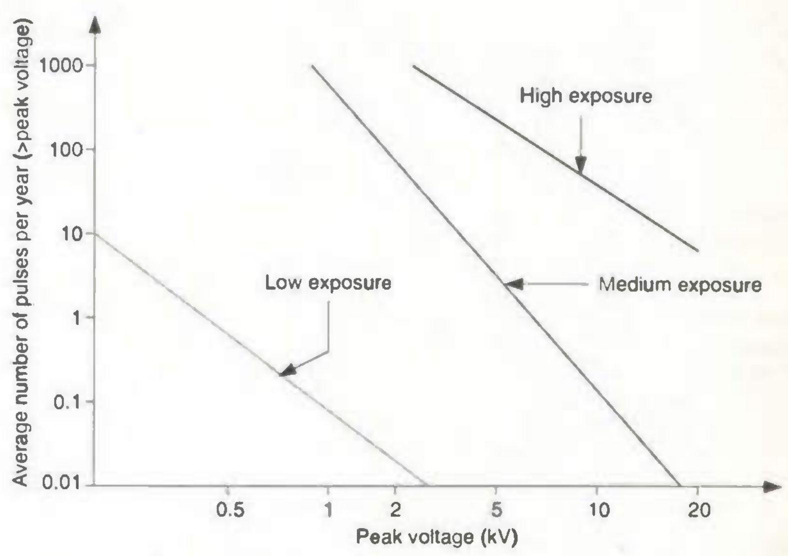

Fig. 2. Average lightning strikes against USA power systems.

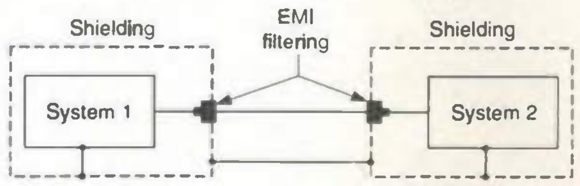

Fig. 3. Shielding and EMI filtering are suitable defenses.

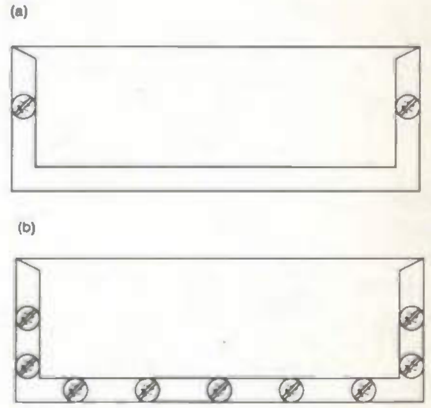

Fig. 4. Simple but effective low frequency shielding, a), in which the lower shell overlaps the upper by at least 4mm. At high frequencies more screws are needed to provide an effective shield.

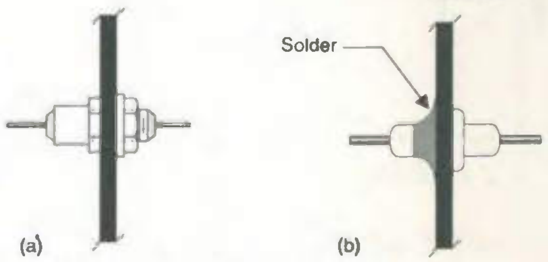

Fig. 5. Two types of feedthrough capacitor.

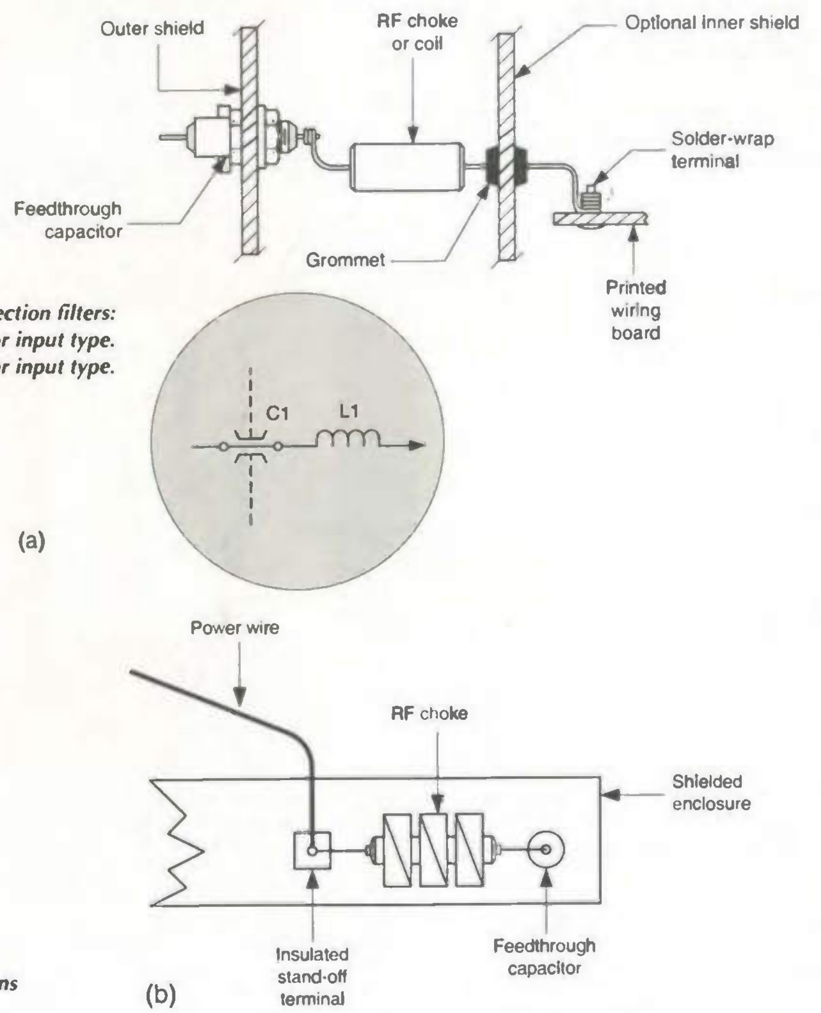

Fig. 6. Simple L-section filters: a) internal capacitor input type. b) external inductor input type.

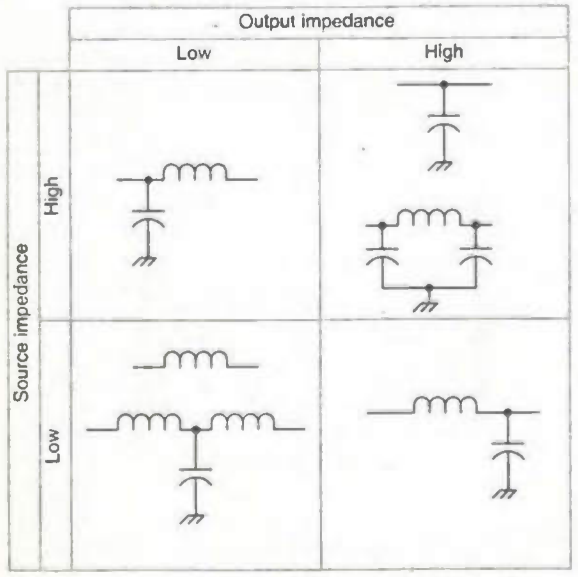

Fig. 7. Several impedance situations and the filters to implement them.

6a). This filter produces a theoretical gain of –40dB/decade, which means 100:1 per decade between input and output signals. Such a filter can be used at any frequency, although the values will tend to differ between, say, LF and VHF. The ideal is to keep the lead lengths as short as possible to prevent radiation of the signal. As an alternative, a higher-order filter can be realised by replacing the grommet in Fig. 6a) with a second feedthrough capacitor. In that case, a theoretical gain of –60dB/decade is realised.

In Figure **6b)** you can see a case where the opposite situation occurs, i.e. the inductor input L-section filter. In this case, the inductor or RF choke is mounted external to the chassis, and directly drives the feedthrough capacitor.

Figure 7 shows a graphic of what filters to use when the source impedances are known. For example, in the case where the input impedance and the output impedance are both low, then use either a single inductor or RF choke, or a T-filter consisting of two inductors and a capacitor.

When the source impedance is high, and the output impedance is low, then use a single capacitor input L-section filter. Similarly, when the input impedance is low, and the output impedance is high, use an inductor input L-section filter.

Finally, when the input and output impedances are both high, use either a single capacitor, or a pi-section filter as shown.

Common-mode and differential currents

Noise currents can flow in two modes: differential and common mode. These are defined as follows:

Differential mode. As shown in **Fig. 8a)** there is a conducted signal on a signal line that is returned along the return line, or via a grounded connection. The noise current is characterised by the arrows flowing in different directions.

Common mode. In common mode, conduction noise appears in multiple conductors flowing in the same direction.

The filtering necessary for differential and common mode filtering is different. In differential-mode filtering, it is necessary to place a filter in series with the 'hot' lead, with suitable bypassing to ground. In the common mode case the same filtering must be applied to all affected leads.

Figure 9 shows a filter that is suitable for both differential and common-mode forms of EMI.

AC power-line filtering

The AC power lines are a source of conducted EMI and must be filtered extensively to make them really clean.

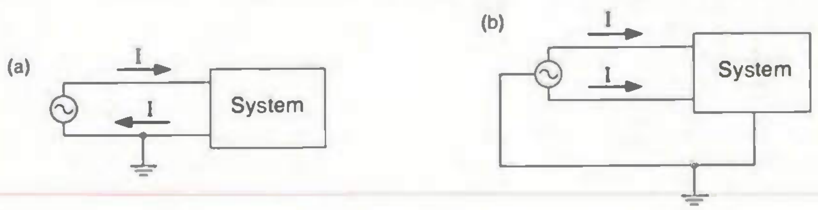

Fig. 8. Differential signals, a), versus common-mode signals, b).

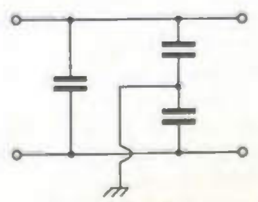

Fig. 9. Filter for both differential and common mode signals.

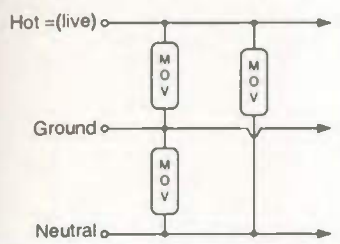

Fig. 10. Metal oxide varistors (MOV) use on AC power mains for clamping spikes.

Not only should the AC lines be filtered against RFI, but also against lightning strikes at a distance. The lightning and other high-voltage transients can be handled as per **Fig. 10** with metal oxide varistors, or MOVs. These devices are made by various suppliers, and can act like a pair of zener diodes back-to-back. In other words, they snip the high voltage transient to a lower level, regardless of polarity.

Basically, MOVs act like an insulator at lower voltages. But when a certain critical threshold voltage is exceeded, the varistor develops a low resistance, shunting the offending voltage transient to a low value. Varistors are used primarily for transient voltage spikes on the AC power line.

For RFI, a filter must be provided to the AC power line. **Figure 11** shows a suitable filter that affects both common-mode and differential RFI on the line. When the values are high enough, it will also protect somewhat against lightning and other transients. This is because those transients have a high frequency component as well as the fundamental frequency.

Suitable filters have been moulded into AC power line sockets.

Special medical EMI problems

EMI problems exist in medical electronic devices such as electrocardiograph (ECG) or electroencephalograph (EEG) machines. These machines have to process signals in the order of a few microvolts to about a millivolt in the presence of strong interfering signals.

In addition to regular RFI problems, there are two additional problems. First, there is the problem of the defibrillator. This is a high voltage – several kilovolts in fact – capacitive discharge device used to 'jump start' the heart of a patient undergoing resuscitation.

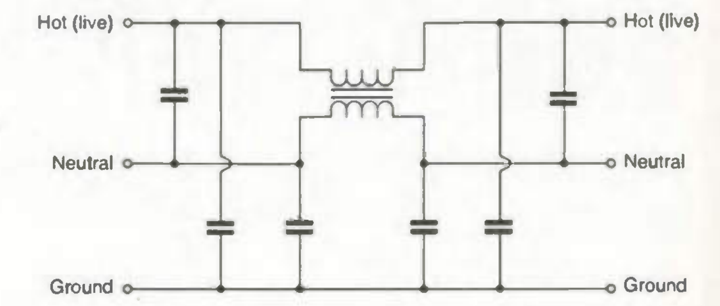

Fig. 11. AC power mains filter.

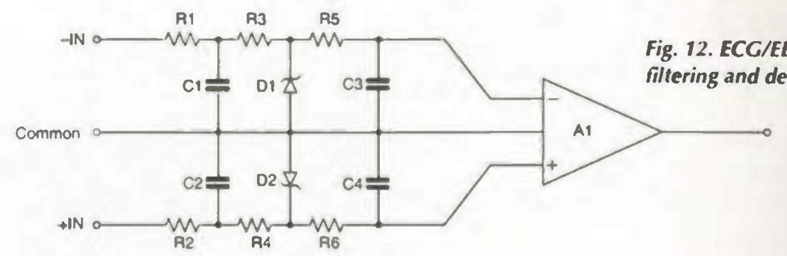

Fig. 12. ECG/EEG preamplifier filtering and defibrillator protection.

The second problem is in the application of 500kHz to 2500kHz high powered RF electrosurgery units just a few centimetres from the site of the EEG or ECG electrodes. As much as 400 or 500 watts of RF could be applied to the 'circuit' only a few centimetres from the pick-up electrodes!

Figure 12 shows the input to a bioelectric amplifier suitable for EEG or ECG use in the presence of high electrical or electromagnetic fields. Both defibrillator and electrosurgery units can be accommodated by the filtering shown.

Signals involved in ECG and EEG are below 100Hz, so an *RC* filter will suffice in this case. The *RC* filter consists of resistors R_1 through R_6, and capacitors C_1 through C_4. This is a low-pass filter.

The defibrillator protection is the zener diodes and series resistors R_1-R_3 and R_2-R_4 between the source and amplifier A_1. Sometimes, in older machines neon glow lamps like the NE-2 are used instead of the zener diodes. The disadvantage of the neon lamps is their relatively high protection voltage.

Computer EMI

The case of computer EMI is very serious. Just place an AM radio anywhere close to a modern computer, and you will hear lots of hash. In fact, with computer clock speeds reaching several hundred megahertz, the interference to FM radios can be tremendous.

Figure 13 shows a method for connecting a digital connector pin that can carry EMI to a printed wiring board. The ferrite beads act like little RF chokes, so will eliminate RFI in the VHF/UHF region. Because the filtering is bi-directional, it will attenuate noise going out as well as coming into the computer.

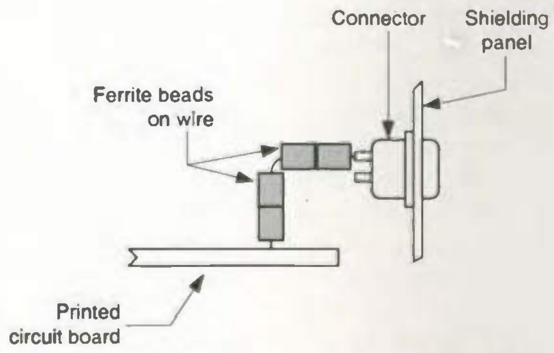

Fig. 13. Computer connector fitted with ferrite beads.

The principal offender with respect to noise from computer systems is the monitor. This is because of two factors. First, the deflection circuits tend to operate in frequency ranges – under 40kHz – that are below many other systems, and they have lots of harmonics. Second, those deflection circuits tend to be high power.

The answer to the problem is to place shielding around the circuits, and placing a common-mode choke in the signal line.

In summary

EMI protection is often an afterthought in the design of electronic equipment. It should be a first requirement, but unfortunately this isn't often the case. The methods that I have just discussed will go a long way towards suppressing the RFI or transient conditions on the signal or power lines. ■

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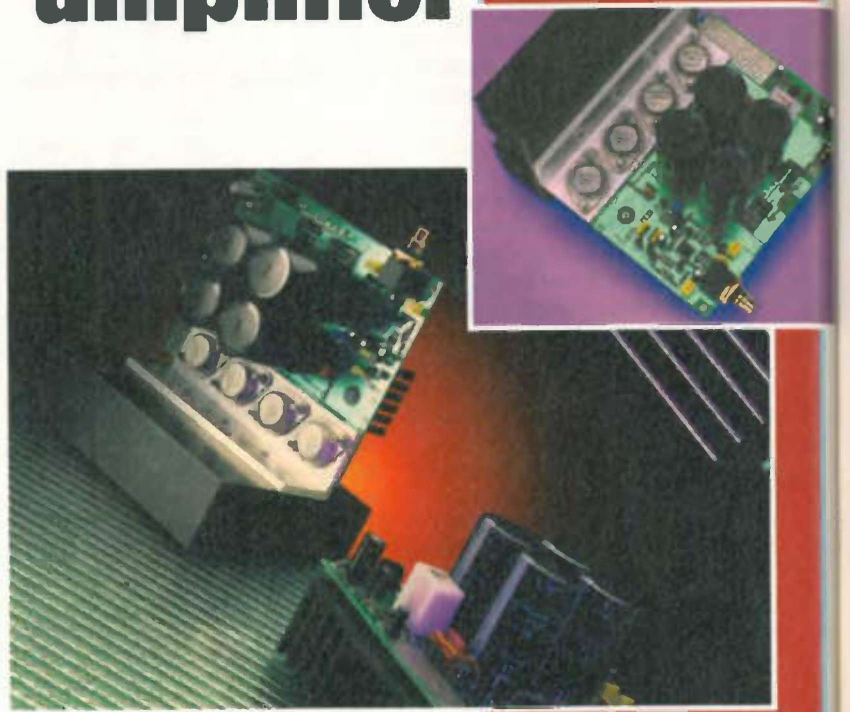

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Analogue I/O for the PC

Plugging into a PC's ISA slot, Tariq Iqbal's analogue i/o card provides eight 12-bit resolution inputs and one 8-bit output. A few QuickBasic routines – including one for displaying thermocouple temperature – demonstrate how easy it is to read and write analogue voltages.

A personal computer can be used for applications including automating machines, process monitoring and control, electronics test, programmable signal generation and data logging. All such applications require an interface with the analogue world.

The analogue i/o card described here has five single-ended channels with 12 bit resolution that can be sampled at up to 100kHz. Analogue output is provided by an 8-bit d-to-a converter with unipolar voltage output.

Two K-type thermocouples can be directly attached to this card. One on-board instrumentation amplifier can be used for many applications.

One spare 8-bit ISA PC slot on your PC is required. My prototype measured 15cm by 10cm and included a prototyping area for extensions. All I/O connections were available on a 25 pin D-type connector.

Hardware details

Interfacing of the a-to-d and d-to-a converters to the ISA bus is done using an 8255A parallel peripheral interface. The d-to-a converter connects to port A of the PPI. Port B and the upper half of port C are used to interface a-to-d converter.

Three least-significant bits of port C are used to switch analogue multiplexer. Bit 3 of port C is used for starting a conversion, **Fig. 1**. A detailed circuit diagram is presented in **Fig. 2**.

Address decoding corresponding to address 792 is achieved by two 74LS85 comparators and a 74LS00 NAND gate. The 74LS245 bus transceiver is meant to provide protection against any bus clash.

A DAC0800 d-to-a converter is interfaced to port A of the 8255 PPI. The a-to-d converter is a 12-bit resolution AD574 from Analog Devices. Lower byte data output of this device is supplied directly to port B of 8255, while a nibble of high byte is supplied to the upper nibble of port C.

A start-of-conversion pulse following the sequence high→low→high is generated and supplied to the a-to-d converter by port C, bit 03 of the 8255. In this card the a-to-d converter is configured to measure input voltages in the range ±5V DC.

To increase the number of channels to 8 inputs a 4051B analogue multiplexer is used at the input of a-to-d converter. Channel selection is made by the lower 3 bits of port C.

Inputs marked CH0-CH4 are directly available on the 25pin connector and they have a signal range of ±5V DC.

Two AD595 thermocouple amplifiers and one AD521 instrumentation amplifier are also present on the card. Channels 5 and 6 are reserved for the thermocouple amplifiers. Two K type thermocouples may be directly interface with this card.

Channel 7 is reserved for the instrumentation amplifier, whose gain can be selected as 100 or 1000 by a slide switch. The AD521 instrumentation amplifier can be used for interfacing with resistance bridges – e.g. strain gauge – and various other sensors like thermistors and RTDs.

On my prototype, +12V, –12V, +5V and –5V supplies from the PC are also available on the 25 pin D-type connector.

Table 1 provides necessary addresses of I/O card and control word. **Table 2** gives the details of I/O available on 25 pin D-type connector.

Table 1. Addresses of the I/O card.

Address	Function
795	Control word
792	Port A
793	Port B
794	Port C

Table 2 Signals at I/O card connector

Pin	Signal	
1	GND	
2	Analogue input CH 0	
3	Analogue input CH 1	
4	Analogue input CH 2	
5	Analogue input CH 3	
6	Analogue input CH 4	
7	GND	
8	$-TC_{in}$	CH#5
9	$+TC_{in}$	CH#5
10	GND	
11	$-TC_{in}$	CH#6
12	$+TC_{in}$	CH#6
13	GND	
14	–IN	CH#7
15	+IN	CH#7
16	GND	
17	Analogue output	
18	NC	
19	NC	
20	NC	
21	NC	
22	–12	From PC bus
23	+12V	From PC bus
24	–5V	From PC bus
25	+5V	From PC bus

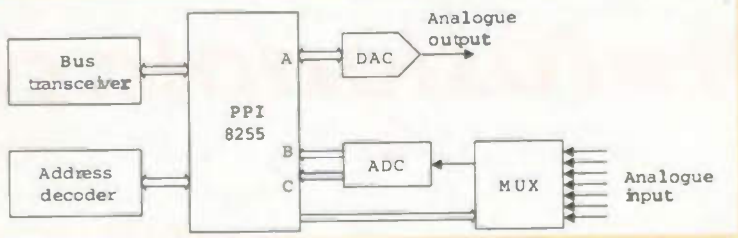

Fig. 1. Block diagram of the analogue I/O card for the PC.

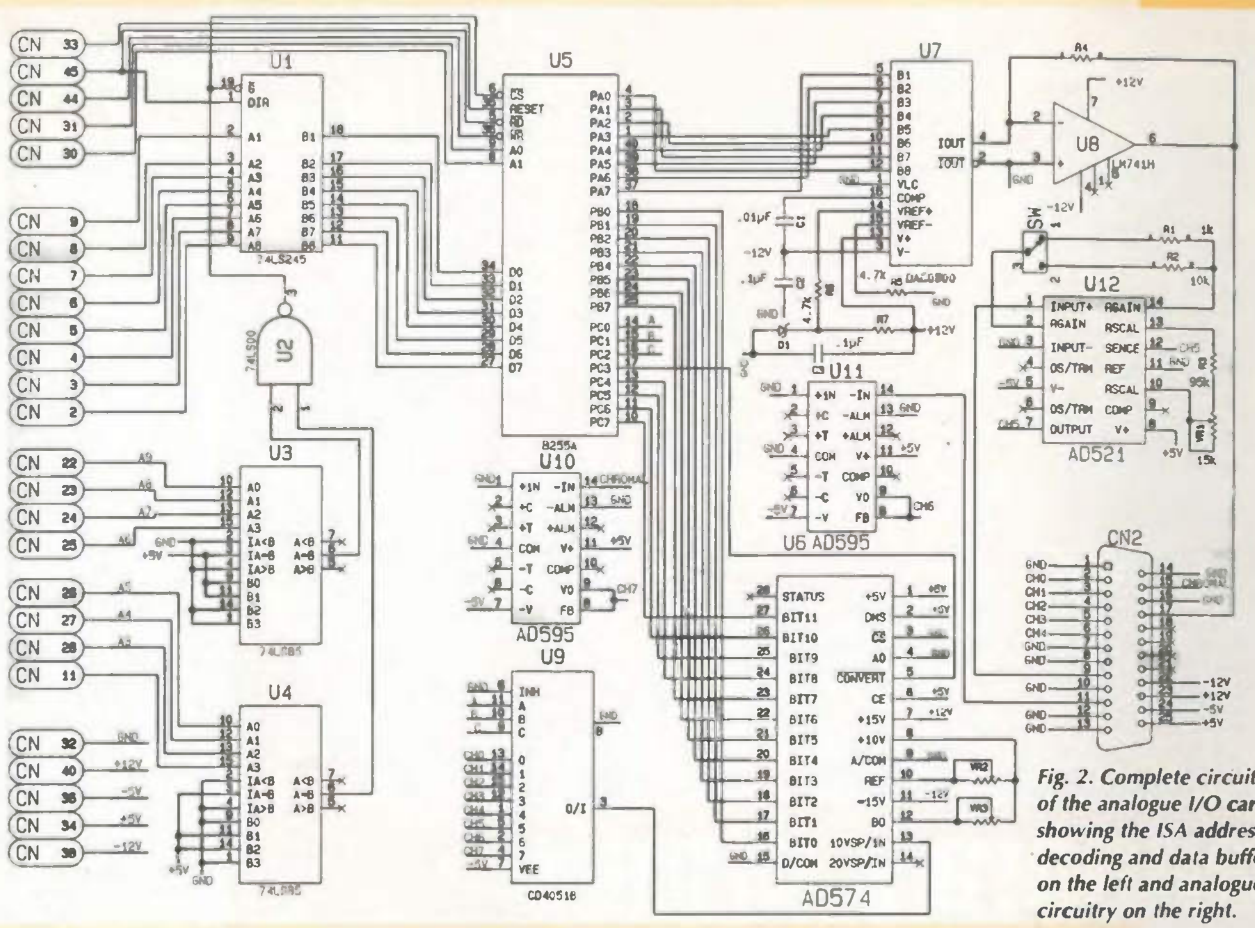

Fig. 2. Complete circuit of the analogue I/O card, showing the ISA address decoding and data buffering on the left and analogue circuitry on the right.

