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REGULARS

Doug Self power amplifier PCBs are now available. Ordering details on page 358

IN NEXT MONTH'S ISSUE
Selective calling for mobile radio. The telephone dial or keyboard and its power to summon up a specific connection is taken for granted. Not so with radio. Individual radio services have each evolved their own calling codes; the aeronautical service uses a different set of codes from maritime networks. James Vincent presents both protocol and circuitry for selective radio calling

THE JUNE ISSUE IS ON SALE FROM MAY 26

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No place for initiative

Industry minister Tim Sainsbury has indicated that there will be no more government cash in the pot to stimulate the UK's microelectronics industry when the £3m *Microelectronics in Business* initiative has run its course. This is both right and wrong.

It is right because the description "electronics industry" covers such a wide range of activity that it would be positively wrong to single out one facet for special economic treatment. After all, "Microelectronics in Business" strikes me as a particularly fatuous subdivision; why not "software design tools for industry" or, perhaps, "tax credits on automated test equipment"?

But why concentrate on the sharp end of the product design process? Companies and consultancies pay their staff to be well informed about design and process technology and, if these people can't hack it, they should consider a job in hair dressing or the legal profession. The Government has no place in persuading people to do what should naturally be part of their job.

It makes sense to stimulate the health of the infrastructure by ensuring a supply of well educated and motivated people who leave college with a training that fits more closely the needs of employers. Seed money should be used to forge additional links between education and industry: personnel exchange programmes between the two, joint sponsorship of final year projects and applications oriented postgraduate research come to mind.

I appreciate that such things already happen but the electronics industry would benefit from more of them, much more so than by the simple funding of a few non-recurring engineering charges.

But if business currently fails to appreciate the importance of modern electronics in general design – which the Government seems to think is the case – then it is for reasons other than a specific lack of knowledge about asic chips. I suggest that industry generally fails to appreciate the worth of competent and qualified electronics engineers, an indifference which it matches in remuneration. This shortfall is definitely the responsibility of industry, not the Government.

It has to look to itself for a solution. It could begin by appointing engineers to control the plague of lawyers and accountants in company boardrooms.

When industry (and government) is prepared to pay engineering graduates (and maths and science teachers) more than this deeply unimaginative group of business professionals, we will begin to see a real change in our industrial fortune. **Frank Ogden**

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NEWS

Radar helps police look for bodies

As the horror of 25 Cromwell Road unfurled, one could not help but be impressed with the apparent ease in which long decayed remains were discovered.

Thermal imaging systems can find living people under mounds of rubble and debris after plane crashes or earthquakes. But in Gloucestershire the remains were long dead.

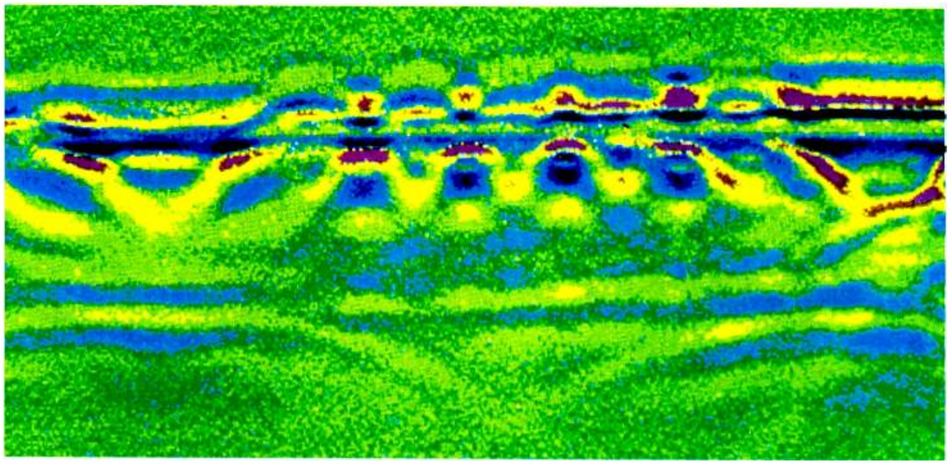
Searchers reverted to a method designed during the Falklands war for finding plastic mines. Called surface penetrating radar it has since found applications in archaeology, civil engineering and searching for victims of snow avalanches.

A major advantage of this type of radar over other nondestructive testing methods such as ultrasound is that an antenna can be used that is not in physical contact with the material being tested. It is therefore possible to scan large areas quickly.

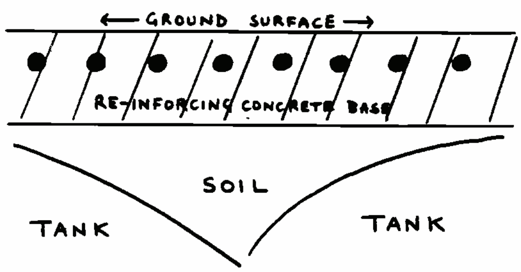
Wideband radar pulses directed into the ground give rise to back-scattered energy from a target buried in a material. Anomalies in density or dielectric constant produce reflections which are detected by the receiving antenna. A conventional radar is stationary and its target mobile. Surface penetrating radar moves and its targets are stationary.

A typical system consists of an antenna head, transmitter-receiver and a purpose designed real-time processor, storage and display unit.

Transmitting and receiving antennas are housed in an ABS plastic antenna head measuring about 50x50cm. This assembly is moved in a regular pattern over the surface of interest. A pulse of energy – as narrow as



A typical colour enhanced image using surface penetrating radar. This civil engineering application shows petrol tanks buried under reinforced concrete. By comparing the line drawing and the colour image you can see the concrete layer shown by the false colour line near the top. The dark patches denote the reinforcement. Below that deviations in the soil (coloured green) provide the outline of the tanks.



1ns – is repetitively transmitted into the sub-surface material at an average power level of 5mW.

A PC based processor provides an image of the internal composition of the structure in real time. A trained operator can see what is happening as the scan progresses. The system produces an enhanced image with different types of structure density being shown in different colours.

This was good enough to give a 100% hit rate in Cromwell Road because the structure of the foundations of the building was known as was the type of soil.

In use, each measurement consists of about 256 samples, each to 12-bit resolution, and each averaged four times before being transferred every 6ms to the processing unit. The time interval between samples may be set to provide a range between 6.4 and 1048ns, corresponding to depths of 0.3 to 53m. In the wet soil conditions of the UK – which are heavily attenuating – a depth of 2 or 3m is achievable.

As well as presenting the data as a cross section, the radar can build up an area scan by taking a grid of correctly registered line scans. Selected depths may also be viewed.

The choice of bandwidth dictates the compromise between depth and detail. The equipment covers the 50MHz to 5GHz range and involves several octaves of bandwidth. Low frequencies are normally used for deep probing, say 50m, and high frequencies for shallow probing.

The wavelength of the transmission decreases as the velocity of propagation slows within the material, a process governed by the relative dielectric constant of the material. If the propagation velocity can be measured or derived, an absolute measurement of depth or thickness is possible.

Flexible battery charged to plastic

Researchers at Bellcore say they have created a revolutionary new type of battery that is as flexible and as light as a plastic credit card yet delivers high-energy and is rechargeable.

The lithium-ion battery offers equal or better performance than rechargeable nickel-cadmium and lead-acid batteries at half the weight. Bellcore says that the flexible battery can be shaped for virtually any application from small hand-held video games to large sizes capable of powering electric cars.

"This is the first plastic, rechargeable battery," says Jean-Marie Tarascon, leader of Bellcore's battery team. "It does not contain toxic metals, like lead, cadmium, mercury or cobalt. What's more, no liquid will leak out if the battery is cut or punctured, making it safe to install and use."

Jeff Dahn, a battery research expert and professor at Simon Fraser University says, Bellcore's plastic lithium-ion battery appears suited for portable electronics applications where a lightweight, high energy, thin battery is preferred."

The Bellcore battery does not have any liquid electrolytes and can be recharged hundreds of times without losing capacity at the same rate as comparable liquid lithium-ion batteries. The batteries' elements are permanently bonded together and covered in a waterproof barrier.

Bellcore, which is funded by US telephone companies, says that one of the first applications of the new battery will be to provide back-up electricity for central switching offices. Bellcore says it will license the battery technology to other firms.

Once the signal has been received, it has to be processed for display. This requires deconvolving the unprocessed waveform in time and space. With surface penetrating radar, the initial transmitted impulse becomes convolved with a series of responses due to the antenna, the ground and so on. The only part of the signal of interest is the target impulse response. The processing techniques used to isolate this are similar to those used in seismic exploration, ultrasonic imaging and medical tomography.

The equipment was designed and operated by a team from ERA Technology. Speaking about the Cromwell Road events Keith Cheshire from the company said: "We knew what the subsoil was like. If you bury something in soil, the soil gets disturbed so its not uniform."

Another advantage the searchers had came from the way a buried body decays. The missing flesh is not necessarily replaced by the surrounding soil. This especially true for areas like the skull where, once the brain has rotted away, a tell-tale void remains which is easily picked up by the radar. When voids are discovered, grid pattern analysis can be used to measure the size of the void and make judgements about its likely cause.

For searches more complex than Cromwell Road, the data can be stored and post-processed to get rid of clutter and enhance interesting parts. Up to 2Gbyte of data can be stored by the assembly. Fully processed data can be continuously displayed at a measurement rate of one point every 300mm at a traverse speed of 25km/h.

Post processing is useful in archaeology for detecting ancient burials and artifacts, or in civil engineering for assessing the condition of roads or tunnel linings.

Looking for pipes and cables may be done on the spot as can finding people trapped under snow avalanches. This last task is particularly easy for the system as snow is a uniform conductor which reflects well from changes in conductivity.

Ground penetrating radar has more difficulty with collapsed buildings because of excessive clutter and an irregular antenna datum.

Report backs digital TV

Channel 5, the proposed new TV channel, should be put on hold while the economics of launching a digital TV service are investigated, according to a report from Convergent Decisions, a specialist consultancy.

The report gives a boost to those in the BBC and elsewhere lobbying behind the scenes for channels 35 and 37 to be allocated to a new digital TV service.

A study produced last year by the Independent Television Commission suggested that the addition of these channels would not make much difference to the potential coverage of digital TV. But digital TV backers point out that channels 35 and 37 are unique because they are available across the whole country. This means they could be used for a 'single frequency network', where one frequency is used to transmit the same digital TV channel across the whole country. Existing analogue terrestrial TV stations each require 44 frequency channels for national coverage.

●IBM Microelectronics unveiled a single chip MPEG-2 decoder at last month's National Association of Broadcasters conference in Las Vegas. The firm joins a highly competitive market already involving several major semiconductor firms including C-Cube, SGS Thomson and AT&T.

"You will not find anyone with a better price at the end of this year," said an IBM spokesman. He expects the price to fall to $30 or less in volume production.

Opto amplifier with bright prospects

BT has claimed a breakthrough in the transmission of broadband services direct to the home with the development of an optical amplifier which will work with optical pulses with a 1.3μm wavelength.

BT claims that the optical amplifier, developed in conjunction with Hewlett-Packard, will boost the optical signal by a factor of 1000 (30dB). According to group leader Dr Colin Millar, that will make it possible to transmit a 5Gbit/s data stream over 100km. "But these are very new results and it will be between two and five years before the amplifier is put into commercial service," said Millar.

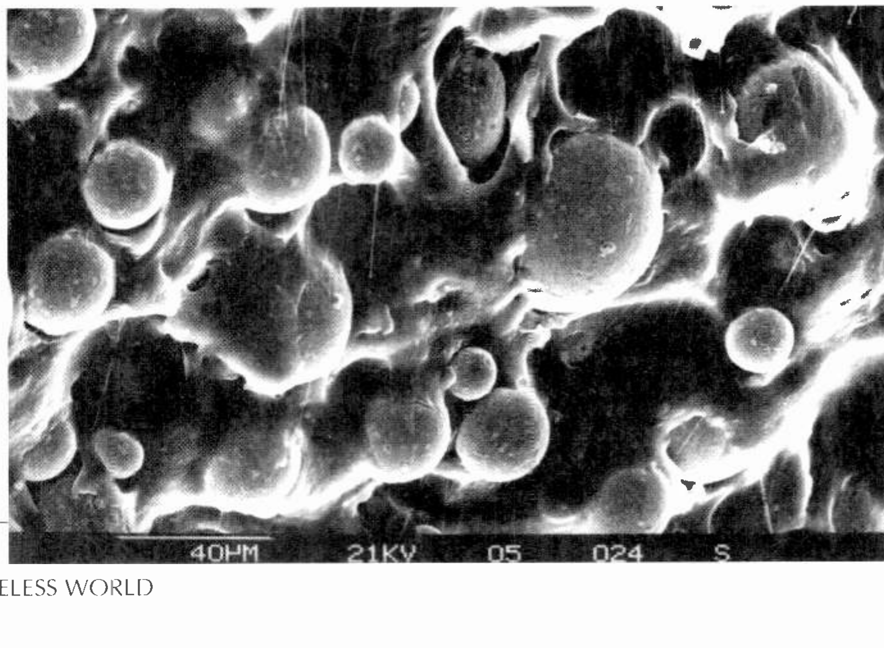
The significance of this development compared to erbium-doped optical amplifiers, which are already in commercial service, is that because it boosts light at a wavelength of 1.3μm, it will work on the optical fibre systems commonly used in BT's national network. Erbium doped amplifiers operate at the less commonly used 1.55μm optical fibre window.

Operation at 1.3μm has been achieved by using a new type of optical fibre known as fluoride fibre. The amplifier incorporates a 20m length of fluoride fibre which has been chemically doped to promote energy transfer from one wavelength to a higher one. The dopant used is praseodymium (Pr^{3+}).

The optical energy is transferred from a local source at 1.047μm to light at a wavelength of 1.3μm which may be used to boost the carrier signal. The energy transfer is mediated by the praseodymium ions which are incorporated into the core of the fluoride fibre. The local source is a Nd:YLF solid state laser, pumped by a low power diode laser, which emits the light at a wavelength of 1.047μm. The energy in light of this wavelength matches the energy needed to excite the Pr^{3+} ions in the fibre.

The two optical signals, one from the local source and the signal to be amplified, travel together along the length of doped fibre. The photon energy of the 1.047μm pump light matches the energy difference between the low and excitation energy levels of the Pr^{3+} dopant and the light is absorbed by the ions in the fibre. The excited ions are unstable, and the presence of the signal light at 1.3μm is sufficient to force the Pr^{3+} ions to fall to a lower level and emit additional light at the

Spheres of influence: This electron microscope view shows an electrically conductive light weight elastomer comprising silver coated glass spheres in a matrix of silicone rubber. The material can be used for screening electromagnetic interference from telecommunications equipment, UHF and microwave radio systems, and computer controlled devices. It is claimed to provide 100dB of shielding between 20MHz and 10GHz. James Walker 0483-757575.



signal wavelength of 1.3μm.

The BT group's amplifier uses 500mW pump power to amplify the information signal by 30dB.

One problem BT found with using Pr^{3+} doping ions in conventional silica fibre was that the excitation lifetime of the ions was too short for the secondary excitation to work. Hence the fluoride fibre with its longer excitation lifetime.

Although the 100μs lifetime in the fluoride fibre is still along way short of the 10ms lifetime of erbium dopant ions in conventional 1.55μm optical amplifiers, it was sufficient to support amplification.

The optical amplifier, which will work on BT's three million km of 1.3μm fibre, will enable the operator to increase capacity and reduce equipment costs on its trunk network. Using the amplifier on a single high bit rate carrier, BT is confident that it can support 5Gbit/s data transmission over 100km of fibre. This would extend the reach and capacity of BT's trunk network which currently uses 140 and 565Mbit/s data rates over 30km spans.

In addition, the amplifier's 40nm bandwidth will enable multiple signals, generated by wavelength division multiplexing, to be amplified by over 20dB. **Richard Wilson**, *Electronics Weekly*

PowerPC: "Pentium performance at half the price"

Growing support for the new PowerPC microprocessor from Motorola and IBM was demonstrated at Cebit last month, with systems on display from Germany's three largest indigenous PC makers, and new motherboard reference designs from Taiwanese and American companies.

The German trio of Vobis, Escom an Peacock are the first mass-market PC makers apart from Apple, to demonstrate systems containing the new chip. Although none is shipping now, all were running early versions of Windows NT on the new microprocessor.

Peacock showed a server machine based on the Sandalfoot reference design from IBM. Asked why Peacock had opted for PowerPC, Peacock's marketing manager Christof Basener said: "We are not married to anyone in microprocessors. The PowerPC gives the performance of Pentium at half the price." **David Darcy**, *Electronics Weekly*

PCBs for Douglas Self's power amplifier series

Circuit boards for Douglas Self's high-performance power amplifier are now available via *EW+WW*.

Detailed on page 139 of the February issue, Douglas Self's state-of-the-art power amplifier is the culmination of ideas from one of the most detailed studies of power amplifier design ever published in a monthly magazine. Capable of delivering up to 100W into 8Ω, the amplifier features a distortion figure of 0.0015% at 50W and is designed around a new approach to feedback.

Designed by Douglas himself, the fibreglass boards have silk-screened component IDs and solder masking to minimise the possibility of shorts. Sold in pairs, the boards are supplied with additional detailed constructional notes.

Each board pair costs £45, which includes VAT and postage, UK and overseas. Credit card orders can be placed 24 hours on 081 652 8956. Alternatively, send a postal order or cheque made payable to Reed Business Publishing to EW+WW, The Quadrant, Sutton, Surrey SM2 5AS.

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RESEARCH NOTES

Smallest piece of metal in the world

A single atom of a metallic element behaves very differently from the bulk metal. But at what size does that transformation start to happen? Workers at the Universities of Leiden in Holland and Essen in Germany have been searching for an answer which could have important ramifications for electronic devices.

Most of us are familiar with some of the strange quantum effects that occur as structures get smaller and smaller. Many electronic and opto-electronic devices depend for their function on such counter-intuitive phenomena as electron tunnelling.

But it is not just that ultra-small structures behave in a different way; in many respects they *are* different.

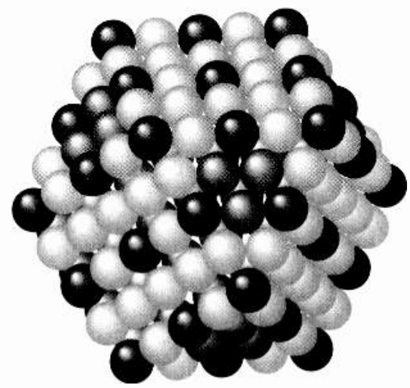
Any school child knows that one atom is the smallest unit of a chemical element. But if scientists were able to isolate a single atom of, say platinum (which is virtually possible thanks to scanning tunnelling microscopy), it would certainly not look or behave like a metal.

Bulk metallic properties depend essentially on the highest energy bands of electrons – the conduction and valence bands – from which electrons are shared freely between all the atoms in the lump. It is these relatively free electrons that are responsible for such features as metallic appearance and electrical conductivity. But if a lump of metal were progressively chopped up into smaller pieces, there would come a point when the electrons would lose their mobility and the metal would stop being a metal.

The team from Leiden and Essen performed experiments (*Nature*, Vol 367, No 6465) to try to discover the number of atoms below which this metal to non-metal transition occurs. Reducing the number of atoms in a lump of platinum was achieved, not with a knife, but by clever chemical synthesis of platinum cluster compounds.

Under professor Leenert de Jongh, the researchers created a whole range of compounds built round a central polyhedral cluster of platinum atoms. The atoms exist only in complete onion-like shells with well defined numbers. The Pt_{309} cluster is a four-shell member of an *n*-shell magic number series that goes 1, 13, 55, 147, 309, 561...

Using Mössbauer spectroscopy to examine each shell of the platinum onion, de Jongh and his team were able to discover something of the electronic environment of the atoms in each shell – this being a sensitive test of 'metallicity'. What they found was that the electrons associated with the 147 atom cluster behaved with the same freedom as they do in a piece of bulk platinum: a 147-atom chunk of platinum behaves as a metal.



At somewhere below 147 atoms, platinum clusters stop behaving like a metal.

But in an earlier experiment, with gold clusters, de Jongh found that a 13-atom cluster did *not* exhibit metallic properties. So he concludes that, somewhere between 147 and 13 atoms, there is a transition between metal and non-metal.

It all may look extremely esoteric and theoretical. Until we remember that the properties of very small quantities of materials are becoming increasingly important, not just for electronic devices, but also for environmentally important chemical devices such as catalysts.

Shooting at the stars

Jules Verne's idea of shooting a satellite into space from the muzzle of a gun, looks like becoming science fact, thanks to the efforts of a group working at the Lawrence Livermore National Laboratory in California.

The continuing attractiveness of Verne's idea, dreamed up more than a century ago, stems from the extraordinary inefficiency of today's rockets, 95% of whose take-off weight consists of fuel. To put a kilogram of human into orbit costs about \$20,000 – or about half that for an unmanned launch.

As John Hunter of LLNL put it rather graphically on a BBC *World Service* programme: "Right now, the cost of putting any object into low Earth orbit is the object's weight in gold."

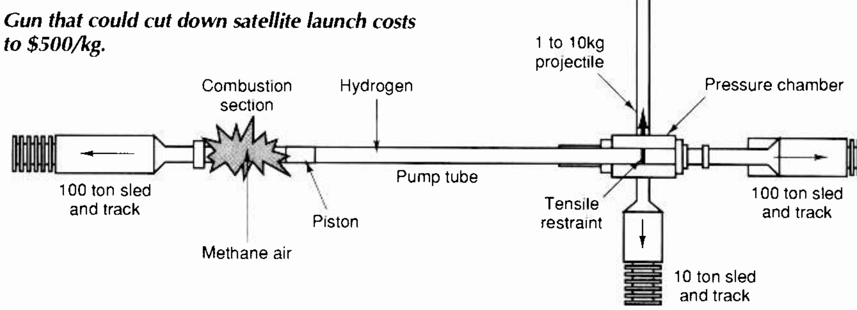
Using a huge gun to shoot an object into space has a wonderful simplicity about it, at least from the theoretical point of view. It would have far fewer moving parts than the now-standard three-stage rocket.