Programming

Programming of the I/O card is relatively simple using Microsoft's QuickBasic. A routine for reading data from channel 0 is provided in **List 1**. After a start of conversion plus first a low byte is read and then low nibble of high byte is read.

Output from the a-to-d converter is constructed by combining low and high bytes. Since the input range of converter is ±5V, a simple calibration equation at the end of program is used to transform the converter's output into equivalent voltage.

A program for providing an analogue voltage output is given in **List 2**. Analogue output is in the range is 0 to 10V. Sending 00 to address 792 results a 0V at the output, while sending 255 to the same address results a 10V at the output.

This routine generates a saw tooth waveform at the output. Any desired voltage can be generated at the d-to-a converter's output by using **List 3**.

A simple program to read K type thermocouple attached to pin 8 and 9 of the D-type connector is in **List 4**. Output of the thermocouple amplifier AD595 is calibrated at 10mV/°C.

It is useful to allow a prototyping area on the card for building additional interfacing circuits. I have used this card to measure parameters such as blood pressure, motor angular velocity, motor position, flow rate, respiration rate, body temperature, force on a load cell, and water level. This is a multipurpose card and has many applications. ■

List 1. A routine for reading an analogue voltage applied to channel 0 of the I/O card could be written as follows.

```
OUT 795, 138'CONTROL WORD
CH% = 0
SC% = CH% +8
OUT 794, SC%:OUT 794, CH%:OUT 794, SC%
LB% = INP(793):HB% = INP(794)
HB% = HB% \ 16
d% = LB% + HB% * 256
VOLTS = d% / 4096 * 10.14 - 5.07
```

List 2. Routine for writing values to the analogue output port to produce a sawtooth waveform.

```
OUT 795, 138
10 FOR N% = 0 TO 255
OUT 792, N%
NEXT N%
GOTO 10
```

List 3. Routine for producing a specific voltage at the d-to-a converter's output.

```
OUT 795, 138
10 INPUT "VOLTS TO BE OUTPUTTED" =";V
OUT 792, CINT(V*25.5)
GOTO 10
```

List 4. Simple routine for reading and displaying the temperature of a thermocouple.

```
OUT 795, 138'CONTROL WORD
CH% = 5
SC% = CH% +8
OUT 794, SC% : OUT  794, CH% : OUT 794, SC%
LB% = INP(793) : HB% = INP(794)
HB% = HB% \ 16
d% = LB% + HB% * 256
VOLTS = d% / 4096 * 10.14 - 5.07
Temp = 100* VOLTS
Print "Temperature C° = "; Temp
```

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4. The Volunteer Organist, Peter Dawson, 1913
5. Dialogue For Three, Flute, Oboe and Clarinet, 1913
6. The Toymaker's Dream, Foxtrot, vocal, B.A. Rolfe and his orchestra, 1929
7 As I Sat Upon My Dear Old Mother's Knee, Will Oakland, 1913
8 Light As A Feather, Bells solo, Charles Daab with orchestra, 1912
9 On Her Pic-Pic-Piccolo, Billy Williams, 1913
10 Polka Des English's, Artist unknown, 1900
11 Somebody's Coming To My House, Walter Van Brunt, 1913
12 Bonny Scotland Medley, Xylophone solo, Charles Daab with orchestra, 1914
13 Doin' the Raccoon, Billy Murray, 1929
14 Luce Mia! Francesco Daddi, 1913
15 The Olio Minstrel, 2nd part, 1913
16 Peg O' My Heart, Walter Van Brunt, 1913
17 Auf Dem Mississippi, Johann Strauss orchestra, 1913
18 I'm Looking For A Sweetheart And I Think You'll Do, Ada Jones & Billy Murray, 1913
19 Intermezzo, Violin solo, Stroud Haxton, 1910
20 A Juanita, Abrego and Picazo, 1913
21 All Alone, Ada Jones, 1911

Total playing time 72.09

21 tracks – 72 minutes of music.
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Being able to create two identical but phase-shifted sine waves is useful for testing, but generators capable of producing them are rare. To rectify this, W Q Yang and T Y Ng* have produced a design using two high-performance direct digital synthesis chips to produce two 1V signals at frequencies up to 16MHz. Under the control of an RS232 port, these signals can be used together, with fully adjustable phase shift, or independently.

Phase-shiftable signal generator

Many applications in industry and laboratories require the use of signals with certain phase difference or of adjustable phase shift between two signals. Phase-sensitive demodulation is one example, whose output depends on the phase difference between input signals.

Other examples include signal mixers, vector generators, selective voltmeters, rotational torque induction, etc. All of these applications lead to the demand for highly versatile signal generators that can provide the user with full control over the signal's amplitude, frequency and phase.

The phase of a signal can only be measured and shifted with respect to a reference signal. There are no simple and direct methods to generate phase shifts.

At present, phase differences between signals are produced using dedicated phase-shifting networks and circuits that provide limited and inflexible phase shifts. Although function generators are well established as the most versatile of signal sources, they do not have the capability to change the phase of the signals. This is a major drawback – especially when phase shifting or phase difference between signals is required.

Direct digital synthesis, or DDS for short, has long been recognised as a superior technique for generating highly accurate signals of low distortion over a wide frequency range. With DDS, the amplitude, frequency and phase of the generated signals can be easily and instantaneously controlled.

Using two DDS chips, we have developed a dual synchronous DDS signal generator. It is small, lightweight and has all the advantages of DDS technology in term of performance and cost.

The generator derives it power through an external 9V power adapter. It can be operated through a PC's RS232 serial port with the developed Visual Basic software and the generated signals are fully user programmable.

Both DDS chips can be operated individually to produce two independent signals, or as a synchronous pair to produce two signals with the same frequency up to 16MHz. But the phase between them can be adjusted. Figure 1 shows the prototype.

In the first part of this article, details of the hardware and the operational principle of the system are given. The second part shows the user interface created using Visual Basic for controlling the signal generator, and lastly test results are given and future developments are suggested.

A block diagram of the generator is shown in Fig. 2 and Fig. 3 shows its detailed circuit. The system is made up of four parts: power supply, serial-to-parallel interface, data latching circuit and DDS chips and clock.

A 9V power adapter forms an external power source and a 78L05-voltage regulator provides the on-board power supply. The generator employs two AD7008 chips as a pair of signal generating sources controlled serially via the PC's RS232 port.

*Dr Yang is with UMIST, as was his co-author Mr Ng at the time of writing.

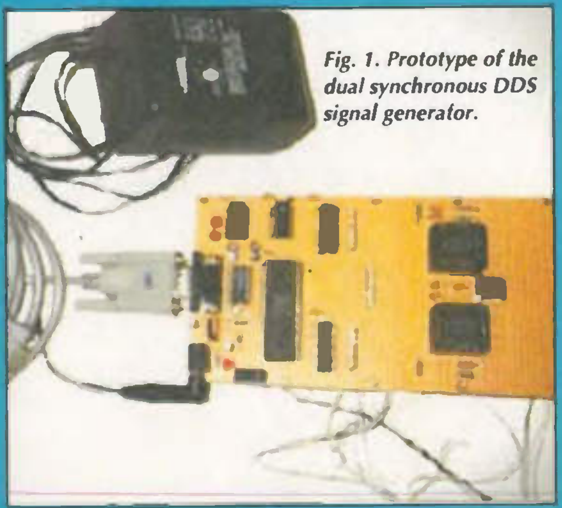


Fig. 1. Prototype of the dual synchronous DDS signal generator.

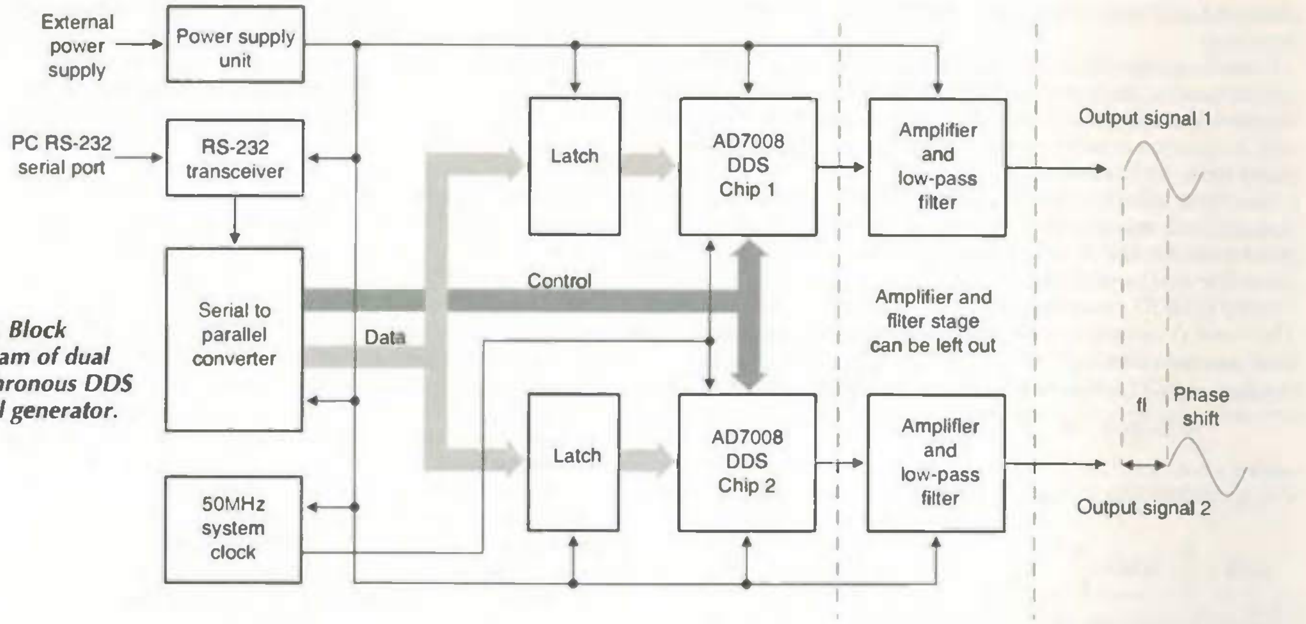


Fig. 2. Block diagram of dual synchronous DDS signal generator.

Digital synthesiser chip details

Analog Devices' AD7008 is a DDS device0 integrated on a single CMOS chip. A detailed introduction to DDS theory and the operation and control of the AD7008 DDS chip can be found in its data sheet.

Figure 4 is the functional block diagram of the chip. It consists of a numerically-controlled oscillator, or NCO, employing a 32-bit phase accumulator, sine and cosine look-up tables in ROM, in-phase and quadrature modulators and a 10-bit d-to-a converter.

Clock rates up to 50MHz are supported and the chip is capable of phase modulation, frequency modulation and both in-phase and quadrature amplitude modulation. The generated signal is controlled by loading values into the chip's registers either through its parallel interface or the serial interface. Registers 'Freq0 reg' and 'Freq1 reg' determine the frequency, 'Phase reg' determines the phase and 'IQMOD' determines the amplitude.

Read-only memory stores the quantised amplitude values of a sine/cosine wave. Given a phase step value – ΔPhase – stored in FREQ0 or FREQ1, the phase accumulator steps through the ROM repeatedly to produce the sine/cosine values. The sum of the two values is input to the d-to-a converter and a sinusoidal waveform of the desired frequency, f_{signal}, is reconstructed at the converter's output. Tuning for the chip is determined as follows.

$$f_{signal} = \frac{\Delta Phase \times f_{clock}}{2^{32}} \qquad (1)$$

Here, $0 \le \Delta Phase < 2^{32}$. Solving for ΔPhase,

$$\Delta Phase = \frac{f_{signal} \times 2^{32}}{f_{clock}} \qquad (2)$$

From the second equation, the ΔPhase required to produce a specific f_{signal} is determined and loaded into either of the FREQ registers.

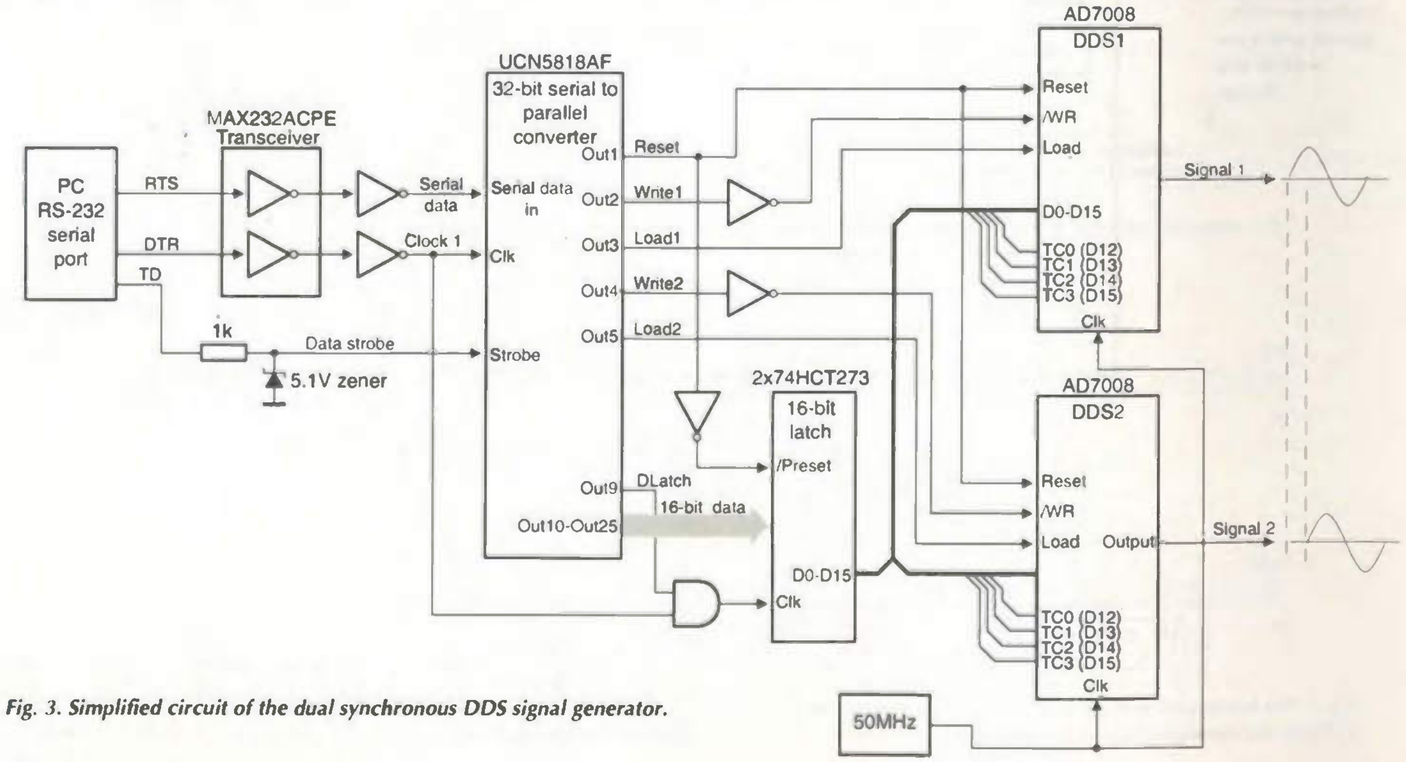


Fig. 3. Simplified circuit of the dual synchronous DDS signal generator.

The FSELECT pin is then used to select the register for generating the signal.

Even though the phase of the output signal can be offset using the 12-bit phase register, accurate phase shift cannot be produced due to the limited phase resolution provided by the register. Quadrature amplitude modulation is performed using IQMOD to generate the desired signal phase shift instead.

The 20-bit IQMOD register is used to control the amplitude of the cosine (*I*) and sine (*Q*) signal components generated from the ROM. Its first 10 bits, bits 0 to 9, define the *I* amplitude. The next 10, bits 10 to 19, define the *Q* amplitude.

Using IQMOD, quadrature amplitude modulation can be performed. The *I* and *Q* components are summed together before entering the DAC and their sum is given by,

$$Icomponent + Qcomponent = A\cos\omega t + B\sin\omega t = R\sin(\omega t + \theta) \quad (3)$$

where,

$$R = \sqrt{(A^2 + B^2)}$$

$$\cos\theta = \frac{B}{R},\ \sin\theta = \frac{A}{R}$$

Equation 3 shows that by summing the *I* and *Q* components together, a signal of amplitude *R* and phase angle *θ* is produced. The values of *R* and *θ* depend on two amplitudes, *A* and *B*. Conversely, given the desired signal amplitude *R* and phase angle *θ*, the required amplitudes of the *I* and *Q* components can be determined by equations 4 and 5.

$$A = R\sin\theta \quad (4)$$

$$B = R\cos\theta \quad (5)$$

Here, $0 \le R \le 1$ Vpk-pk and $0° \le \theta \le 360°$. By loading the determined *A* and *B* values into IQMOD, the generated signal will have the desired amplitude and phase shift.

Capable of producing low distortion signals on its own, the AD7008 chip does not require additional output signal conditioning. It has an output compliance of 1V pk-pk.

Serial-to-parallel interface

In this design, a PC's RS232 is used to control the generator. **Figure 5** shows that RS-232 port has only three output pins for sending signals, namely RTS, DTR and TD. These are not enough for sending data to and controlling the AD7008 chips.

Therefore, a serial-to-parallel interface is used to produce a 16-bit data bus and a 5-line control bus to both AD7008s. The AD7008 chips are configured to operate in 16-bit parallel-loading mode. Serial signals sent from the RS232 port to the generator are converted into parallel form for use as either control signals or as data to be loaded into the chip's registers.

Data control signals RTS and DTR are first converted from RS-232 voltage level into TTL using a MAX232ACPE low power RS-232 transceiver. The transmit-data pin TD is converted into -0.6V to +5.1V using a voltage clamping circuit consisting of a resister and a zener diode, Fig. 3.

The UCN5818AF chip is a 32-bit serial-to-parallel converter chip that forms the heart of the serial-to-parallel interface. It is basically a 32-bit shift register with parallel latched outputs as shown in **Fig. 6**. RTS is used for sending serial data to the converter. DTR is used as clock signal (Clock 1) to shift in the data and TD is used as a data strobe to internally latch the data stored in the shift register to the chip's output.

Referring to **Fig. 7**, serial data presented by RTS at the UCN5818AF's input is shifted into the shift registers bit by bit towards the serial data output at the positive transition of each Clock 1 (DTR) pulse. Data stored in the shift registers are then internally latched to their respective outputs in parallel when the STROBE signal (TD) is high. The internal latches continue to accept new data as long as STROBE (TD) is high.

Data in 32-bit serial form sent from the PC to the signal generator are converted into parallel form for use as 16-bit direct digital synthesis data or five control signals to both AD7008 chips by the UCN5818AF chip.

Serial data is in one of two forms. These are control and DDS data, as shown in **Fig. 8**. This data is shifted into UCN5818AF starting from the most significant bit (MSB).

Bit 8, also called the DLATCH bit, distinguishes the two formats. Serial format DDS data is represented by a high DLATCH bit and control data is indicated by a low DLATCH bit.

Corresponding to the serial data format, UCN5818AF's parallel outputs are classified into respective control and DDS data lines, **Tables 1 and 2**. Output OUT 10 to OUT 25 of the serial-to-parallel chip are used for 16-bit DDS data while OUT 1 to OUT5 are used for control signals.

The DDS data has to be present when control signals are sent to the AD7008 chips. Therefore, the UCN5818AF's outputs can only be used for sending DDS data or control signals at any one time. This is accomplished by using an external data latching circuit.

Output OUT 9 corresponds to the DLATCH bit of the serial data. It is used for enabling the data latching circuit each time DDS data are sent and disabling it whenever control signals are sent. Once the DDS data is latched by the data latching circuit, the UCN5818AF outputs are then used for sending control signals to either one or both AD7008s.

Data latching details

Two 74HCT273 8-bit latches combine to form the 16-bit data latch. The latches are enabled only when OUT 9 (DLATCH) is high and DDS data will be latched for every subsequent clock pulse produced by DTR.

Serial interface line DTR is also used as Clock 1 for shifting serial data into UCN5818AF. For every clock pulse, a new serial data bit is

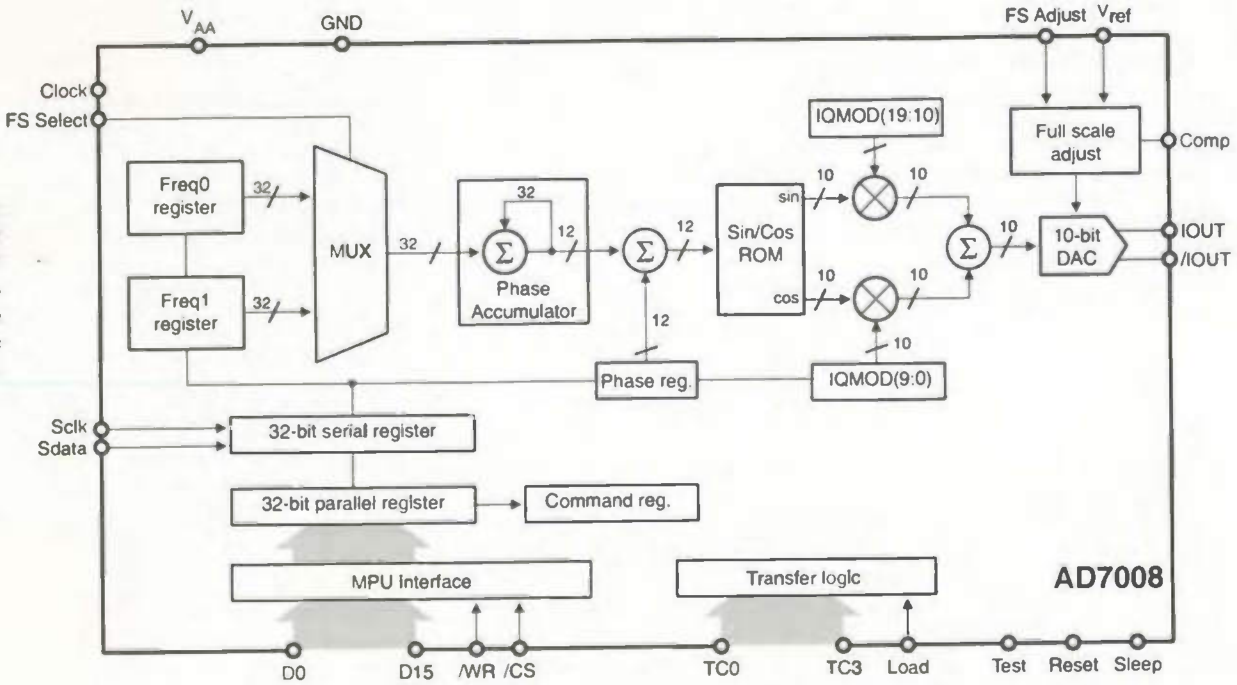
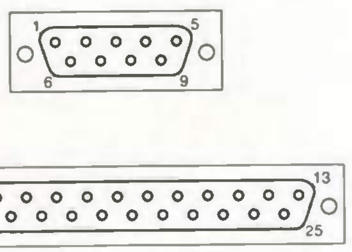
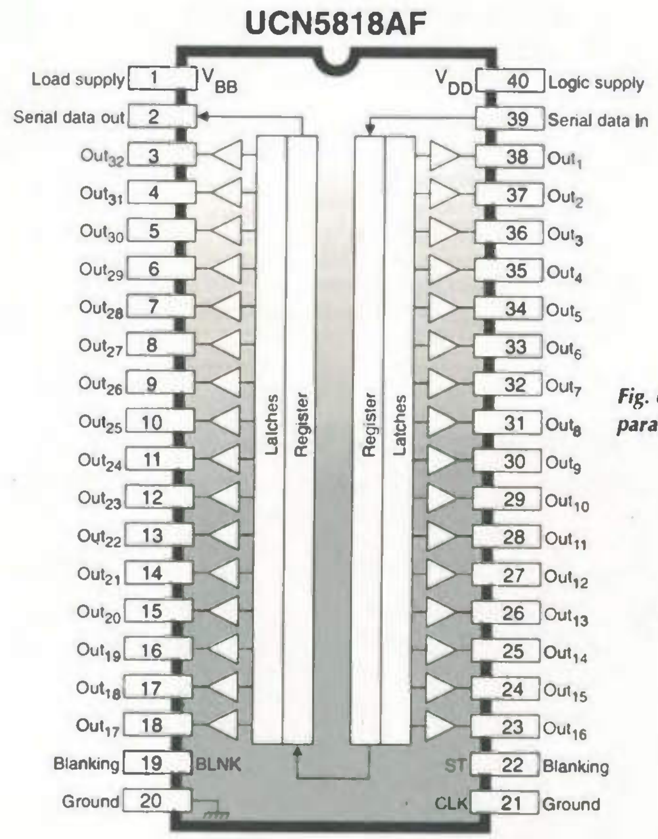


Fig. 4. Block diagram of Analog Devices' AD7008 synthesiser chip – two of which are used in this design.



No. of pins 25	9	Name	Direction (for PCs)	Description
1		Pro		Protective ground
2	3	TD	Output	Transmit data
3	2	RD	Input	Receive data
4	7	TRS	output	Request to send
5	8	CTS	input	Clear to send
6	6	DSR	Input	Data set ready
7	5	GND		Signal ground (common)
8	1	DCD	Input	Data carrier detected
20	4	DTR	Output	Data treminal ready
22	9	RI	Input	Ring indicator
23		DSRD	I/O	Data signal rate detector

Fig. 5. Pin layout and functions of the RS232 connectors on PCs, a) 9-pin male socket. b) 25-pin male socket.