Verne clearly had in mind something like the ill-fated Iraqi supergun, powered by conventional gunpowder. All very simple — until you try the calculations. Even back-of-envelope sums show that to reach escape velocity with a conventional type of gun would require impossibly-oversized engineering.

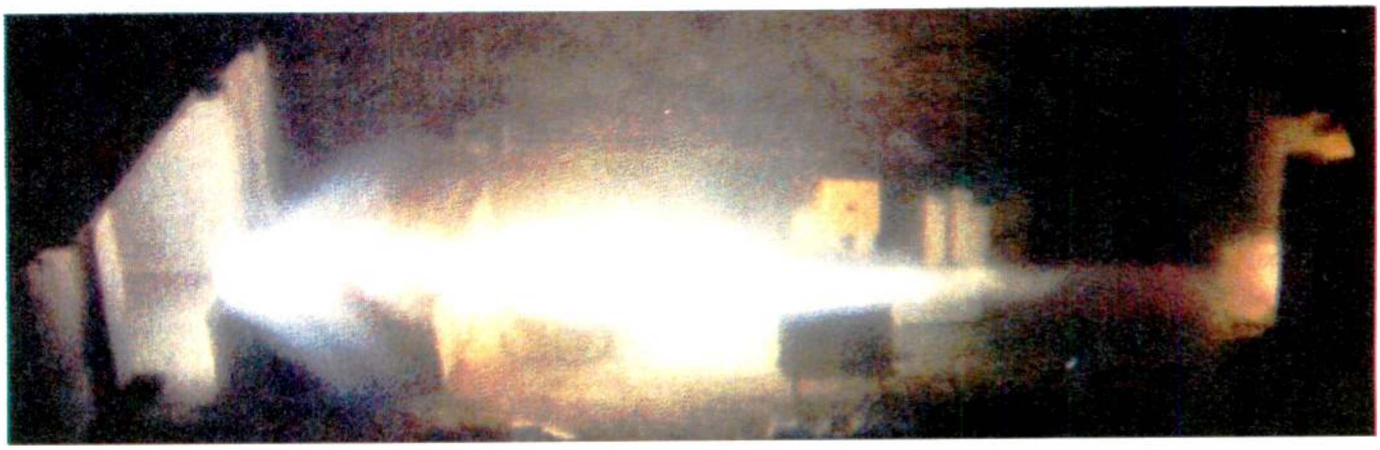
So Hunter and his team at LLNL have adopted a different approach, albeit with slightly more modest immediate objectives. They have developed what amounts to the biggest air-pistol in the world, extending the principles used in current high-velocity laboratory gas guns.

The basic idea behind such guns – many of which can accelerate a rifle bullet-sized object to 25,000mile/h – is to compress hydrogen gas behind a rupture valve which suddenly releases the gas, propelling the projectile down a barrel.

The LLNL team, realising that a scaled up version of such a gas gun would be too heavy to point skywards, have developed a two-stage version called Sharp (super high altitude research project). Sharp has an 82m pump tube along which a one tonne steel



Gun that could cut down satellite launch costs to \$500/kg.



Supersonic launch from a light-gas gun, right and a projectile – a scramjet – in flight at Mach 8 just before entering a retaining bunker, above. In-flight glow is caused by the engine's hydrogen fuel and air friction. Flames on the right are hydrogen used to accelerate the jet along the 4in launch tube.

piston is propelled by a methane/air charge. This rapidly heats the hydrogen in the tube, compressing it to 7000lb/in^2 and rupturing a tensile restraint. The hydrogen then rushes into the launch tube, driving the projectile up the tube at 9000mile/h (10 and 100t sleds are designed to absorb the recoil forces).

So far projectiles up to 10kg have been flung into the hillside with the launch tube horizontal, though the team are hoping to conduct vertical launch tests at sites such as the Vandenberg Air Force Base in California. In the course of these tests they want to discover, for example, how their projectiles fare during their high velocity transit through the atmosphere. There would be little point in shooting satellites into space if they burnt out on the way up!

Parallel research is aimed at scaling up even this monster gun to something that would enable the team to launch a missile into orbit. Success will demand a yet-newer and more advanced approach, such as the development of a piston-less pump tube. Calculations show that while the necessary 3km-long launch tube could be made, the existing style of pump-tube exceeds current engineering limits. One idea being explored is a launch tube in which hot compressed hydrogen is injected at various intervals along the tube, just behind the passing projectile. This would provide progressive acceleration to around 25,000mile/h without exceeding engineering limits.

Obviously such a system, if it ever comes to fruition, would not be able to launch people (or even fragile satellites) into orbit. The acceleration would be destructively great. But as Hunter points out, much of the material that would be needed to construct a space station would not be at all delicate. He also estimates that, at a rate of one launch per day, the cost of shooting material into orbit with a gun would be a mere $500/kg – vastly less than with today's rocketry – with less than 10s needed to get into orbit.

Simple satellites, says Hunter, will be shot into space by the year 2001.

Electrons exhibit brittle behaviour

Metals are soft and malleable: non-metals on the other hand – whether elements or composites – are comparatively hard and brittle. Now scientists in California think they can use electronic behaviour to explain why.

At a superficial level, hardness is explained by the mobility of electrons associated with atoms. In metals these move relatively freely, whereas in covalent solids the electrons are tightly paired up.

At a structural level, the hardness of a material such as silicon is a feature of the slowness with which natural dislocations propagate through the material. Dislocations are lines in a crystal where the atoms are not perfectly arranged – a bit like a wrinkle in a carpet. For a material to deform plastically without breaking, dislocations must propagate relatively quickly, as they do in most pure metals. But, until recently, no-one has been able to understand (at the electronic level) why the propagation of dislocations is so slow in hard materials like silicon.

John Gilman of the Lawrence Berkeley Laboratory in California has now analysed how the electronic structure of silicon changes as a dislocation slowly moves its way through the crystal.

Gilman says (*Science*, Vol 261, 1436) that dislocation lines do not move in a concerted fashion; instead they move through kinks that lie along their length. The overall rate at which a dislocation moves is determined by the mobility of the kinks. For a kink to move, says Gilman, it has to separate a pair of electrons that lie in its way.

In the case of a hard solid like silicon, the relative slowness with which dislocations propagate is now seen to be a feature of the strength with which pairs of electrons are bound together. The greater the bonding energy, the harder the material.

By calculating the relationship between kink mobility and the electronic structure of a material Gilman has now been able to determine in detail why other important materials such as germanium and silicon carbide have the physical properties they do, and how these properties vary with temperature.

Disk technology's resistance to change

Scientists at Argonne National Laboratory in Illinois have established what they claim is a record for 'giant magnetoresistance'. The effect – an unprecedented change in bulk resistance as a magnetic field is applied – has been produced by manufacturing a superlattice of iron and chromium consisting of alternate layers of each material. The layers are about a million times thinner than a sheet of paper and are produced by magnetron sputtering — a process already employed industrially.

Using a cone-shaped electromagnet to apply a strong magnetic field to Argonne's record breaking giant magnetoresistant material. A 250% decrease in resistance is the result.

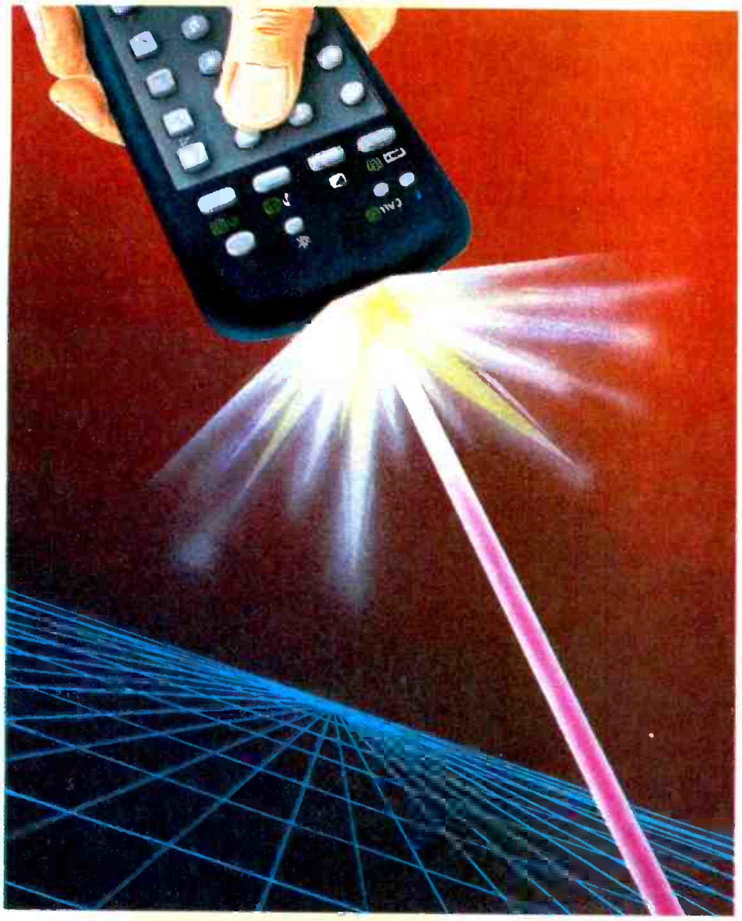
When a magnetic field is applied to the structure, cooled with liquid helium, its resistance drops by a factor of 2.5, beating the previous record of a two-fold reduction. Argonne's figure can also be compared with the 1-2% change typical of most magnetorestrictive materials.

The potential value of material exhibiting such a large change of resistance is in the measurement of tiny magnetic fields. Any given change in magnetic field produces a proportionately bigger electrical signal.

'Read' heads for extracting data from magnetic disks is one of the applications envisaged by the Argonne team. For this purpose, the new material is said to be about 75 times more sensitive than the nickel-iron alloys currently in use. The benefit would be that disks could be spun faster and memory access time reduced. Other possible applications include position sensors in robotics and industrial automation.

The Argonne researchers are now concentrating on creating materials that will exhibit the same giant magnetoresistive effect with lower magnetic fields. ■

Research Notes is written by John Wilson of the BBC World Service.



Opto-electronic components find wide usage in measurement, production control, scientific and consumer applications. Nearly all uses have a common requirement for photodetection in the presence of ambient lighting while others must preserve system bandwidth. Tore Nielsen presents a circuit designer's guide to opto-electronics.

OPTO-ELECTRONICS BY DESIGN

Light sensitivity is common to all semiconductors but the effect is optimised in the photodiode. It may be thought of as a current generator controlled by light. The current generator aspect is almost ideal: the main limitation is the parallel capacitor intrinsic to the relatively large semiconductor junction, and the diode characteristic if the diode is forward biased. The latter may even be considered a feature in some applications.

Silicon photodiodes are sensitive to wavelengths from 400nm to 1100nm which takes in the visible spectrum from 400nm to 700nm, but the sensitivity peaks at 900nm in the near infrared spectrum. Germanium photo devices operate from 400nm to 1800nm with a peak around 1500nm. They exhibit a lower sensitivity in the visible range than the silicon diode.

The photo current is proportional to the illumination for eight or more decades, and is downward limited only by the dark current (leakage current) which may be as low as 5pA for some diodes. **Fig. 1** shows the equivalent circuit for a photodiode

Photodiode interface

The photodiode is easily interfaced to an inverting amplifier (**Fig. 2**). The low input resistance at the inverting input effectively shorts the diode thus reducing the influence of the parallel capacitor. The output voltage, the product of $R_1 \times I_p$, is proportional to light and positive with respect to ground.

Typical device sensitivity would be about 80nA/lx (*see box for an explanation of the lux unit*). With feedback resistor R_1 set at 120kΩ the output voltage equals 10mV/lx, saturating the amplifier at 1000lx . The output could feed a comparator to initiate some action when the input exceeds a certain level, or the circuit could be used in a feedback loop to control illumination.

The classic use for an infrared receiver is to detect an optical pulse stream from a remote control handset or whatever. The absolute level of the illumination is unimportant and is rejected with a differentiating network. This is the idea behind **Fig. 3** where the photodiode is reverse biased and capacitively coupled to the amplifier. High pass filter time constant CR_2 could also be used to suppress low frequency interference. Reverse biasing may increase the sensitivity slightly for some photodiodes, but the main purpose is to accept the photo current emerging from the absolute level of illumination without affecting the function. If R_2 is 10kΩ , the circuit would accept some 15000lx before forward biasing the diode.

The bias resistor may inject noise from the power supply into the amplifier input with a gain of R_1/R_2. The photodiode power supply should be decoupled to reduce this effect.

The transient response of the circuit tends to be limited by amplifier performance rather than the response of the photodiode; risetime of the photodiode is typically 20ns when feeding a 50Ω load.

If the photodiode is connected to the high impedance input of the amplifier (**Fig. 4**), the photo current develops a voltage across the diode, creating a forward voltage across the junction. This varies logarithmically with current – some 60 to 100mV/decade – and would typically be about 450mV at 1000lx effectively compressing the useful luminance range to a convenient scale. A microprocessor fitted with a suitable a to d converter can convert this directly to light measurement units.

The following design originated from this basic circuit.

Light meter

The light meter in **Fig. 5** uses two photodiodes with different optical filters, to create an instrument for measurement in the visible light or near infrared spectrum. The sensitivity characteristics of the two photodiodes are shown in **Fig. 6**. The optical bandpass filters are centred at 550nm (green) and 950nm (infrared) and are approximately 200nm wide.

The instrument features a peak hold rectifier for pulse measurement and an output for an oscilloscope. The logarithmic conversion could have used the photodiodes, but the pulse response was too slow to be of practical use. Logarithmic conversion diodes were used instead. If the instrument is intended for operation below 1lx then the *1N4148* diode should be substituted with a low leakage type. The instrument is calibrated by the injection of a known (negative) current at the virtual ground node of the input amplifiers.

The Siemens *BPW21* has a sensitivity of 10nA/lx, so 1mA should read 100000lx on the instrument scale. The current is reduced in decade steps for the other readings down to 1lx. The *BPW21* is specified to 0.01lx.

The Siemens *BP104* has a sensitivity of 17μA at 0.5mW/cm^2 and 950nm. This corresponds to 3.4μA at an irradiance of 1W/m^2. For 1kW/m^2 the current is 3.4mA, and readings are obtained in decade steps down to 1mW/m^2.

Bright sunlight has the equivalent radiance of 1kW/m^2.

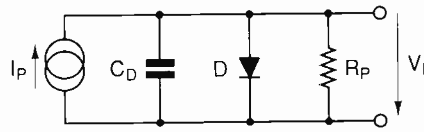


Fig. 1. Equivalent circuit for the photodiode. The capacitance is 10 to 500pF, and the resistance represents the leakage current.

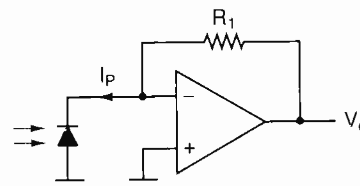


Fig. 2. Photo detector for high speed operation. Output is proportional to illumination level.

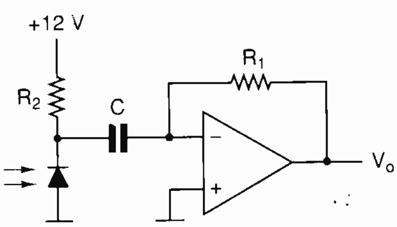


Fig. 3. Photo diode capacitively coupled to increase the amount of extraneous light accepted and to suppress low frequency interference.

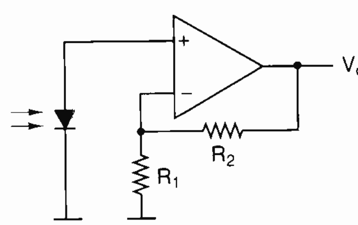


Fig. 4. Photo detector with output voltage proportional to the logarithm of the photo current for eight or more decades.

The circuit exhibits a temperature coefficient of approximately –0.5 %/°C, primarily caused by the *1N4148* diode forward voltage change with temperature. It thus requires some form of temperature compensation for serious use. The basic instrument read correctly over the range 1 to 5000lx compared to a Gossen luxmeter using a tungsten filament lamp source. Fluorescent lamps illuminating 2000lx read some 50% low.

This is because the *BPW21* is specified for use with a tungsten filament lamp. The photodiode sensitivity above 700nm adds extra energy from the filament lamp infrared radiation. This radiation is missing in the fluorescent lamp, leading to under reading.

No comparison against a reference has been performed in the infrared spectrum.

The infrared emitter diode

Popular infrared emitting diodes come in two sorts: 880nm GaAlAs diode and the 950nm GaAs diode. The emitted spectrum (**Fig. 7**) indicates a good match between a GaAlAs emitter and the unfiltered photodiode, and between a GaAs emitter and the IR filtered photodiode.

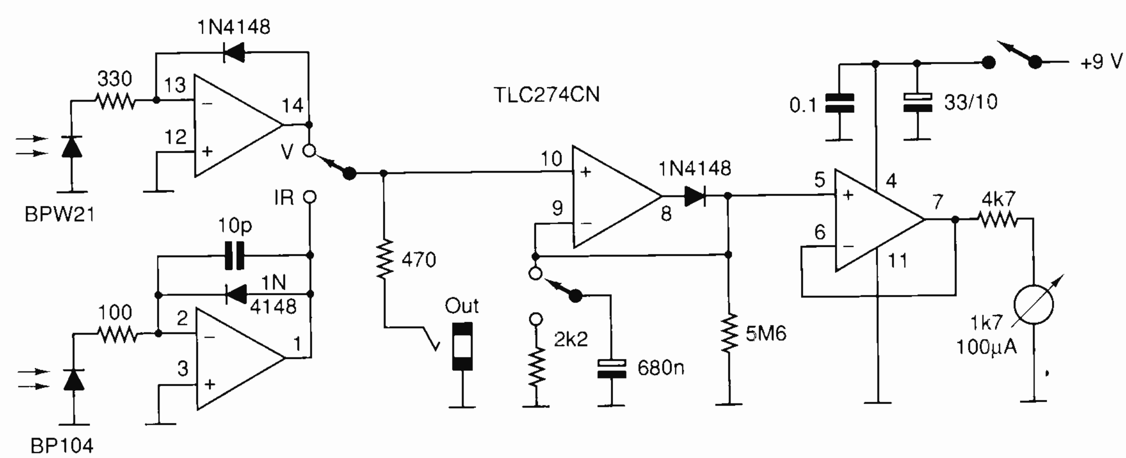
Visible leds may of course be used as light sources, but the photodiode efficiency is reduced significantly below 900nm and the current rating of the visible light emitting diode is lower. In addition the interference from visible light sources, such as fluorescent lamps, are hard to eliminate without filtration.

Detection of objects on a conveyer belt represents a typical application. **Fig. 8** shows a minimalist interrupter circuit. The Telefunken CQX47 IR led develops a radiant intensity of 0.033W/sr at 950nm at 100mA, giving an irradiance of 0.033W/m^2 at 1m distance.

The receiving diode photocurrent develops a voltage across the 33kΩ series resistor. To reach the 3.3V needed to switch the *74HC14* Schmitt gate, the photocurrent should reach 100μA. The Telefunken *BPV23F* photodiode sensitivity is 6μA at 1W/m^2. This equates to an irradiation of 16W/m^2 to produce 100μA photocurrent.

The required distance between emitter and photodiode is easily found by iteration. Reducing the distance to a tenth of the previous distance increases the irradiation one hun-

Fig. 5. Battery powered light meter with selectable visible light or near infrared light filters. Optional peak hold rectifier (minimum 50μs pulse width) and output to oscilloscope.



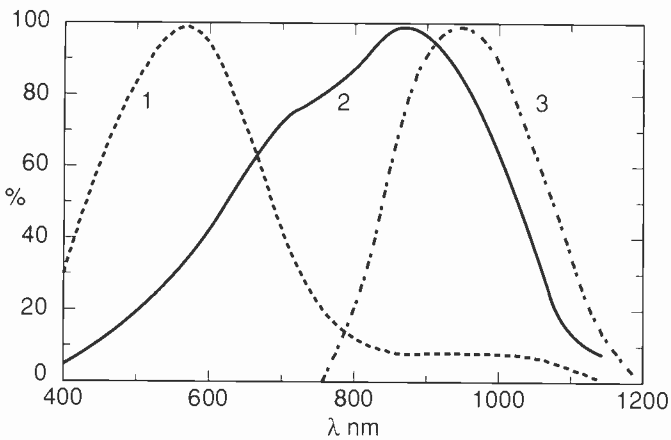


Fig. 6. Relative sensitivity of (1) BPW21 photodiode with daylight filter, (2) BPX43 photo transistor without filter and (3) BP104 photodiode with infrared filter.

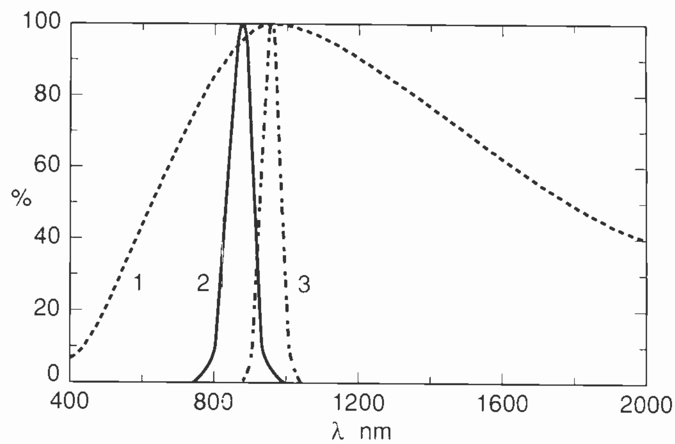


Fig. 7. Power distribution of (1) a tungsten filament lamp, (2) GaAlAs infrared emitting diode at 880nm and (3) GaAs infra red emitting diode at 950nm.

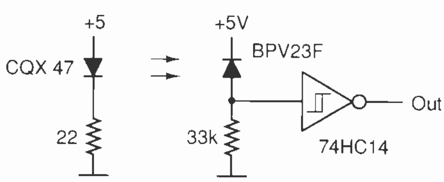


Fig. 8. Optical reader with 45mm detection range. See text for a discussion of the power dissipation.

dred times to 3.3W/m^2 at 100mm. Reducing the distance by a half increases the irradiation four times to 13.2W/m^2 at 50mm. Reducing the distance 10% increases it by 21% to the required 16W/m^2 at 45mm distance.

Calculated distance is based on typical values of emitter diode efficiency and photodiode sensitivity. The tolerance is approximately +50, −30% for both the led and photodiode.

Increasing separation distance between led and photodiode requires a square law power increase. Two series connected diodes would double the radiated power, increasing maximum separation by 1.4 times the distance available with one diode. Doubling the separation requires four diodes, etc.

The speed of the reader is set by the diode capacitance (20pF), the *74HC14* input capacitance (7pF) and the load resistor (33kΩ), a time constant of 1μs, so the reader should catch up with almost any moving object.

With 100mA forward current and a terminal voltage of 2.8V the power dissipation in the led is 0.28W, and with a thermal resistance of 270°C/W, the resulting temperature rise is 76°C. The led exhibits a power output temperature coefficient of −0.8%/°C resulting in a loss of some 60% of the emission. The diode should not normally be used in continuous mode at 100mA forward current.

At higher output intensity the led must be pulsed. This opens possibilities for modulated carrier systems.

Pulsed transmitter

The transmitter circuit shown in **Fig. 9** produces high intensity pulses. The radiant intensity of the Siemens *LD274-2* IR led is between 50 and 100mW/sr at 100mA, producing a typical 0.33W/sr at the peak current level of 450mA. The pulse width of 5μs is a practical lower limit for pulsed operation since the

Power, radiation and calibration

The earth receives approximately 1 kW/m^2 from the sun at the surface of the earth. Photodiode conversion efficiency is approximately 0.6A/W at wavelengths approaching the maximum sensitivity. If the photodiode has an active area of 1mm^2, the received power would equal 1mW, generating 0.6mA of current.

A *BPX43* photo transistor with a glass lens generated 2mA in the collector-base diode (and 1mA in the emitter-base diode) when subjected to direct sunshine. With an active area of 0.675mm^2 this indicated an effective magnification of about five times due to the lens action.

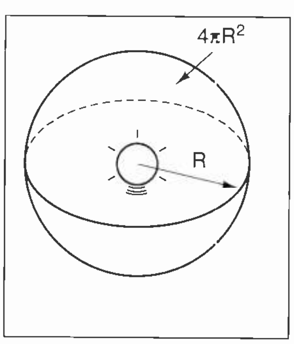
A tungsten lamp filament is heated with the intention of creating visible light. Most of the radiation is infrared, invisible to the human eye. Just one tenth falls inside the visible range. A third is radiation below 1100nm, useful to the photodiode, while the remaining two thirds are wasted as heat.

The radiated power from a domestic 40W bulb can be considered equally distributed in any direction. If this light source is placed at the centre of a sphere of radius *R* (Fig. a), the power *P* is distributed across the sphere surface of $4\pi R^2$, creating an irradiation of,

$$Ee = P/4\pi R^2 = 3.2\text{W/m}^2$$

at a radius of 1m. At wavelengths visible to the photodiode the useful irradiation is approximately one third of the total or 1W/m^2.

The illumination from the 40W bulb was measured to 400lx at 0.3m, indicating a correlation between illumination and irradiation of 40lx to 1W/m^2. This correlation is valid only for filament lamps operating at the same filament temperature, approximately 2500K.



A higher filament temperature will produce more visible light. At 3000K the correlation is 60lx to 1W/m^2, at 4000K 100lx to 1W/m^2. Photo components are usually specified at a filament temperature (colour temperature) of 2856K (standard light A) and the above correlation enables one to perform an approximate conversion between lx and W/m^2. These figures should not be used for sources other than incandescent lamps.

Mains ac modulates lamp output to a depth of approximately 30% with a ripple frequency of 100Hz. This should be borne in mind when designing opto-electronics.

The fluorescent lamp produces mainly visible light although some infrared radiation will be present, probably below 10% of the total. The light may be amplitude modulated by a frequency of 5 to 40kHz from electronic ballasts. Interference from fluorescent lamps may be reduced by optical filters.

device switching time is 1μs.

The led has a half power angle of ±10° but the main intensity is concentrated in a narrow beam within ±4°. This often causes alignment problems at extreme range.

Input gating switches the transmitter on or off. The 4.7μF filter capacitor may be omitted for decreased switching time. However, this results in a slightly distorted pulse waveform.

With the receiver shown in Fig. 8, the pulsed transmitter switches the *74HC14* buffer at 140mm (170mm actually measured) as compared to the calculated 45mm distance attainable with the linear system. The pulsed operation will of course introduce some circuit complications notably some sort of sample/hold which increases the system response time to 500μs.

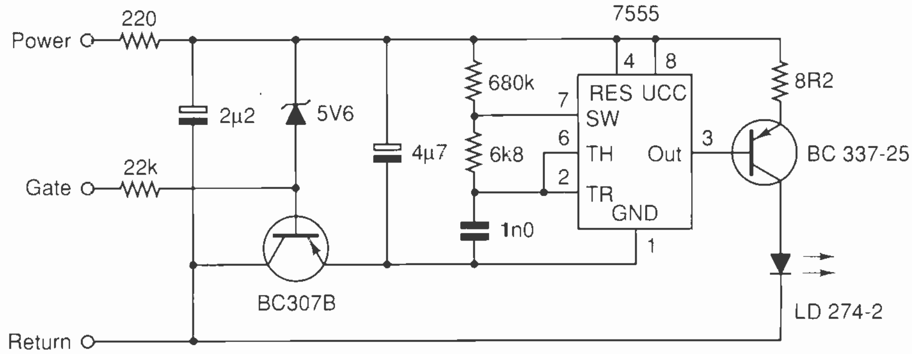


Fig. 9. Pulsed transmitter for 10 to 30V operation with 6mA current consumption. The 555 cmos timer is oscillating at 2kHz with 6μs pulse width and 450mA peak led current producing some 0.33W/sr radiant intensity at 950nm. The gate input is connected to the return path to operate the transmitter.

Active loads

A simple resistive load's maximum value is restricted by bandwidth limitations and overload problems from background light sources. Background light can cause enough photocurrent to forward bias the diode leading to severe losses in sensitivity.

Since most background light sources are either steady-state or have outputs which flicker at low frequencies, a load which is low impedance at these frequencies but high impedance at the desired signal frequency would be ideal. Such characteristics can be achieved with an inductor.

The inductive load shown in (**1**) can provide a signal frequency impedance of 100kΩ while giving a very low resistance path for background lighting photocurrent. Consequently it will operate over a wide range of light levels.

The circuit's output waveform (**2**) consists of a damped sine-wave whose frequency is dependent on the inductance used and the sum of the photodiode capacitance, the inductor's stray capacitance as well as the indicated capacitor. This ringing can cause multiple pulse detection if delays are not included in the receiver logic.

Inductors have been used in remote controls where good sensitivity and high background light tolerance has been required but the problems highlighted have limited the popularity of this approach.

The optimum characteristics provided by the inductive load can be obtained without the disadvantages mentioned by using an active circuit. Two configurations of the same basic active load circuit but with differing output polarities are given in (**3**) and (**4**).

In the first, photocurrent from the BPW41 raises the base voltage of the low noise ZTX384 via the 330kΩ resistor until the transistor's base-emitter voltage reaches about 0.7V and it starts to conduct. An equilibrium point is quickly reached where the transistor holds its collector voltage at around 0.8V by acting as a current generator that matches the photodiode current. This equilibrium is maintained for DC or slowly varying photocurrents thus providing the photodiode with a low impedance load at these frequencies. For the component values shown in (3), the load impedance presented to the photodiode changes from around 1kΩ at DC to approach 250kΩ at 50kHz.

Load impedance of this circuit falls a little at high light levels but its main disadvantages are noise and interference rejection. Although at first sight it appears that high-frequency inputs to the transistor are shorted by the capacitor, this does not apply to the small noise voltage generator within the device. These low level signals are amplified by the transistor. Voltage gain is given approximately by:

$$\text{Voltage gain} = \frac{R_c}{r_e} = \frac{\text{collector load impedance}}{\text{intrinsic emitter resistance}}$$

$$\text{where} \quad r_e = \frac{26}{I_e} \quad (I_e \text{ in mA})$$

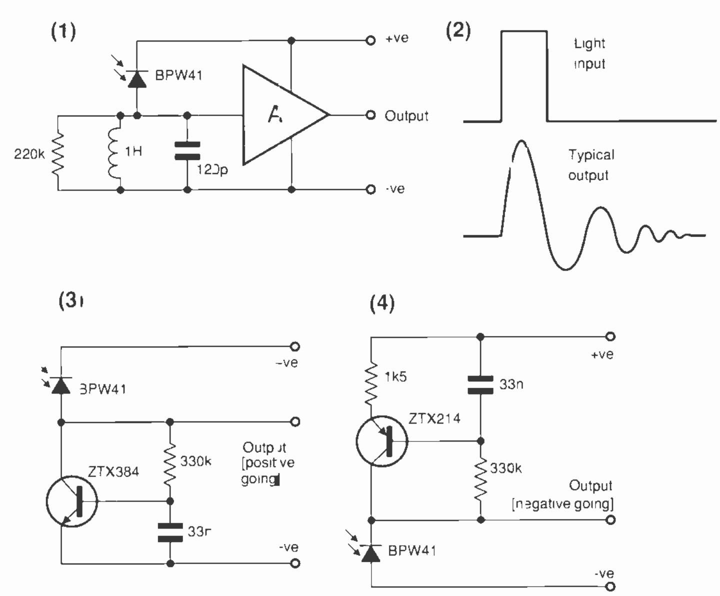
with low background light levels. For instance, in a dimly lit room yielding 5μA of photocurrent, the voltage gain given to these signals will be around 50, leaving their level too low to be of consequence.

Unfortunately, at light levels approaching direct sunlight, the resulting background photocurrent of 1mA will raise the circuit's voltage gain up as high as 10,000 making the noise and interference significant.

However, the problem is easily dealt with. The extra emitter resistor included in the second active load circuit changes this behaviour dramatically. The voltage gain of the circuit now approximates to:

$$\text{Voltage gain} = \frac{R_c}{r_e + R_e}$$

At 1mA the voltage gain of the circuit has been reduced to less than 160, leaving the noise contributions from other sources larger than those generated by the load circuit. The added resistor does increase the low frequency impedance of the circuit a little but it will still operate in direct sunlight. **David Bradbury, Zetex plc**



The photo transistor

In principle the photo transistor corresponds to a photodiode (collector-base diode) with a series connected transistor as amplifier. Gain is normally between 100 and 1000. Rise time is significantly slower than the photodiode due to reverse transfer capacitance which is multiplied by the gain of the device (Miller effect).

A typical photo transistor circuit for direct microcontroller interface is shown in **Fig. 10**, a photo-interrupter. The IR emitter is gated from the microcontroller and the current is limited by a series resistor. The microcontroller repeatedly activates the led and, after a short delay, the photo transistor status is read back, and the transmitter is deactivated to reduce the power dissipation. A delay of 100μs is necessary to switch the Siemens *SFH303F-3* photo transistor.

The transistor develops a collector current of 0.5mA at 1W/m^2 and 870nm and is reduced to 0.35mA at 950nm. To switch the transistor requires an irradiation of 3W/m^2. Since the led is the same as that used in Fig. 8, the maximum separation between transmitter and receiver readily calculates out as 100mm.

The photo transistor is a slow switch, because the photodiode capacitor C_{cb} is amplified by the transistor current amplification factor B. The resulting time constant equals $B \times C_{cb} \times R_c$, where R_c is the collector load. For the Fig. 10 circuit the time constant is approximately 30μs.

The only accessible factor in the equation for the photo transistor time constant is the collector load, R_c. Sensitivity too is proportional to R_c, so a trade off between distance and speed is necessary.

Fibre transmission

Data transmission with fibre optics is useful because of the freedom from electrical connection, crosstalk between channels and electro magnetic interference. The Fig.10 circuit provides a useful basis for experiments with fibre technology if the series resistor is changed to 100Ω to reduce the transmitter diode current to a safe level. The Siemens 660nm transmitter diode *SFH750V* and photo transistor receiver *SFH350V* were used, connected to a 3.5m cable of 2.2mm diameter (1mm core diameter).

The system operated with a minimum pulse width of 50μs, indicating suitability for data transmission to 9600 bit/s. The speed limita-

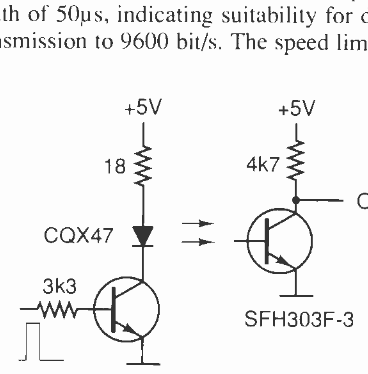


Fig. 10. Optical reader with 100mm detection range. The circuit could be used for data transmission with fibre optics.

Units and meanings

Illuminance (E_v) is expressed in lux (lx). The illuminance is related to the spectral sensitivity of the human eye, and is zero by definition outside the visible range. An illuminance of 1lx is sufficient to read a paper and 100lx is a common indoor illumination level. Starlight is some 0.001lx and sunshine is approximately 100000lx. At 0.25m distance from a 60W tungsten filament lamp (with reflector) the illumination is approximately 1000lx.

For components intended to operate outside the visible spectrum, or at specific wavelengths, the irradiation (E_e) is expressed in W/m^2.

Photo current (I_p) from the photodiode or photo transistor, when exposed to an illumination E_v or irradiation E_e, is:

$$I_p = S \times E_{v \text{ or } e} \quad \dots (1)$$

where S is the sensitivity of the photo component. A typical photodiode sensitivity is 80nA/lx or 5μAxm^2/W (both Siemens *SFH206*).

Radiant intensity (I_e) is the radiant power from an infrared emitting diode solid angle and is expressed in watt/steradian (W/sr).

The solid angle is A/R^2 where A is area and R radius of the sphere (see Fig. a). The full sphere is $A = 4\pi R^2$ and 1sr corresponds to a cone angle of approximately 60°.

The radiant intensity I_e is power P per solid angle A/d^2 (Fig. b), the power is thus the intensity multiplied by the solid angle: $P = I_e \times A/d^2$. The irradiation E_e is power P per area A, thus the power is: $P = E_e \times A$. The relation between radiant intensity and irradiation is thus:

$$E_e = I_e/d^2 \quad \dots (2)$$

The irradiation is inversely proportional to distance squared, and at a distance of 1m the radiant intensity in W/sr equals the irradiance in W/m^2. The distance d should be at least 10 times the diameter of the diodes to minimize errors.

The radiant intensity for an IR led is approximately proportional to the forward current and is usually specified at 100mA. The diodes can typically accept pulse currents at 1A, but light emitting diodes exist with ratings above 10A.

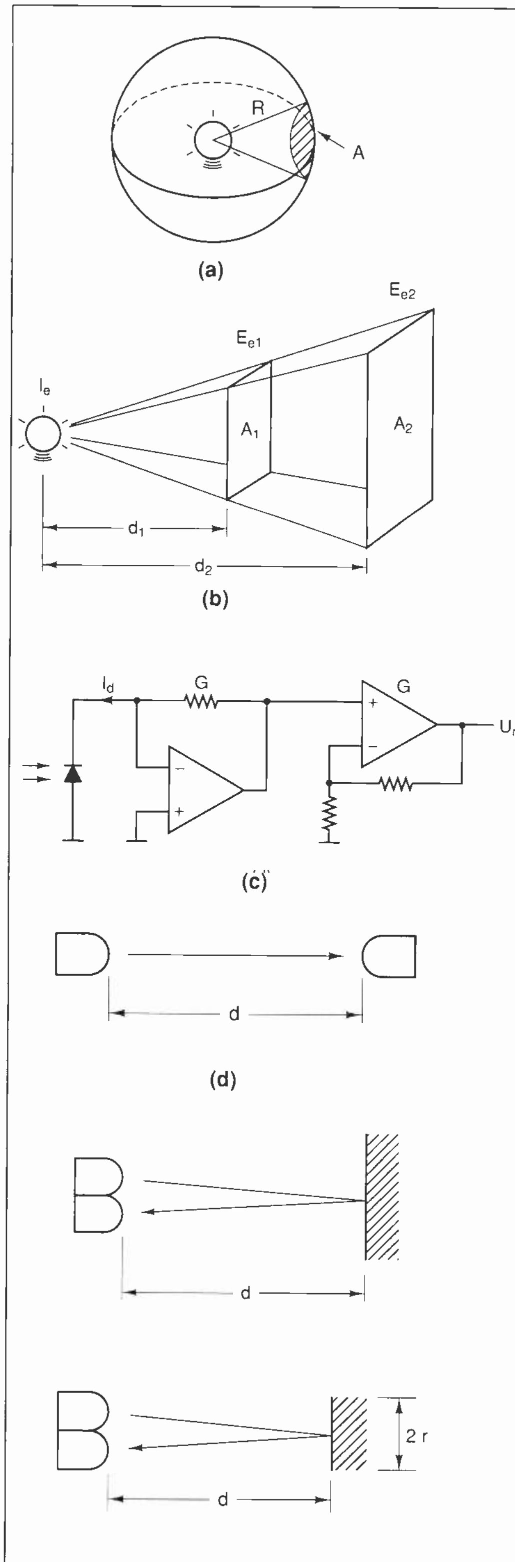
The led does not distribute the emitted power evenly within the 1sr solid angle. The half power angle is typically ±15°.

The radiant power declines with increasing length of operation. The life of the component is defined as the time after which the radiant power has fallen to half the initial value. The average life, dependent on the operation current and ambient temperature, is approximately 10 years.

The photo receiver output voltage (U_r) (Fig. c) is the voltage amplitude of the pulsed signal at the output of the amplifier. The photo current I_p is converted to voltage, and amplified to an output voltage of $I_p.R.G$. With equations (1) and (2) the receiver output voltage becomes;

$$U_r = S.R.G.I_e/d^2 \quad \dots (3)$$

With a receiver sensitivity of 5μA at 1W/m^2, a photodiode load resistance of 100kΩ, an amplifier gain of 100, the receiver delivers 1.5V at an led radiant intensity of 0.03W/sr and a distance of 1m (Fig. d).



tion is the photo transistor collector-base capacitor. For high speed operation a photodiode must be used at the receiver.

Photo transistor receiver

Photo transistors may be used in fast circuits if the impedance level at the basis is relatively low, as shown in the typical input stage of **Fig. 11**. The base network has a dual purpose. Besides biasing, the network resistance provides the collector-base photodiode load (150kΩ). The transistor acts as an impedance transformer reducing the output impedance below 1kΩ.

Using the *BPX43* photo transistor the value of C_{cb} is 20pF at 2V leading to a bandwidth of 35kHz if the stray capacitance is assumed to be 10pF. The bandwidth could be increased if the network resistance is reduced. A similar design operated at 100kHz with 20kΩ network resistance and a Honeywell *SD5443* photo transistor.

The circuit exhibits a sensitivity of 5mV/lx or 330mV at 1W/m^2 with the resistors shown. The photo transistor will accept 500lx before saturation (the base is driven above the positive supply forward biasing the collector-base diode).

The output voltage at the emitter replicates the intensity of the incident light and should be amplified further before detection. The pulsed transmitter of Fig. 9 produces an output voltage of 110mV at 1m distance.

The collector of a metal can photo transistor is electrically connected to the case. This provides a potential pickup point for interference in collector output circuits and this should be taken account of at the design stage. Grounding the can to rf with the transistor configured as an emitter follower overcomes this potential problem.

Photo diode receiver

A photodiode receiver intended for analogue signal processing or microprocessor interface is shown in **Fig. 12**. The design operates at 5V and connects to the system with a screened coaxial cable.

The first transistor acts as a common emitter amplifier with a boot strapped collector load, providing a voltage amplification of 200. The amplifier presents an input impedance of 600Ω providing a fast diode interface. The amplifier limits bandwidth to approximately 80kHz.

The collector of the first transistor is fixed at 0.6V providing a bias for the two-transistor output stage, the open loop gain of which is 650. This is reduced to 100 by feedback. The emitter of the pnp transistor is fixed at 1.2V providing a bias for the photodiode without interference from the power supply. The relatively high value of the biasing resistor limits the extraneous light to some 100lx and thus the application area to dim environments. The resistor value could be lowered at the expense of sensitivity or replaced by a suitable inductor. An inductor of suitable impedance at the prf allows the full ac signal voltage to be developed across it while providing a sink for the static photocurrent due to ambient light.

A suitable detector for a simple microprocessor interface could be a comparator adjusted to accept pulses 1V above the amplifier dc level. The irradiation needed for 1V output is 0.014W/m^2 and with the Fig. 9 transmitter radiant intensity of typically 0.33W/sr, the system should work comfortably at 5m distance.