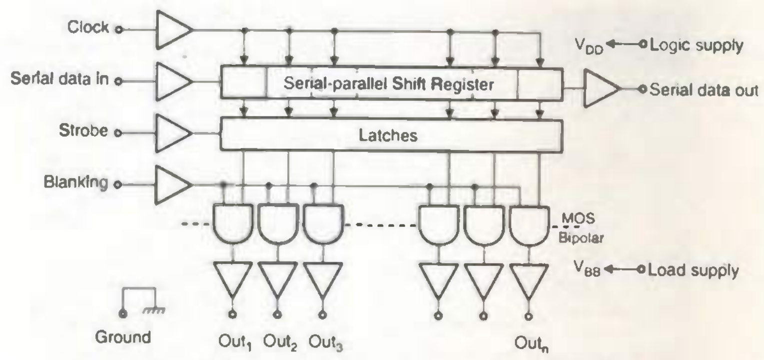


Fig. 6. Pin-out and internal block diagram of the UCN5818AF serial-to-parallel converter.

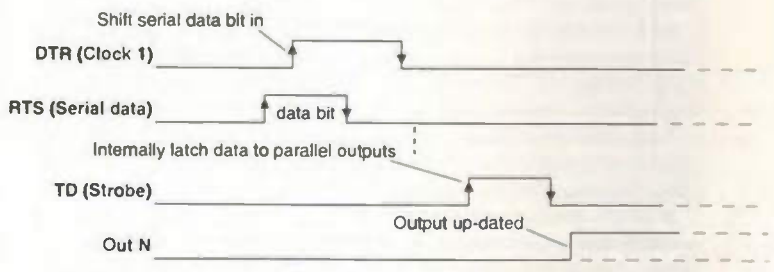


Fig. 7. Timing sequence of the serial-to-parallel converter chip.

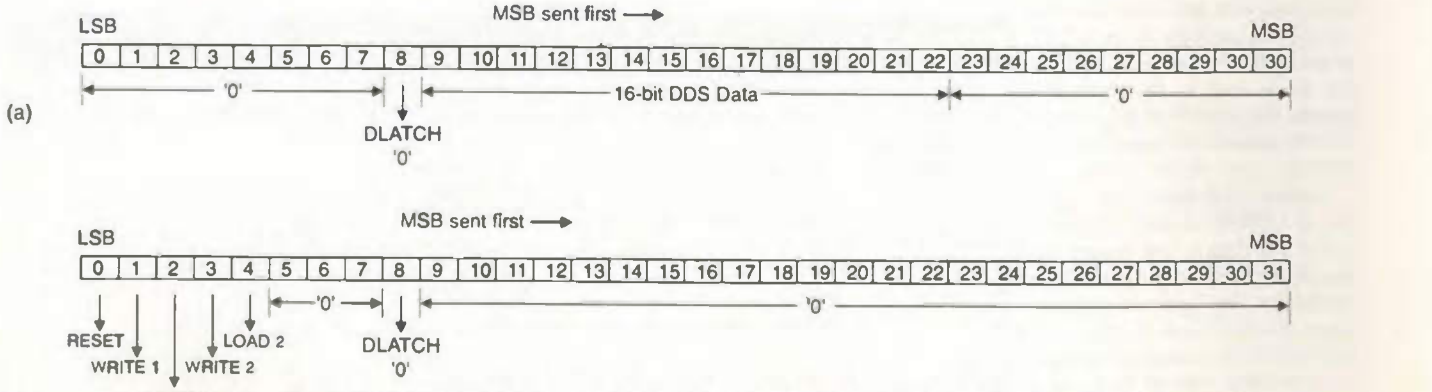


Fig. 8. Format of serial data sent to the serial-to-parallel converter a) for DDS data and b), for control.

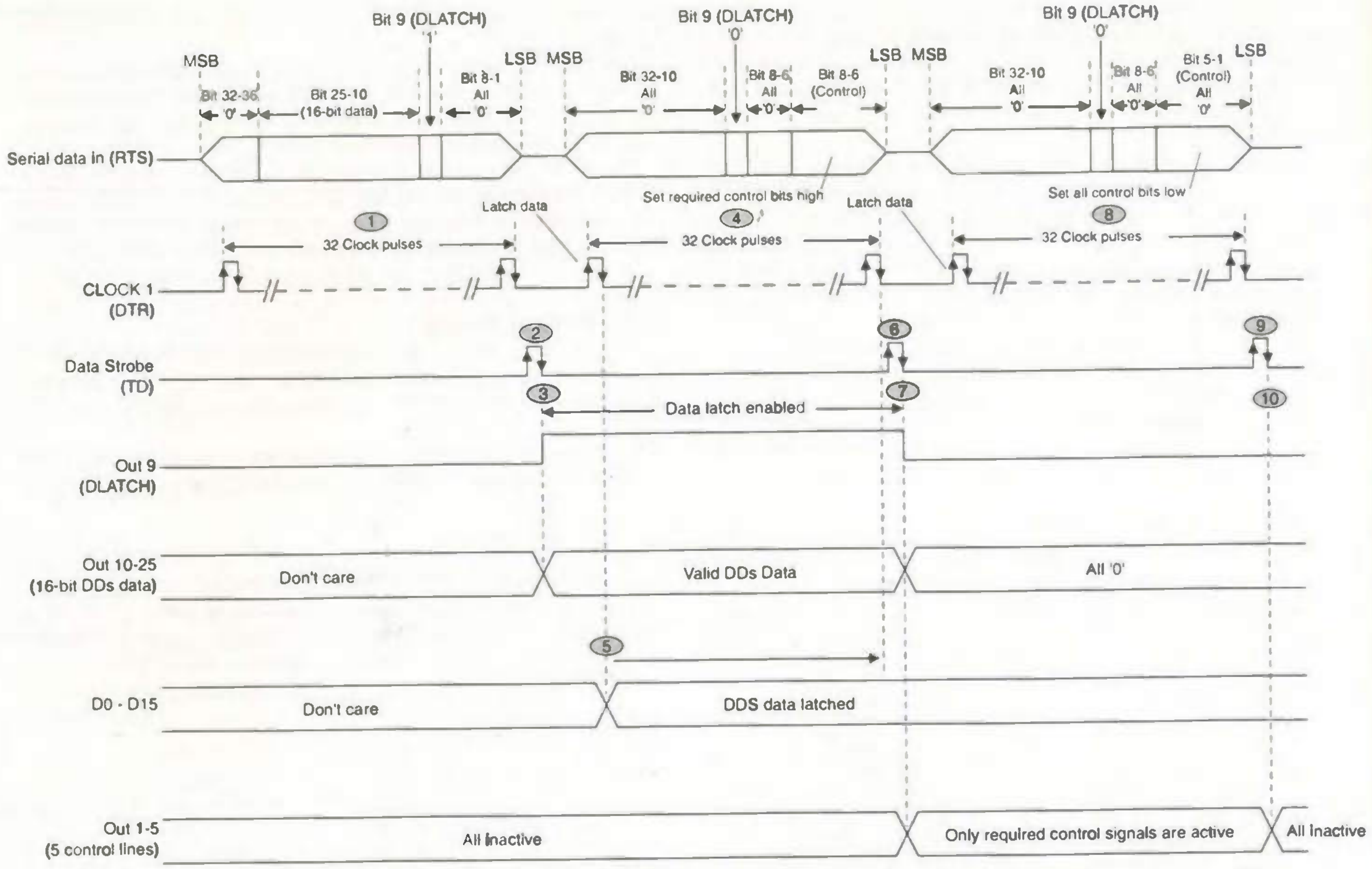


Fig. 9. Sequence of a WRITE or LOAD operation to the generator.

shifted into UCN5818AF but will not be sent to its parallel output until STROBE (TD) is high. At the same time, if OUT 9 is high, data will be latched into the 16-bit data latch.

All DDS data are sent serially to UCN5818AF with a high DLATCH bit. Once converted to parallel, the DDS data is presented to the 16-bit latch through OUT10 to OUT25 with OUT9 high. The data is then latched when control serial data is next shifted into UCN5818AF.

As control serial data is associated with a low DLATCH bit, the parallel form disables the latch and control signals are sent to AD7008 chips without affecting the data.

With this arrangement, the UCN5818A's parallel outputs are used for sending data to and controlling the AD7008 chips independently at any time.

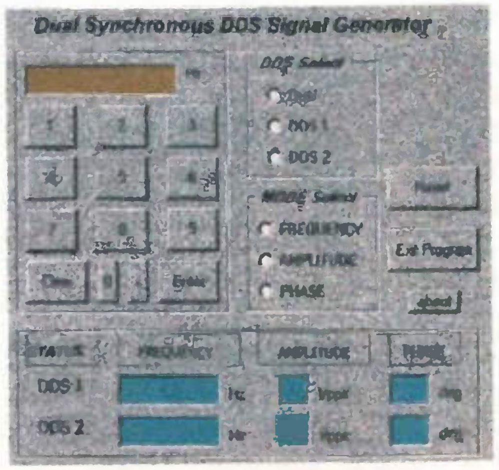


Fig. 10. This user interface has been developed by the authors to help get the design up and running.

Implementing the DDS chips

Control signals from OUT1 to OUT5 require that control serial data be sent twice. The required control signals are activated by first sending serial data with the corresponding control bits set high. A second set of serial data is then sent with all bits set to low to deactivate the signals. Each AD7008 has its own WRITE and LOAD lines, but they share the same RESET signal.

Loading DDS data to the registers of AD7008 requires two operations. The data is first written into the chip's parallel register and then loaded to the selected register. Since the DDS data is latched to both AD7008s, it can be loaded to an individual chip or both chips simultaneously, depending on the settings of control bits in the control serial data sent. To load the data into a particular chip, only that chip's control signals are activated. If both chips are to be loaded at once, their control signals are activated simultaneously.

On the synthesiser chips, TC_0 to TC_3 represent the transfer control address bus connected to D_{12} to D_{15} of the 16-bit latch respectively. The transfer control address is sent in the same way as DDS data.

Only the four most-significant bits contain the address of the destination register into which the written data is to be loaded. This address has to be present before a load can be executed.

Both WRITE and LOAD operations are similar. The exception is that the DDS data sent in a LOAD operation is used as a transfer control

Table 1

Parallel outputs from the serial-to-parallel converter perform the following functions, needed to control the DDS chips.

Serial data bit	Parallel output	Control lines	Functions
0 (LSB)	OUT 1	RESET	RESET data latching circuit & both DDSs
1	OUT 2	WRITE 1	Write to DDS 1
2	OUT 3	LOAD 1	Load to DDS 1
3	OUT 4	WRITE 2	Write to DDS 2
4	OUT 5	LOAD 2	Load to DDS 2
5	OUT 6	Spare	Not used
6	OUT 7	Spare	Not used
7	OUT 8	Spare	Not used
8	OUT 9	DLATCH	Enable data latching circuit

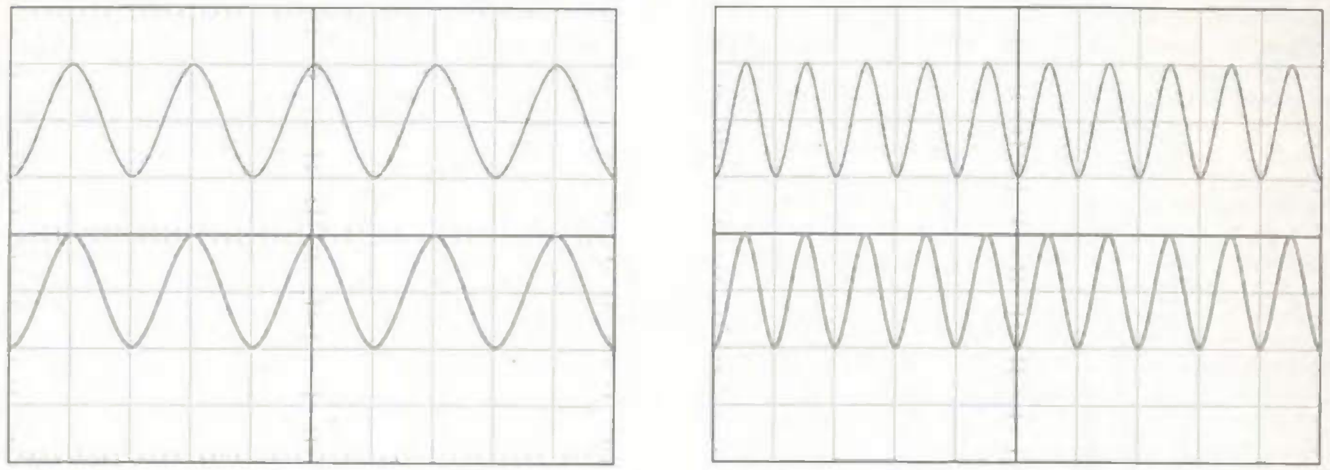


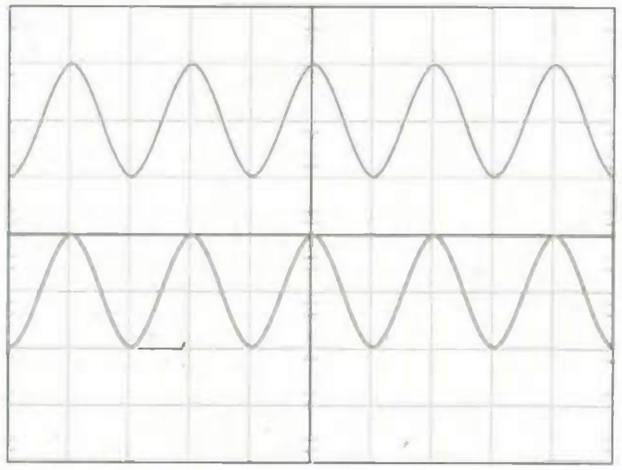
Fig. 11. Signals generated in 'Dual' mode with fixed amplitude and phase of 1V pk-pk and 0° respectively, and varying frequency of 50kHz, left, and 1MHz, right.

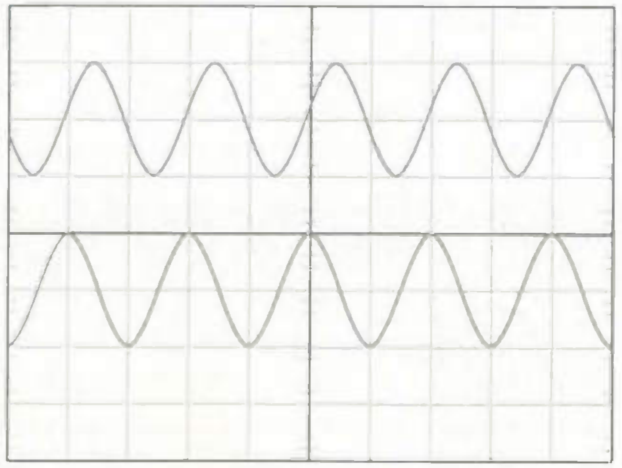
Table 2

Data outputs from the serial-to-parallel converter to the synthesiser chips.

Serial data bit Functions*	parallel output	Data lines	
9	OUT 10	D0	D0
10	OUT 11	D1	D1
11	OUT 12	D2	D2
12	OUT 13	D3	D3
13	OUT 14	D4	D4
14	OUT 15	D5	D5
15	OUT 16	D6	D6
16	OUT 17	D7	D7
17	OUT 18	D8	D8
18	OUT 19	D9	D9
19	OUT 20	D10	D10
20	OUT 21	D11	D11
21	OUT 22	D12	D12
22	OUT 23	D13	D13
23	OUT 24	D14	D14
24	OUT 25	D15	D15
25–31 (MSB)	OUT 26–32	unused	–

*DDS data bits to both synthesiser chips





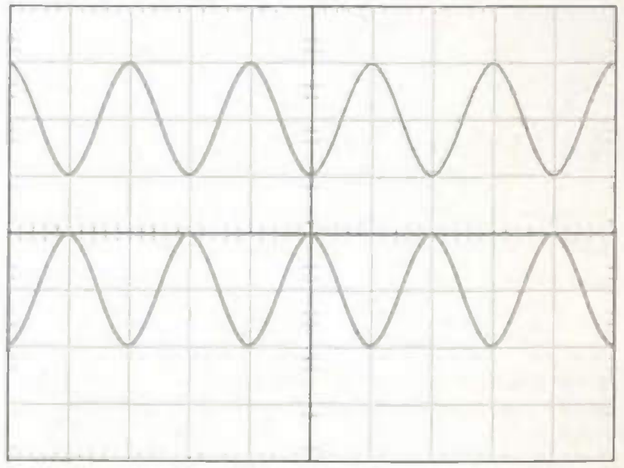


Fig. 12. Signals 1 and 2 were first generated identically and in-phase, top left, in 'Dual' mode with amplitude, frequency and phase of 1V pk-pk, 10kHz and 0° respectively. Next, the DDS2 signal was phase shifted relative to DDS1 for out of phase values of 90°, bottom left, and 180°, bottom right.

address to both chips. It is not used for writing into the parallel registers. **Figure 9** is a sequence diagram of a WRITE or LOAD operation.

With this design, each WRITE or LOAD operation to the DDS chips requires three 32-bit serial data transfer cycles. One sends 16-bit DDS data and two activate and deactivate the control signals.

After powering up or resetting the generator, both AD7008 chips are synchronously configured by setting their command register bits as in **Table 3**. Unlike other registers, the command register only requires a WRITE and a LOAD operation.

The 16-bit DDS data are written into the parallel register, but only the least-significant 4 bits (D_0 to D_3) are loaded into their command registers for chip configuration.

System overview

This generator is capable of producing two signals of up to 16MHz with maximum amplitude of 1V pk-pk. However, the signals can be externally amplified and filtered if required. It has two modes of operation determined by the control serial data received from the PC.

Table 3

DDS configurations used in the system.

Command register bits	Configuration data bit	Setting	Mode
CR0	D0	1	16-bit parallel loading enabled
CR1	D1	0	Normal operation
CR2	D2	1	Amplitude modulation enabled
CR3	D3	0	Synchroniser logic enabled

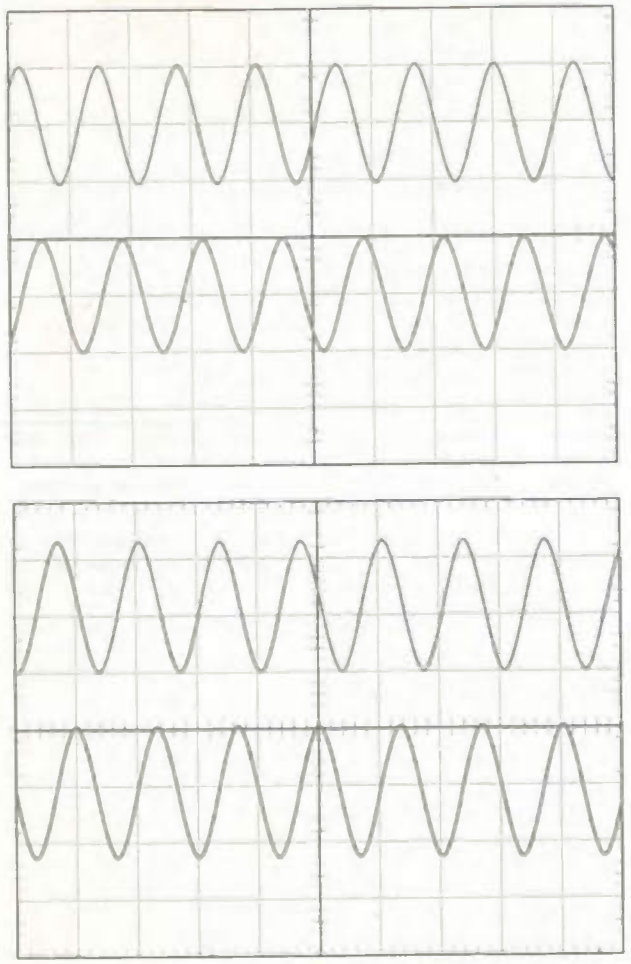


Fig. 13. Top: dual 30kHz, 1V pk-pk signals with DDS2 signal shifted 130°. Bottom: dual 150kHz signals, 0.5V pk-pk signals with DDS1 shifted by 80°.

In 'individual mode', DDS data are loaded into the selected AD7008 and only the respective output signal is affected. In 'dual synchronous DDS mode', the data is loaded into both AD7008 synchronously and both output signals are affected in similar ways at the same time.

Signal frequency is determined by the value loaded into FREQ register. Both amplitude and phase are determined by that loaded into the IQMOD register.

The key feature of this signal generator is its capability to shift the generated signals over the range of 0° to 360°. Two identical signals of zero phase difference are generated in dual mode, and the desired phase difference is then set by phase shifting either signal in individual DDS mode. This allows the phase of either signal to be shifted accurately with reference to the other as desired.

User interfacing

We have developed a user interface and control program using Visual Basic 6. It allows the user to operate the generator from a PC. The developed user interface is shown in **Fig. 10**.

There's a numerical keypad in the graphical display. The user can enter the values for the signal's frequency, amplitude and phase and the unit automatically changes to 'Hz', 'Vppk' or 'Deg' respectively, depending on which component is selected.

The 'DDS Select' allows the user to select which AD7008 chips the values are entered to: DDS1, DDS2 or both. By selecting 'Dual', the entered values are loaded to both AD7008s synchronously to generate two identical and in-phase signals. One signal's phase then be shifted with respect to the other by entering the phase value for either DDS1 or DDS2. Each time a frequency, amplitude or phase value is entered, the affected AD7008 chip status is updated respectively.

From frequency, amplitude and phase information entered by the user, the control program uses equation 2 to calculate the ΔPhase value to be loaded into AD7008's FREQ register. It also uses equations 4 and 5 to determine the A and B values to be loaded into IQMOD register. The values are then converted into DDS data and loaded into the respective registers of the select AD7008 to generate the desired signals.

Test results

We have tested the generator using the software just mentioned for different values of frequency, amplitude and phase in various 'DDS Select' modes. The generated signals were examined using an oscilloscope, **Figs. 11, 12, 13 and 14.** Signal 1 was generated by DDS1 and signal 2 by DDS2.

The results show that the signals are generated accurately in amplitude, frequency and phase in relation to the values entered. With this generator, it is now possible to produce fully-adjustable signals in terms of frequency, amplitude and phase.

Future developments

The signal-generator prototype was developed for use with a PC. By replacing the PC with a microcontroller, the generator could be turned into a stand-alone instrument.

In addition, amplifiers and filters could be incorporated to increase the amplitude range of the signals and to remove high frequency harmonics. ■

Software availability

While developing this signal generator, the authors produced rudimentary Visual Basic software that provides an interface between the PC and generator. To obtain a copy of the software free of charge, e-mail j.lowe@cumulusmedia.co.uk.

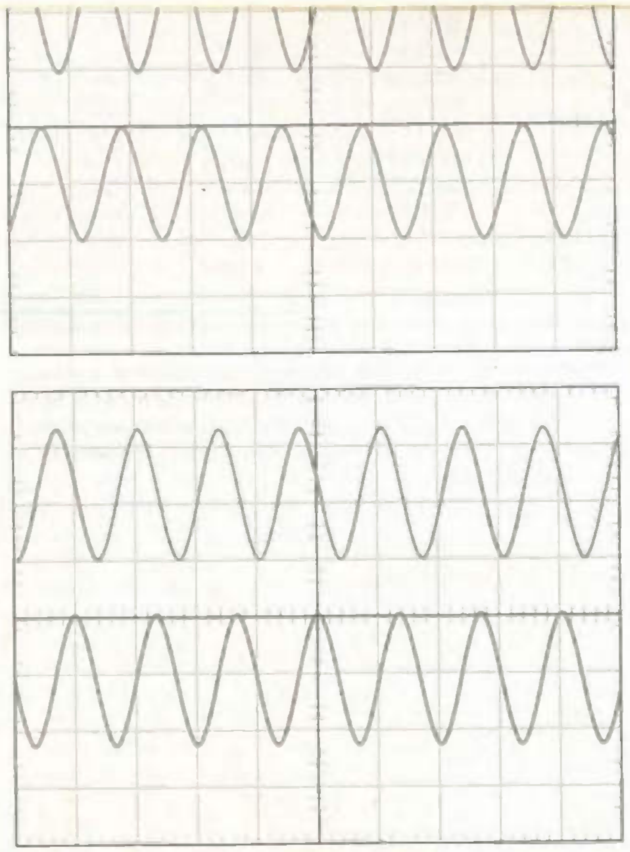


Fig. 13. Top: dual 30kHz, 1V pk-pk signals with DDS2 signal shifted 130°. Bottom: dual 150kHz signals, 0.5V pk-pk signals with DDS1 shifted by 80°.

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Designing with DSP

Having covered the hardware needed to implement a simple DSP system in last month's article, Patrick Gaydecki now explains how to produce the software. "The idea that DSP chips are more difficult to program than the average microprocessor is a myth," he argues.

In this third article explaining how to implement digital signal processors, or DSPs, I will be discussing the DSP56000's language. This language is used by all members of the family including the DSP56002. I will also explain the steps involved in coding, simulating and executing programs for this DSP device.

To give the subject a hands-on flavour, I will be presenting a very simple program to emphasise the link between the language and the hardware sub-systems of the DSP56002. Finally, I will be exploring the way in which system interrupts are implemented, as interrupts are crucial to the way the device operates in many environments.

Knowing the language set enables the system described in last month's article to perform some truly impressive real-time processing of live audio signals. Before you begin though, it is worth bearing in mind three points.

Firstly, the *results* obtained from certain operations in this language might initially appear surprising and counter-intuitive in comparison to those from general-purpose microprocessors. This is because the language involves fixed-point, fractional arithmetic.

Secondly, in common with other modern DSP devices of its kind, this processor has an extraordinarily wide-ranging syntax. There is only enough space to cover part of its functionality here.

Thirdly, it is a myth that the DSP device is difficult to program; it is quite straightforward, and the basics given below provide a good springboard to explore the full breadth of this device's capability.

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Table 1. Fractional format data representation in DSP56000 language.

Number range	−1	-2^{-23}	+0	$+1-2^{-23}$
As power	2^{23}	$2^{24}-1$	0	$2^{23}-1$
Hex number	800000_{16}	$FFFFFF_{16}$	000000_{16}	$7FFFFF_{16}$

The arithmetic and logic unit

The ALU has ten data registers. Data input register X is subdivided into X0 and X1 while data input register Y is subdivided into Y0 and Y1. In addition, the data accumulator register A is subdivided into A0, A1 and A2 while the data accumulator register B is subdivided into B0, B1 and B2, making ten registers in all.