The output noise voltage is 0.1V peak, mainly due to the illumination level. It could be reduced by increasing the cut off frequency of the high pass filters (hum suppression) and by reducing the amplifier bandwidth. Semiconductor noise is not usually a limiting factor in opto-electronics. Detection at a 0.1V threshold voltage (which corresponds to a range of 15m) would be possible using a synchronous detector to integrate several received pulses.

Proximity sensors

Sensitive receivers are used in proximity sensors (the optical equivalent to radar). The sensor detects reflections from objects within the sensor proximity.A combination of the Fig. 9 transmitter and the Fig. 12 receiver provides a useful proximity detector. Using reflection from the roof the pair produced an output voltage of 0.2V at 1.8m distance.

The circuit of **Fig. 13** is a complete proximity sensor with 0.35m detection range and low power consumption. The *4069UB* hex-inverter is used as six inverting amplifiers self-biasing at half supply voltage.

The transmitter is a free running oscillator with 1% duty cycle and a pulse amplitude of 100mA.

The photo transistor is used as a photodiode with internal lens, and with the metal enclosure connected to ground for shielding purpose. The collector and emitter leads are shorted to use the photo current from the emitter diode as well, increasing the sensitivity some 50% to approximately 3μA at 1W/m^2.

The first amplifier stage performs the conversion from current to voltage while the second stage provides a gain of 30. The amplified pulse is rectified and smoothed, a pulse amplitude of approximately 1V turning off the output transistor.

This pulse amplitude was reached at 0.35m distance with a large object. As shown in the panel the receiver voltage should reach the required pulse amplitude at approximately 0.25m. The transmitter half power angle should be 15°. ■

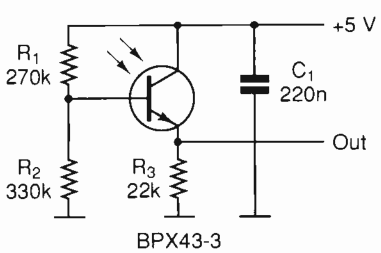


Fig. 11. Photo transistor input stage for high speed operation. The photo transistor enclosure is grounded to reduce sensitivity to electro magnetic interference.

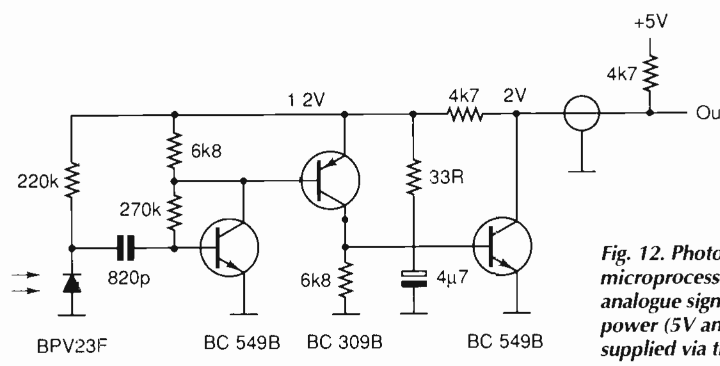


Fig. 12. Photo receiver for microprocessor interface or analogue signal processing. The power (5V and 0.6mA) is supplied via the coax cable.

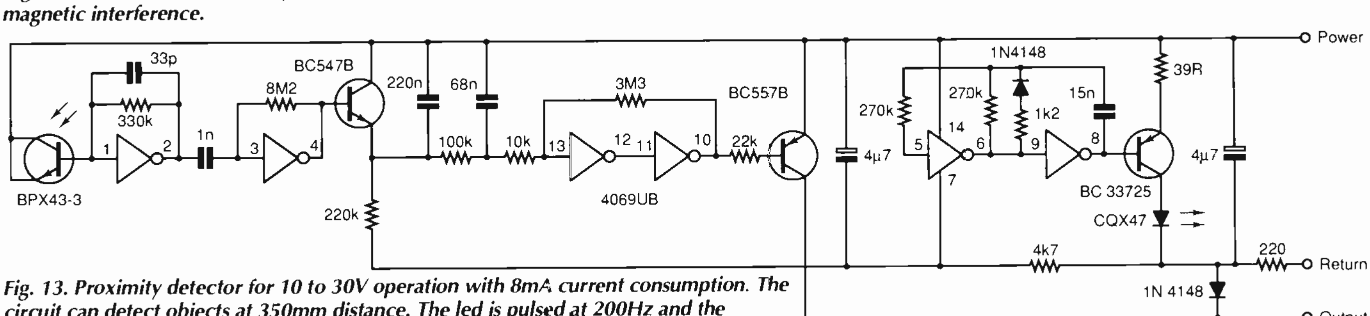


Fig. 13. Proximity detector for 10 to 30V operation with 8mA current consumption. The circuit can detect objects at 350mm distance. The led is pulsed at 200Hz and the reflected pulse is amplified and rectified. The output is active in absence from light and has a reaction time of 30ms.

LOW DISTORTION AUDIO OSCILLATOR

It is a chicken and egg situation. You may wish to design the ultimate in distortion free audio equipment but can you be sure that you are measuring a residual from the equipment under test or an artifact from the test oscillator (or distortion meter) itself? This test oscillator design presented here by Ian Hickman could break the dilemma.

This magazine has a long tradition of interest in audio design and measurement. Central to this are low distortion oscillators and distortion meters, and examples of both of these have appeared from time to time[1,2].

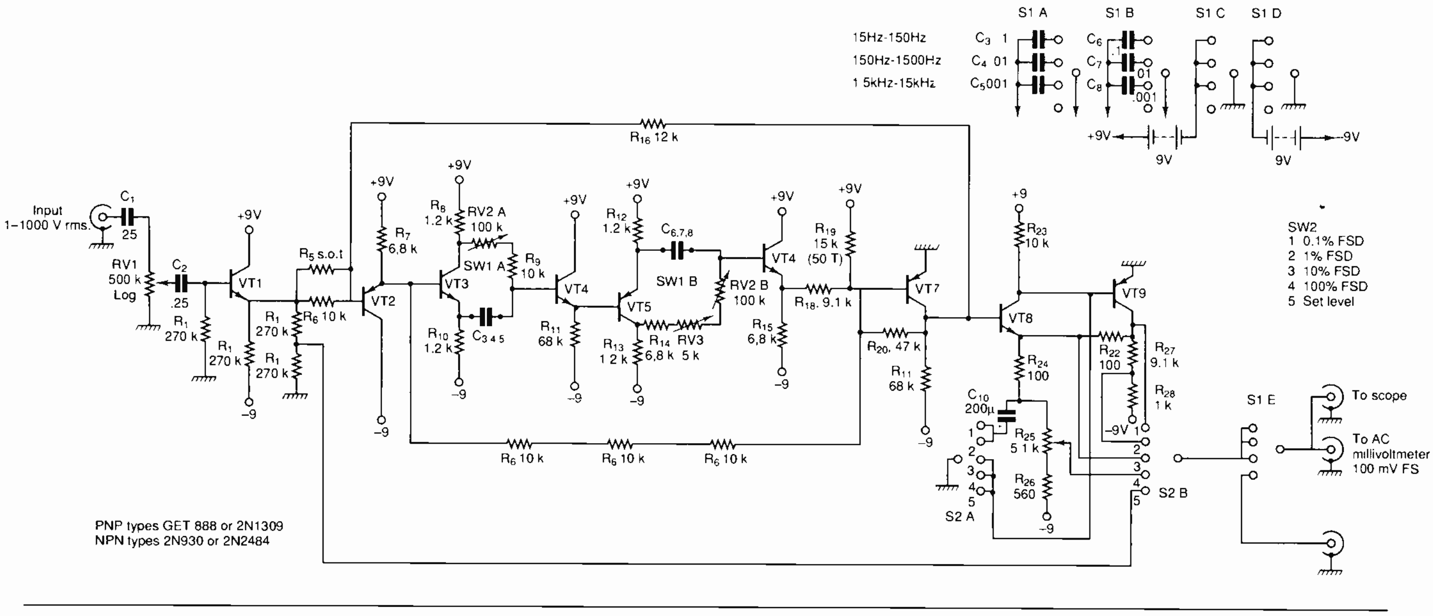
The oscillator in the first reference has a claimed performance of <0.005% THD from 20Hz-20kHz and as low as 0.0005% typical (or even 0.0002% at 1kHz, actually lower than the commonly accepted typical distortion figures for the *NE5532* op-amps used), would be very suitable for evaluating the performance of a THD measuring system as described in Ref. 2. In fact, low distortion oscillators and matching THD meters form a chicken-and-egg pair, with each being (ideally) tested with a sample of the other having a level of internal distortion much lower than its own.

This aspect has prompted me to experiment with various AF oscillators and THD meters. A trawl through my files unearthed a circuit from 1966, shown in **Fig. 1**; even the required layout, using turret tagstrip, is there. This circuit was designed for use with a separate external AF millivoltmeter, which would have been average responding, scaled to rms on a sinewave. Thus the indicated distortion would only approximate the true figure. See Box.

A new THD meter was designed around a state variable filter some ten years ago. This offered ranges in a 30 - 10 - 3 - 1 sequence down to 0.01% fsd permitting measurements down to 0.001% or less. Its residual noise level corresponds to 0.0009% in a 20kHz bandwidth, 0.003% in 80kHz, far too high) and it had a built-in true rms responding indicating section using the *AD536* rms to dc converter. This instrument worked well and is still in use although, when making measurements at the 0.001% level, it requires suppression of the fundamental in excess of 100dB. This is a tall order for a single notch.

This is especially the case in a THD meter where the notch must not depress signal level at twice the notch frequency where the second harmonic is encountered. The result is that any slight frequency instability in the test source, appearing as noise sidebands, will cause energy to appear either side of the notch and raise the level of the measured residual. Negative feedback, necessary to ensure a virtually flat response at second and higher harmonics of the rejected fundamental, actually causes a bit of a design problem. The flattening of the fre-

Fig. 1. Circuit of a THD monitor designed by the author and dating from the mid sixties. Like that in Ref. 2 it was designed to be used with a separate external AF millivoltmeter



quency response is bought at the price of an increase in gain at the fundamental at an internal circuit node, as shown in **Fig. 2**. This is based around a Wien bridge, though a very similar argument applies whatever circuit is used to implement the sharpened-up notch.

The level of the enhanced fundamental component must of course be kept well below clipping level, limiting the permissible input level to the notch stage. This in turn reduces the dynamic range due to the reduced clearance of the signal above the wideband noise floor, providing another factor, in addition to the instrument's own residual internal distortion, limiting the lowest level of distortion that can be observed. Despite this, a high performance THD meter *should* be easier to design than an oscillator, since all circuitry within the former can be linear, whereas the latter requires an amplitude control mechanism which involves some non-linearity.

The SVF-based instrument mentioned above actually, like many THD meters, responds to harmonics of the input, hum and the wideband noise floor, as well as significant spectral noise sidebands surrounding the fundamental which fall outside the notch. Useful as this instrument proved, initially there was no way of knowing what was the level of residual distortion in the instrument itself. Thus a very low distortion oscillator was designed to provide a test source.

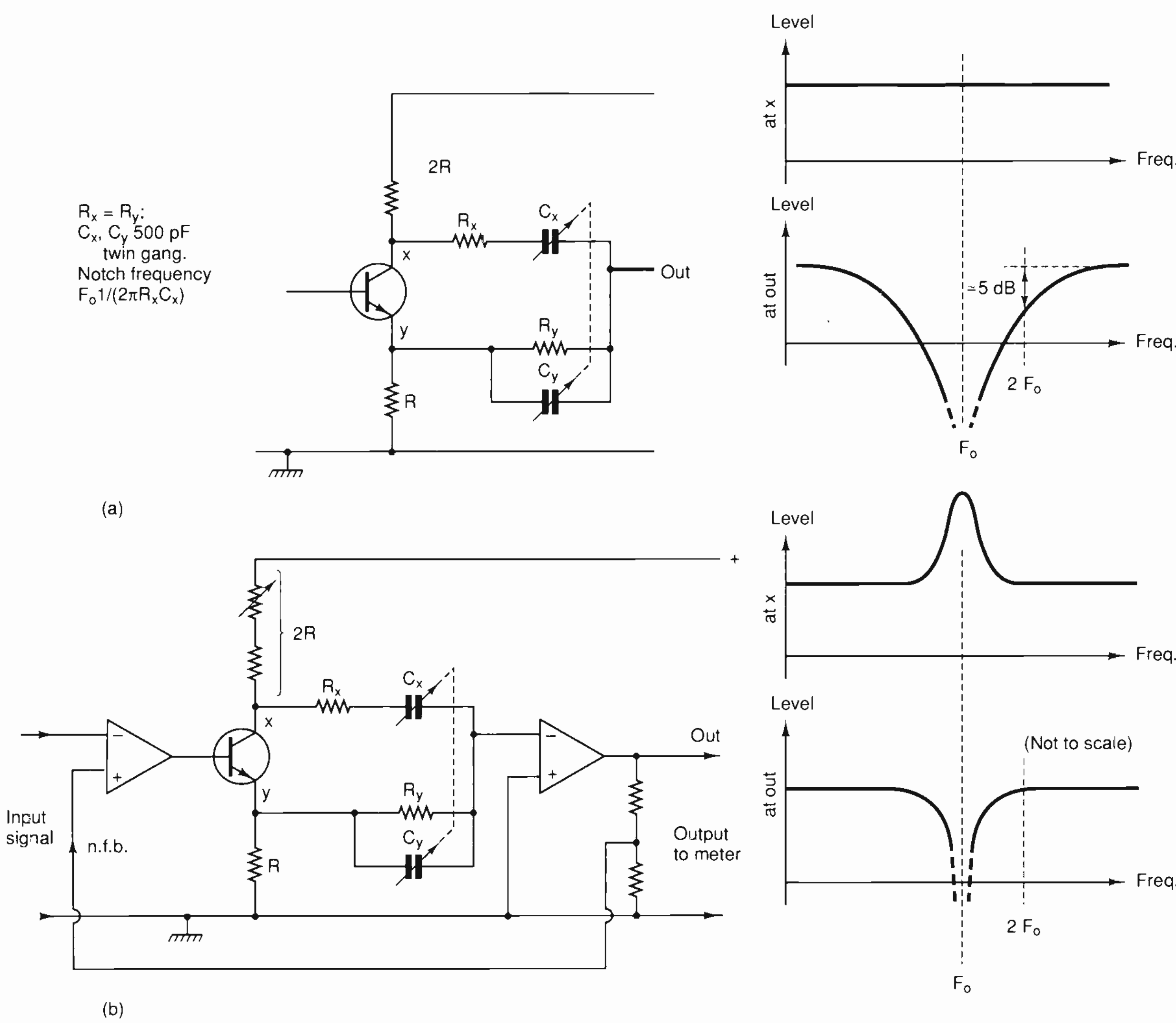
Some nonlinearity is needed to constrain an oscillator's output amplitude at a suitable constant level. This non-linearity can operate on a cycle-by-cycle basis or over many cycles as with thermistor control. Having a thermal time constant of nearer to a second than a millisecond, the thermistor's resistance remains sensibly constant over each cycle of the output, except at frequencies below 100Hz. Here, it leads to 20Hz distortion figures in the range 1% or worse for a typical design, down to 0.1% in the case of Ref. 1 (reduced at the output to <0.005% by a distortion cancelling technique described in the article).

Fig. 2a Circuit providing a notch at one frequency f. The response at 2f is still 4.77dB down, eventually returning to the dc value at a much higher frequency.
b. By enclosing the notch circuit within an NFB loop, the response can be sharpened up to be no more than a fraction of a dB down at second harmonic. However, internal to the loop, the response at the fundamental is actually peaked up significantly.

Chicken and egg measurement

It is very difficult to design a really low distortion oscillator using this approach. It was therefore decided to use an SVF-based system incorporating nonlinearity operating over many cycles, effectively adjusting the loop gain. In fact, it is actually the loop phaseshift which is adjusted since, with the two integrator loop of the SVF, there is always a frequency at which the loop gain is unity. This contrasts with the approach in Ref. 2 which uses an all-pass filter based oscillator: here there is always a frequency at which the loop phase shift is 360°, so it is the loop gain which has to be adjusted to obtain a stable output. The circuit adopted is shown in block diagram form in **Fig. 3a**, with the operating principle explained in **3b**.

Initial breadboard results were promising, which immediately reintroduced the chicken and egg problem of measurements. This required a further piece of equipment to resolve it. As **Fig. 4** shows, this comprises a passive twin-T notch filter followed by a second order Chebychev active highpass filter, forming a fourth order elliptic highpass filter. The peaking of the Chebychev highpass filter is set to compensate for the attenuation of the



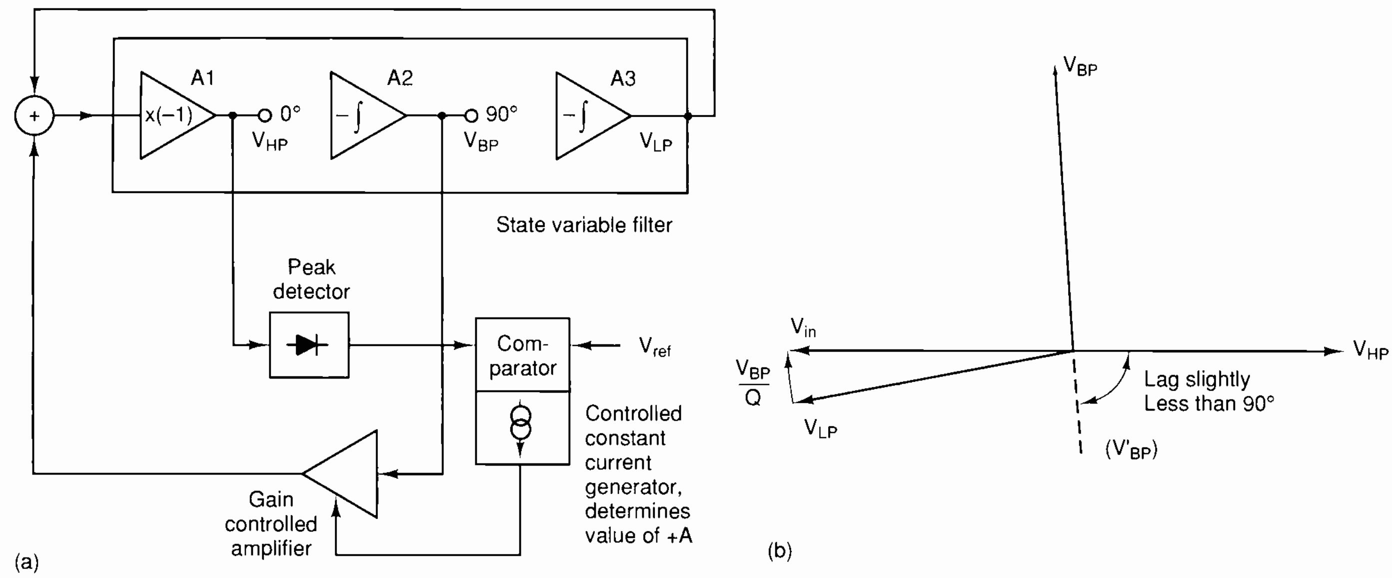


Fig. 3a. Block diagram of the SVF-filter based low distortion oscillator described in this article. The filter selectively amplifies the fundamental component of the output of the variable gain amplifier, discriminating against any harmonic distortion present.
b. Showing how the BP output, lagging the HP by a fraction less than 90° (much exaggerated for clarity), looks as though it is leading by just over 90°, as A2 is an inverting integrator. Similarly LP with respect to BP, so LP lags HP by slightly less than 180° and cannot by itself provide the necessary input to the filter V_{in} to produce the output shown. Addition of a fraction 1/Q of the BP output increases the phaseshift to 180° and gives a voltage equal to the required V_{in}, causing oscillation.

passive notch filter at twice the notch frequency, amounting to some 8dB relative to the 0Hz and far-out high frequency response. In the case of channel 2 (600Hz) shown, the LF roll-off of this elliptic highpass filter also discriminates against 50Hz and its harmonics. The Chebychev highpass filter section is followed by a lowpass filter designed, in the case of the 600Hz notch channel, to cut off beyond 3kHz. With this in circuit, THD up to and including the fifth harmonic is measured, very low levels being easily seen due to the reduction in wideband noise afforded by the lowpass section. In the case of a low distortion oscillator circuit, the design will be such that often only the lower orders of harmonics are significant. But in other cases, such as a high power amplifier using a push-pull output stage, higher order distortion components due to, say, crossover effects may be present. In this case, the lowpass section would be inappropriate, so provision is made to switch it out.

The adjustments shown provide a sensibly flat response from the second harmonic of 600Hz upwards (or from the second to fifth harmonics, inclusive). The through gain in the flat position is approximately unity (actually somewhat higher), while it is arranged that with a notch channel in circuit, the gain over the specified range of harmonics is increased by exactly 40dB. Thus when used in conjunction with an existing THD meter, the 10% distortion range becomes 0.1% fsd, the 1% range becomes 0.01% etc.

Clearly, when making measurements using a notch channel, the effect of any internal noise or distortion in an associated THD meter is reduced by a factor of 100, while due to the protection afforded by the passive twin T notch there should be a negligible contribution to distortion from the notch channel itself, though it will of course contribute its own noise.

The 20Hz channel uses exactly the same circuit arrangement as the 600Hz channel, with again the option of switching the lowpass section in or out as desired. However, the 10kHz channel is somewhat different. The twin T section is followed by a buffer/10dB gain stage like the others, but this drives two state variable filters, one tuned to 20kHz and the other to 30kHz. The outputs of these are combined in the output buffer amplifier.

With this notch box in use ahead of a conventional THD meter, the measurement is no longer an "everything else" measurement, especially with the lowpass section in circuit. The great disadvantage is that measurements are confined to the three test frequencies catered for. Of course, additional channels could be provided, but the added complexity rapidly becomes cumbersome.