The X and Y registers are each 48 bits wide. Hence X0, X1, Y0 and Y1 are each 24 bits wide. Both X and Y can be addressed as 48 bit or 24 bit registers.

Both A and B registers are each 56 bits wide. Registers A0, A1, B0 and B1 are each 24 bits wide while registers A2 and B2 are each 8 bits wide. Both A and B can be addressed as a whole or according to their subdivisions, as in the case for X and Y.

The DSP56000 family represents its data using two's complement fractional format, in which only numbers (operands) within the range ±1 may be represented. The resolution depends on the word length.

If the operand is represented as a word, it has 24-bit resolution – or one part in16777216. If it is represented as a long word, it has 48-bit resolution. **Table 1** shows how the DSP56002 uses 2's complement arithmetic to represent words. The same scheme may be applied to long words.

From the Table 1, you can see that positive numbers from 0 to $1-2^{-23}$ are represented in hexadecimal from 0 to $7FFFFF_{16}$, and negative numbers from −1 to -2^{-23} are represented from 800000_{16} down to $FFFFFF_{16}$ respectively. Note that in word format, six hexadecimal symbols are required, since each symbol requires 4 bits and a word is 24 bits wide.

Single-statement syntax

In common with other DSP languages, the DSP56000 assembly code allows certain multiple statements to be included within a single instruction. Due to the nature of its architecture, it can execute these in parallel.

In this section, I will deal only with single-statement instructions. These follow the general syntax,

```
<OPCODE> <SOURCE
  OPERAND>,<DESTINATION
  OPERAND>      ; comment
```

A space is required between the opcode – i.e. the instruction – and the source operand, but no spaces are allowed between operands. Comments are preceded by a semicolon.

Within the instruction, a '#' symbol preceding a number (operand) implies that it is a literal value. If this is omitted, the number is a memory reference. A '$' symbol preceding a number means that it is expressed in hexadecimal form. If it is omitted, the format is decimal.

It is *very* important to remember that since the processor uses fractional arithmetic, it loads literal values into its arithmetic and logic unit registers with left justification. Hence the command,

```
MOVE #$20,X0
```

places the value 200000_{16} into X0, not 000020_{16}. The best way of thinking about this is to assume a decimal point to the left of the register. If you want to force a number to be right justified, you use the right-caret thus:

```
MOVE #>$2,X0.
```

This would load X0 with 000002_{16}. One of the nice things about DSP56000 assembly language is that you can use decimal numbers directly. Thus the instruction,

```
MOVE #0.25,X0
```

achieves exactly the same result as,

```
MOVE #>$200000,x0
```

Those of you who are new to this language may well consider a processor that can only handle numbers less than or equal to ±1 to be extraordinarily limited. In fact, nothing could be further from the truth; it is simply a question of scaling. Its all-important properties are the resolution and speed at which it can conduct multiplications, additions and shifts.

The above rule for left justification does not apply when performing register-to-register moves, where data position is preserved. Neither is it the case for moving literals to peripheral (control) registers as opposed to ALU registers. This is because these registers perform control rather than arithmetic functions.

In addition to the data registers, the address-generation unit (AGU) has 24 16-bit address registers that are used to hold the addresses of data referenced by the instructions. These are:

```
Rn,    n=0...7 (address)
Nn,    n=0...7 (offset)
Mn,    n=0...7 (modifier)
```

Each R register is associated with an N and M register according to its subscript *n*. Register R is used to locate operands in memory, while N provides an offset to an R address and M specifies the kind of addressing that is being performed.

Three types of addressing are possible: linear (default), reverse carry (for FFTs) and modulo, for accessing circular buffers. Modulo addressing is used extensively for real-time filtering operations.

Addressing modes

A DSP56000 instruction, or op-code, consists of one or two 24-bit words. There are fourteen addressing modes possible, and some of the more important ones are summarised in Tables 2 and 3.

A number of points concerning Table 2 is worth noting. First, with simple immediate addressing (2nd listed), only the sub-register specified, in this example A0, will have its contents changed. Secondly, with immediate to 56-bit (3rd listed), the data are left justified as discussed above. Furthermore, in this case a negative number has been loaded. This is because it is greater than

Table 2. Summary of key DSP56000 'register-direct' addressing modes.

Sub-type	Instruction	Operand before	Operand after
Ordinary	MOVE X1,A0	X1=$000123	A0=$000123
Immediate	MOVE #$818181,A0		A0=$818181 A1, A2 unchanged
Immediate to 56 bit	MOVE #$818181,A		A=$FF818181000000
Immediate short	MOVE #$81,A1		A=$00000081000000
Immediate short to A	MOVE #$81,A		A=$FF810000000000
Absolute	MOVE X:$200,A0	X:$200=$123456	A0=$123456
Absolute short	MOVE A1,X:$2	A1=$123456	X:$2=123456
I/O short	MOVEP A1,X:$FFFE	A1=$123456	X:$FFFE=3456

Table 3. Summary of key 'register indirect' addressing modes for the DSP56000.

Subtype	Instruction	Operand before	Operand after
No offset	MOVE B1,Y:(R0)	B1=$123456 R0=$1200	Y:$1200=$123456
Post increment by 1	MOVE B0,Y:(R0)+	B0=$123456 R0=$1200	Y:$1200=$123456 R0=$1201
Post decrement by 1	MOVE B0,Y:(R0)-	B0=$123456 R0=$1200	Y:$1200=$123456 R0=$11FF
Post increment by offset register	MOVE B0,Y:(R3)+N3	B0=$123456 R3=$1200 N3=3	Y:$1200=$123456 R3=$1203
Post decrement by offset register	MOVE B0,Y:(R3)-N3	B0=$123456 R3=$1200 N3=3	Y:$1200=$123456 R3=$11FD
Index + offset register, indirect	MOVE B0,Y:(R3+N3)	B0=$123456 R3=$1200 N3=3	Y:$1203=$123456 R3=$1200

$7FFFFFFFFFFF_{16}$, and so register A2 sets all bits to 1, i.e. $A2=FF_{16}$ to indicate this.

Finally, with I/O short addressing (bottom row), a peripheral register is being addressed. Several DSP56002 peripheral registers are 16-bit or smaller, and are not arithmetic. Hence the data are right justified.

There's a number of other variants of these addressing modes, but the ones listed above are certainly the most commonly used.

Instructions and parallel operations

There are 62 instructions in the DSP56000's language. These are divided into the following groups: move, arithmetic, logical, bit manipulation and program control.

I have already mentioned single statement instructions. However, there are thirty instructions that can specify one or two parallel data moves in one instruction cycle in parallel with the opcode.

An instruction with the same source transferring to several destinations is permitted. However, an instruction with several sources transferring to the same destination is not. **Table 4** below is by no means an exhaustive list, but it does provide details on some of the more frequently used instructions.

For further information, take a look at the DSP56000's user manuals, details of which are given later.

The really impressive thing about these instructions is that they are carried out in a single instruction cycle. Look, for instance, at the final instruction, i.e.

```
MAC X0,Y0,A X:(R0)+,X0 Y:(R4)+,Y0
```

This is performing a multiplication, a summation, two data moves and two register updates. For a DSP56002 clocked at 60MHz, this entire list of actions takes a mere 33ns.

Developing a program

Once you have written a DSP56000 routine using any standard text editor, it may be assembled using an assembler program supplied by Motorola. Typically, the instruction might be,

```
asm56000 -a -b -l myprog
```

where the '–a' option specifies the code is to be generated using absolute addresses, the '–b' option specifies the generation of a .CLD Motorola object file, and the '–l' option specifies that the assembler is to generate an assembled program listing.

If the assembled program contains no syntax errors, it may then be downloaded to the target processor and run. However, just because a program assembles correctly, it does not mean that it will do what it is supposed to. Hence it is often a good idea to load the program into a DSP56002 simulator. This allows you to step through the code, instruction by instruction.

The simulator shows how the memory and register contents change as each instruction is executed. **Figure 1** below shows a display from the simulator provided by Motorola. If the program performs correctly on the simulator, you can be fairly sure that it will also perform correctly in the target system.

Simple square-wave oscillator

Take a look at the following code fragment:

```
        MOVEP #$1FF,X:$FFE3
              ; enable port C all
output
LOOP  MOVEP #$1FF,X:FFFE5
              ; pulse all pins
high
        MOVEP #$0,X:$FFE5
              ; pulse all pins low
JMP LOOP
```

The first line places $1FFF_{16}$ (111111111_2) in the peripheral register located at X:$FFE3. This is the port C data direction register, and this configures all the pins for output.

In the next line, all the output pins are taken high because X:$FFE5 is the location of the port C data register. In the following line, they are returned to zero and then a return is made to pulsing the pin high, with the JMP instruction.

You can calculate very precisely the frequency of this oscillator, since each instruction takes a fixed number of clock cycles. There are three instructions in the loop, i.e. two MOVEP instructions and one JMP instruction. Each takes 4 clock cycles to execute. Hence if the DSP56002 is clocked at 32MHz, the above code will generate a pulse train at 2.66667MHz. The frequency can be altered by including NOP instructions (no operation), which require two clock cycles.

Using interrupts

DSP56002 interrupts may be generated from many sources, for example, the SCI, the SSI, the timer, from a hardware input signal, and so on. When the processor receives an interrupt, it completes its current instruction and places the contents of the program counter (PC) and status register (SR) in an area called the stack (i.e. this represents work in progress). The stack is a reserved block of internal memory, 15 x 32-bits wide.

Now, a jump to a specific address in memory called an *interrupt vector* occurs. In turn, this vector holds an instruction to jump to the address at which the interrupt service routine (ISR) starts. It jumps to the ISR, executes it and after completion, restores the PC and SR, continuing from where it left off within the main program. This sequence of events is shown in **Fig. 2**.

All the DSP56002 interrupt vectors are located in program memory space between P:$0 – P:$3F, and each vector is associated with a specific type of interrupt. Each vector is always only two words in length. For example, the interrupt associated with the SCI resides at P:$14 – P:$15. This is fixed by the design.

It is up to the programmer to ensure something sensible is programmed here if an SCI interrupt occurs. This is usually a jump to subroutine instruction (JSR), followed by its address, for example,

```
JSR mysub
```

Here, *mysub* is the label at the start of the ISR. The ISR must end with the RTI instruction (return from interrupt) if JSR is specified in the interrupt vector. The DSP56002 will only jump to an interrupt vector if that particular interrupt has been enabled. Otherwise, it ignores it and carries on with the main program.

It is not mandatory that a JSR instruction be specified in the interrupt vector. However, in this case the programmer will be required to know the address to which program control should be returned once the ISR is complete.

In summary

In the next and final article in this series, I will be looking at how to program the system to do so real-time processing, and how to use interrupts to communicate with a PC through the serial interface. ■

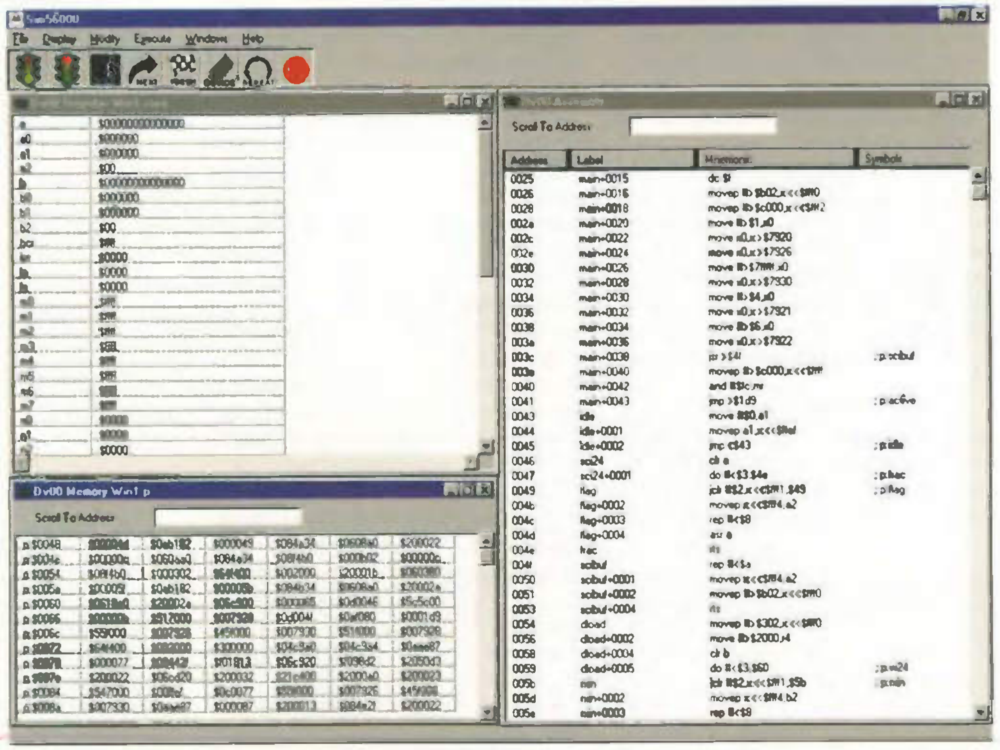


Fig. 1. Screen shot of the Motorola DSP56000 family simulator.

Table 4. Some key multiple-statement instructions in the DSP56000 language set.

Description	Instruction	Operands before	Operands after
Add B to A and move $81 to B0 in parallel, unsigned	ADD B,A #$81,B0	A=$00111111000000 B=$00222222FF0000	A=$00333333FF0000 B=$00222222000081
Add B to A and move $81 to B in parallel, signed	ADD B,A #$81,B	A=$00111111000000 B=$00222222FF0000	A=$00333333FF0000 B=$FF810000000000
Add B to A and move X1 to B in parallel	ADD B,A X1,B	A=$00111111000000 B=$00123456111111 X1=$900000	A=$00234567111111 B=$FF900000000000
Add B to A and update R1 in parallel	ADD B,A (R1)+N1	R1=$1000 N1=$4	R1=$1004
Add A to B and move A to memory in parallel	ADD A,B A,X:$1000	A=$00123456000000 B=$00111111000000	B=$00234567000000 X:$1000=123456
Multiply Y1 and X1, place result in A and move X0 to Y0 in parallel	MPY Y1,X1,A X0,Y0	Y1=$400000 X1=$400000 X0=$123456	A=$00200000000000 Y0=$123456
Multiply-accumulate X0 & Y0 to A, place content of X:(R0) into X0, update R0, place content of Y:(R4) into Y0 & update R4, all in parallel	MAC X0,Y0,A X:(R0)+,X0 Y:(R4)+,Y0	X0=$400000 Y0=$400000 A=$00123456000000 R0=$4 R4=$3 Y:(R4)=$565656 X:(R0)=$232323	X0=$232323 Y0=$565656 A=$00323456000000 R0=$5 R4=$4

Further reading

DSP56002 Digital Signal Processor User's Manual Motorola Inc. Document DSP56002UM/AD (1993)

DSP56000 Digital Signal Processor Family Manual Motorola Inc. Document DSP56KFAMUM/AD (1995)

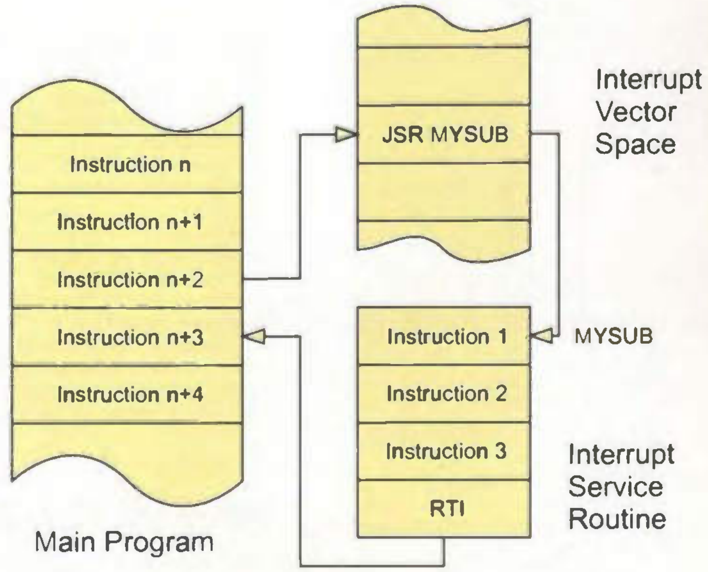


Fig. 2. Sequence of events involved in an interrupt.

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

Intelligent UPS with regulated output

New from Amplicon Liveline are King uninterruptible power supplies with regulated output voltage within ±25 per cent of the input voltage. If this is exceeded, the battery supplies the output voltage. A battery capacity of up to 7Ah is available and can filter spikes from the supply. Batteries provided are hot swappable and user replaceable. Parallel operation is available for larger loads. The units have automatic and manual control facilities, and internal clock for scheduled power up and power down.

Amplicon Liveline
Tel: 01273 570220
www.amplicon.com

Efficient battery-voltage step-up converter

A single-cell IC and discrete combination from Zetex is for DC/DC applications where step-up conversion from a low input voltage is required. The ZXSC100 operates from a NiCd or NiMH cell. Efficiency is above 82 per cent, maintained even as the battery voltage falls below 1V. The circuit will start under full load conditions and maintains regulation down to an input voltage of 0.926V. It generates the control signals for the ZXT14 Super-SOT4 pass transistor, which has a saturation resistance down to 13mΩ. Quiescent current is typically 150μA. The chip has a programmable output voltage, which can be set to anywhere between the input battery voltage and the maximum V_{ce} rating of the pass transistor.

Zetex
Tel: 0161 622 4460
www.zetex.com

Digital multimeter for cat IV safety ratings

Fluke has introduced the 170 series of digital multimeters for category IV safety ratings. They are for front-line industrial, electrical and electronic technicians and engineers for use up to 1kV and 10A AC or DC. Models 175, 177 and 179 are for use from relatively clean bench areas to harsh work environments. The true-rms engine has 0.09 per cent accuracy (177 and 179) with 6000 counts of resolution on a backlit display. The meters provide minimum, maximum and average recording, auto-hold and display hold for troubleshooting. The meters emit an audible tone to signal that a new minimum or maximum reading has been sensed and stored. Frequency, capacitance and resistance ranges are up to 100kHz, 10 000μF and 50MΩ.

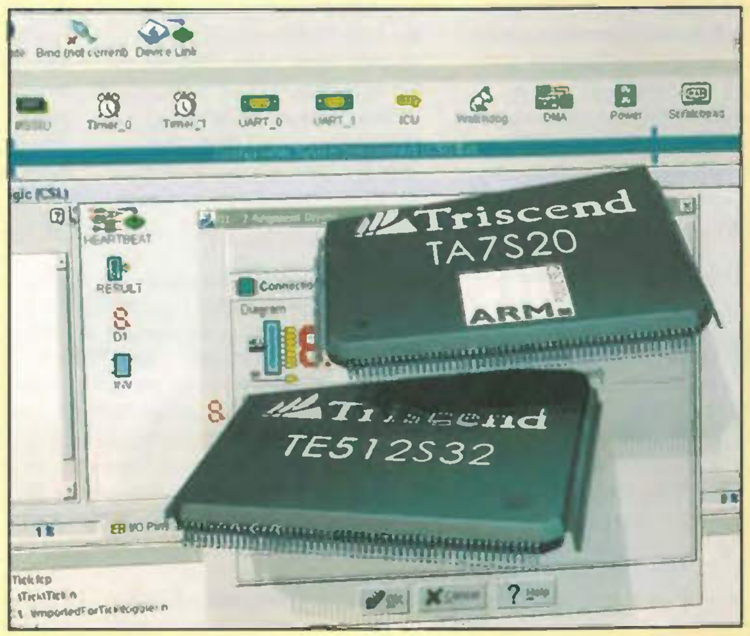
Fluke
Tel: 01207 9420700
www.fluke.com

SM inductors benefit from protection case

Cooper Electronic Technologies has expanded its Coiltronics

IP software is compatible with third party EDA and debug tools

An IP module library and compatibility with third party EDA and debug tools are provided by Triscend's Fastchip software. Windows based, it manages the configurable system-on-chip (CSoC) design process for the firm's 8-bit E5 and 32-bit A7 products. The library of pre-licensed, pre-verified drag-and-drop IP software modules includes a two-wire serial I²C-compliant interface, a quarter VGA LCD controller and a triple DES controller. The modules can be parameterised to meet application needs. The basis to the IPMI protocol, the I²C module is for enclosure management applications, operating in multi-master, slave only or master-only modes. By adjusting settings, the QVCA LCD controller module supports various single drive STN panels and colour and monochrome displays. The triple DES controller module is for POS terminal work. It also supports the Synplify logic synthesis tool from Synplicity. Operability with Wind River's Visionprobe II design tool and Visionclick source level debugger has been changed so logic designers can access and control the entire CSoC .



Amplicon Liveline
Tel: 01273 570220
www.amplicon.com

range of Uni-Pac surface mount inductors with the UP1B, UP2B, UP3B and UP4B with protective cases to reduce the risk of core breakage. They are for use within energy storage applications and DC/DC converters. Operating temperature is –40 to +85°C ambient with an rms current rating of 19.2 to 0.47A. Standard inductance is set at 0.47 to 100μH; modified values for higher applications are available on request.
Cooper
Tel: 01373 472148
www.cooperindustries.com

Solid-state relays

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 500 or 600mW respectively.
Crydom
Tel: 01444 473555
www.crydom.com

Programmable ASSP with a PLD core

Altera has announced the availability of its Mercury programmable ASSP. It integrates the functions of a transceiver ASSP with a PLD core. The clock data recovery (CDR) transceivers within the device eliminate frequency barriers by providing data rates up to 1.25Gbit/s and a CDR bandwidth of up to 45Gbit/s. It has a distributed multiplier capability. Uses include serial backplane, chip-to-chip and line side applications. Versions provide 8 or 18 channels of CDR capability on one device. The device is made on a CMOS process.
Altera
Tel: 01494 602000
www.altera.com

16-bit micro with 32kbyte on-chip SRAM

Hitachi has introduced a 16-bit microcontroller with 32kbyte on-chip SRAM. Versions include the H8S/2329F with 384kbyte on-chip flash (3.3V, 25MHz), the ROMless H8S/2324 (3.3V, 25MHz) and the ROMless H8S/2394 (5V, 20MHz). At 25MHz, these devices provide

about 12Mips Dhrystone if executing from on-chip memory. They are almost pin-compatible with the H8S/2357F, 2350 and 2532. Each device has the same peripheral set including a DMA controller with four channels and a data transfer controller. A bus state controller divides the 16Mbyte memory into eight areas. Parameters can be programmed into each to allow glueless access to external memories and peripherals. A six channel 16-bit timer provides 16 input capture and output compare registers. The chips also have two 8-bit timers, which provide a reduced timer functionality during low activity. A programmable pulse generator has 16 real-time outputs that drive data patterns onto the pins under timer control. This eliminates the jitter that would normally be associated with interrupt-driven operation. Three serial ports provide asynchronous, synchronous and multi-master operation, as well as support for a subset of ISO7816-3, while an eight channel 10-bit ADC provides results within 7μs at 20MHz. The ADC is complemented by a two-channel 8-bit DAC.
Hitachi Europe
Tel: 01628 585163
www.hitachi-eu.com

Wireless platform ups 36 channel density

Blue Wave Systems has launched the Comstruct CPCI/C6800 wireless platform, which will let 3G developers increase base-station transceiver channel density. It is based on the TMS320C6415 and C6416 DSPs from Texas Instruments. The open architecture lets it be closely coupled with a user's Asic technology, letting wireless equipment suppliers develop combined DSP and Asic base-station transceivers. The C6416 includes coprocessors and I/O interfaces targeted at 3G wireless base-stations. On-chip coprocessors for Viterbi and turbo coding eliminate the need for external error correction coprocessors. Data paths on the platform provide up to 6Gbit/s rates to support multiple receivers. Interprocessor communications let developers specify a DSP for chip rate or symbol rate processing to ensure the system is balanced, depending on the ratio of voice and data services offered.
Blue Wave Systems
Tel: 01509 63444
www.bluews.com

16-bit microcontroller operates at 40MHz

STMicroelectronics has presented a 16-bit automotive-grade microcontroller chip with embedded flash memory based on the ST10 core. For applications such as engine control, the chip operates at up to 40MHz. Memory embedded in the ST10F269 is based on single-voltage flash technology operating at 5V.