The low distortion oscillator

With means to hand to examine distortion levels below −115dB relative to the fundamental (<0.00018%) work on the development of the low distortion oscillator became possible. When working at low levels of distortion, layout becomes just as important as it is in an RF circuit, so breadboarding was abandoned in favour of discrete wiring on a commercial prototyping PCB.

The intention was to use the state variable filter running at the highest practicable value of Q, to filter a low level, low distortion sinewave drive. The SVF circuit was therefore tested running without any intentional quadrature feedback, either negative or positive, providing in theory a filter with an infinite Q. At low and middle frequencies, the inevitable fractional shortfall of phaseshift in each integrator meant that the loop phaseshift fell fractionally short of 360° and the circuit was stable. But on switching to the 2-20kHz range, at the higher frequency settings the three op-amps in the loop were beginning to contribute a mite of phaseshift each, leading to oscillation when the frequency was set to 17kHz or higher. This was compensated out by providing a touch of phase advance, which turned out to be not quite so simple.

A single capacitor between the lowpass (LP) output and the virtual earth of the highpass (HP) stage will certainly stabilise the circuit, suppressing any tendency to oscillate at 17kHz or above, but the damping it provides becomes quite excessive at 20kHz, leading to a low Q at this frequency. The phase advance was therefore distributed between all three stages within the basic loop, each capacitor being provided with a series resistor to limit the phase advance as shown in the full circuit diagram, **Fig. 5**. The required value of capacitance at the HP stage virtual earth, **5a** was so small that it was provided by two 20mm lengths of wire-wrap wire, twisted more, or less, as required.

Now, a small quadrature (negative damping) term was taken from the BP output via an operational transconductance amplifier (OTA) and added in at the HP stage virtual earth. When this is just sufficient to bring the loop phaseshift up to 360°, the circuit will oscillate (Fig. 3b). A current source provides the I_{abc} bias current required by the OTA, but it is arranged that as the amplitude of the oscillation increases beyond a certain threshold, the peak detected voltage causes a reduction of I_{abc}, reducing the loop phaseshift back to exactly 360° and stabilizing the amplitude at that level.

OTAs are not renowned as the lowest of low distortion devices, so the voltage applied to its non-inverting input is kept to a very low level by the 100kΩ/27R potential divider R_{17}, R_{18}.

The OTA's output current is divided between two 100kΩ resistors, one grounded to ensure that its output voltage remains centred within its voltage compliance range, and one providing dc-blocked "negative damping" to the HP stage virtual earth. The voltage set by the 10kΩ/4.7kΩ resistors R_{15} and R_{16} at the base of the pnp long-tailed pair provides a +5V reference voltage. If the detected peak voltage is more or less than this value, I_{abc} will be reduced/increased respectively, adjusting the negative damping term as required to maintain a constant amplitude of oscillation.

The peak detector time constant of 1s (10MΩ x 1µF) is long enough to do duty on all three frequency ranges, covering 20Hz-20kHz, though it is responsible for some rise in distortion at 20Hz. However, increasing this time constant is not without its problems, so the values shown have been retained.

Assuming E_1 stands for the rms value of the fundamental of a distorted sinewave, E_2 for that of the second harmonic component, etc., then the total harmonic distortion is given by:

$$\text{Total THD} = \frac{\sum E_2^2 + E_3^2 + E_4^2 \ldots}{\sum E_1^2 + E_2^2 + E_3^2 + E_4^2 \ldots} \qquad \ldots 1$$

By setting a reference level with a flat frequency response, corresponding to the denominator of (1), and then measuring the relative level of the signal with the fundamental component E_1 notched out, this is exactly what a THD Meter measures – except that in the measurements, any noise, hum or other signal present is also inevitably included.

Ideally, we might prefer to measure:

$$\frac{\sum E_2^2 + E_3^2 + E_4^2 \ldots}{E_1} \qquad \ldots 2$$

but if all components E_2, E_3, E_4 etc. are less than 10% of the amplitude of E_1, then the difference due to each is less than 1%. In low distortion measurements, where the harmonics are all much less than 1%, then in practical terms (1) and (2) are identical.

Estimating the Q

To see just what Q the filter runs at, I_{abc} was reduced to zero and the value of resistance required between BP output and HP virtual earth input to just cause oscillation was determined: this value divided by the value of the resistance from LP output to HP virtual earth (R_3, 100kΩ) gives the filter Q. The circuit oscillated over the whole 20Hz-20kHz range with quadrature feedback via 10MΩ; over most of the range with 20MΩ and some of the range with 30MΩ. Thus the operating Q of the loop, considered as an SVF, is generally between 200 and 300.

In an active filter operating at such a high Q, the output amplitude would normally be very sensitive to temperature and other external factors. Here, however, the narrow range of level needed to change I_{abc} from zero to maximum results in very tight amplitude control. With a gain at the fundamental of, say, 250 times and an attenuation of three at third harmonic in each of the integrator stages, any third harmonic in the signal from the OTA should be reduced by a factor 2250 or 67dB (a factor of –60dB at second harmonic). So if the distortion introduced by the OTA can be held below 0.1%, the output distortion should be <0.0001% at second harmonic, and the third even smaller... provided the op-amps in the loop have zero distortion themselves. The op-amp A_{6a} with its gain of 11 was included to provide a buffered amplified version of the OTA output as an aid during circuit development and testing. The voltage measured here confirms the estimate of operating Q, being 0.5Vpk/pk at 10kHz. This corresponds to 45mVpk/pk at the OTA output as against 10Vpk/pk at the LP output, indicating a Q of 220.

The oscillator lowpass output labelled *LP* in Fig. 5a was used to drive an output section as shown in **5b**, being fed to the virtual earth of op-amp A_{2b}. It can be seen that there is also a contribution from the HP oscillator output, via $R_{31} + R_{35}$. This was included so that R_{31} could be adjusted to suppress any third harmonic component in the output, as described in Ref. 3. The output of A_{2b} drives A_{3b}, whose feedback resistor is variable, providing fine output level adjustment. Coarse adjustment in 10dB steps is provided by the 600Ω bridged T attenuator associated with S_4, and by the 0 or 50dB pad, S_5. The odd value resistors R_{44-48} and R_{53-57} were made up by paralleling standard values to get within 1%.

S_3 provides a choice of a 600Ω output impedance or (with the step attenuators set to 0dB) a low impedance. This provides twice the maximum available peak to peak output voltage into 600Ω, but if the step attenuators are used, the first 10dB step will no longer be accurate. Note that great care is required in the earth routing, as indicated in Fig. 5b, if the distortion is to go down pro rata with the output. screened lead should be used between the main output socket, S_5, S_4 and the main circuit board.

Performance

When set up, the oscillator provided the following performance, all measurements being taken with S_3 set to low output impedance, S_4 and S_5 to 0dB, and using the notch box.

20Hz	0.00062% (20kHz bandwidth, notch box LPF out, **Fig. 6a**.
20Hz	0.00059% (20kHz bandwidth, notch box LPF in)
600Hz	0.00034% (20kHz bandwidth, notch box LPF out)
600Hz	0.00022% (20kHz bandwidth, notch box LPF in)
10kHz	0.00092% (80kHz bandwidth, **Fig. 6b**)
10kHz	0.00062% (20kHz bandwidth, **Fig. 6c**)

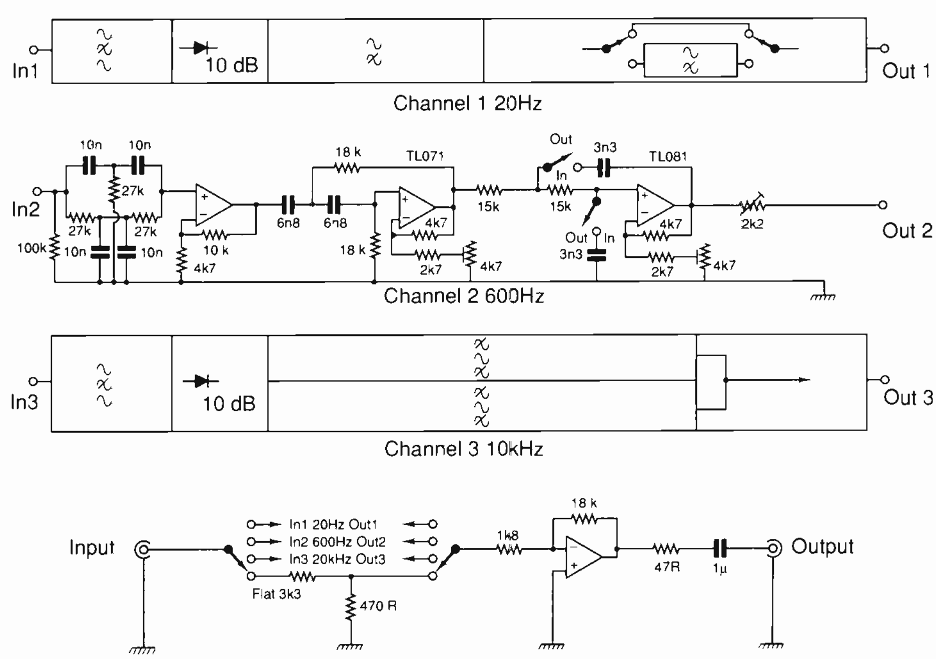


Fig. 4. Circuit of the "Notch Box" used to extend the measurement range of a THD Meter, providing three spot measurement frequencies of (approx) 20Hz, 600Hz and 10kHz.

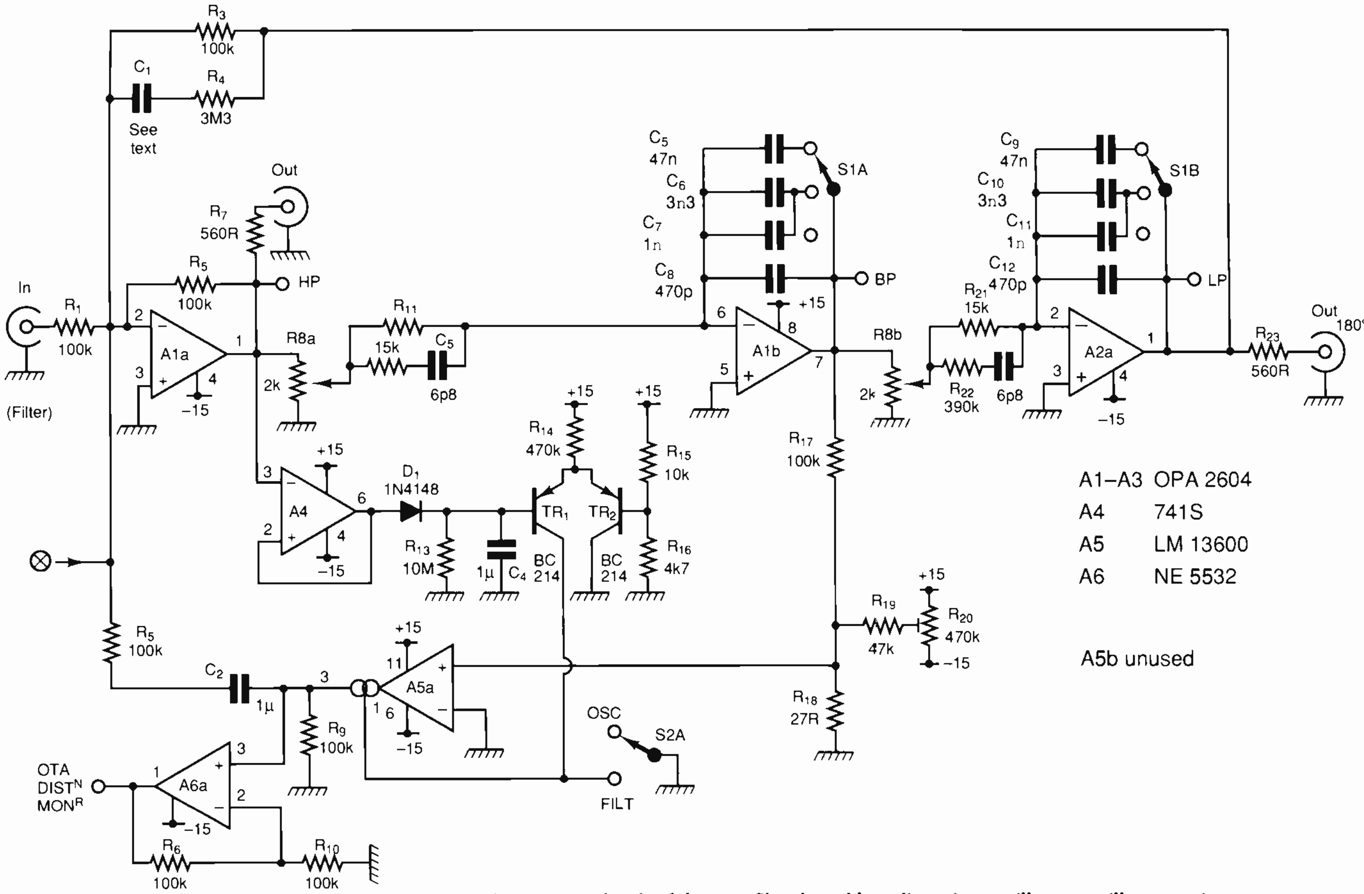
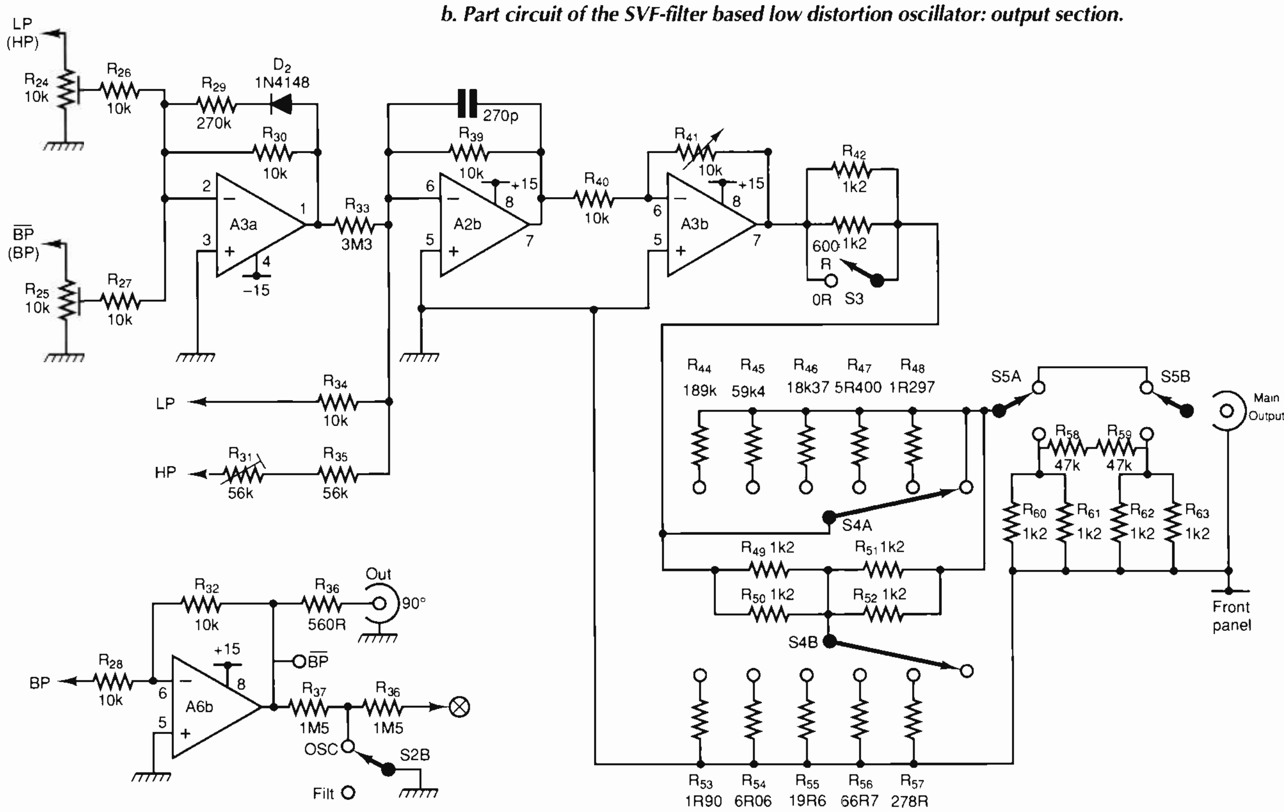


Fig. 5a. Part circuit of the SVF-filter based low distortion oscillator: oscillator section.

(a)

b. Part circuit of the SVF-filter based low distortion oscillator: output section.



(b)

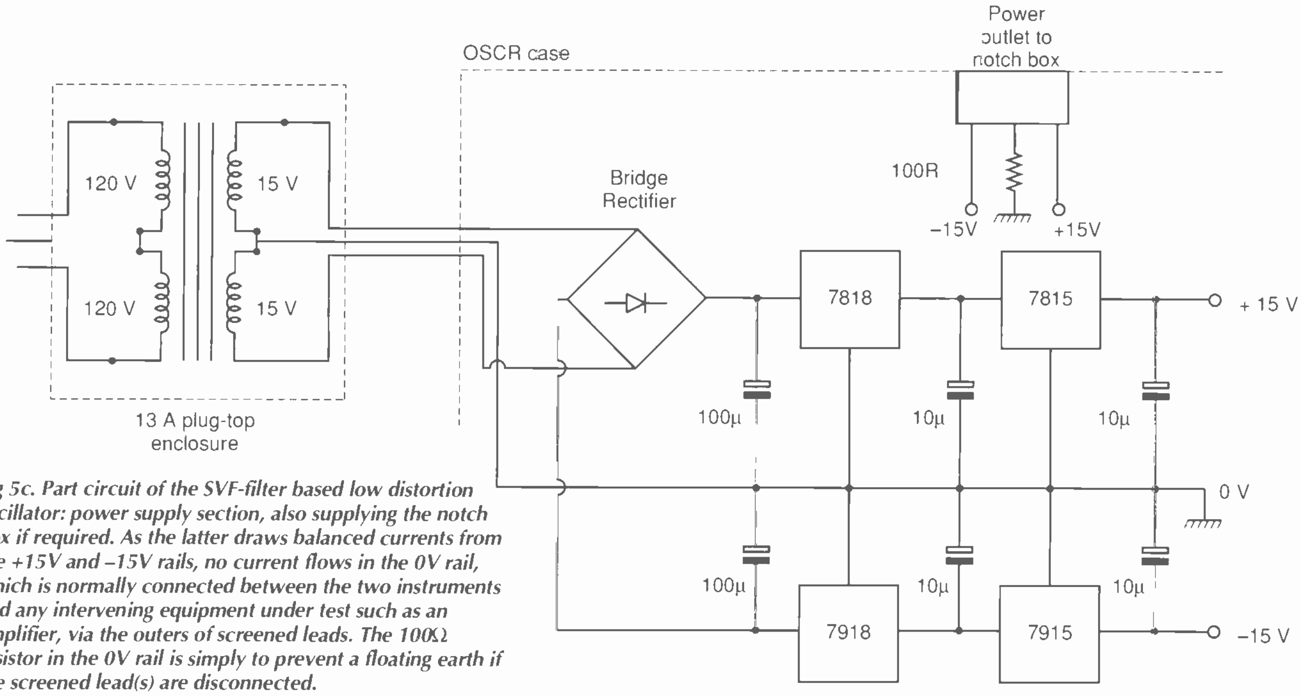


Fig 5c. Part circuit of the SVF-filter based low distortion oscillator: power supply section, also supplying the notch box if required. As the latter draws balanced currents from the +15V and −15V rails, no current flows in the 0V rail, which is normally connected between the two instruments and any intervening equipment under test such as an amplifier, via the outers of screened leads. The 100Ω resistor in the 0V rail is simply to prevent a floating earth if the screened lead(s) are disconnected.

At 600Hz, the residual distortion can be clearly seen when directly viewing the THD meter's *residual* output on an oscilloscope, but with the 20s exposure required by the scope camera, it was totally obscured by the broadband noise when photographed. In fact, the distortion is visibly virtually pure second harmonic, the figure of 0.00034% being higher due solely to the extra noise admitted in the absence of the notch box's 3kHz lowpass filter.

I will not pretend that the above performance will automatically result from the circuit shown: a number of stages of setting up are required. Firstly, the Burr Brown *OPA2604* op-amps were not specially selected – only three samples were to hand – but they were swapped around for best results. (With a manufacturer's quoted typical THD + noise of 0.0003% at 1kHz, the *OPA2604* fet input dual op-amp is not the lowest distortion device available. Analog Devices *AD797* is supposed to deliver 0.0001% THD (−120dB) at 20kHz. But this is a bipolar input type, optimized for source impedances less than 1kΩ, which is not convenient in the present application.

Next (with the wipers of R_{24} and R_{25} set to ground), R_{20} was adjusted for minimum distortion at 20Hz, this control proving the most critical at low frequencies. It compensates for the input offset in the OTA, centring the signal in the latter and thus minimizing second harmonic distortion. It should not be left too far off centre, as there is then the possibility of control signal breakthrough appearing at the OTA output, leading to instability of the control loop in the 2-20kHz range.

The next adjustment was of R_{31}, which as already stated was included as a way of outphasing any third harmonic at the output. However, as noted in an earlier design brief[4] the main distortion mechanism in the op-amps used is second order, and there was no visible third harmonic to cancel R_{31} was therefore set to minimise the second harmonic content at 10kHz. This just leaves the circuitry associated with R_{24} and R_{25}, which has not been mentioned up till now.