Reprogramming flash content is controlled by a dedicated erase and program controller that lets flash content be changed through simple commands. This controller also allows new operating modes such as a read-program operation of a second block during the erase suspend of a different block. Each of the seven embedded memory blocks can be separately protected against programming and erasing. Memory read protection prevents attempts to read firmware from the chip. Data retention time is claimed to be 20 years and 100 000 programming cycles over the –40 to +125°C automotive temperature range. Inside are 256kbyte of embedded flash memory and 12kbyte of internal RAM. Current consumption in power-down mode when system clocks are stopped is 100μA. Peripherals provided on chip include standard microcontroller functions, such as interrupt handling, timers, serial channels, analogue-to-digital converter and pulse width modulators. There are two CAN interfaces and two watchdog timers, one to monitor the clock oscillator and one to monitor the program execution.
STMicroelectronics
Tel: 01628 890800
www.st.com

Transceiver uses direct conversion

Hitachi has introduced the HD155141F, or B4, integrated transceiver using direct conversion technology. The receiver incorporates a flexible, configurable analogue IQ interface and an integrated and automated DC offset cancellation circuit. These enable baseband compatibility without the need for special hardware or complex DSP algorithms. The device supports dual-band applications in the GSM900 and either DCS1800 or PCS1900 bands. It is GPRS compliant up to class 12. Support tools, including evaluation boards and application notes, are available. The transceiver incorporates on-chip LNAs. All filtering required after the LNA in the receive path is integrated inside the device. An offset PLL transmitter reduces TX filtering and insertion loss. State machine control is through a standard three-wire bus. It uses a 0.35μm BiCMOS process and is housed in a 56-pin TQFP with a 9 by 9mm footprint and 0.5mm pin pitch.
Hitachi Europe
Tel: 01628 585163
www.hitachi-eu.com

LVDS transmits at high speed

National Semiconductor unveiled several low voltage differential signalling (LVDS) data transmission products at DesignCon 2001. They include devices with on-chip test capability for telecoms and datacomms systems, mobile phone base-stations and Internet infrastructure equipment. LVDS technology is suitable for point-to-point and multipoint data transfer. Also on show were IEEE1149.1 JTAG compliant LVDS devices with boundary scan test access to improve fault coverage by providing test access to the boundaries of a chip, enabling structural and connector test capability. IEEE1149.1 defines a four-wire digital interface to a standard test access port on each compliant device. The DS92LV16 is a single-chip 16:1 LVDS serialiser and 1:16 deserialiser. The 1.28Gbit/s design can be used to construct 16 bidirectional point-to-point links across two pairs between two of the parts. The clocking scheme allows variable input rates from 30 to 80MHz with a ±5 per cent clocking disparity between chips. Built-in local and line loop-back modes facilitate segregation of pre-specified parts of the system by the repetition of signals back to the board (local) or back to the cable or backplane (line).
National Semiconductor
Tel: 00 49 8141 351443
www.crydom.com

Power-management IC for broadband

Intersil has announced a home gateway power management IC for residential broadband communication devices. Part of the Endura family, the range includes single and multiple-output DC/DC power controllers and USB host hot-plug ICs. They are for broadband appliances such as cable modems, set-top boxes and broadband wireless-access subscriber terminals. The ICs combine several PWMs and linear regulators on one chip. The ISL6520 is a single output synchronous buck PWM controller for embedded communications processors and DSP chips. The output voltage is lower than the input voltage.
Intersil
Tel: 01344 350250
www.intersil.com

Microwave analyser from 10MHz to 46GHz

IFR has announced a 46GHz microwave system analyser. The 6845 MSA is a 10MHz to 46GHz spectrum analyser with a full bandwidth, independently controlled tracking generator. It includes a synthesised source and three-channel scalar analyser, and is an aid for the manufacture, installation and maintenance of radio links, satellite links and microwave systems. It includes a 46GHz tracking generator, which acts as transmitter and receiver when testing up and down links. The instrument is aimed at manufacturing, installation and maintenance of point-to-point and point-to-multi-point radios and satellite communications equipment. It is suitable for testing components and assemblies used in radio and military systems, and can measure return loss and fault location on antennas and feeders at mobile communications base-stations as well as verifying radio performance through the modulation spectrum. For insertion and reflection measurements, the tuned input gives a dynamic range of 70dB at 40GHz. The ability to set the spectrum analyser and source frequencies independently in CW and swept modes simplifies the testing of mixers, up and down converters and frequency multipliers and dividers.
IFR Systems
Tel: 01438 772087
www.ifsys.com

Voice coil actuators come in flat format

Densitron has developed voice coil actuators with a flat format for packing in confined places. Applications include printing heads where an array of flat coils is stacked close to emboss coded information in tickets for security. Other applications are

in locks, security equipment and textiles machinery. They can be driven electrically in either direction, so a spring is not required to act in the return direction. Devices can have a profile of less than 1mm and their action is the same as that used in cylindrical actuators.
Densitron Control Systems
Tel:01959 642000
www.densitron.com

Embedded PGA cores for Asics and ASSPs

Actel has introduced its Varicore embedded PGA IP cores for Asic and application specific standard product systems on a chip. The unit is based on technology from Prosys Technology and Gatefield, which Actel bought last year. These blocks have been designed using 0.18μm CMOS SRAM technology. Additional cores are being developed for smaller process geometries.
Actel
Tel: 020 7823 3226
www.actel.com

JTAG emulators for config processors

Arc Cores has announced two JTAG emulators that provide development and test capabilities for configurable processors. The emulators from Corelis use Arc's on-chip debug capabilities to support any processor clock speed at which the target system runs. The Scanice and Netice emulators use the JTAG port on Arc's Tangent processor to provide access to its on-chip debug facilities. Both include the Metaware Windows 95, 98, NT and 2000 compatible Seecode source-level debugger. For JTAG emulation access, the Scanice uses a PCI bus JTag controller that is installed in a PC and the Netice a LAN-based Ethernet JTAG controller. On-chip debug capabilities provide non-intrusive memory access and multi-core debug, and do not need debug firmware. To support real-time debugging, the on-chip logic lets the user examine memory locations without stopping the processor. Debug operations can be achieved with no resident code running on the CPU. Programs and data can be downloaded to any part of the system RAM through the JTAG port without a resident loader program or a ROM emulator.
ARC Cores
Tel: 0208 236 2800
www.arccores.com

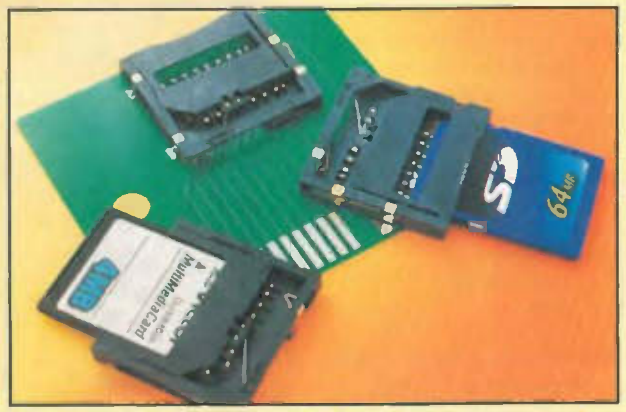
Digital pressure sensor weighs 30g

The UZU3 digital pressure sensor from Matsushita weighs 30g and is for robotics or automated manufacturing systems. A panel-mounting bracket lets multiple sensors be mounted directly on top of each other without a space between adjacent sensors. The device has a two-colour (red and green) display that changes colour if the detected pressure exceeds the user-defined set point. This lets the operator confirm output status at a glance. The display can also be changed from digital to analogue, indicated as a bar display. Response time is 2ms. A chatter prevention function lets the response time be adjusted so insignificant pressure fluctuations can be ignored. There are four output modes - hysteresis, window comparator, automatic sensitivity setting and forced output. Vacuum, positive and compound pressure types are available. The compound type serves pressure requirements between ±100kPa.
Matsushita
Tel: 01908 350700
www.matsushita.co.uk

Rad hardened mosfet at 1000V

International Rectifier has introduced a 1000V radiation-hardened (rad-hard) power Mosfet. The IRHY7G30SE 1000V device is a significant rating improvement on previously available 600V rad-hard Mosfets, said the company. As a result, said the company, it can be used to replace less efficient bipolar transistors in high-voltage applications, and also enables designers to accommodate safe DC-rating conditions without losing functionality. Typical applications include travelling

wave tube amplifiers, which are used to amplify microwave signals in satellite communications systems. The 1000V Mosfet is an enhancement-mode n-channel device, made with IR's proprietary radiation-hardened gate- and field-oxidation process to achieve single-event upset and total ionising dose hardness requirements. The device is single event effect hardened with linear energy transfer of 37MeV/(mg/cm^2) and retains virtually identical electrical performance up to 100Krads (Si) total dose, said the firm.
International Rectifier
Tel: 020 8645 8003
www.irf.com



Connector for secure digital memory cards

Honda has launched a range of low-profile slot connectors for use with the latest secure digital (SD) solid-state, removable memory cards. Designed for use in telecoms and datacoms applications as well as hand-held devices, the connectors incorporate a card retention mechanism for additional security. The connectors accept standard SD memory card sizes and feature a patented 'push-push' eject mechanism, which provides a reliable alternative to snap-in alternatives, said the supplier. The SD card is inserted into and ejected out of the connector body by a simple tactile push action. SD memory cards provide removable storage for up to 64Mbyte of encrypted data and feature mechanical card detection and write protect. Incorporating a three-stop sequential contact design, the series accepts standard 7-position multi-media cards as well as the thicker, 9-position SD cards. An optional card detect feature is also available and SM terminations are housed beneath the connector body to reduce overall size. Furthermore, the low profile design meets published MMC and SD specifications.
Honda Connectors
Tel: 01793 523388
www.hondaconnectors.com

Please quote *Electronics World* when seeking further information

FR-V chip based on VLIW architecture

Fujitsu is releasing its Leo FR-V VLIW chip designed around the company's FR500 core for the European navigation market. The 32-bit, four-way VLIW has media and graphics enhancements, and is software compatible with the firm's MB93501. It is modular and expandable, with the memory management unit being ARM compatible with Windows CE support. Internal frequency is 266MHz, and the chip has an external flexible 32-bit system bus operating at 66MHz, 1.8V, and a set of generic and automotive specific peripherals. The architecture can execute parallel instructions. The compiler rather than the processor is responsible for guaranteeing simultaneous issuing of each packet. There are six execution units on the processor – two integer units, two floating point units and two media execution units, all running in parallel. Its compiler decides which unit is going to carry out a command using VLIW. The CPU core provides a peak performance of 1064Mips, 1.064Gflops and 4256Mops at 266MHz, integer operation, floating-point and media.

Fujitsu Microelectronics
Tel:01628 504600
www.fujitsu-fme.com

32.768kHz SM crystal is stable to ±20ppm

The ACT135 is a surface mount 32.768kHz crystal, from Advanced Crystal Technologies. It has stability to ±20ppm and operates from –40 to +85°C. Applications include portable equipment for industrial, consumer and white goods markets. Measuring 6.8 by 3.0 by 1.7mm, it can also be used in personal equipment such as watches and other timing or data logging instruments. Ageing stability is 13ppm at 25°C (±3°C), and shock resistance is ±5ppm maximum. It is suitable for automatic placement from tape and reel.

ACT
Tel: 0118 979 1238
www.actcrystals.com

Miniature coaxial connectors on the edge

Developed for those PCB applications where space is at a premium, Radiall said its MC-Card can be used in either edge card or on board applications (i.e. PCMCIA or ISA cards). The connector operates from DC to 8GHz with an impedance of 50Ω. Specified for over 5000 matings, it features snap-on mating for security of the connection. A switch version is available and various interface lengths can be produced to allow products not to mate with the standard connector, in conformance to Section 15.203 of FCC standard requirements.

Radiall
Tel:020 8997 8880
www.radiall.co.uk

SM electronics are hot hat devices

Available from Nichicon (Europe) is the UN series of bipolar, "Top Hat" surface mount electrolytic capacitors. Available in diameters: 12.5, 16,18 and 20mm, the capacitors operates across a temperature of –55 to 105°C. Suitable for automatic insertion using carrier tape and tray the series has a working voltage of 6.3 to 100V and capacitance 22 to 3300μF 20 per cent tolerance. Load life of the SM electrolytic capacitors is 64 000 hours at 55°C. Allowable ripple is 830mA max (at 105°C) depending on capacitance and working voltage. Size range is 12.5(D) by 13.5(L) mm by 20 by 21.5mm.

Nichicon
Tel:01276 685393
www.nichicon.co.uk

32-bit Risc MCU with on-chip mask ROM

Epson has developed a 32-bit Risc MCU integrating peripheral circuits with built-in ROM based on its own E0C33209. Two models, the E0C33264 and the E0C332129, include a built-in high-speed MAC (multiplication & accumulation) instruction, a multiplexed A/D converter, DMA, HDMA, 4 channels, SIO and various timers. These functions make them suitable for classic DSP applications especially for software codec and signal processing (voice recognition and text-to-speech), said the company.

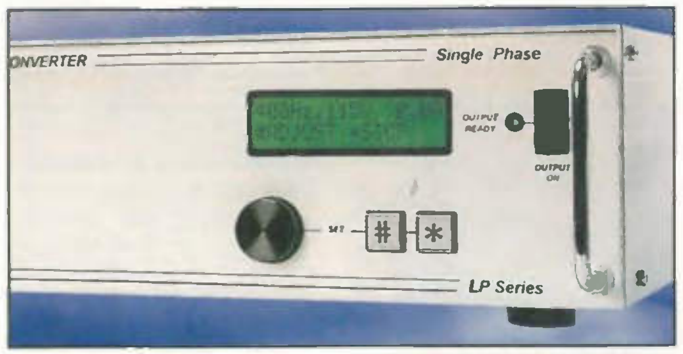
Epson Electronics
Tel:00 49 89 14005 227
www.epson-electronics.de

Programmable caps to replace inductors

Maxim Integrated Products has introduced its first electronically programmable capacitor with high self-resonant frequency and 32 discrete capacitance steps. Housed in an SC70 package (2.2mm x 2.4mm), the MAX1474 is designed to replace mechanical inductors in common tuning circuits.

It can be set to any one of 32 discrete capacitance steps ranging from 6.26pF to 12.74pF in 0.2pF increments. Application circuits can be used to increase the capacitance range, increase capacitance steps, or to multiply the Q-value to fit a particular application. Because the capacitors are actually implemented in silicon, there is virtually no drift over time, said the firm and the temperature drift coefficient is less than 33ppm/°C. A Q of 50 is possible. Two pins are used to program the MAX1474. Capacitance values are selected by raising the enable pin and sending the device a discrete number of pulses corresponding to the desired capacitance value. It does not require a clock and will retain the programmed capacitance value for as long as power is supplied. If the system is dynamic and has the ability to retune itself, the capacitance value can easily be changed at any time by simply raising the enable pin and sending the desired number of pulses. Applications include post-trim of low-cost regenerative stages,

tuneable RF stages, garage door openers, keyless entry, industrial wireless control, and precision trimming of capacitive-based sensors.
Maxim
Tel: 0118 930 3388
www.maxim.com



Frequency converters have variable outputs

Magnus Power has launched the redesigned LP series of frequency converters, which is offered with either single or three phase output. The single phase range consists of the LP300, 600 and 1000 and is rated at 300, 600 and 1kVA respectively. Each model has an output frequency that is variable from 45Hz to 440Hz. Output voltage varies from zero to maximum in two ranges, usually 0-135V, and 0-270V. Output current capability is doubled on the lower range.
Magnus Power
Tel: 0116 2672856
www.magnus-sales@kvaerner.com

TFT-LCD module with A/D conversion board

Customised versions are available on request. Semicom is offering a 15in TFT-LCD module that incorporates the A/D conversion board. Eliminating the need for a discrete A/D circuit and the associated interfaces to the LCD module, Hyundai's HM15X11 and HM15X12 LCD modules support XGA resolution, 16.7 million colours and a refresh rate of up to 75Hz, without external frame memory. Features also include automatic detection of changes in input display mode and frame rate and power management capabilities that meet DPMS and VESA requirements. The CCFL backlight delivers a typical brightness of 200cd/m^2 (at 6.0mA), whilst an option allows luminance to vary according to the ambient brightness.
Semicon
Tel: 01279 422224
www.semicon.co.uk

NEW PRODUCTS

Please quote *Electronics World* when seeking further information

USB 2.0 controller with serial interface engine

Cypress Semiconductor has added to its recent USB 2.0 controller announcement with a serial interface that provides a 480Mbit/s USB 2.0 connection to a peripheral containing a microcontroller. It does this without draining the controller's resources, said the firm. The EZ-USB SX2 integrates a USB 2.0 transceiver, a PLL, an SIE, a 4kbyte FIFO, and a local bus interface. The integrated SIE takes care of USB housekeeping chores, so the microcontroller can spend its resources on the peripheral application.

Cypress Semiconductor
Tel: 01707 378700
www.cypress.com

10-bit Gunning logic transceiver

Fairchild Semiconductor has added the GTLP10B320 to its range of Gunning logic transistor (GTLP) products. It is a 10-bit universal bus transceiver, with separate LVTTL inputs and outputs, as well as a feedback path for diagnostics, that provides LVTTL to GTLP signal level translation. High-speed backplane operation is a direct result of GTL's reduced output swing, high drive, reduced input threshold levels, and output edge rate control, said the firm. GTLP is the firm's derivative of the Gunning Transistor Logic (GTL) JEDEC standard JESD8-3. All devices are process, voltage and temperature compensated. While many transceiver technologies are designed and specified for lumped-load environments, this device is optimised for distributed-load environments that characterise backplane applications. Designed using BiCMOS process technology, the GTLP devices operate at 3.3V and provide 50mA drive for backplane environments. ■

Fairchild Semiconductor
Tel: 001 888 522 5372

Development environment for Motorola MSC8101 DSP

Green Hills Software has introduced the Multi 2000 integrated development environment (IDE) for Motorola's MSC8101 DSP based on the Starcore SC140 core. The IDE has a C and C++ optimising compiler for the Starcore SC140 EFR benchmarks and provides editing, debugging, profiling and project management capabilities. It automates software development. The source-level debugger, with incremental debug capability that supports process- and system-level debug, is RTOS aware, which lets designers debug and tune applications at a task level. An instruction set simulator lets programmers develop and test code on a PC or workstation without the target hardware. The compiler automatically partitions C programs for execution on the DSP's Mac, ALU and BFU processors. It also provides more than 100 optimisations to boost performance and code efficiency. The compiler increases parallelism by reordering operations within loops, and reduces looping overhead by supporting zero-overhead hardware looping for loops that are nested up to four deep. It collects frequently used data and places it in the lower 16 bits of the address space. This lets the data be accessed using 32-bit instructions, thereby improving VLES packing efficiency.

Green Hills Software
Tel: 01494 429336
www.ghs.com

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 second article looks at the multiplexer's software.

Audio specifications

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

In the first article describing this professional audio-visual router, I looked at the hardware side of the design. Many of the router's functions are carried out in software by code in the router microcontroller's on-chip PROM. The program that forms the user interface to the PC can be used to add further features.

Software, and router configurations used in a typical studio environment, are the subjects this second article.

Software for the router's microcontroller

It is not possible to publish the whole firmware assembler code for the VRS-8x4 video switcher, the ARS-8x2 audio switcher and the PVRS-1 keyboard, due to the limited amount of space available. However, the source code can be obtained via e-mail, as can the object code, which is available free of charge.

Object code for those of you who want to use it without any modifications is given in **Listings 1-3**, presented later in the article.

Code for the video and audio switchers has to be burned into a standard 27256-type 32K PROM.

The code for Atmel's 89C2051 micro can be programmed into the controller's onboard flash with a parallel port programmer. You can buy such a programmer or build one yourself using the documentation, provided on Atmel's web-site (http://www.atmel.com).

Configuration commands

Usually, the user will control the complete routeing system through its own keyboard or via a PC serial link. If you want to build your own interface or control the system through a terminal emulation program (set to 9600/8/1/N) you need to know the serial commands that the firmware can understand. These are listed in **Table 1**.

All commands are ASCII-strings. Each command ends with a carriage return – not with a carriage return plus line feed.

It is worth mentioning how the keyboard retrieves the active configuration after power is applied. Imagine you have a valid start-up

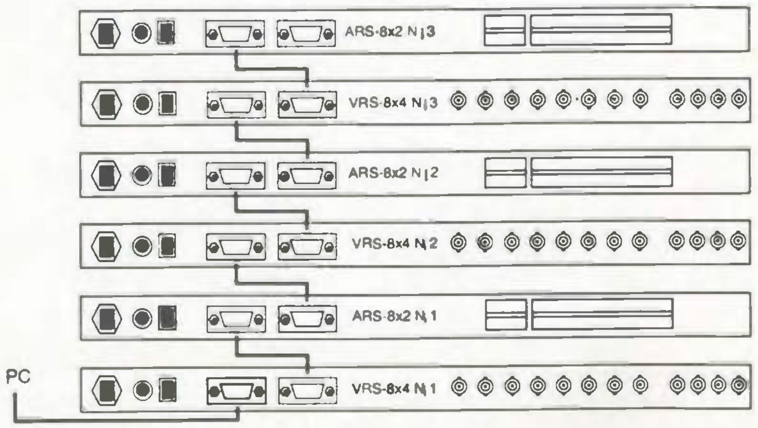


Fig. 1. Three video and audio router pairs daisy chained and controlled via a PC.

Table 1. These are the serial commands to which the audio-visual router system responds.

Audio and video router commands:

Command	Description
Out1InX	Connects Output 1 to Input X (number 1 to 8)
Out2InX	Connects Output 2 to Input X (number 1 to 8)
Out3InX	Connects Output 3 to Input X (number 1 to 8) - ignored by ARS-8x2
Out4InX	Connects Output 4 to Input X (number 1 to 8) - ignored by ARS-8x2
Out1InD	Disable Output 1
Out2InD	Disable Output 2
Out3InD	Disable Output 3 - ignored by ARS-8x2
Out4InD	Disable Output 4 - ignored by ARS-8x2
Out1234InXXXX	All outputs configure command, X is a number between 1 and 8 or D for disable
SOut1234InXXXX	Save StartUp configuration command, X is a number between 1 and 8 or D for disable
DIXXXXDIXXXXDIXXXX...	Configure command for stacked devices. Up to 4 devices can be connected, so the command can contain up to 4 sections. The first section is for the first device in stack, the second for the second device and so on. X is a number between 1 and 8 or D for disable.
SOut1234InXXXXInXXXX...	Save StartUp configuration command for stacked devices. X is a number between 1 and 8 or D for disable.
GetConf	Gets the current configuration command. The switcher returns the configuration for Output 1 and Output 2 in the format Out1InXOut2InX. This command is issued by the PVRS-1 keyboard.
GetConfF	Gets the full current configuration command. The switcher returns the configuration for Output 1, Output 2, Output 3 and Output 4 in the format Out1InXOut2InXOut3InXOut4InX. This command is issued by the PC interface software.

Keyboard commands:

Command	Description
Out1InXOut2InX	Configure the keyboard command - internal registers are set up and the corresponding LEDs are enabled according to the current configuration.

configuration, or you have set up the connection paths through a PC interface and you change the control method from PC to keyboard on the fly. This is possible by the way.

The corresponding LEDs on the video router have to be lit and that the internal configuration registers of the keyboard software need to know the current configuration. This is done in as follows.

After power is applied the keyboard, the microcontroller waits 100ms and then issues the 'GetConf' command. The router acknowledges this command by returning the active configuration in an 'Out1InXOut2InX' format. This format is a command for the video router and is used by the 89C2051 code to set up the internal registers and to light the appropriate LEDs.

An interesting situation arises if you control the 8-by-4 video matrix using an 8-by-2 keyboard. You will not be making full use of the crosspoint switch in this case, but if you do not need the capacity, there is little point in implementing the full keyboard.

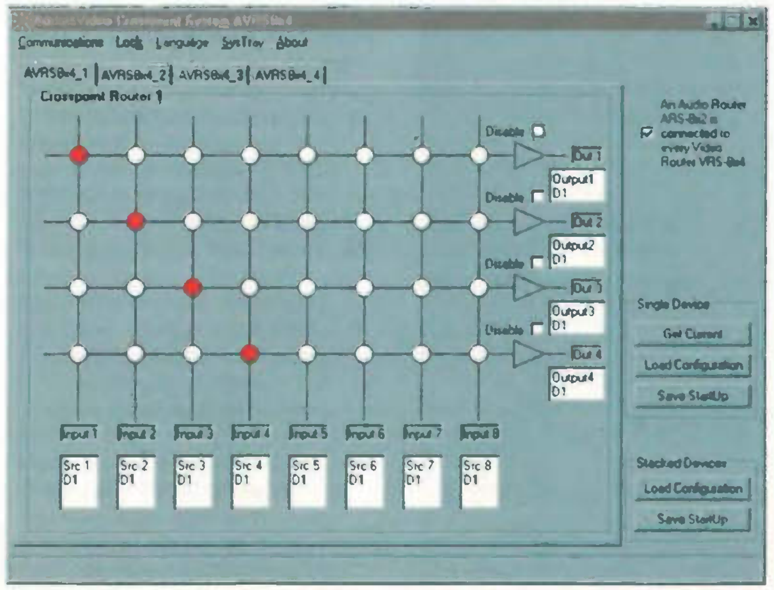


Fig. 2. Screen shot of the author's Windows software for controlling single or multiple audio and video routers.

From the keyboard, you can specify connections only for two outputs – outputs 1 and 2 in this case. However, the keyboard subsystem is designed so that it also configures output 3 to mirror output 1 and output 4 to mirror output 2. These additional outputs can be used for monitoring purposes.

Typical configurations

This section outlines some typical configurations of the routeing system that you might find useful in studio applications.

Audio follows video router, PC control. In the configuration shown in **Fig. 1**, the whole system is controlled via a host PC using appropriate software.

My PC graphical user interface software supports up to four stacked pairs of audio and video routers. Using this software, **Fig. 2**, the 'Audio' check box in the control window of the PC interface (details later) has to be checked for correct operation.

In this configuration the user can control up to four audio and video switcher pairs completely independently. Each device pair responds only to serial commands from the PC that are designated for it. You don't need four serial ports or four PCs to control the four switchers in your studio – you have only one single interface.

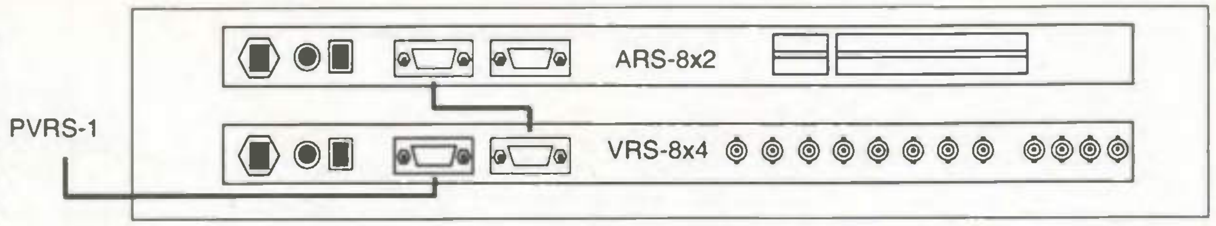


Fig. 3. In this configuration, an ARS-8x2 audio and a VRS-8x4 video router are controlled via a PVRS-1 keyboard.