Op-amp A_{3a} provides a means of deliberately producing an element of second harmonic distortion of any amplitude and at any relative phase angle over the full 360° of the fundamental to cancel residuals. This is added in via R_{33} to the virtual earth at the input of A_{2b}. With suitable settings of R_{24} and R_{25}, the distortion at the main output at 600Hz can be driven down to the point where the residual reads the same as circuit noise but this is an academic exercise. A better overall result was obtained by a compromise setting which resulted in somewhat more distortion at 600Hz but a substantial improvement (down from 0.0016%) at 10kHz.

It might be thought that the presence of D_2 would make the performance unduly temperature sensitive, so a test was undertaken with the oscillator's top cover removed. The 600Hz distortion residual was monitored and hot air blown into the case, monitoring the temperature adjacent to D_2. At 50°C the distortion was unchanged from the room temperature value. The particular settings of R_{24} and R_{25} used resulted in 1.5V pk-pk at the output of A_{3a}, the waveform exhibiting a percent or so of second harmonic distortion. This is diluted by a factor of about 300 by the ratio of R_{33} to R_{34} and a

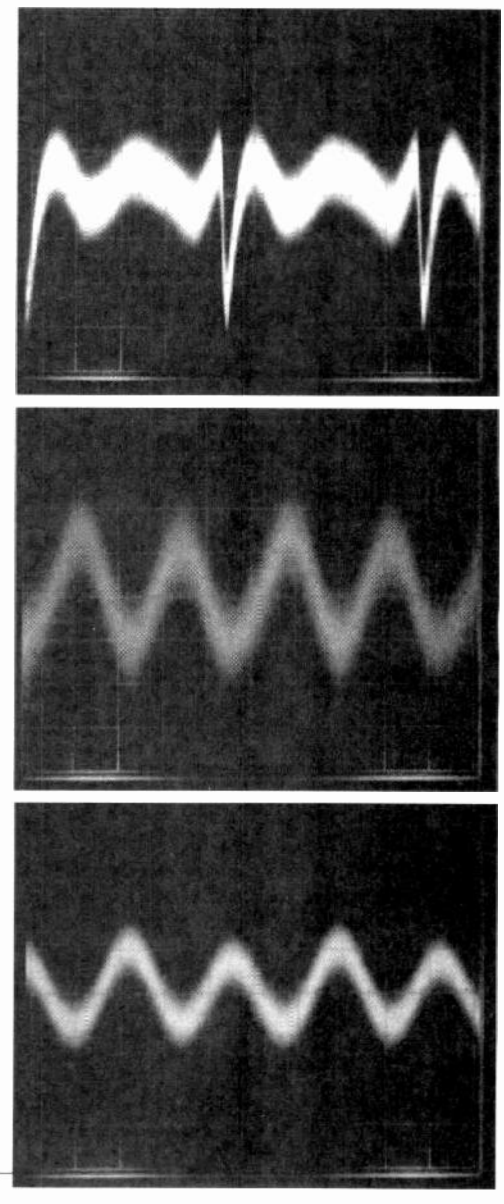


Fig. 6a. Residual distortion plus noise at 20Hz is 0.00062%.
b. Residual noise and 2nd + 3rd harmonic distortion with THD Meter bandwidth set to 80kHz measured 0.00092%. Note: no significant 3rd harmonic observable.
c. As b, but THD meter bandwidth set to 20kHz; reading 0.00062%.

further factor set by the ratio of 1.5V pk-pk at A_{3a} output to the 10V pk-pk at LP, to around 0.0003%.

Other matters

A number of practical details which arose may be of interest. An attempt to include a power supply within the case of the oscillator was a resounding failure, due to hum from the mains transformer's stray magnetic field.

Having produced a basic oscillator, it seemed sensible to include other facilities. The first of these enables the unit to be used as a quadrature oscillator – very conveniently done with an SVF based circuit. The HP output forms the 0° phase output, taken via a 560Ω resistor to a front panel BNC socket. The inverted BP output, obtained from BP by inverting stage A_{6b}, provides an output lagging by 90° – note that although A_{1b} and A_{2a}, the two integrators within the loop, each provide 90° lag, the BP waveform actually *leads* that at HP, as they are *inverting* integrators.

Another 560Ω resistor driven from LP provides the 180° output at a third front panel BNC socket. The instrument is also provided with an input, via a front panel socket and R_1. This may be used in two different ways. Firstly, with the oscillator running, its frequency can be locked to a low level signal injected via R_1, simply by tuning it to the frequency of the injected signal.

Secondly, the oscillator can be disabled by setting S_2 to the "filter" position. This has the effect of spilling I_{abc} to the negative rail, disabling the positive feedback via the OTA and stopping the oscillation. At the same time, damping is applied from inverted BP via R_{37} and R_{38}, defining the Q of the filter as $(R_{37}+R_{38})/R_3=30$. Since $R_1=R_3$, the bandpass gain from IN to the 90° OUT socket (inverted BP) should simply equal Q, the measured value at 600Hz actually being 27.

R_8 is a 2K wirewound two gang potentiometer fitted with a ten turn counting dial. As 10 on the dial corresponds to 200Hz, 2kHz or 20kHz it can act as a frequency readout if the dial reading is simply doubled. A dial reading of 10 can be arranged to correspond exactly to 2kHz by fitting a select on test resistor between the top end of R_{8a} (and R_{8b}) and the op-amp output driving it. By selecting different a resistor for each frequency range (using additional poles on S_1), the top frequency on each range can be correct, without requiring exact close tolerance values for C_5–C_{12}. Frequency readout will be almost linear, with a maximum error of –3% at five turns, due to the loading of R_{11} (R_{21}) on the wiper.

This parabolic error can be reduced to a much smaller cubic error by connecting a 15K resistor from the top of R_{8a} to its wiper (and likewise R_{8b}). Some further development of the design is planned, notably affecting the fine output level control. Even using a high quality conductive plastic or cermet type, R_{41} is a possible source of excess noise in the output. A better arrangement would be a switch providing 1dB steps, with a 0-1dB continuously variable control provided by a potentiometer varying the reference voltage at the junction of R_{15}, R_{16}. ■

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2. *Portable Distortion Monitor* J L Linsley Hood *Wireless World* July 1972 pp. 306-308
3. *Design Brief: Low distortion audio frequency oscillators* Ian Hickman *EW+WW* April 1992 pp. 345, 346.
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Power switching in SM packaging

Superior bipolar chip technology and enhanced lead-frame design have contributed to Super SOT – a new surface-mount transistor family with almost double the performance of its contemporaries. Martin Eccles reports.

Working towards maximum reliability, designers try to minimise the number of moving parts in their products. Nevertheless there is boom in tiny mechanisms. Hard and floppy disk drives, laptop printers, CD drives, cassette players and fax machines are just a few examples where miniature motors and solenoids are found in abundance.

Proportional to the boom in miniature moving electrical parts is the rise in demand for transistors to switch them. Even small motors, solenoids and lamps can demand significant power, particularly at switch on. High power dissipation and surface-mount PCBs however are incompatible. In SM applications, adding heat sinking or moving the power switching transistor to a cooler location is expensive. As a result, demand is highest for SM transistors that can switch more power without dissipating more heat.

A perfect transistor switches gigawatt loads without heating up. In practice, a number of parameters influence heat dissipation in switching transistors, saturation voltage being the primary factor. Mounted in traditional SOT23 surface-mount packaging, a transistor with a typical 0.5V saturation voltage is limited to a continuous current capability of less than an amp. Any further current causes the device's dissipation capability of less than half a watt to be exceeded.

By reducing saturation voltage and increasing power dissipation, Zetex has produced surface-mount transistors capable of switching resistive loads up to 100W with almost double the efficiency of conventional SOT23 devices.

Gain hold-up can also affect dissipation. As current through a transistor increases, its gain falls. This means that base current needed to achieve full saturation rises disproportionally with increasing collector current. If base current falls below the level needed to turn the transistor hard on at high currents, dissipation increases rapidly. The table and **Fig. 1** show that gain of the *FMMT* transistors is significantly higher than other popular devices which also have much larger packages.

	FMMT618	FMMT619	BCX54	BCP54
Package	SOT23	SOT23	SOT89	SOT223
BV_{CEO}	20V	50V	45V	45V
I_C	2.5A	2A	1A	1A
$I_{C(max)}$	6A	6A	1.5A	1.5A
hFE(min)	200@2A	200@1A	25@0.5A	25@0.5A
$V_{CE(sat)}$, max @ I_C	50mV @1A	200mV @1A	500mV @0.5A	500mV @0.5A
P_{tot}	625mW	625mW	1W	1.5W

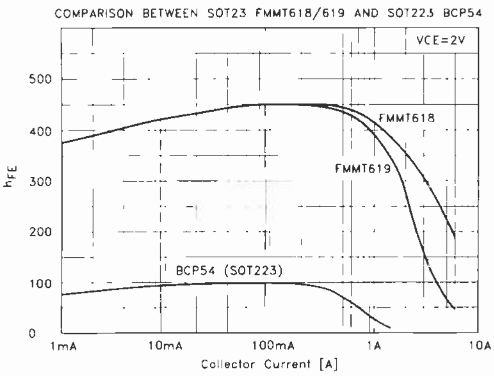


Fig. 1. Large-chip SM transistors like the BCP54 can handle the current but at a little over an amp their gain is no longer useful. The curve for the BCX54 is similar to that of the BCP device shown.

Advanced lead-frame design

By optimising thermal conductivity of the SOT23 package lead frame, Zetex has arrived at a SOT23 package with performance. To achieve this, the standard silver or gold patch plating of the lead form has been replaced by highly conductive plating. This reduces junction-to-case thermal resistance from 280°C/W to only 100°C/W.

This improvement in thermal resistance allows the package to dissipate 625mW when mounted on a ceramic substrate measuring only 15 by 15 by 0.6mm, or 1.25W on an infinite heatsink. The industry norm for SOT23 packaged devices is 300mW. For a given power dissipation, the devices run cooler than comparable products due to their low saturation voltage. Cooler operation allows increased packing densities, and in turn higher reliability and lower costs.

Matrix design

Pioneered by Zetex, the matrix chip design has had an important influence on performance, achieving even better results than large-emitter designs but without the inherent disadvantages.

In the matrix design, current distribution is uniform across the chip but there is no loss of emitter area. The ideal matrix chip is composed of an infinite matrix of vanishingly small current sinks distributed evenly throughout the emitter region, **Fig. 2**.

Base current is extracted through the current sinks. It is possible to fill over 90% of the base region with active emitter, achieving at the same time uniform current distribution. This reduces the saturation resistance of the transistor to such an extent that previously insignificant components of resistance become dominant. Detailed analysis of these components has led to reduction in saturation resistance.

Applications

Having a low saturation voltage and high current capability, the 20V *FMMT618* is useful in battery-powered applications. It can carry up

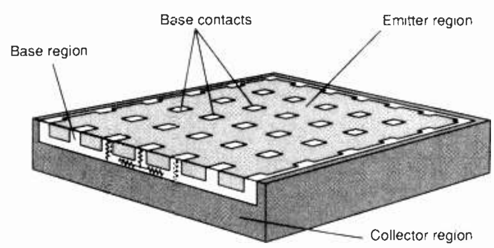


Fig. 2. In the matrix design, tiny current sinks are distributed evenly throughout the emitter region to ensure uniform current distribution without loss of emitter area. Saturation resistance is significantly reduced.

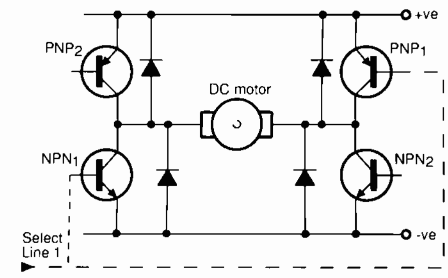


Fig. 3. Typical H-bridge driver allows full single-rail supply voltage to be applied to the motor in either polarity. Diodes may be needed to protect the transistors from transients.

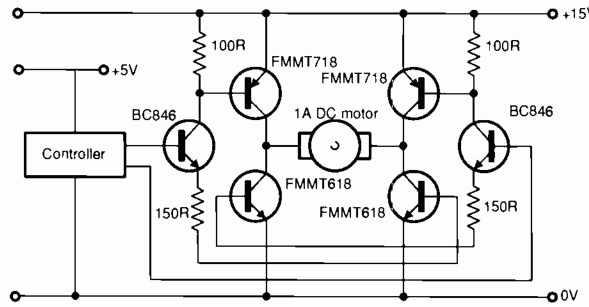


Fig. 5. A transistor's gain falls as its V_{CE} rises so current needed to drive the H-bridge pairs can increase above that available from logic ICs. With this arrangement, only two extra transistors are needed to buffer all four bridge devices.

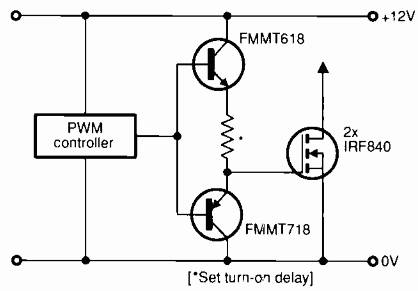


Fig. 6. Real input capacitance of a power mosfet gate can rise to nanofarad levels, causing significant problems if fast switching is needed. A driver pair combining high gain and high current capability can overcome this problem with a minimum of components.

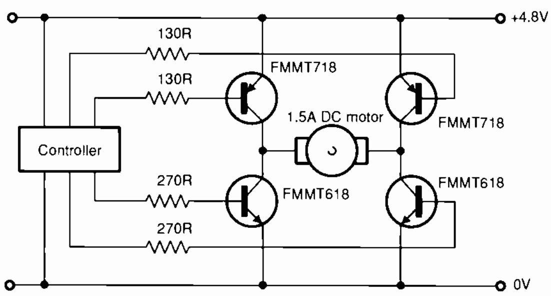


Fig. 4. Base resistors in this H-bridge motor controller are selected to suit maximum output current expected. Protection diodes are not needed since the transistors can absorb inductive transients.

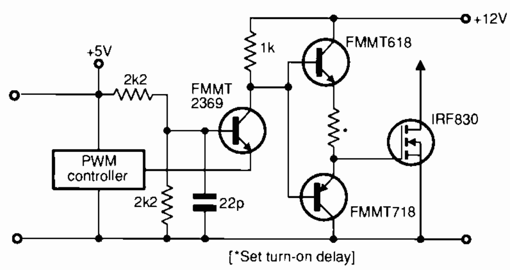


Fig. 7. In some cases, 5V logic levels cannot provide enough gate voltage swing for efficient switching. The pre-driver transistor has to have a very low storage time, resulting in a total-circuit turn-on time of only 20ns when driving a 2nF-input mosfet.

to 2.5A continuously and has a typical saturation rating of 130mV. Minimum gain at 2A is 200. Even at the device's peak current rating of 6A, gain is still more than 100.

Further devices in the current *SuperSOT* range are the 50V *FMMT619* and four pnp types with 20, 40, 70 and 100V ratings and current handling capabilities between 1 and 1.5A.

H-bridge motor drive. Motor drivers in H-bridge format are used in a wide range of products such as disk drives, toys, coin-control mechanisms and servo systems. They provide bidirectional outputs from single-polarity supplies, usually under digital logic control.

Two npn/pnp pairs are normally used, all operating in grounded-emitter mode. By turning on one npn device and its diametrically-opposed opposite pnp device, virtually all the supply voltage can be applied across the motor. Switching the second pair reverses the supply to the motor, **Fig. 3**. Often, H-bridge transistors need collector-emitter diodes to protect them from regenerative currents and transients stemming from the motor.

In battery-powered applications, it is vital that as much of the supply voltage as possible is applied across the load. This maximises battery life through greater efficiency and minimises the effects of falling battery voltage.

With an *FMMT618/718* pair, the circuit of **Fig. 4** will handle load or stall currents to 1.5A. Adapting the circuit for lower current motors is simply a matter of increasing the base resistor values. Base current at maximum load should be 2% of collector current for the pnp types and 1% for the npns.

Saturation losses at 1.5A total only 0.3V and can easily be halved for lower load currents. If the circuit is used to deliver high current, logic drivers need to be able to deliver 30mA. The diodes are unnecessary since the transistor reverse voltage rating is high enough to handle transients.

Figure 5 is for higher supply voltages or applications where logic output drive is insufficient to drive the previous circuit. Low-cost buffer/level-translators have been added. These compensate for the fall-off in gain with increasing supply voltage, which is inherent in all transistors.

Mosfet gate drive. Input capacitances of power mosfets and igbts can rise as high as tens of nanofarads. When Miller effects – i.e. amplification of feedback capacitance – are taken into account, by using the more valid method of evaluating gate charge as opposed to C_{iss} for calculating effective input capacitance, values around three times higher are obtained.

To minimise switching losses, particularly in high frequency converters, it is vital that the gate capacitances are charged and discharged as rapidly as possible. Consequently driver circuits must act as low impedance voltage sources, capable of supplying large transient charge currents.

Since standard switching power supply control ICs are rarely able to drive larger capacitance mosfets adequately, a high speed buffer is often used.

Complementary emitter followers as shown in **Fig. 6** can act as an ideal buffer provided that transistors of high current capability combined with high f_T are used. Combined, the *FMMT618* and *FMMT718* have these characteristics. The 10nF effective capacitance of two *IRF840* mosfets in parallel can be charged to 12V in under 30ns – a feat requiring a peak current of around 4A. Included in the the circuit is a resistor for introducing a turn-on delay without affecting turn-off performance. This component is sometimes needed to avoid cross conduction problems in push-pull output stages.

Where 5V logic provides a pulse-width modulation, a buffer can be necessary to translate level, giving 10V or greater gate drive for the power switches. By using an *FMMT2369* switching transistor, the circuit shown in **Fig. 7** converts 5V logic drive to a 12V gate drive signal. Driving the emitter of the *2369* from the logic output avoids signal inversion.

Storage time of the switching transistor is very short. This combines with the high gain *FMMT618* to give the circuit a turn-on time of only 20ns when driving a mosfet with an

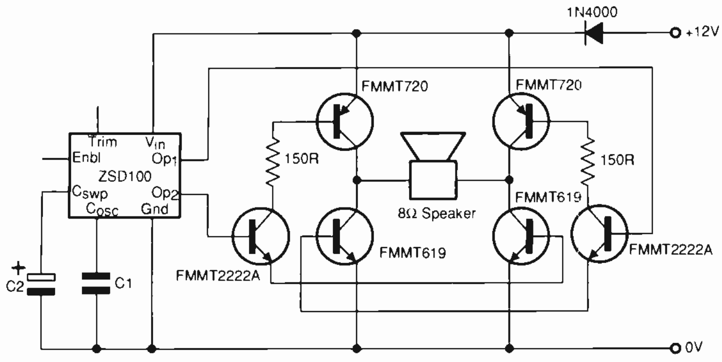


Fig. 8. Burglar-alarm siren drive for an 8Ω loudspeaker is normally provided by large TO126 or TO220-packaged transistors. Using FMMT transistors not only reduces cost and size but also eliminates the need for protection diodes.

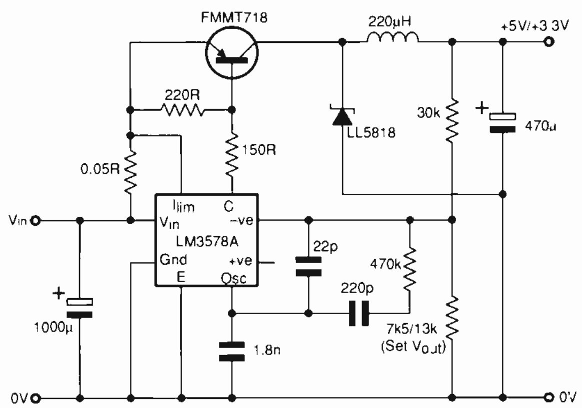


Fig. 9. In battery-powered DC converters, the key to maximising efficiency is to minimise voltage drops over all high-current paths. Next to the inductor, saturation voltage in the switching transistor is usually the biggest loss contributor.

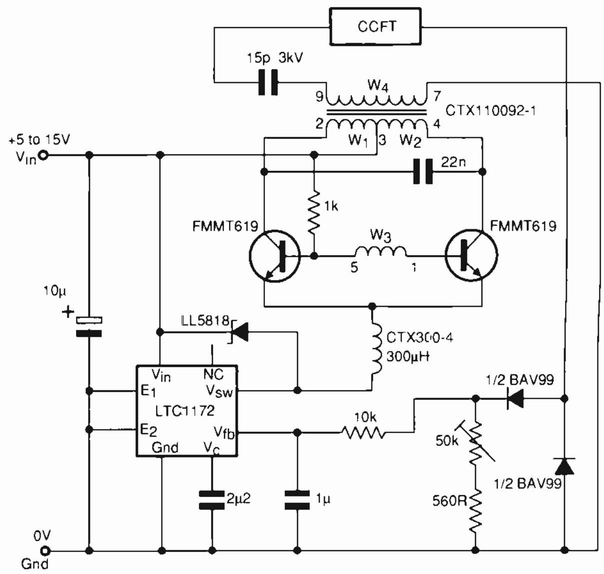


Fig. 10. Driving a cold-cathode fluorescent lamp used in, say, computer backlighting, needs an efficient, compact converter. This design originally had SOT223 BCP65 transistors. Replacing them with smaller FMMT types reduced saturation voltage by half.

effective input capacitance of 2nF. The *FMMT718* helps make turn-off times even shorter, leading to reduced cross-conduction problems in bridge or push-pull converters.

By providing excellent high-current performance in a SOT23 package, the *FMMT618* and *FMMT718* replace SOT223 and SOT89 transistors in these gate drive circuits leading to cost and PCB area savings – particularly in very high frequency converters.