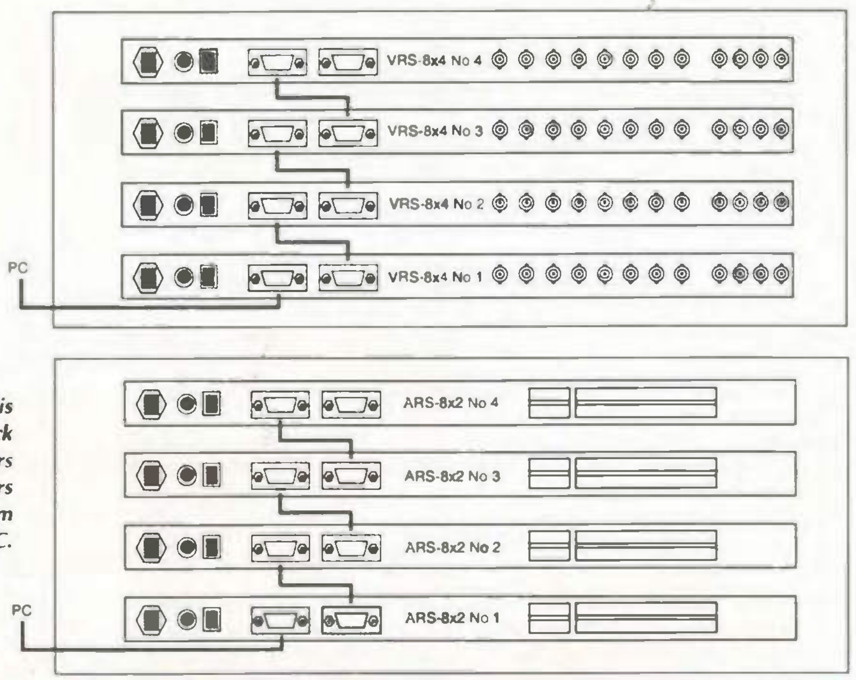


Fig. 4. It is possible to stack just video routers or audio routers and control them via a PC.

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.

Audio follows video router, keyboard control. In the situation depicted in Fig. 3, control over the system is accomplished via the PVRS-1 keyboard. More than one pair of video and audio devices will not usually be stacked in this case, unless identical operation of all stacked pairs is desired. This is a typical situation for a small studio.

Independent video or audio stacks, mostly PC control. In this case, Fig. 4 , all devices in the stack are of the same type. They are either VRS-8x4 video routing systems or ARS-8x2 audio routing systems. Control is usually performed via a host PC.

As in the case of the keyboard control, all devices will be loaded with the same configuration. Up to four devices can be controlled through the GUI application software that I developed.

The 'Audio' check box in the Windows interface is not checked in the depicted configuration – even if you are using only audio routers. This check box is only meaningful when you are using the audio-follows-video configuration.

Creating a graphical user interface

My user interface, Fig. 2, is compiled to run in Windows 95/98 environment. It has the following menus:

- Menu 'Communications' – the user can select the COM port of the PC, connected to the system. The software supports COM1 to COM4.
- Menu 'Lock' – the application software is locked and unlocked with this menu. When locked, the user can change the connection patterns of all stacked devices, but can not download the selected configuration to the target device. In this way, accidental switching and possible errors and program drops are prevented. This function is the same as the hardware lock switch on the PVRS-1 keyboard.
- Menu 'Language' – the application software supports two languages – English and my native language Bulgarian. The source code for the GUI interface can be obtained, details later, so if another language is desired, the modifications can be easily done.
- Menu 'SysTray' – the application is minimised as icon in the system tray and can be activated every time the need arises. This feature is especially useful if the host PC is used for controlling other equipment at the same time. In this case, the operator does not want to be distracted by too many virtual control panels. The software for this menu is written by E. Spencer (elliot@spnc.demon.co.uk) and is public domain.
- Menu 'About' – gives information about the current version of the software.

The software also has the following ancillary control functions:

- 'AVRS8x4_1, AVRS8×4_2, AVRS8×4_3, AVRS8×4_4' tab strip – this strip gives the configuration area the same look as the dividers in a notebook. So multiple configuration planes, one for each stacked AV system pair, are defined in the same area of the window. The maximum number of stacked systems is four, so there are four tab strips.
- 'Crosspoint Router 1, 2, 3, 4' frames – the frames for the stacked devices in sequential numerical order are switched by the tab-strip controls. In every frame, a visual representation of the 8×4 connection matrix is given. For the audio router, only rows 1 and 2 have meaning. Rows 3 and 4 can be specified as connections, but this will not have any effect on the audio router, since it has only two outputs. A valid connection can be performed by clicking on the connection circle. The active circle is highlighted in red. Clicking on the red circle again will disable the activated connection. The selected configuration pattern(s) becomes active after the corresponding 'Load Configuration' buttons are pressed. So the operator has first to specify the configuration pattern and then to load it into the system.
- 'Disable' check box – this box is used to disable the corresponding output of the matrix.
- 'Output-, Input-' text box labels – using these controls, the user

Listing 1. Video switcher firmware.


```
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Software availability

Object code contained in Listings 1-3 is available free of charge. The same code together with the author's Protel PCB layout and Windows 95/98 GUI is available for £15. Source code for the design is also available direct from the author. Its price will depend on whether your application is for commercial or domestic use. E-mail Ms Jackie Lowe at j.lowe@cumulusmedia.co.uk with your requirements.

can specify a custom name or label to every output or input for every of the stacked devices in the corresponding frames. The information is stored in the Windows registry, so it is preserved after the application is quitted and started again. This feature gives the operator the possibility to work not with numbers, but with meaningful sources and destinations labels like 'Studio 1', 'Camera 1', 'Monitor 3', and so on.

- 'An Audio...' check box – this control is used to specify if an audio-follow-video stack is realised or not. It has to be checked if an audio routing system is connected to a video routing system and the user wants to route not only the video signal, but the whole signal path – plus audio – from the source to the specified destination.
- Frame 'Single Device', button 'Get Current' – this control is used to retrieve information about the current active configuration of the first, or master, routeing system connected to the PC. The circles, representing active connections, are represented in red. The status of this button is not affected by the lock menu, because there is no danger that the active configuration in the router can be destroyed. This button is used only if a single video and audio router system is being controlled. Information about slave devices in the stack can not be retrieved in this way.
- Frame 'Single Device', button 'Load Configuration' – the function of this control is to transfer the specified connection pattern to the target device. This function is only used if a single audio-video router system is being controlled. So the connection pattern for the first AV router group only has to be specified.
- Frame 'Single Device', button 'Save StartUp' – this button is used to transfer the start-up configuration pattern to the target device. This start-up configuration is stored in the EEPROMs of the respective systems – with several images to increase the reliability. When the switcher is next powered, this configuration will be loaded. The active configuration on the target device is not affected by this operation. The control is used only if a single audio-video router pair is being controlled. So the connection pattern for the first router group only has to be specified again. The start-up feature is useful in critical areas, where power supply failure can occur. After normal working conditions are restored, the active signal path configurations must be restored too immediately.
- Frame 'Stacked Devices', button 'Load Configuration' – the function of this control is to transfer the specified connection patterns to the target devices. The control is used for stacked AV router systems. The connection pattern for every group of routers has to be specified. If you have only two AV router systems connected to your PC, you don't need to specify all the four patterns of course. You should do it only for the first two router pairs.
- Frame 'Stacked Devices', button 'Save StartUp' – this button is used to transfer the start-up configuration patterns to the target devices. Active configurations on the target devices are not affected by this operation. The control is used for stacked audio and video routers. A connection pattern for every group of routers has to be specified.

The AVRS-8x4 audio and video routing system with keyboard.

Back panel view of the video switcher.

Back panel view of the audio switcher.

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

In summary

The AVRS-8×4 audio and video routeing system has been designed for studio environments, where the requirements for reliable 24 hours a day, 7 days a week operation are essential. As a result, the firmware has been designed extremely carefully with much redundancy, so that every situation is counted for.

The system has the advantage that it uses no special cabling for realising the necessary frame configuration. Standard serial cables can be used to connect between audio and video switchers, or stacks.

Having the opportunity to organise large stacks of switchers can prove to be extremely useful when expanding a current studio configuration, or building a large television production facility.

The system described in this article is in uninterrupted use in the Sofia and Pleven studios of EVROCOM – the largest cable TV provider in Bulgaria. According to the staff operating the equipment, it is a pleasure to work with it, because it is never failed and the interface is intuitive and easy to use. ■

References

1. MAX458/MAX459 8x4 Video Crosspoint Switches with Buffers Data Sheet, Rev. 3; 8/99, www.maxim-ic.com.
2. National Semiconductor, LM1881 Video Sync Separator Data Sheet, February 1995, www.national.com.
3. Atmel 2-Wire Serial EEPROM AT24C02 Data Sheet, www.atmel.com. Atmel 8-Bit Microcontroller with 2K Bytes Flash, AT89C2051 Data Sheet, www.atmel.com.

Listing 2. Audio switcher firmware.

```
:030000000200807B
:08000300323232323232323265
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:08001300323232323232323255
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Listing 3. Keyboard firmware.

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```

CIRCUIT IDEAS

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

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

Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

Don't forget to say why you think your idea is worthy.

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

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

Glitch detector and delay

Restoration of supply after a power cut or switching of heavy duty equipment can cause power glitches that may be detrimental to some types of equipment.

One example is the ubiquitous PIR/security light. In some cases a multiple glitch can replicate the manual switching mode initiating a permanently on condition. If left, this is wasteful of power, causes unwanted light pollution and in a lengthy absence of the occupant can indicate empty premises to a potential burglar.

This circuit produces a delay on initially switching on and then monitors the supply to detect such glitches, introducing a delay and isolating the power supply. In my case this is to several security lights but the circuit may be of use to protect other types of equipment where sudden surges may cause a fault, damage or mis-operation.

A two-diode full wave rectifier circuit produces a waveform which is shaped to produce narrow negative pulses at 10ms intervals. These are applied to a MC 14528B re-triggerable monostable with a time constant slightly longer than 10ms.

A missing pulse such as that caused by a mains glitch or drop-out will fail to extend the monostable period just before the end of its natural period. It therefore times out and the edge triggers a second monostable with a time constant of around 15s.

Diode LED2 is illuminated indicating that the power supply to external equipment is switched off. A positive level is available from pin 10 or a negative level from pin 9 of the MC 14528B to suit any switching requirement. Mechanical relays, solid-state relays and triacs can be accommodated.

In my case, a small 12V relay switches others with higher contact ratings, and manual switching is added to control the security lights. The delay remains until such time as normal 10ms pulses return and both monostables return to their initial state.

Apart from any desired changes to the final delay period, the only adjustment is that of the 100kΩ 10-turn preset VR_1. This should be adjusted until LED_1 just turns off, or pin 7 may be observed on a scope until the 10ms pulses disappear. If other than an 18V transformer is used, it may be necessary to adjust the value of the bottom left 4.7kΩ resistor, just under point a, to produce suitable 10ms pulses.

A normally-closed push-button switch in one leg of the rectifier circuit is used to test the system, changing the rectified output waveform from 10ms to 20ms.

N L Smith
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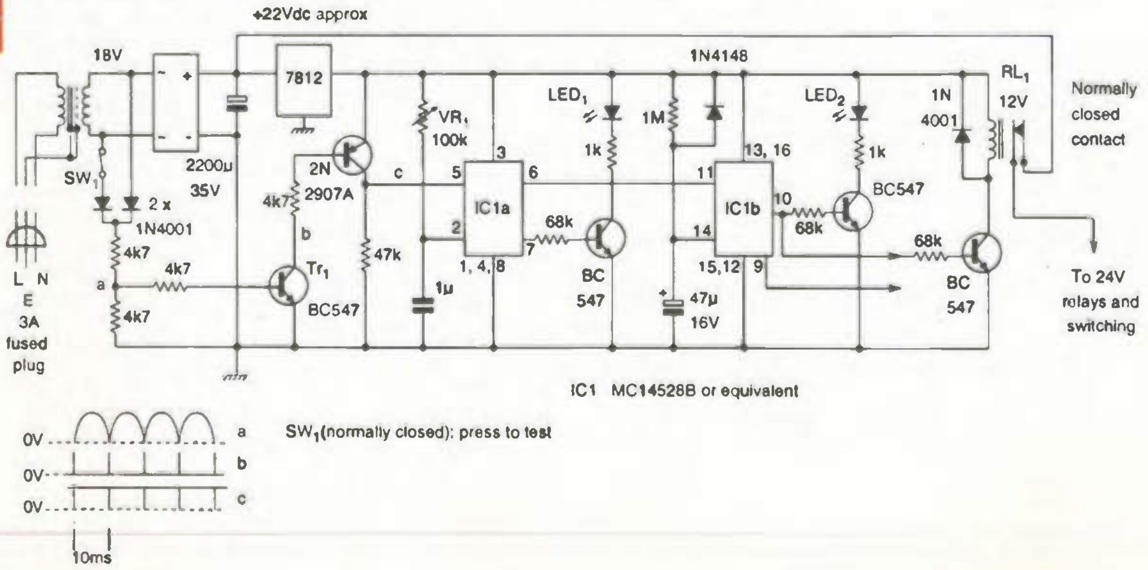
£50 winner



Unity-gain phase-shifter

An operational-transconductance amplifier – OTA for short – is at the heart of this phase shifter.

The relationship between e_o and e_i can be derived by using the formulae below. If e_- and e_+ are the voltages at the negative and positive inputs of the OTA respectively, and assuming $\gg R_2$ and that the gain of the output buffer is unity, then:

$$e_- = (e_i + e_o)\frac{R_2}{R_1} \qquad e_+ = 0$$

$$i = (e_+ - e_-) = -(e_i + e_o) g_m \frac{R_2}{R_1} \tag{1}$$

$$e_o - e_i = i\frac{1}{j\omega C}$$

$$i = (e_o - e_i) j\omega C \tag{2}$$

Since (1)=(2) then,

$$-(e_i + e_o) g_m \frac{R_2}{R_1} = (e_o - e_i) j\omega C$$

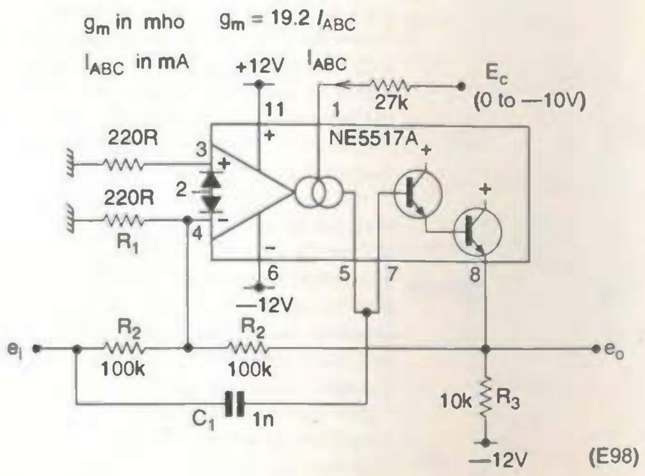
$$-\frac{e_o}{e_i} = \frac{R_2 g_m - R_1 j\omega C}{R_2 g_m + R_1 j\omega C} = A$$

The absolute value of A is 1; the phase-shift is found to be:

$$\sin\varphi = \frac{-2R_1 R_2 g_m \omega C}{R_2^2 g_m^2 + R_1^2 \omega^2 C^2}$$

In practice, with the components shown in the diagram and a source frequency of 1000Hz, the phase can be shifted between 210° and 330°. The 1nF capacitor can be scaled as required for other frequency ranges.

W. Dijkstra
Waalre
Netherlands
E98



£50 winner

A versatile interface for a narrow frequency sweep display

Frequency-sweeping signal generators – or sweepers – can be used with an oscilloscope to display frequency response characteristics of a device under test. Often they have a provision for an external drive to the frequency sweep circuitry.

If a ramp output from the sweeper is available to feed the X-amplifier of the oscilloscope, the scan rate may not be low enough to display accurately a steep-sided narrow bandwidth response, such as that of a crystal filter. In such cases, a separate variable frequency ramp generator has to be used for the sweeper's external input.

If an oscilloscope having an X-sweep output (ramp output) is used with the sweeper, then this ramp can drive the external input of the sweeper, thus controlling the scan rate from the scope. However, it is likely that a direct connection of the ramp output to the sweeper's external input would be unsatisfactory – possibly even catastrophic.

Some form of signal conditioning/buffering between sweeper and scope is essential. This could provide a suitable polarity and amplitude of ramp voltage from a low source impedance to drive the sweeper for a linear frequency/X-sweep display on the oscilloscope.

An inverting/non-inverting amplifier, having a gain continuously variable between +1 and –1 has been described previously in Circuit Ideas[1]. An adaptation of this circuit could form an interface unit between the sweeper and scope, to provide flexibility of ramp output amplitude and polarity.

In a conventional swept display, frequency increases left-to-right along the X-scan. A ramp having either a positive or negative slope will be required, according to the sweeper's external input characteristics of frequency *versus* voltage. This polarity selection, together with control of ramp output from the interface, is provided by the continuously variable +1 to –1 feature.

Ramp output voltage can be made to have nominally equal excursions either side of zero. An additional control offsetting the zero allows you to centre the display on the scope screen.

A similar shift of the display by a pre-determined fixed amount, as and when required, can be derived from a stabilised voltage source in the interface. This gives a frequency calibration in terms of displayed frequency sweep, i.e. kilohertz per division on the scope screen graticule, at any particular sweep width. This facility is useful if no frequency markers are available or they are too widely spaced to cater for narrow frequency sweeps.

The circuit shown is of a battery powered interface for a specific application, using a Marconi TF2008 swept signal generator and a Gould OS250 oscilloscope. The oscilloscope ramp output is 0 to +10V from a nominal 18kΩ source. The sweeper external input characteristic is 35kHz per volt on the particular RF frequency range in use.

Calibration voltage, derived from a 1.23V band-gap reference IC_2, is set by choice of R_2 to produce a 10kHz shift of the display when the push-button switch is operated. A ramp output of +2.5V is obtained with the value of R_1 shown, but this can be adjusted to suit other source resistances and voltages of a scope ramp output.

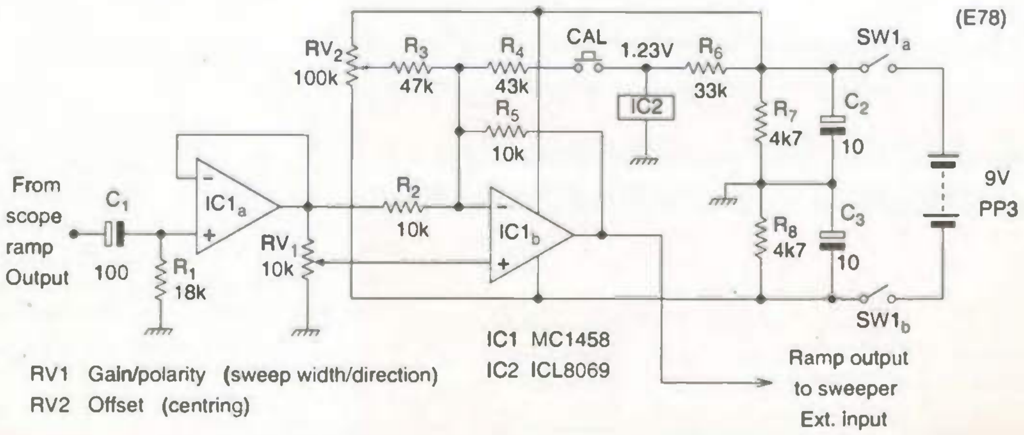
The time-constant of $C_1 \times R_1$ must be large enough to pass without distortion the lowest sweep frequencies used. In the application described, this was 5Hz.

The polarity of C_1 must suit the oscilloscope's ramp polarity. If a greater output than +2.5V is required, then a higher rail voltage and/or an IC_1 capable of a peak-to-peak output swing closer to the rail voltages than that provided by the MC1458 will be needed to avoid limiting.

The interface provides smooth control of sweep width between maximum and zero, so very narrow sweeps are easily set up. These, in conjunction with an 80dB dynamic range log-amp detector feeding the scope Y-amp, enable crystal filter response characteristics to be displayed over a large portion of the scope screen, making it easy to measure bandwidths and shape factors.

Ross Muddell
Malvern
Worcestershire
E78

Reference
1. *EW/WW* Aug. 1994 p. 688.



Simple stepper-motor drive

This stepper motor control circuit – or 'translator' – comprises a shift register and clock source, Fig. 1. Outputs Q_A-Q_C are gated so that if any one is high, a zero is clocked into the first stage output Q_A. When the first three stages all hold zeros, a one appears at the right-shift input. On subsequent clock pulses this is gated into Q_A, and subsequently propagates through the other stages. Thus a continuous circulating wave drive sequence is obtained.

The outputs Q_A to Q_D drive four power amplifiers, each as in Fig. 2. The wave drive sequence can be gated with NOR gates as in Fig. 3, to provide a two-phase drive sequence. In order to change the direction of the stepper motor, inputs at S_0 and S_1 are set high or low with switch SW_1. R_1 resets the translator.

The advantage of this circuit is its immunity to noise. If a noise pulse disrupts the any of the Q_n outputs, the sequence is automatically re-established by succeeding clock cycles. R_2 is a reset for the clock circuit.

V Gopalakrishnan
Bangalore
India
E79

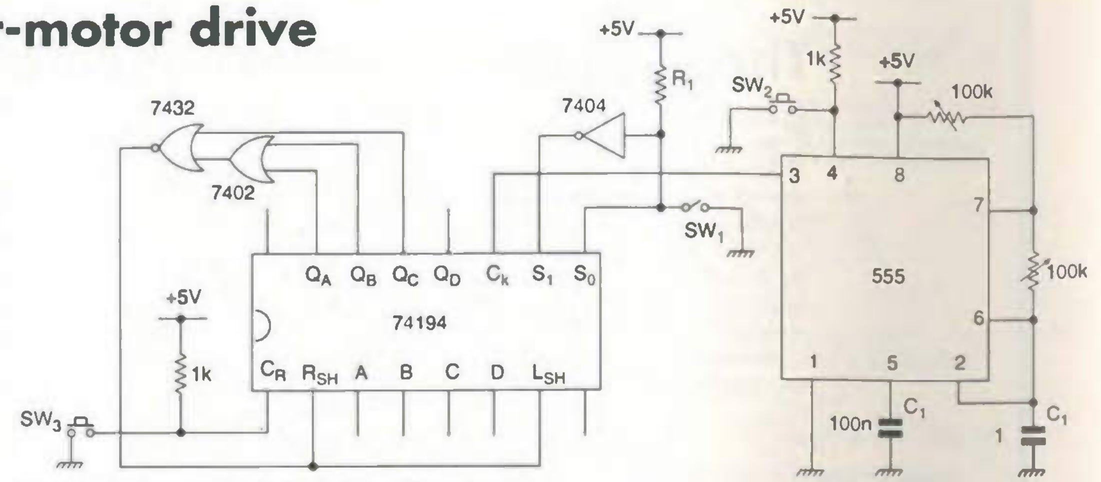


Fig. 1. Stepper motor controller using a shift register.

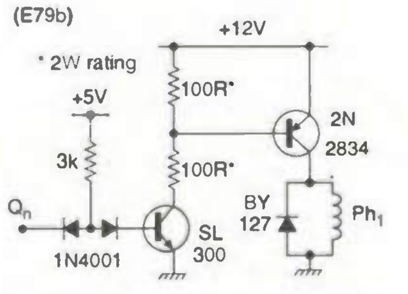


Fig. 2. Power driver for one stepping coil.

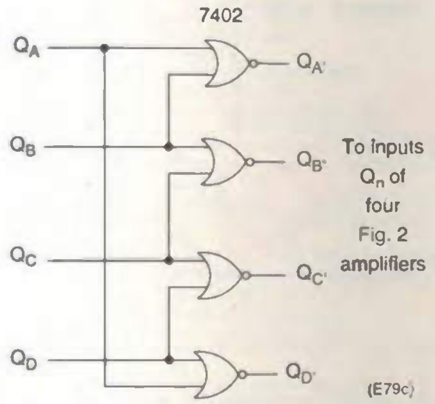


Fig. 3. Gating to convert the controller output from wave drive to two-phase drive.

Table. Wave drive sequences without the modification in Fig. 3, and two-phase drive, as obtained with the modification.

Wave drive sequence				Two-phase drive sequence			
Q_A	Q_B	Q_C	Q_D	Q_A'	Q_B'	Q_C'	Q_D'
1	0	0	0	1	1	0	0
0	1	0	0	0	1	1	0
0	0	1	0	0	0	1	1
0	0	0	1	1	0	0	1

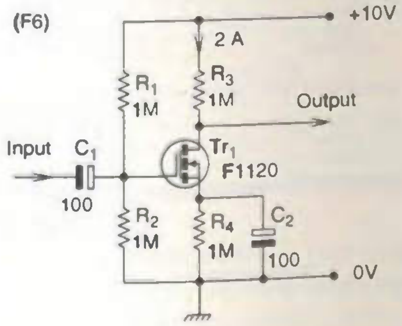
A microcurrent amplifying stage

A common-source amplifying stage using a single field-effect transistor can ensure enormously large voltage gain, if the field-effect transistor's drain current is very small. An example circuit is shown in here.

PSPICE simulation of the circuit has shown that the voltage gain factor of the amplifying stage is equal to 5500. But that is not the limit. Replacing resistors R_4 and R_3 by 10MΩ types makes the voltage gain factor equal to 17000. The maximum possible value of the voltage gain factor depends on the drain-source and drain-gate leakage currents of the field effect transistor and can be very large.

A similar arrangement was used in the days of valves, with a pentode used in the 'starvation' mode. In both cases, the large gain is bought at the expense of limited bandwidth.

S Chekcheyev
Tiraspol
Moldova F6



£75 winner

Three-phase sequence indicator

Phase rotation or phase sequence is an important parameter of a three-phase supply, as the direction of rotation of induction motors will depend on it. In some cases, serious damage to machinery may result if the driving motor is reversed. In such cases, local protection against phase reversal is essential.

The circuit given here is probably the simplest and the lowest cost solution possible for providing an indication of the direction of phase rotation. It is readily adapted to provide an interlock, to prevent operation with the wrong phase sequence.

The circuit makes use of the fact that in a three-phase system, the voltage vectors of two phase terminals with respect to the third phase have a phase difference of 60°. Whichever voltage leads the other is an indication of phase sequence.

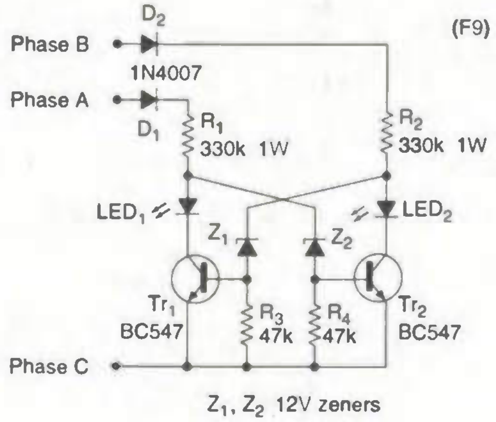
If the phase A leads phase B, the transistor Tr_2 receives base current first and LED_2 will light when phase B becomes positive. At that time Tr_1 will not receive base current since Tr_2 is conducting, and therefore LED_1 will remain switched off even when the phase B becomes positive later.

Should the phase sequence be reversed, Tr_1 receives base current first and only LED_1 will light in the same manner. Therefore LED_1 and LED_2 will indicate clockwise and anti-clockwise phase rotations of the supply respectively.

If one of the *LEDs* is replaced with an opto-coupler, a signal can be obtained at a potential close to the ground level. This may be used to interlock the motor starting switch gear so that motor will not operate in the reverse direction.

C Palihawadana
Dehiwala
Sri Lanka

F9



Important note

- All the components of the circuit will be live at near phase voltage, so the circuit is potentially lethal.
- The circuit can be used on supplies up to 400V phase-to-phase voltage.

Keyboard has serial interface

This circuit is based on the UCN5833 IC, which is a 32-bit serial-input latched driver with 32 open-collector outputs. Control pins are clock, serial data in, serial data out and strobe, i.e. output enable.

The keyboard is controlled by a microprocessor or microcontroller using three output pins and two input pins. The software should implement the following steps.

At power on, the software should disable the output enable line and zero all 32 shift registers stages by clocking in zeros until the serial data out line reads zeros continuously.

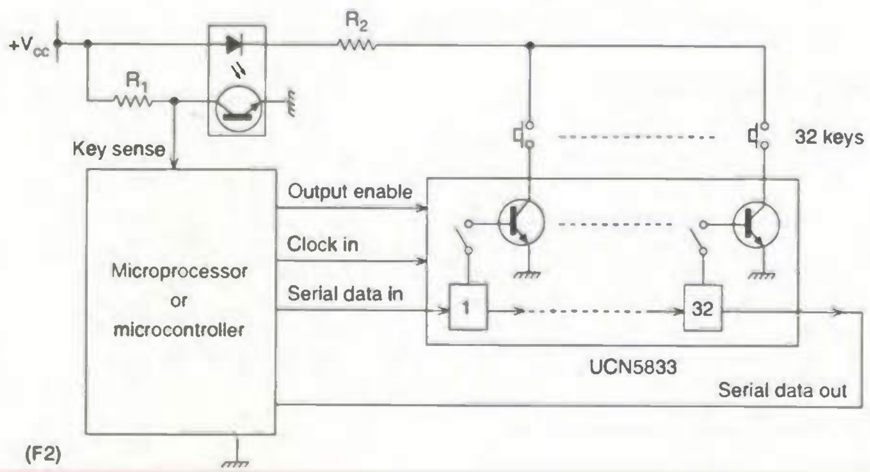
Now the serial data in line is taken high for one clock, and this is then circulated through the shift registers on subsequent clock pulses. When the circulating high bit reaches any pressed key latch, the open-collector transistor of the latch conducts and the controller is informed of the key closure by the opto-coupler output 'key sense' signal.

This information is stored along with the clock pulse count – which is reset every 32 counts. The same key is then repeatedly sensed for key release when corresponding clock period is reached. Once the key release is detected, the program can jump to the appropriate action routine.

Note that the clock frequency should be chosen to ensure that there are several samples of any key pressed. Precautions should be included to deal with two key roll-over; a detected second key-press being ignored until the first key release is sensed.

Jayant Kathe
Mumbai
India

F2



Letters to the editor

Letters to "Electronics World" Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS e-mail jackie.lowe@rbi.co.uk using subject heading 'Letters'.

Input-filter distortion

I thank Graham Maynard (*EW* Jan. 2001) for taking the time to reply to my letter. Unfortunately, he has failed to explain the quantum of distortion claimed and his analysis of the cause is specious.

As all aspects must be addressed technically, competently, and rationally, please allow me to explain, and hopefully promote a more accurate understanding of the relevant issues.

To the best of my knowledge and understanding, linear distortion is measured in terms of amplitude, phase, and time, with the units used being decibels (or ratios of power, voltage, pressure, etc), degrees (or radians) and seconds.

I have yet to see explained how it might be expressed as a percentage like non-linear distortion, as Graham does. He gives me no formula or other clue as to how this might be done.

My understanding of linear fundamentals appears not to be the same as Graham's. For audio it is pointless to consider the effects of a circuit at one frequency for the purpose of determining the extent of any linear distortion.

Graham is correct in that the filter "introduces a tiny delay", but he is incorrect in suggesting that it "increases sharply with frequency", certainly within the range of human hearing.

What in fact happens is that the filter introduces a pretty uniform 2.2μs delay over the entire audio band, with a maximum variation of less than 160ns. This is nearly identical to the effect that would be produced by sending the signal down a perfect transmission line with a transit time of 2.2μs. This would be inaudible since the same time delay is forced on all signal components.

I know of no reliable evidence that the effects above 20kHz are audible. I reject any and all claims not backed by properly conducted research.

This can be easily demonstrated with a very simple PSPICE simulation.

```
* LP RC circuit for Group Delay
  Measurement
* rc_gd01.CIR
*
Vin 1 0 AC 1 0
R1 1 2 10k
C1 0 2 220pF
.AC DEC 500 100Hz 1000kHz
.PROBE
.END
```

By running PROBE and plotting vg(2), you will get a plot of the group delay through the circuit. The delay at 100Hz is 2.2μs and the delay at 20kHz is just over 2.04μs.

Over the range of human hearing the variation is smaller. To put this in some perspective, if you moved your head back about two thirds of a millimetre you would get pretty much the same audible effect – virtually none. I don't think anyone is likely to think that was important.

When one considers common listening

Tracking down mains earth leakage

A while ago, I suffered from my regular tripping of my domestic mains earth-leakage detector.

The trips occurred about every month or two – but sometimes two or three times in a 24 hour period, and the problem could have been on any of eight lighting circuits or six ring mains. Almost invariably the breaker would hold in immediately after a trip, and frequently, nobody had been turning anything on or off in the house.

I concluded, therefore, that there was earth leakage current running all the time, but that random fluctuations were causing it to just reach the trip threshold. In order to measure this, initially I tried winding a few turns of wire round both line and neutral feeds to the fuse box, and connecting a multimeter on AC volts. Not sensitive enough.

After further head scratching, I remembered that I had an old line-output transformer in my junk box. By demounting the clamp and re-assembling the core round the incoming mains, the EHT winding produced an output of about 45mV with the house fully powered up – which is enough to measure.

By turning off all the trips, the voltage reduced to virtually zero, suggesting that I was now indeed measuring the residual current. Bringing the circuits back on one at a time, I found that each one added a few millivolts to the reading, but that one ring main added about 15mV.

First I unplugged all the earthed appliances, judging that a device with a two-core mains lead could not possibly leak to earth. Wrong. It turned out that a television was leaking out to its aerial lead – not much, measured at about 5mA – but we'd be better off without it.

Interestingly, I had noticed that trips seemed more common in wet weather, and I suspect this has something to do with the conductivity of the chimney on which the aerial is mounted...

But there remained the puzzle of why the earth current rose a little with each circuit. Surely they couldn't *all* have faults on them? To find this out, I replaced the multimeter with an oscilloscope. To my horror, I found that the waveform coming out of the EHT winding looked like a field of long grass – not the sinewave I was expecting. A moment's thought then told me that any capacitance to earth would provide a low impedance path to any noise on the incoming mains.

My first question, then, is whether or not this would contribute to RCCB trips?

Deciding that I was more interested in resistive – rather than reactive – leaks to earth, I now added a simple *RC* filter of 10k/330nF to allow me to see the 50Hz component. I was rewarded with a slightly rough sinewave.

Using the second channel on the 'scope to monitor the line, I could see that there was a phase lead. This confirmed my suspicion that it is capacitance between the line and earth that accounts for most of the current measured.

I should mention that my household is littered with kit with switch-mode PSUs – e.g computers, TVs – all of which no doubt have filter capacitors down to ground in an attempt to clean up the resulting mess.

So here is my next question. Is it true that mains cabling capacitance and filter capacitors at SMPS inputs will permit a certain amount of standing 'leakage'. By this I mean is the actual earth fault current required to trip the breaker is considerably less than the 30mA rating on the breaker?

Chris Miller
Via e-mail

environments, one sees how insignificant this is. Hardly anyone hears anything in a free acoustic field. Most of us spend our lives in an acoustic near field of some description. Those that spend all their listening time in anechoic chambers might be excepted.

This means that there are peaks and troughs across the audio band, and these will change with position. For example, my favourite chair has a moderately high back, which ends a few centimetres below my ears – even if I slump down. I notice that there are distinct – but not serious – changes in the spectral quality if I move my head forward or back by 3 to 5cm.

Clearly, to avoid any linear distortion in that chair – well, I'm assuming it is the chair's fault – I obviously need to have a frame fitted that rigidly fixes my head on one position relative to the speakers.

But this would do nothing to address other problems caused by opening or closing doors and windows, rearranging the furniture, or the presence of any friends I might have over to enjoy the music with me.

These effects are not caused by signal delays but by reflection and diffraction of the sound by objects in the listening room. These all affect the sound field in one way or another. They would even do so in an anechoic chamber, unless their shapes and surfaces were treated.

I think if anyone does try switching Graham's filter in and out, and they do detect some audible difference, then it is likely to be a product of their imagination and not any physical or auditory effect. Double blind testing is about the only way I know of to avoid the bogeys of human bias and misconception whatever their sources.

Cyril Bateman's contribution on the matter (*EW* Jan. 2001), while informative, is irrelevant. Having spent some years working with banks of aging Plessey Ducon electrolytic capacitors, I am quite aware that real components are not ideal.

It was pretty clear from the context of Graham's letter that he was talking about linear distortion, to which Cyril's comments are not germane. But thanks for sharing it Cyril.

Phil Denniss
Sydney, NSW
Australia

Graham replies...
What a real shame it is that Mr Denniss did not follow my suggestion and try these values for himself, for they are audible on a system that is not already filtered. No amount of theorising or self opinionation can alter reality.

Mr Denniss' forthright 2.2μs group delay explanation shows that he firmly believes the filter cannot affect audio dynamics. Unfortunately, he has incorrectly assumed that an electrical *RC* filter behaves as if it is a propagational transmission line.

As I wrote, these filters do not distort LF, but they do reduce clarity by having a HF effect. This effect is initially asymmetrical and lasts for longer than the spot frequency 'group delay' period. This is especially so at frequencies where the phase-lag 'delay' is no longer insignificant.

The greater the *RC* phase lag, the greater the initial asymmetry and waveform distortion, which becomes severe on sharp leading edges. See the 20kHz simulation, **Fig. 1**, for clarification; at LF there is similar delay but minute and quite insignificant error.

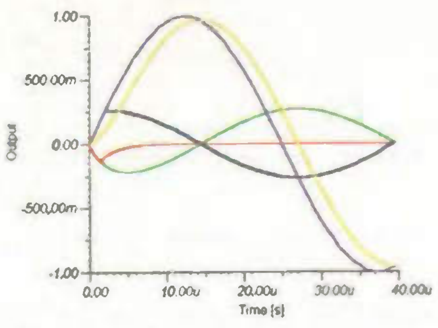


Fig. 1. 20kHz simulation highlighting RC phase lag.

The non-linear phase lagging AF error wave components that I wrote about can be much greater than people realise.

At the end of a delay line my input-output difference curve would be an error free negative maximum at 2.04μs. See the 43.5mV difference between too small a negative peak, when compared to the normal positive (group delay) peak, and this within a 40μs time frame – most significant when compared with the 1V, 50μs original – but then I never did want to 'fuss' this much as Mr Denniss has already accused me.

No filter. No real-life error. Thus I'll leave Mr Denniss to ponder what happens to leading edges of waveform transients when they pass through his delay line. We cannot have one examination for LF and another for more dynamic signals.

Incidentally, the filter proposed by Mr Ellis (it is not mine as Mr Denniss writes) limits slew in a 50W, 8Ω amplifier to 17V in 2.2μs. This is slow by today's standards, and I'm sure that Mr Denniss is as able as I am to understand the error that extends to 10μs, and the perceived lack of transmission line delay. He probably has already PSPICEd this aspect in the normal course of his work, but not yet thought this far forward.

Also, I have checked both my letters and thus Mr Denniss must have misread me to misquote. It is he who has written 'delay' where I wrote of a "...'phase shift' that increases sharply with frequency...".

He infers human bias and imagination where I have actually taken the trouble to check these filters in isolation for their additional input-n.f.b. impact on amplifier input circuitry. I have no reason to make false report about my real-life findings of the 1970s and 1990s, which computers are now able to examine.

He also intimates that I was talking about linear distortion. That was not clever.

Stranger than fiction...

In his article 'Dome myths exploded' in the February 2001 issue, John Watkinson said that the old air-cooled Volkswagen bus had a better drag coefficient than the Jaguar E type. Where did he get this information? I am not doubting him; I just would like to see published data.

Jose Senna
Via e-mail

John replies...
My figure for the drag coefficients of the E-type Jaguar and the VW bus came from tests done at MIRA. These were quoted in a set of books called 'Our Four Wheels' published by Orbis in 1973.

The VW bus (Cd=0.42) was described for low drag and includes a number of features, based on Kamm's work, which was well ahead of its time.

The nose is curved and the front bumper acts as a spoiler. The underbody is flat and smooth. The air intakes for engine cooling are at the rear of each side, behind the windows. These scoop in the boundary layer, leaving more energetic air to negotiate the change of cross section at the tailgate. This reduces turbulence.

The E-type (Cd=0.44) was designed to be pretty, which it is. But aerodynamically it was poor. The long pointed nose forced air under the car and there was no spoiler. The long tapered tail meant that flow separation occurred well forward at a variable point and the large surface area of the tail was just producing profile drag. The contemporary Citroen DS19 had a Cd of 0.31.

Incidentally, the XJS which followed the E-type was wind tunnel tested and put all this right. There is a spoiler and a large smooth undertray at the front, and the famous 'flying buttresses' at the rear follow the path of the vortex structure leaving the roof and sides.

The sharp inner lip of the buttresses and the end of the rear roof produce a defined separation point and I wouldn't part with mine.

A VW concept van based on the old 1950s Ti split-screen van – dubbed the Microbus and hailed as the bus of the future.

Sprint timer

The 'Infra-red sprint timer' in the April issue is not new. In January 1985 I designed the STAR System – and acronym for 'Sprint Timer Analyser & Recorder' – for the Sports Institute of South Australia. It used a Tandy TRS80 Model 100, modified intruder-alarm beam break detector, a TRP-100 printer plotter and a CCR-82 cassette recorder.

The system was used to plot a sprinter's startup, and 'finish' velocity and acceleration/deceleration. The implication is that sprinters – runners, skaters and cyclists – all slow down before they reach the finish line.

Manager of the Sports Institute Mike Nunan suspected this, but couldn't prove it until the timer became available. So the philosophy is now to train athletes to treat the finish line 100m ahead, so they don't slow down at a critical phase.

Accuracy of the system was about 100mm or 1/100 second. Printouts were: distance versus time, acceleration versus time, velocity versus time.

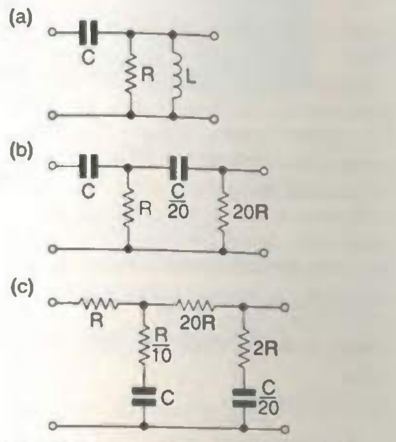
John Ingram
Australia
Via e-mail

Sub-bass challenged

I am afraid that Mr Maynard, with his Bass Boost circuit in *Electronics World's* February 2001 issue, has been labouring under a misapprehension.

To compensate for the natural fall-off in response of loudspeaker systems at low frequencies, we do not need phase advance as the frequency is reduced – the loudspeaker is already doing that! We need to compensate for what is actually happening.

The voltage to acoustic-pressure low-frequency equivalent circuit of a typical loudspeaker system will be similar to that shown in **Fig. 1a)**. This is a two-pole filter whose resonant peak will be determined by the damping resistor *R*. Unfortunately, the vast majority of loudspeaker systems are far too under-damped - hence the familiar 'boom boom boom' sound.



The equivalent circuit of a well-damped system, where the resonance effect can be discounted, will approximate to Fig. **1b)**. Resistor values are shown increasing in value to prevent the later stages appreciably loading up the earlier ones.

The transfer-function of the individual stages is of the form $ks/1+ks$, where $k=C.R$ and s can be replaced by $j\omega$ when you are looking at the sine wave response.

To give good transient response, a bass-boost system, must correctly compensate for this effect down to as low a frequency as is practicable.

Mathematically, you need to solve the equation $T.(ks/1+ks)=1$, where T is the transfer function of the compensating network. Multiplying both sides of the equation by $1+ks$ gives $T.ks=l(1+ks)$. Hence $T=(1+ks)/ks$. This simplifies to $T=1+1/ks$. This is unity gain added to the output of a perfect integrator.

An approximate integrator can be made from a RC network as shown in Fig **1c)**. This will need amplification to restore unity high-frequency gain. This is the same circuit

What does 'engineer' mean?

The shortage of engineers in Britain is primarily due to the fact that the title, 'engineer' has no meaning in Britain, and also to the low salary levels offered to well qualified engineers. If you look at the number of engineering graduates who actually take up engineering careers, you will probably discover that it is less than 50%.

The shortage of engineers with a hands-on approach is primarily due to the decline in manufacturing, research and design, that has occurred in recent years in Britain. This decline has meant that these people have also taken jobs in more secure occupations.

The various training schemes, that have been introduced have done little to resolve this problem. For the most part they do not meet the needs of the electronics industry, as the course content is theoretical and often several years out of date.

The Society of Engineers, which is Britain's third oldest professional engineering body, realised this problem. It has restructured its examinations to meet the requirements of modern electrical and electronic engineering.

The design and project papers are completed by the candidate at home over a specified period of time. These provide the candidate with the opportunity to demonstrate that he or she has a practical approach to engineering as well as a theoretical knowledge. This means the candidate has often to research topics or regulations to ensure that the paper is completed successfully.

If any of you want to find out more about The Society of Engineers and its examinations, telephone 01206 263332 or e-mail secretary@society-of-engineers.org.uk.

David Purnell
via email

that Mr Maynard dismissed in referring to his Fig. 1.

In fact Mr Maynard's circuit will make the bass transient response worse, not improve it. The simple circuits, with a well-damped loudspeaker system, can do all the necessary compensation.

Readers would be far better advised to use simple bass-boost circuits and, if necessary, put some good acoustic damping material into their loudspeaker cabinets. Dare I suggest long-fibre wool?

Arthur R Bailey PhD MSc FIEE
Ilkley
West Yorkshire

Graham replies...

I thank Arthur Bailey for his letter in response to my sub-bass equaliser article. Mr Bailey has many decades of hands-on audio experience and I much respect his work, but my article approaches sub-bass reproduction from an historically unconventional starting point.

I too have done my work carefully, and thus I trust that he will understand differences which at first sight are not clear, for my equaliser does do exactly as he suggests. It compensates for what is actually happening; theoretically as well as practically.

'E-bass' is not a phase-altering bass boost system. It is a phase-linear and adjustable sub-bass equaliser that has been specifically designed for use with strongly constructed bass drivers. Because these drivers are driven only at frequencies below their deliberately raised cabinet resonance, they transduce with much less phase error than when mounted in our conventional cabinets.

I agree with Mr Bailey that we do not want any phase advance with reducing frequency. Nor do we want even the changes that occur due to loudspeaker resonance. But cabinet damping alone cannot prevent these.

Long-fibre wool might reduce system 'Q' and the sharpness of the resonant phase change, but the phase change still occurs. In conventional cabinets the effects are always audible at sub-bass frequencies.

Dare I say that we have allowed ourselves to become used to it, because of the difficulties presented in overcoming the problem! Bass boost cannot help either, for it only adds to the overall phase distortion; it might improve some sounds at some frequencies, but it is deleterious to others.

Unfortunately, Mr Bailey has introduced theoretical representations that are not always properly understood. These place me in a position where I must respond.

I agree that his Fig. **1c)** circuit will equalise his Fig. **1b)** characteristics. I also agree that his Fig. **1a)** is a fair representation; but where has the driver-cabinet combination gone in Fig. **1b)**?

Applying C=22µF and R=150Ω to Fig. 1b) leads to a –12dB roll-off below a –3dB turnover at 80Hz. But the real-life cabinet resonance that produces a peak and tightens the phase change about its frequency of occurrence is missing!

Yes, his Fig. **1b)** does create a leading characteristic. In real life though, the compact sub-bass loudspeaker is electrically connected such that it runs properly in phase at sub-bass frequencies. It actually develops a lag only at frequencies close to the deliberately raised resonance, which is not driven.

Also, with a steep-cut electronic crossover – having phase changes of its own that can never be properly avoided, as with all crossovers – the amplifier is then able to directly exert a high degree of linear cone control all frequencies. This means that it will electrically dampen loudspeaker system induced, resonant frequency resultants where they have not been fundamentally energised. Such resultants are unavoidable within all loudspeakers.

Thus I have not been labouring under a misapprehension. I have actually tried my Fig. 1, which is Mr Bailey's Fig. 1c). In real-life it sounds atrocious when applied to a phase linear sub-bass reproducer.

I sought other 'ears' to confirm whether my own observations were correct. It took me ages to get my head around what was actually happening; also to get beyond conventional circuits.

My real-life findings appeared to conflict with what I had read through the decades. But I could not be so brash as to say 'if the theory does not fit, then change the theory'. My findings could not be disputed either. It was my own interpretation and application that had to be re-evaluated.

In common with other integrator circuits, Mr Bailey's Fig. 1c) network shifts the entire sub-bass spectrum into what is virtually a lagging quadrature. This would then re-mix with bass harmonics plus other mid and treble signals that have been simultaneously amplified.

There is a distracting 25ms delay at 20Hz. Overall reproduction literally falls apart because a resistively-operating sub-bass loudspeaker cannot be made to compensate by transducing electrical waveforms before they arrive.

I wrote that I used a dual-beam oscilloscope to observe and measure these time differences, and I worked away until they were minimised by a circuit that would not overdrive at infra-sound frequencies. My own hands-on work was completed without the aid of computers, but these were later used to confirm measurements and then to draw up the article.

My equaliser circuit offers an alternative possibility for level and flat-phased sub-bass reproduction. I implore that anyone who might feel a need to comment to please audition what is possible before putting pen to paper.

Just because I prefer compact sub-bass does not mean that I advise readers not to construct heavy and large conventional cabinets that leave drivers at risk of over excursion and introduce a ludicrous phase change at sub-bass frequencies. When it

Sampling 'scope memories

Firstly, one issue without an article by Ian Hickman is bad enough, but two issues without is a disaster!

Secondly, in the the March 2000 issue, we were presented with a most interesting insight into sampling techniques – sufficient to divert my attention from trying to improve the performance of a TDA8703 a-to-d chip.

Not having such a well filled spares box I used a BFR93A with BAT83 diodes. Coupled with a PIC16F877 and a bit of programming this reproduced a 50MHz square wave quite well on the computer. Unfortunately it had to be driven from a low impedance as the suggested active probe (May 1996) using a MAX4005 converted square waves above some 20MHz to fairly good sine waves!

However, with the original article we were admonished to 'Watch this space...' Well I have but to no avail. Is it not about time that the recent gaps were filled with the promised follow up article?

John Kanaar
Via e-mail

Ian replies...

John, I am afraid that readers have been "watching this space" for far longer than I intended. The missing part of the design is the trigger and incremental delay processing, to reconstruct the sampled waveform.

I had in fact done quite a lot of work on this part of the design before developing and publishing the sampling circuitry. But for various reasons, it has languished on the back burner.

At the moment, apart from my many non-electronic activities, I have a new edition of one of my books to complete by late summer, but after that, I intend to return to and finish the sampling 'scope design.

comes down to it, neither system is anywhere near perfect.

What I would suggest though is that you might try both, and then see which you would rather live with. It is possible that anyone then building up a new system might actually choose to use much smaller conventional stereo loudspeakers as head height satellites, and then take the overall system response down to the lower limit of human hearing by using compact 'e-bass' technology.

Graham Maynard
Newtownabbey
Nothern Ireland

EMC and the DIY PC

Regarding John Woodgate's letter in the May 2001 edition of *Electronics World*, unless the requirements have changed in the past few years, he is incorrect to say that CE certification is required for an individually built PC.