H-Bridge Siren Driver. Many modern burglar and automotive alarm sirens employ an 8Ω moving coil loudspeaker driven by a bipolar H-bridge. Handling peak output currents of 2A, traditional TO126 or TO220 packaged output transistors require parallel collector-emitter diodes. These are needed to divert destructive reverse transients generated by the inductive load.

In **Fig. 8**, *FMMT619* and *FMMT720* transistors in SOT23 packaging replace these bulky and expensive leaded transistors, giving other savings too. High reverse h_{FE}, inherent in matrix technology, eliminates the need for parallel collector-emitter protection diodes Both *FMMT619* and *FMMT720* transistors conduct reverse collector current sufficiently well to clamp any inductive transients generated by the load.

A specially designed Zetex asic provides a variable frequency drive to the SOT23 H-bridge ensuring a very loud and irritating noise. The combination of an ASIC signal generator and SOT23 H-bridge produces a compact and inexpensive module.

DC-DC Converter. Using standard PWM controllers, it is easy to construct buck-type step-down converters with low component counts. Harder to achieve are designs that are both simple and efficient – as required for modern battery-operated equipment. The key to maximising efficiency is eliminating voltage drops in all high current areas.

In the buck converter shown in **Fig. 9**, the high current paths are via the 50mΩ sense resistor, the series switching transistor output inductor L_1 and the schottky diode. Once resistance of the output inductor has been minimised the most critical aspect is the saturation voltage drop of the switching transistor. Saturation is particularly important when V_{IN} approaches V_{OUT}.

By using an *FMMT718*, which drops only 200mV @ 1.5A, this converter can operate at an efficiency of over 90% at minimum input voltage and an I_{OUT} of 1.5A. Even when output current has fallen to 200mA, efficiency is still around 80%.

As input voltage increases, the operating gain of the switching transistor becomes more important. The high gain of this transistor minimises base drive losses, leading to high efficiencies over a wide supply range.

Fast rise and fall times of the *FMMT718* allow the converter to operate at 50kHz with minimal switching losses. At this frequency it is essential to use low ESR input and output capacitors and keep any wires carrying switched high currents very short so as to minimise rfi and output ripple.

The converter will operate from a supply of V_{OUT}+0.5V up to 16V. Its output voltage can be set to 5V or 3.3V by altering the value of R_1. The circuit will supply loads from 0 to 1.5A, current limiting to around 2A with a shorted output.

LCD backlighting converter. Cold-cathode fluorescent lamps used for portable computer lcd backlighting and similar applications require a converter generating between 1 and 2kV to strike and run.

Standard circuits provide control of tube brightness against input supply variations and other factors such as temperature, tube ageing etc. These circuits commonly use SOT223 transistors in the high voltage converter. This is because high currents must be passed with minimal saturation losses if good efficiency is to be achieved.

In **Fig. 10**, SOT223 *BCP56* transistors have been replaced with smaller *FMMT619* SOT23 types. Exhibiting a saturation voltage of only 125mV at 1A – less than half that of the *BCP56* – the *FMMT619* not only reduces cost and PCB area, but also raises efficiency of the converter. ■

USING RF TRANSISTORS

Getting a grip on instability and load mismatch

***Instability in solid state amplifiers may well be the most difficult problem a designer must face. Norm Dye and Helge Granberg review some common causes. The authors also explain the practical implications of using the vswr bridge to handle load mismatching. From the book* RF Transistors: principles and practical applications.**

Engineers sometimes talk about unconditional stability, where no matter what the load is, the amplifier does not exhibit spurious oscillations even with drive levels and supply voltages outside nominal values. But such conditions rarely exist with rf amplifiers – except possibly in low power class A designs.

Instabilities can be observed in several ways. Some types may be too low in amplitude to be detected with anything but a spectrum analyser, or may appear only outside the nominal level of drive power and supply voltage. But others may become apparent through erratic tuning (if tuning elements are provided) or if current is being drawn when the drive power is removed.

At least three variables affect amplifier stability: the load (*R* and *X*); drive level and supply voltage. In mobile communications the nominal supply voltage is 12.5V, but can vary down to as 10.5V or as high as 16V.

With low level amplitude modulation and ssb, effective drive power varies with modulation. High level AM collector/drain voltage varies between zero and the maximum.

Taking into account the two latter variables, plus the *R* and phase angle (*X*) of the load, we can see how difficult it is to design a stable rf amplifier operating under these conditions. Many designers have spent hundreds of hours trying to get an amplifier to meet a stability specification even at 3:1 load mismatch.

Putting aside the drive level and supply voltage variables for a moment, it is relatively easy to reach stability in an amplifier operating into a resistive (usually 50Ω) load.

But in real life, there is always some load mismatch. In communications, for example, the load is an antenna connected to the amplifier output through a harmonic filter. In industrial and medical applications the load can consist of various types of matching networks presenting, at least momentarily, an undefinable load to the amplifier.

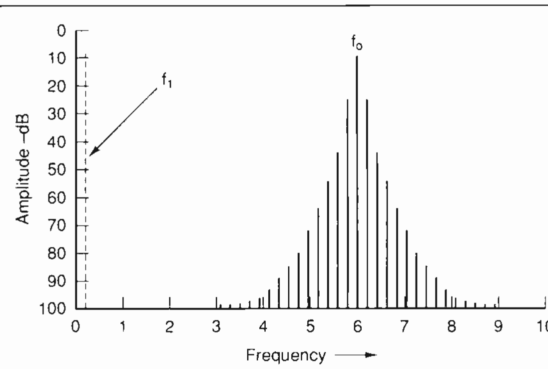
Amplifiers designed for frequency modulated communications have one fewer variable because the power input remains constant. This should make design of stable amplifiers for FM somewhat easier than for AM. Several types of instability can be identified in rf amplifiers, some are circuit or layout oriented; some are device oriented or may be a combination of both.

Many oscillation modes depend strongly on non-linear effects. So they can be very difficult to analyse compared with the small-signal feedback type oscillations that are adaptable to linear circuit analysis.

Ratio of feedback capacitance to input impedance (feedback capacitance to input capacitance in a mosfet) in a power transistor determines much of its stability criteria. The higher the ratio, the greater the possibility for device stability. Normally feedback capacitance introduces negative feedback, reducing power gain. But at certain frequencies the feedback will turn positive due to phase delays. Clearly, devices with higher ratios of feedback capacitance to input impedance will exhibit the most stability.

Transistors processed for low voltage operation generally have lower ratios, making stability in 12.5V systems more elusive than in, say, a 50V design. For bipolar transistors, feedback capacitance is not given in data sheets because it is not easy to measure. It is a function of many parameters such as device geometry, types and values of emitter ballast resistors and the silicon material resistivity. The ratio of input impedance/capacitance to feedback capacitance is somewhat higher with mosfets rather than bjts, and so mosfets are more stable in this respect.

Fig. 1. Low frequency instability. Frequency f_1 is mixed with the carrier (f_0) which produces a series of sidebands around f_0.



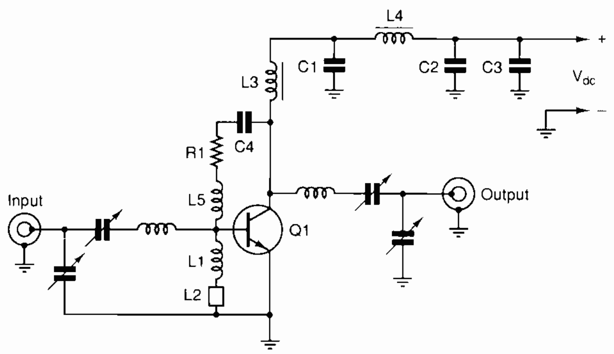


Fig. 2. Typical rf amplifier showing the appropriate dc feed structures and a negative feedback network.

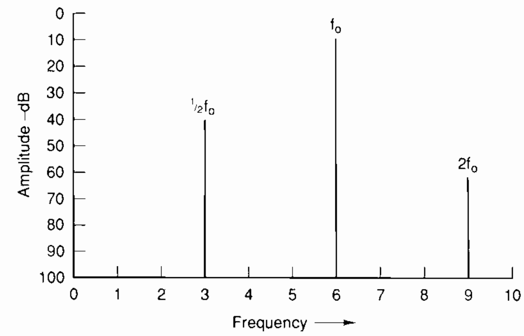


Fig. 3. Display of half f0 instability, caused by varactor effects primarily in the base-emitter junction.

Low frequency factors

One of the most common instabilities occurs at low frequencies (1-10MHz), where device power gain – assuming use of vhf or higher frequency transistors – can be as high as 30-40dB.

Oscillation can be strong, but noticeable only by its mixing products with fundamental f_0 (**Fig. 1**). In professional circles this display is referred to as a Christmas tree, the width of its skirts depending on amplifier bandwidth.

Frequency of oscillation, f_1, (Fig. 1) can be high in amplitude, but may not be detected on a spectrum analyser because of bandwidth limitations of the circuit. In some instances this low frequency oscillation can be strong enough to cause a transistor to exceed its dissipation limits and destroy itself.

It is a mode of instability that is almost completely circuit-oriented and is mostly preventable by controlling the low frequency power gain of the amplifier.

Selecting a transistor with low h_{FE}, controls low frequency gain, but has a minimal effect at high frequencies. Conversely, the emitter/source inductance and resistance have a larger effect in the device's high frequency gain and a lesser effect at low frequencies. Keep these values at their minimum for best amplifier high frequency performance.

Low frequency gain of the amplifier can be lowered by certain simple design practices – though the available gain of the device itself remains unchanged. In a typical rf amplifier, values of the most critical dc feed elements (L_1 and L_3, **Fig. 2**) should be selected to block the low frequencies, where device gain is considerably higher than it is at the frequency to be amplified. Values should be as low as possible without resulting in any loss in power gain or efficiency.

To be on the safe side, reactances should not exceed 5-10 times the impedances at the base and collector.

Q values of L_1 and L_3 must also be controllable. If this cannot be done with lossy ferrite beads (L_2 for example), parallel resistances can be used.

Common practice is to wind the chokes (L_1 and L_3) over low valued (10-50Ω) non-inductive resistors. This is particularly so for the collector feed choke (L_3) where a lossy ferrite bead may face excessive heating at the high power level.

C_1 (Fig. 2) must have a large enough value to bypass these low frequencies to ground. To avoid possible resonances, multiple capacitors of different values (0.01 and 0.1μF for example) are sometimes paralleled.

Reactance of L_4 is high enough to pass a minimum of the low frequencies. In fact, it should be as large as possible up to the point where ir losses start producing excessive dc voltage drop.

Capacitors C_2 and C_3 again are two different values and paralleled to avoid resonances. Their values should be large enough to bypass all low frequencies to ground.

The $L_4/C_2/C_3$ network should prevent any rf from feeding back to the dc power source.

Another more effective means of reducing low frequency gain of an rf amplifier is to introduce negative feedback – the purpose of $C_4/R_1/L_5$ network in Fig. 2. C_4 is merely a dc blocking capacitor. Its value is not critical, except it must be large enough to provide a low reactance at low frequencies.

Feedback slope is controlled by L_5, whose function is based on its increasing reactance with frequency. L_5 is determined to produce minimum feedback at the operating frequency and maximum feedback at low frequencies, where its reactance is low.

Resistor R_1 controls the overall level of the feedback, and its value is normally very low, except where it is used in conjunction with L_5 to control the gain slope.

Another cause for low frequency instabilities can be the physical layout of the circuit.

Benefits of good layout

The most important point in rf amplifier layout is to provide a solid ground plane. A good ground plane will minimise generation of rf ground loops that can feed rf energy back to the input in a phase that would make an oscillator out of the amplifier.

In most cases this problem can be fixed only by making a new circuit layout – which generally proves costly.

Excessively high matching network Qs, and high Qs of the dc feed networks (if the input and output networks happen to resonate) can also result in self-oscillations at some intermediate frequency. But these high Qs can be prevented by following proper design guidelines.

Proper design can also prevent inductively induced feedback. Such instability is more common in hf and vhf amplifiers, with lumped constant matching elements than in, for example, microstrip designs. Enough rf energy can be coupled to the inductor(s) of the input matching network, from the output, to trigger an oscillation.

Oscillation occurs at a frequency where the input-output phases approach 360° – though it may not be enough to destroy the transistor. It usually disappears and snaps to the driven frequency when input drive is applied.

A way to prevent oscillations, is to locate the input and output matching networks physically as far apart as possible, while orienting the inductors of the input and output networks in 90° angles. Electrostatic shielding may also be useful, though it has only proven effective in small-signal designs.

Varactor effect instability

In addition to multiplication, a varactor multiplier can also generate sub-frequencies if a selective circuit is provided for those frequencies. The effect is known as varactor instability, and implies that there are $2.f_0$ and $3.f_0$ products present, though they would fall on the harmonics and would be hard to distinguish. Their amplitudes are probably much lower than that of the 0.5 x f_0, since system power gain is much lower at these higher frequencies.

Instead, a stronger 0.5 x f_0 spur is generated since in most cases the bandwidth extends to those frequencies and sufficient power gain is available (**Fig. 3**). The 0.5 x f_0 oscillation is usually of a fairly low amplitude and does not noticeably affect amplifier performance. There is no real cure for it, and it is most likely to occur in low gain amplifiers of classes B and C. One possibility may be to add a half frequency band reject filter to the amplifier output. But this only works in relatively narrow band designs.

In classes AB and A, the diode junctions do not go out of forward-conduction and the 0.5 x f_0 phenomenon does not usually occur.

The ideal approach is to use a diplexer to provide a proper resistive load for the amplifier at its harmonic frequencies even if there is a load mismatch at f_0.

Another common practice in rf amplifier stabilisation is to insert a resistive attenuator between amplifier output and load. The attenuation need only be 1-2dB. But there is always a power loss and so the technique is practical only where stability is more important than system efficiency.

Advantage of resistive loading is that in addition to isolation, the resistive load (although not 50Ω) is provided for the amplifier at all frequencies.

For 1dB attenuation, the resistor would have a value of 440Ω, while for 2dB the value would become 220Ω.

Testing stability

Testing an amplifier for instabilities can be accomplished by using a spectrum analyser to see the spurious responses (if any). An *LC* network can simulate a load mismatch having a reflection coefficient of near unity in magnitude and all possible phase angles. Any value of load mismatch can be realised by inserting an attenuator between the amplifier output and the complete mismatch simulator (**Fig. 4**).

Generally, at uhf and microwave frequencies, the desired magnitude of reflection coefficient is achieved by a transmission line attenuator terminated in a short circuit. Variation in phase angle of the load reflection coefficient is accomplished with a line stretcher.

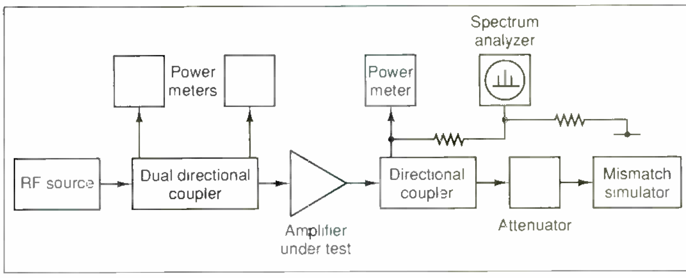


Fig. 4. Test set-up for measuring stability of rf amplifiers under various load mismatch conditions.

Class C amplifiers operating at low voltages are the least stable and high voltage units of classes A and AB exhibit the best stability. Complete stability (no spurious oscillations) of a low voltage class C amplifier into a 3:1 load mismatch would be acceptable, and into 5:1 would be excellent.

Stability of an amplifier can be analysed using large-scale *S*-parameters. But this will not help in design. Many variables affect stability and since it is largely circuit-layout and component dependent, computer aided design gives no guarantee of a stable amplifier. If stability measurements are unacceptable, there is little to be done except go back to the drawing board. Circuit design must be re-examined, including board layout, following the proven guidelines.

VSWR protection in load mismatching

Most transistors fail in solid state amplifiers at load mismatch phase angles that present a high current mode of operation to the output transistor(s). The result is an increase in power dissipated by the transistor(s).

Since the temperature time constant of a typical rf power transistor die is 0.5-1.0ms, any protection system (including all delays in the agc/alc loop) must react faster than this.

Several methods can be used for protecting solid state rf power amplifiers against load mismatches, but reflectometer vswr sensing is most common.

The reflectometer is usually located in series between amplifier-output and load, and produces a voltage proportional to the amount of output mismatch. This voltage is processed and then fed back to either the power amplifier input, or one of the preceding stages, gradually reducing the power gain or completely shutting down the amplifier.

The standard reflectometer principle used here to detect rf power amplifier output mismatch is commonly known as a vswr bridge. With proper mechanical design, its use can be extended to microwave frequencies (greater

Matching networks in practice

Design calculations for matching networks can become completely meaningless unless the network components are measured at the operating frequency. For example, a 100pF silver mica capacitor that meets all specifications at 1MHz can have an effective capacitance of 300pF at 100MHz, due to its series inductance. At some frequency, this inductance will tune out the capacitance altogether leaving the capacitor with a net inductive reactance.

Values of inductance in the low nanohenry range are also difficult to achieve, since the inductance of a 25mm straight piece of AWG #20 solid copper wire is approximately 20nH. Component tolerances have no meaning at vhf frequencies and above, unless they are specified at the operating frequency.

Unencapsulated mica capacitors – Unelco, Underwood, Standec, Imenco, Semco – are widely used in rf designs, from low band to uhf. They are more rugged than ceramic chip capacitors but have higher series inductances.

Unelco is a common name for these capacitors, coming in two basic physical sizes. At vhf or uhf, their real values must be adjusted according to the frequency of operation. Parasitic inductances for Unelco and Miniunelco are 1.5-2nH and 1-1.2nH respectively.

The following equation has sufficient accuracy for determining the required low frequency value when the effective value and frequency are known:

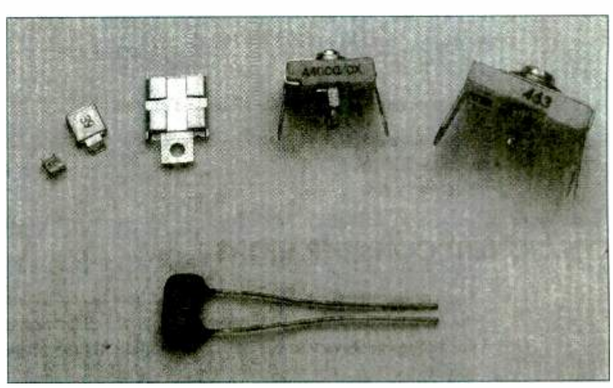
$$C_{nom} = C/(1 + ((2f\pi)^2 LC)10^{-9})$$

where C is the effective capacitance in pF, L is the parasitic series inductance in nH and f is frequency in MHz. Assuming a capacitance of 100pF is required at 400MHz, and we wish to use a Miniunelco:

$$C_{actual} = 100/(1 + (2512^2 1.0 * 100 * 10^{-9}) = 61.3\text{pF}.$$

So the actual low frequency value of the capacitor required for a 100pF effective value is almost 40% lower. The nominal value at 150MHz would be 91.8pF (as a comparison), which is well within the standard 10% tolerance limits for these components.

Any type of capacitor's nominal value can be calculated with this equation, as long as its parasitic inductance is known.



Various types of capacitors used in rf power circuits. From left to right (upper): a multi-layer ceramic chip, Miniunelco, standard Unelco, and two types of compression mica variables. Lower centre is a dipped mica or silver mica suitable for use up to vhf with very short leads.

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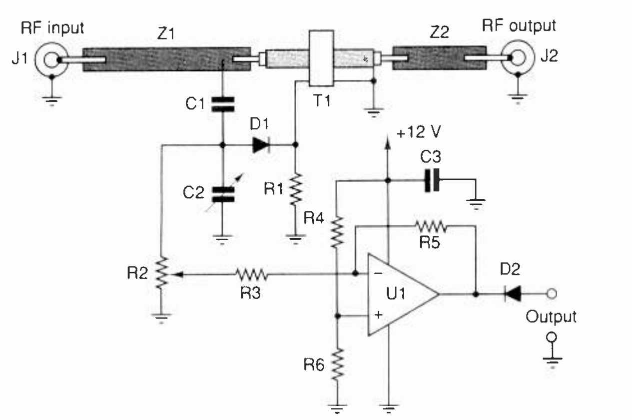


Fig. 5. VSWR protection circuit, referred to in subsequent figures as the VSWR sensor. The forward port has been omitted since the system is not intended for forward power measurements.

than 1GHz). UHF and microwave designs usually employ microstrip transmission line techniques, whereas lower frequency circuits favour lumped constant implementations.

In fact, up to uhf, the lumped constant concept is probably the most practical way to approach the coupling coefficient required between current line (amplifier output) and sample line to produce an output voltage of moderate amplitude. Passing the current line through a multi-turn pick-up coil achieves a tight coupling in a lumped constant system. The effect is of a transformer, with the current line as the primary and the multi-turn coil the secondary. The multi-turn winding is usually toroidal allowing magnetic material to be used as the core, increasing low frequency response.

Inductive reactance of the multi-turn winding must be greater than the load impedance of the current line at the lowest frequency of operation. But the ports are usually terminated into low reactance dummy loads equal to the characteristic impedance of the system – in most cases 50Ω.

High frequency limits of the vswr bridge are determined by leakage inductances and physical length of the multi-turn winding. Whenever the length of the secondary winding (sample line) equals a wavelength divided by 2^n, where n is an integer 1, 2, 3, etc., there will be resonances. But their amplitudes diminish as n increases. Keeping sample line length shorter than $\lambda/16$ ensures amplitudes of any resonances are negligible.

Bridge operating principle

Voltage across the multi-turn secondary of the transformer is proportional to the current passing through the current line and number of turns in the sample line. When the amplifier has a perfect load (at *J2*, **Fig. 5**) the rms rf voltage measured across the forward port (terminated with 50Ω) would be V_{rms}/R_L decreased by the amount of coupling (C_p) between the current line and sample line.