While it is true that the European EMC Directive applies to all electronic equipment, whether 'placed on the market' or 'brought into service' (without being first placed on the market), the requirement for CE marking, and hence declaration and hence probable need for testing, only applies to products placed on the market. In other words your home built PC is required to meet the regulations but you are not required to demonstrate that is does.

In effect the Directive empowers the 'Powers that be' to do something if you are causing a problem – to require you to fix it at your own cost. Otherwise, it only requires you to not knowingly cause a problem.

John comments that he has had PC suppliers respond to his querying PC EMC performance by saying that 'all the parts were CE marked'. It is important to understand that this is not a defensible position on behalf of the manufacturers. At the very least they must obtain the 'Declaration of Conformity' from the manufacturers of the component parts to ensure that the declaration which the CE marking represents actually covers the requirements that the PC manufacturers need to meet.

The classic example for this is the power supply module. Within the context of the European Directives this is treated as a component, not a product. As a component, it is exempt from the EMC Directive, but is covered by the Low Voltage Directive (think of this as the safety directive).

The power supply has to have CE marking affixed to show compliance with the safety requirements, but this says *nothing* about its EMC performance. The onus is on the PC manufacturer to place a commercial requirement on the power supply manufacturer that the power supply should be designed so that when the PC manufacturer affixes CE marking to the PC that the EMC requirements will be met.

Philip Williams BSc CEng.
Harlow
Essex

In the May issue, Robert Atkinson rightly picked me up on a statement I had made concerning the EMC requirements for home-built pcs. I had said "...but there are no requirements for home built machines. (If a company builds a non-compliant computer it can be prosecuted, but an individual can buy the same parts, make the same machine, and not have a problem.)" Mr Atkinson was right in the sense that the passage was poorly worded and could easily be misinterpreted.

This is a very difficult area for all concerned. In the UK we first have the 51 page Statutory Instrument 1992: no 2372 "The Electromagnetic Compatibility Regulations". Then we have S.I. 1994: no 3080, the 8 page amendment to the regulations. These are the key legal enactments of the 89/336/EEC, the EMC directive. And then we have the official guidelines on the application of the directive. These run to 59 pages in Word format.

For those interested in the guidelines, they are available from several sites on the web. The DGIII site has a very long URL so you might prefer either www.eucs.com/news.htm or www.emc-journal.co.uk.

Now these guidelines are official, but not legally binding. This in itself is tricky. The bottom line is that regardless of the expertness of the person you consult on this matter, the only final arbiter is a court of law.

Bearing this *caveat* in mind, clause 6.4.1 of the guidelines allows an end user to construct a computer system from CE marked parts with no further testing, marking or declaration. This is intended to be for separately buying and plugging together a main box, a monitor, a printer, a scanner, etc. This is analogous to making up a hi-fi system from parts. However, a retailer selling this system as a complete unit is in a different position.

Now making the main computer unit from a mother board and plug-in cards is arguably stretching this clause somewhat. Nevertheless it is done extensively. Another clause is also of interest to the home pc builder; clause 7.3. "Modifications carried out by the end-user (under his responsibility)" covers a multitude of sins. They are not covered by the EMC directive, but by "product liability and other pertinent legislation". My current pc, for example, was a store-bought 486DX-33 until 'upgraded' by new motherboards, hard drives, ram, processors, etc., to a Pentium III. Only the monitor and the sound card are now original in fact! At least I had enough inside knowledge to buy a new case with decent EMI protection.

However, what about poor Joe Public? (S)he is not required to produce a certificate of competence to build a pc, and yet can buy all the bits to do so. The only unambiguous point is that if you make a pc for someone else, regardless of whether you get paid for it or not, you are definitely technically committing an offence if it is not CE marked as a system. If you make one for yourself, that seems relatively ok, provided you never sell it or give it away – without first CE marking it.

The idea of making Joe Public guilty of the crime of not CE marking his own equipment seems untenable. He can of course be prosecuted if he causes actual, as opposed to possible, interference. Does Joe Public know that he can be prosecuted for buying non-CE marked relevant equipment? (Clause 29 of the first mentioned Statutory Instrument.) Of course not, and I bet more than 30% of people in the trade don't realise this either!

I hope this adequately explains the point. My previous 'aside' was too brief to fully illuminate this murky corner of the legislation.

Leslie Green CEng MIEE
Via e-mail

Get in Linux

Regarding John Jameson's letter in the April issue, I certainly can commiserate, being in the midst of a programming project myself. But there is an alternative that may please his programmers – though his customer service department may wince.

There are two related projects, Wine on Linux and Odin on OS/2, that are API redirectors/replacements. These allow Win32 programs to run on their respective host systems.

While they are definitely works in progress, they already allow many Windows 95/98/2000/ME/NT programs to run.

I have Mr Jameson's Proteus electronics design system running on a preview version of eComStation – the upcoming version of OS/2 – with a daily build of Odin. Is it perfect? No. I have to manually refresh sometimes as repainting still has bugs, and panning by moving the mouse while holding down the shift key doesn't work. But otherwise it functions. Not bad for an eComStation beta and an Odin pre-alpha. Can't wait for the GA's.

A second purpose is to provide a common API that programmers can use to produce programs that run on Win32, OS/2 and Linux. Attempts at this sort of thing have been tried in the past, but there are a lot of volunteers working on this *and* several large companies cheering and throwing services their way.

Daniel Carroll
Principal Cyclotron Engineer
King Faisal Specialist Hospital & Research Centre
Riyadh

John replies:
Daniel is quite correct in saying that multi-platform GUI/API solutions have been tried in the past. In the days when there was need to support MS-DOS, Windows and Mac OS we looked at various commercial solutions.

Unfortunately, everything on offer was utterly dire – either failing to work properly, or restricting the programmer to a poor *subset* of features available on the various platforms. We ended up establishing our own abstraction of the Windows API.

It is the experience of the time/effort devoted to this that makes me dread a return to a multi-platform marketplace.

On the other hand, were Linux to end up providing a Win 32 compatible API, this would give it a realistic chance to become main-stream because users and developers alike could then migrate without major disruption.

Perhaps the first step would be for the US authorities to force Microsoft to make Win32 an open standard? Such a development would also force a degree of stability which would do wonders for the reliability of application software.

John Jameson
Managing Director
Labcenter Electronics

About valve substitutes

In the May issue's letters column, Ranulph Poole asked if anyone knows about commercial solid-state valve substitutes.

In the Italian magazine *Sperimentare*, November 1977, page1141, I found the following replacement suggestions for the BC-221 apparatus and others, to be made placing the semiconductors inside the empty base of an old tube.

"For the tube 77 use a 2N3085 JFET, source to pin 5 of the tube base, drain to the pin 2, gate to a lug (that I think that it is applied to the tube base, and is intended to replace the tube cap); other pins not connected.

Again for the tube 77, using a 3N128 MOSFET, source to pin 5 of the tube base, drain to the pin 2, gate to a lug, and screen/substrate to the pin 1; other pins not connected.

For the tube 76, use a JFET C610, source to the pin 4, gate to the pin 3, drain to the pin 2; others pin not connected.

For the tube 6A7 place a 2.2kΩ resistor to the pin 6, a 270pF capacitor to the lug, join together the free ends of the resistor and of the capacitor.

Place a first MPF107 JFET, source to the joint just made, gate to the pin 5, drain to the pin 4. Place a second JFET MPF107, source to the pin 6, gate to the lug, drain to the pin 2; others pin not connected."

However, I think that these are neither commercial replacements nor truly direct replacements, as the power supply will have to be reduced.

I think that 'direct' replacements for rectifier tubes have been made by International Rectifier, but I could not find any reference to them.

● **Battery pack short-circuit protector:** the circuit published on page 382 of the May 2001 issue does not seems to work correctly as it is drawn.

The current of the main cell pack is flowing through the 3.6V Li-ion battery. The "Cellpack V_{cc}" connection should go to the left terminal of R_{sense}, that is, to the negative pole of the Li-ion battery.

Ezio Rizzo
Italy

High-voltage amplifier

With reference to Arni Ingvarsson's letter in the May issue describing his high-voltage transistor power amplifier to drive electrostatic speakers, why use transistors?

If you want to run a 5kV HT line then surely a valve output stage would be a more appropriate choice? After all, the Quad 22 valve amplifier was designed to drive the complex load presented by electrostatic speakers without the threat of instant failure posed by those expensive three-legged fuses.

With a 5kV HT supply and a push-pull output stage you could dispense with the output transformer and use AF chokes for anode loads, the speaker feed being taken directly from the anodes. However, I must admit I don't like the idea of external speaker wiring carrying five thousand volts!

Quad amplifiers had the usual low impedance output intended to drive either conventional speakers or the Quad ES units. With the electrostatic speakers, the low signal voltage from the amplifier was stepped up to the required level by a transformer in the ES speaker unit.

Regarding Ranulph Poole's letter in the same issue, I remember plug-in solid-state replacements being available for the valved video output stages in early seventies Pye colour TVs. The video output PCB was densely populated with components including hot running power resistors.

The heat from those together with the four closely spaced valves fitted on the panel gently toasted the Paxolin until it was sufficiently charred to allow heavy leakage currents to flow.

It was not unknown for the solder securing the valve holders to the print to melt. The only option was to replace the complete panel.

The solid-state replacement modules were supposed to be cooler running, and used a high-voltage FET. The module was built on a B9A plug and intended to be quite literally a plug-in replacement.

Unfortunately they weren't very reliable; although it must be borne in mind that high-voltage semiconductor technology was in its infancy – well, OK, its adolescence – at the time. Not only that, the dissipation of the power transistor was the same as the anode dissipation of the valve it replaced. The only saving was in heater power, which in the application under discussion made only a minor contribution to the problem.

When I was in my teens back in the sixties, I used to play around with what were even then old valve radios. When the 6V6 failed in my bedroom set I simply plugged in one of its cousins – a 6L6. The 6V6 shares an identical pin-out with the 6L6 and 6F6.

While the 6V6 and 6F6 are pentodes the 6L6 is a beam tetrode, indeed the famous 807 transmitter PA bottle is basically a 6L6 with a top-cap.

Substituting one of the alternatives for a dead 6V6 should be done with care, as the heater currents are somewhat higher. The cathode bias resistor's value also needs to be changed. Heater current and grid bias voltages, for 250V on the anode, are as follows: 6V6, 0.45A, –12.5V; 6L6, 0.9A, –4V; 6F6, 0.7A, –16.5V.

It's even possible to fit an EL34, but the heater current is a hefty 1.6A and as well as changing the cathode resistor value, pin 1 (g3) must be connected to the cathode.

Pete Roberts
Via e-mail

Mysterious EMC

I would like to commend Ian Darney on his attempt to demystify EMC in the May 2001 edition of *Electronics World*.

It is a subject that for too long has been viewed as a black art. Ian is quite right when he identifies that we need to focus attention on the too often neglected electrical properties of the circuit interconnections. I would commend readers to develop the type of thinking that Ian recommends and I, for one, will seek to learn from his evident experience in the field

However I want to take issue with him over his view of the equipotential plane and particularly his assertion that there is a voltage drop along the ground plane equal to that in the signal conductor. It is the correct approach to consider the action of the ground plane by imagining an image of the conductor, but this actually leads to the opposite conclusion.

The image conductor is to be imagined to exist the same distance from the ground-plane surface, but on the opposite side of it.

At every point on the image conductor a potential exactly opposite to that on its real counterpart and a current equal in magnitude and flowing on the opposite direction is visualised. If you now consider the electric field that exists between this image conductor and the real conductor, it will be at a potential of zero midway between the conductors throughout their length. This confirms that there is not potential difference across the ground plane.

Indeed it is actually the need to fulfil the requirement that there be no potential across the ground plane that leads to the development of the image representation. It is also incorrect to assign the image inductance and resistance values to the current flowing in the ground plane. They do have to be assigned to the image conductor, but only along with assigning an image voltage source and load resistance hence maintaining the resultant current flow in the real conductor to the same as it would be if there were a true short circuit between the ground plane end of the load impedance and the ground plane end of the source.

Ian is correct in saying that the return current flowing in the ground plane is restricted to an area immediately beneath the conductor – at high frequency only. This would suggest some inductance associated with this flow, but not equal to that of the forward conductor. It may well be that Ian has experienced voltage drops between points designated as ground. If so this represents an inadequacy of the ground plane and in these circumstances the image representation would break down.

A particularly important property of the ground plane is that it should be continuous under the forward conductor's path. Any break in the ground plane continuity will force the return current to take a path other than directly under the forward conductor this can greatly increase the area enclosed by the forward and reverse currents and hence compromise the EMC performance.

At first I was surprised to read Ian's recommendation that the situation with a conductor over a ground-plane be improved by connecting a ground return wire in parallel with the ground plane. In my minds eye I visualised a conductor running, say, 4in (200mm) over a ground plane with source and load connected. I then envisage laying a conductor from ground plane end of the load across the surface of the ground plane to the ground plane end of the source. I am pretty sure that a negligible current would flow in this wire. It would be easy enough to check this with a suitable set up including an rf current probe if someone wanted to try it.

It then occurred to me that Ian probably intended that the return conductor be run immediately adjacent to the original conductor and not along the ground plane. The effect of this is indeed interesting. This second conductor now forms a short circuited secondary of a transformer of which the original conductor is the primary. As a result a current will be induced in this conductor opposite to that in the original conductor hence generating an opposing radiated field and reducing emissions (and by reciprocity improving immunity).

The effect will be limited by the effectiveness of the coupling between the two conductors. If perfect there would now be no current flowing in the ground plane at all. Indeed the effectiveness would be improved by disconnecting the ground plane at the load (or source) end and so forcing all the return current through the second conductor. A further improvement can then be achieved by twisting the conductors together.

Could I appeal to Ian to consider writing a further article – or the editor to reference an existing article if such exists – dealing with the importance of the areas being enclosed by the loops formed by forward and reverse connections from source to load? Visualising these can be of great help in understanding what reduces emissions and improves immunity. ■

Philip Williams BSc CEng.
Harlow
Essex
Via e-mail

E-series anomalies

Referring to the letter on E series resistors in the March issue, I do not understand how Mr Gundry calculated the errors in the E12 series. I think that the best method to clarify things is to reproduce the calculations here.

The first column shows the theoretical values using the 12th root of 10, rounded to 2 decimal places. In the second column is results from the 1st column rounded to 2 digits. The third column contains the E12 series. Numbers in bold type are the ones that differ.

10	10		10	
		–0.95%		
12.12	12		12	
		3.17%		
14.68	15		15	
		–0.95%		
17.78	18		18	
		0.88%		
21.54	22		22	
		–2.45%		1.3%
26.10	26		**27**	
		1.588%		–0.88%
31.62	32		**33**	
		–1.98%		–2.45%
38.31	38		**39**	
		–0.083%		–0.528%
46.42	46		**47**	
		0.48%		–1.65%
56.23	56		56	
		0.228%		
68.13	68		68	
		0.748%		–0.47%
82.54	83		**82**	
		–0.55%		0.66%

The percentages in between are the ratio errors between neighbours compared with the 12th root of 10. As you can see, there is no advantage in the E12 series compared with the mathematically built series. On the contrary, the choice was unfortunate because it avoids easy computing of resistor values in some programs.

The E24 series has the same problem, as it was made compatible with E12. Series E48 and E96 were built from scratch and are mathematically correct.

As an example, a very useful program that runs on an HP48 or similar in shown as **Fig. 1**. It transforms any number no the nearest value of the E48 series. For E96 change "48" to "96":

Francesc Casanellas Chartered Engineer, MIEE, Senior MIEEE
Barcelona
Spain

Fig. 1. Program for determining the E48 value nearest to a given number using an HP48 or similar.

/'10^(1/48)' ® x y 'y^IP(LN(x*√y)/LN(y))' EVAL -3 RND ™

With reference to the discussion in your journal about E-series resistors, I first made the acquaintance of this series of values about 54 years ago. They were manufactured by Erie and supplied in long flat green boxes in 1/2 and 1 watt ratings and were made of carbon.

I have always understood that the E designation originated because the resistors were made and supplied by (E)rie and that those particular ohmic values were arrived at by the constraints imposed by the post-war period.

I believe that although a full range was desirable, it was not practicable. An excellent compromise was derived by some genius at Erie and marketed accordingly.

As an aside, I have in my possession a tattered and fragile copy of P. H. Brans' Radio Valve *Vade Mecum* dated 1946. Amongst many other things, it gives details of some German valves that contained, within the one glass envelope, three valves coupled internally by capacitors (anode to grid) and decoupled by built in resistors (grid to cathode).

Can these be considered to be the first attempt at integrated circuits? As well as having a manufacturers designation they were called Wunderlich valves.

Jim McDermott
Via e-mail

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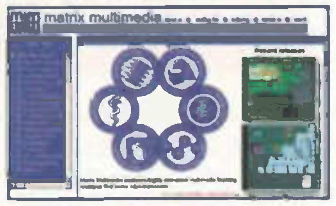

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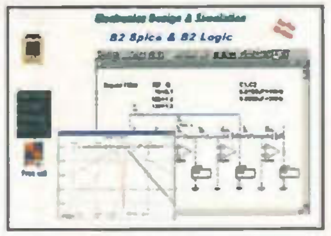

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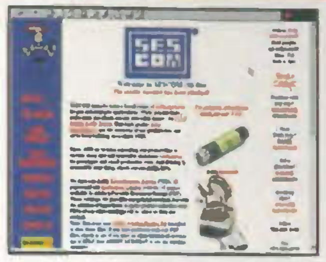

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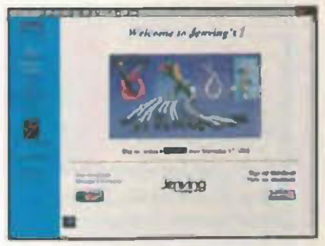

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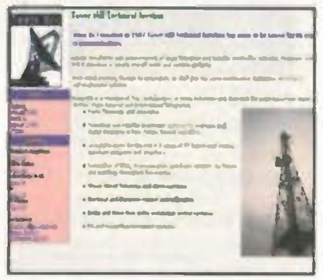

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UK MAILING LIST GROUP
http://www.egroups.com/list/uk tvrepair

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

REED CONNECT
http://www.reedconnect.net/

Another free Internet access site, this time from Reed Business Information. However the site possesses a useful UK People and Business Finder, with an e-mail search. There's also business news and local information, and some good links to directory sites.

REPAIRWORLD
http://www.repairworld.com

Repairworld is a sophisticated US based fault report database which is updated bi-weekly. It operates on a subscription basis and describes itself as an "affordable solution for all technicians". You can see some samples of the material for free, monitors, VCR, DVD and Camcorders being of particular relevance to UK users. The site also provides a "chat room".

Pat Bunce on 020 8652 8339 or fax on 020 8652 3981. or e-mail: pat.bunce@rbi.co.uk

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Company name		Web address		

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*Additional pairs as part of the same order, only £19.24 each pair.

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- two x1, x10 switchable probe bodies
- two insulating tips
- two IC tips and two sprung hooks
- trimming tools

There's also two BNC adaptors for using the cables as 1.5m-long BNC-to-BNC links. Each probe has its own storage wallet.

To order your pair of probes, send the coupon together with £21.74 UK/Europe to Probe Offer, Jackie Lowe, Cumulus Business Media, Anna Boleyn House, 9-13 Ewell Road, Cheam, Surrey, SM3 8BZ

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Probes

Total________

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Address ____________________

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Please allow up to 28 days for delivery

Specifications

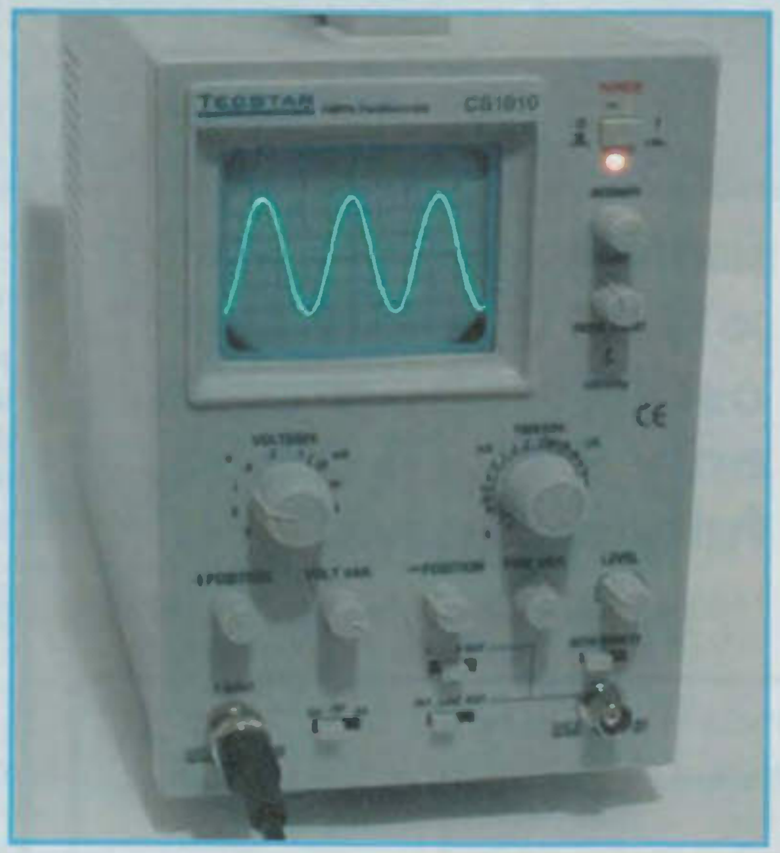
Switch position 1	
Bandwidth	DC to 10MHz
Input resistance	1MΩ – i.e. oscilloscope i/p
Input capacitance	40pF+oscilloscope capacitance
Working voltage	600V DC or pk-pk AC
Switch position 2	
Bandwidth	DC to 150MHz
Rise time	2.4ns
Input resistance	10MΩ ±1% if oscilloscope i/p is 1MΩ
Input capacitance	12pF if oscilloscope i/p is 20pF
Compensation range	10-60pF
Working voltage	600V DC or pk-pk AC
Switch position 'Ref'	
Probe tip grounded via 9MΩ, scope i/p grounded	

Reader offer

10MHz single-trace oscilloscope for under £200

The Tecstar CS1010 oscilloscope is easy to operate and highly reliable. It is an ideal instrument for research, production, education and development work. This is a compact and portable oscilloscope with a frequency bandwidth of 10MHz and sensitivity of 5mV/div to 5V/div. It is equipped as standard with a switchable x10:1 probe, extending the sensitivity up to 50V/div. The horizontal timebase is 0.1s/div to 0.1μs/div.

- Simple operation
- Portable, small and lightweight
- Excellent Triggering



Features

- 10MHz bandwidth
- High brightness
- Calibration output
- Variable-Y attenuation
- AC DC and Ground
- Auto, Normal and TV triggering
- Variable trigger level
- +ve or –ve Ext trigger modes
- Internal, External or Line trigger
- Complete with x1/x10 probes
- Low cost

Vertical system	
Sensitivity	5mV/div-5V/div±3%
Trimming ratio	2.5:1
Rise time	35ns
Bandwidth (–3B)	DC:0-10MHz AC:10Hz
Trigger system	
Int trigger sensitivity	1 div
Ext trigger sensitivity	0.3V
Ext. trig. input impedance	1MΩ, 30pF
Ext. trig. max. input	400V pk
Trigger sources	Int, Line, Ext
Trigger mode	Norm, AUTO, TV
Horizontal system	
Sweep time	0.1s/div-0.1ms/div, ±3%
Trimming ratio	≥2.5:1
X-Y mode	
Sensitivity	0.2V/div-0.5V/div
Bandwidth (–3dB)	DC: 0-1MHz AC:10Hz-1MHz
Calibration signal	
Waveform	Symmetric square wave
Amplitude	0.5V±2%
Frequency	1kHz±2%
CRT	
Display area	8 x 10div (1div=6mm)
Accelerating voltage	1200V
Display colour	Green
Power source	
Voltage range	220V ±10%
Frequency	50Hz ±2Hz
Power consumption	25W
Physical features	
Weight	3kg
Dimensions (H x W x D)	190 x 130 x 270mm
Working environment	
Working temperature	5°C to 40°C
Storage environment	–30°C to 60°C, 10-80%RH
Working altitude	2000m

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*£159.00 ex. VAT, £6 UK carriage, £14 overseas.

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Software Engineer

PIC programmer required with 2+ years experience of embedded control applications. Salary to £25,000

For further details please send a CV or contact us in confidence as below:

Matthew Bramble, Technical Director, Audio Partnership PLC, Gallery Court, Hankey Place, London SE1 4BB.

t: 0207 940 2200 f: 0207 940 2233 e: matthew@audiopartnership.com w: www.audiopartnership.com

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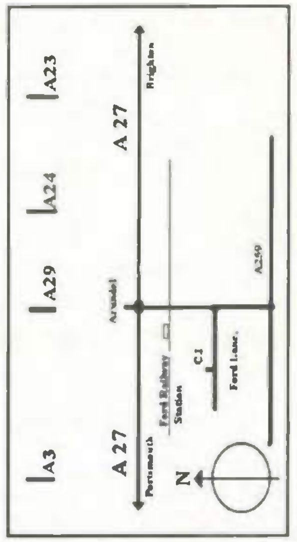

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112	113	114	115	116	117	118	119	120	121	122
123	124	125	126	127	128	129	130	131	132	133
134	135	136	137	138	139	140	141	142	143	144
145	146	147	148	149	150					
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505	506	507	508	509	510	511	512	513	514	515
516	517	518	519	520	521	522	523	524	525	526
527	528	529	530	531	532	533	534	535	536	537
538	539	540	541	542	543	544	545	546	547	548
549	550	551	552	553	554	555	556	557	558	559
560	561	562	563	564	565	566	567	568	569	570
571	572	573	574	575	576	577	578	579	580	581
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