Similarly, the rf voltage at the reflected port relates to the rf power reflected at the output port (*J2*).

If the load at the output port is totally mismatched – open or shorted – voltage at the reflected port will equal the voltage at the forward port.

To produce a useable voltage in either the forward or reflected port, coupling coefficient C_p should not be higher than 30-40dB for power levels in the main line of 100-1000W. For example, if C_p = 30dB and we have a 100W amplifier, the power appearing at the forward port of the sample line will be 100mW and the voltage will be $(PRL)^{0.5} = (0.1 \times 50)^{0.5} = 2.24$V.

Coupling coefficient C_p can be figured as $C_p = 20\log[(1 + [1/(2N^2)])/(1/N)]$. N is the number of turns in the sample line.

Reversing the equation gives $N = 10^{C_p/20}$ where C_p is the required port coupling coefficient in dB.

Then if C_p = 30dB: $N = 10^{30/20}$ = 32 turns (for one-turn primary).

Input-output insertion loss = $20\log[1+(1/2N^2)]$=0.0042dB, and input return loss = $20\log[2N^2 + 1]$ = 66dB.

In addition to the voltage derived from the secondary of the toroidal transformer, a voltage sample is taken from the current line using a capacitive divider (C_1-C_2, Fig. 5). Half wave rectified voltages are created at the junction of C_1 and C_2 by the transformer secondary, 180° out of phase in the case of a non-mismatched load.

Amplitudes are made equal with C_2, reducing the voltage to near zero at the junction of C_1 and C_2 until a mismatch in the load causes the phase shift to deviate from 180°.

Mechanical restrictions clearly place a limit on the circuit bandwidth. Design of extremely wide-band and high power systems is difficult since, for high frequencies, the toroidal pick-up coil should be as small as possible, while low frequencies require it to be large enough for the minimum reactance required.

High permeability (μ = 100 and higher) ferrites in the toroid are usually too lossy at high frequencies and will heat up even at moderate power levels. For example, at 150MHz, materials with μ = 15 or less have been found acceptable.

A Faraday shield employed between the current line and toroidal winding, preventing capacitive coupling between the two, can best be accomplished with a length of coaxial cable of proper characteristic impedance. The inner conductor forms the current line and the outer conductor the Faraday shield.

Normally, only one end of the Faraday shield is grounded to prevent formation of a shorted rf loop. But if the length of the Faraday shield is considerably smaller than the ground loop, it can be grounded at both ends – for mechanical reasons, for instance.

Good high frequency performance calls for a solid ground plane in the 'current line' and 'sample line' area. Otherwise the resulting ground loops may reduce the circuit frequency response or produce uneven response characteristics as a function of frequency.

Practical circuits

The circuit shown has been tested simulating a load mismatch of 5:1 at a power level of 1kW at 30MHz, and up to 200W at 220MHz (**Figs. 6** and **7**).

A fast operational amplifier such as the *MC34071* – 13V/μs slew rate – will allow output switching at 2ms, rapid enough to protect the majority of rf power amplifiers.

Most operational amplifiers can sink currents up to 20mA, sufficient to turn off directly the bias voltage of an enhancement mode mosfet for example; or an emitter follower can be added for higher current requirements. In fact the op-amp output can be made the main bias source to provide the mosfet gate bias voltage.

Controlling the gate voltage of a mosfet for a gradual gain reduction would not be possible in linear operation since a steady idle current has to be used. Such cases require use of some type of voltage or current controlled rf atten-

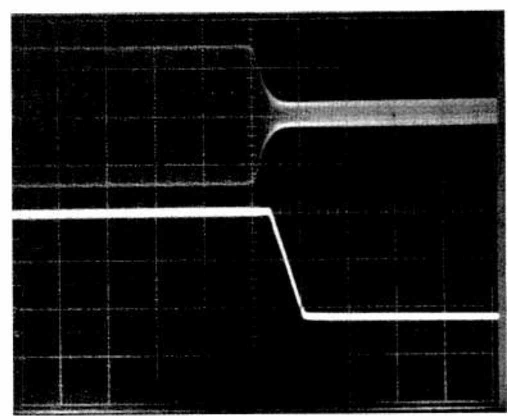


Fig. 6. RF envelope and amplifier output of vswr sensor. Horizontal scale is 2μs/div. Vertical scale is 5V/div.

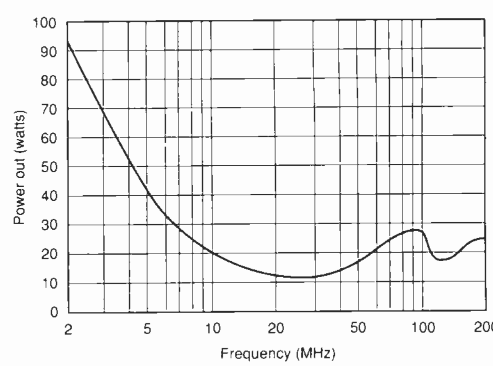


Fig. 7. Sensitivity vs frequency response of the VSWR sensor at 5:1 load mismatch.

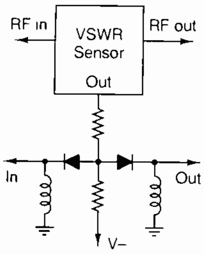


Fig. 8. A pin diode switch is used here and is adaptable with either mosfet or bjt amplifiers. A current boost for the diodes may be necessary to drive them into full conduction, depending on the type of diodes used and the signal level.

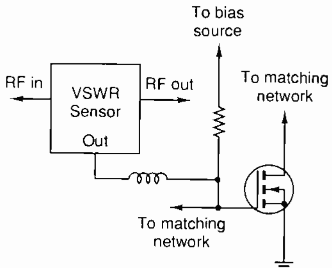


Fig. 9. Typical application for the VSWR sensor (shown in Fig. 5), where its output directly controls a mosfet gate bias voltage at a low level stage.

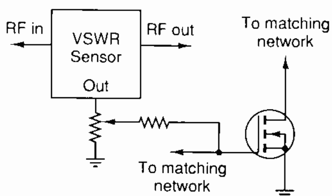


Fig. 10. Similar to Fig. 9, except that the VSWR sensor acts as the gate bias voltage source for the fet in addition to providing a shut-off function.

uator, preferably in the low level pre-stages (usually operating in class A), which are insensitive to variations in output load.

Attenuators, such as a pin diode (**Fig. 8**), are the only way to control the power gain of bipolar transistor amplifiers since the mosfet agc function is not available.

Depending on attenuator characteristics and power level, the power output can be adjusted, for a given output mismatch, using a combination of R_2 and R_5 (Fig. 5). For this, and the circuit given in **Fig. 10**, D_2 must be shorted to use the output of IC_1 for a voltage pull-up function.

Fast shut down of the amplifier, without linearity requirements, can be achieved simply and adequately with the circuits given in **Figs. 9** and 10. For best results, an early stage in the amplifier chain should be controlled, since low power mosfets have low gate input capacitances, speeding up shut off.

Circuit testing

Specific amounts of load mismatch must be developed for testing a system. For example, in some applications a fold-back may be desired at 3:1 or 5:1 output vswr, whereas in others a load vswr of 10:1 may be tolerable.

Component losses make an infinite mismatch impossible, so a value of 30:1 has been adapted as a standard for "infinite mismatch" by the industry. The scale is logarithmic so there is not much practical difference between 30:1 (or even 20:1) and infinite.

A 30:1 mismatch covering all phase angles and R from nearly zero to open circuit can be simulated with an LC network (**Fig. 11**). In the diagram, C consists of two similar variable (air) capacitors whose voltage ratings depend on rf power level. C can also be a butterfly dual capacitor, where the wiper can be used for the ground contact.

Minimum-maximum capacitance ratio should be at least 5-6 to obtain a coverage for all phase angles and values of R. The initial maximum capacitance values are not critical and will only slightly affect the circuit Q and the values of L. Typical values for 30MHz are 300-400pF; for 100MHz 40-50pF; and for 200MHz 10-15pF. Ls are usually air-wound inductors, physically large enough to handle rf currents at the power level in question. L values can be calculated as:

$$L_1 = 1/[(2\pi f)^2 C_{(min)}]$$
$$L_2 = 1/[(2\pi f)^2 C_{(max)}]$$

For example, if f is 100MHz, $C_{(min)}$ is 7pF and $C_{(max)}$ is 40pF, then: $L_1 = 1/[628^2 \times 7] = 362$nH; and $L_2 = 1/[628^2 \times 40] = 63$nH. The same function can be accomplished with a single inductor and a differential capacitor, where one section is at its minimum capacitance while the other is at its maximum. Their capacitance ratios are roughly the same as the inductance ratios in the network previously described.

The circuit is in the familiar pi network configuration and is widely used for testing amplifier stability. 30:1 mismatch can be reduced by inserting a power attenuator between the circuit and the amplifier output through the vswr sensor.

The attenuator must, of course, be able to handle the power level in question. But remember that an attenuator only dissipates part of the power fed into it. A 1dB attenuator, for example, dissipates only 10% of the power. So one with a 100W rating could be used at a power level of 1kW, providing its resistor elements can handle the current.

Attenuation in dB to produce a specific vswr between a signal source and a 30:1 load mismatch can be calculated. First obtain a value for the magnitude of the voltage reflection coefficient (Γ) as $|\Gamma| = (VSWR-1)/(VSWR+1)$. Then, $RL = 10\log_{10}(1/|\Gamma|^2)$, where RL is the return loss.

For the condition of a load return loss of 0dB (load is open or shorted), the value of attenuation in front of the open/shorted load needed to achieve a particular vswr is equal to one-half the return loss created by the desired vswr. For example, if we wish to create a 5:1 'load' vswr when the actual load is a short circuit, $|\Gamma|=4/6=0.67$, and $RL=10\log_{10}[1/(0.67)^2]$ = 3.5dB. Then the value of the attenuation is 3.5dB/2 = 1.75dB.

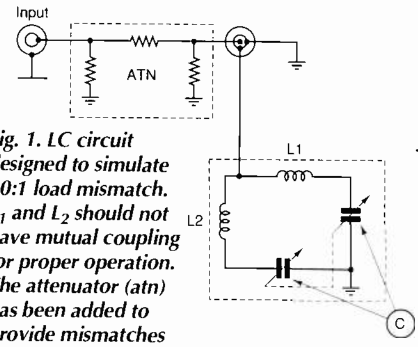
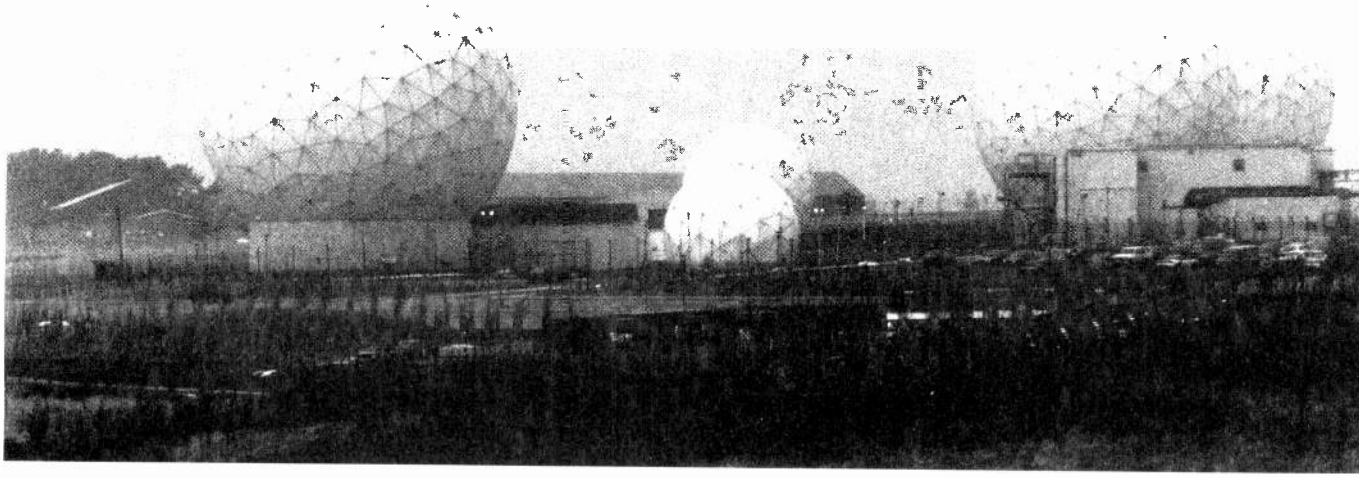


***Fig. 1.** LC circuit designed to simulate 30:1 load mismatch. L_1 and L_2 should not have mutual coupling for proper operation. The attenuator (atn) has been added to provide mismatches at various levels of swr.*

The vswr simulator can be made to function over about 30% bandwidth, and with such a broad band design, L_1 should be calculated for the high frequency limit and L_2 for the low frequency limit. Several networks would be needed to test multi-octave bandwidth amplifiers.

An advantage of this type of set-up to create load mismatches is that it can be adjusted to any phase angle: different phase angles are needed to simulate loads in laser drivers, plasma generators, communications equipment, and certain medical instrumentation. ■



Jon Mitchell

Big Brother's protection racket

Protecting computer data from hackers and tappers is a difficult task. But there's one method of protection that is just about unhackable and the major financial companies have been using it for years: DES encryption.

Any hacker who would tap a phone line or hack into a computer knows that it's not worth trying to unscramble information so encrypted. Algorithms like DES (the Data Encryption Standard, developed by IBM and much-used in government and financial circles) and RSA (also known as public-key encryption) are still as near to 100% secure as it is possible to get.

The beauty of DES and RSA lies in the fact that everyone knows how they work, but no one can crack them – not even government agencies. For a simple explanation of how RSA works, think of two prime numbers. Make them big ones, with at least 500 digits. Now multiply them together and you get a very large number. Now try reversing the process. Take the very large number and try to find out which two 500-digit prime numbers were used. There's only one solution (that's a feature of prime numbers) and it would take literally hundreds of years of computing power to do. Double-encryption (encrypting already-encrypted data a second time, with a different password) adds to the security.

With DES-like encryption available easily and cheaply to anyone who wants it, it was only a matter of time before criminals came to appreciate its advantages. Any drug dealer worth his salt now keeps his customer list on a PC and protects it with an encryption program. And any gang that's planning a bank job via electronic mail makes sure that all its messages are so encrypted.

Not surprisingly, governments all over the world are worried. It used to be possible for organisations like GCHQ or the NSA to tap almost any voice or data conversation and decipher the contents. Nowadays, even the most sophisticated systems can't cope with DES encrypted information.

So, in the US, the Clinton administration has come up with a plan. If the President can get the necessary legislation through Congress, further use of DES, RSA and any other encryption system within the US will be outlawed. There's even talk of Europe going down the same route.

Under the proposed new laws, if a user wants to encrypt data or voice traffic, they will have to do so with a system based on a new device called the Clipper chip. As the name suggests, Clipper is a hardware device which, according to the US government, provides totally unbreakable encryption... unbreakable that is by anyone other than the Government. It will be made available to any manufacturer wishing to incorporate encryption in their products.

So far so good. But the computing and electronics community is objecting to two aspects of Clipper. First, it's a proprietary algorithm designed by the intelligence agencies in the US. No one knows the algorithm, or anything about its functioning. Potential users are simply being asked by the National Security Agency to trust them when they say it's safe. So is it really secure? Will someone be able to crack it? No one is telling.

The second objection is more serious, and it is all to do with Clinton's desire to stop criminals encrypting their data. Basically, there is a special password which can crack all Clipper-encrypted data. Any data whatsoever – without access to the original password used for encryption.

This "back door" key will, say Clinton's team, be held securely by the US Government and used only when a judge grants a warrant. But the mere fact that such a key exists is worrying a lot of people. What if the key should fall into the wrong hands? What if someone else manages to make their own version of the key? Can any encryption algorithm be considered secure if it's possible to crack it with a back-door key?

> ...governments all over the world are worried...

Even now, those who will potentially be forced to use the system are investigating possible methods of beefing up the protection. There are ways of doing this, of course, such as encrypting the data with DES before passing it through the Clipper chip. Then, even if the Clipper encryption is removed, the resulting information will still be safe. But as intentions currently stand, this process will become illegal in the US.

Many US citizens and companies are surprised that, in a country which espouses human rights, the Clinton administration is now attempting to deny the basic right of privacy.

There's much legal argument currently taking place. As yet, it's unclear whether the legislation will make it through without amendment. Most observers suspect that, if Clipper becomes mandatory, it will be in a watered-down form. Its opponents, which include many of America's largest and richest companies as well as its drug dealers, have not yet given up the fight for totally secure data.

Robert Schifreen

Security not guaranteed

There are, of course, encryption algorithms other than DES or RSA which are less secure. Vendors of PC application software packages have been using these for a long time... For instance the encryption system which password-protects wordprocessor files. These are anything but secure. There are at least two commercial cracker products available which will break files saved with the password feature on the industry's leading wordprocessor package. One costs $15 and takes around half a second to do its job while the other costs $185 and takes half a minute. The delay in the second program is probably deliberate, as no one would pay $185 for a job which takes half a second.

There is a third class of encryption algorithm: proprietary systems. They are typified by the adverts in the computer press for a certain software protection dongle. "What makes our system so secure?", asks the headline. "We'll never tell", it goes on to say. Maybe they won't, but as soon as someone finds out, or a disgruntled employee leaks the secret, the product, and everything protected with it, becomes open to the world.

Contrast this with DES and RSA, whose algorithms are public and which can thus be proved to be secure.

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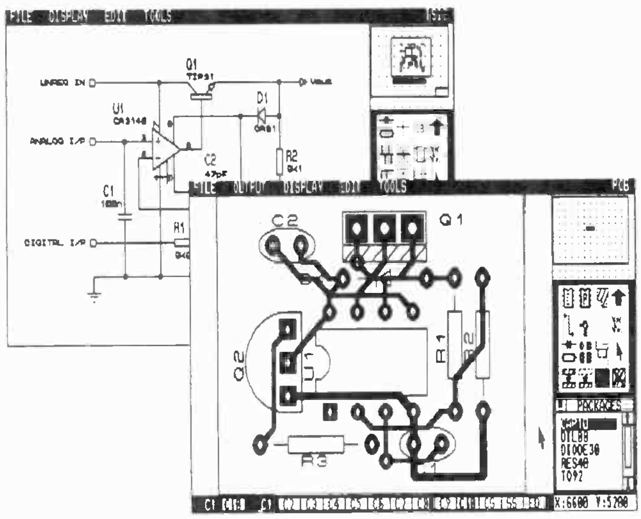

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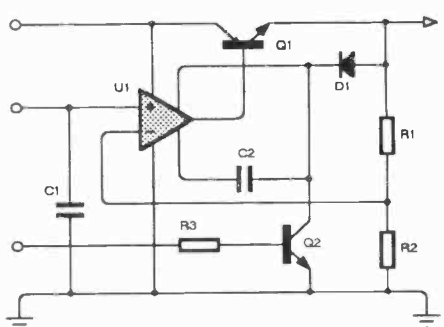

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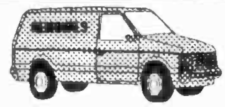

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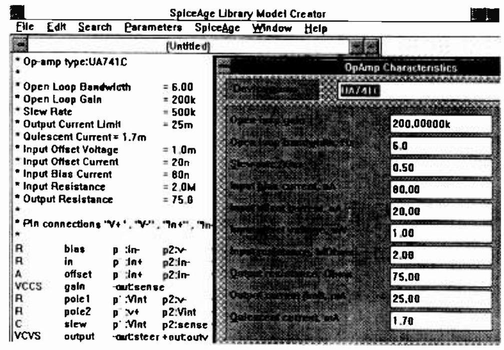

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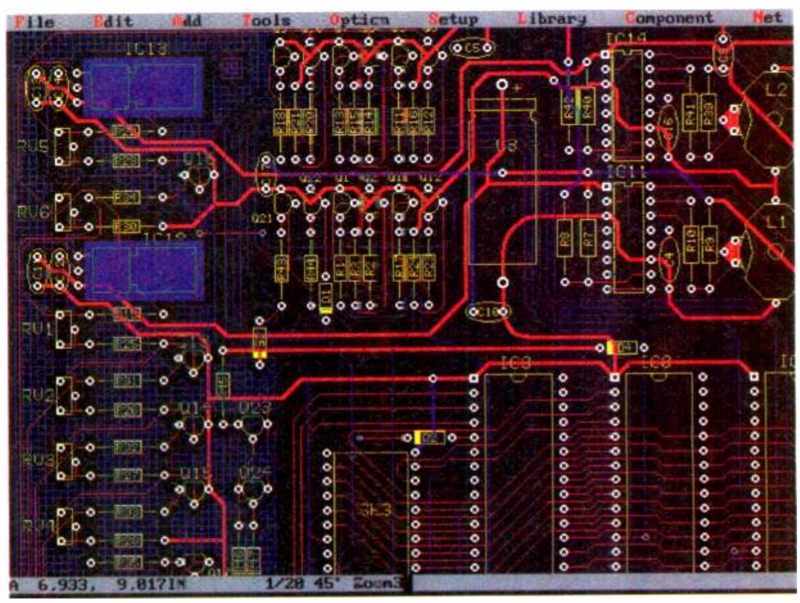
CIRCLE NO. 117 ON REPLY CARD

PC ENGINEERING

Circuit capture and pcb design: a winning combination?

The inclusion of schematic capture with PCB edit provides a powerful productivity design tool. ***John Anderson*** *reviews the latest addition to Tsien's Boardmaker software which adds full schematic capture.*



Boardmaker PCB edit screen.

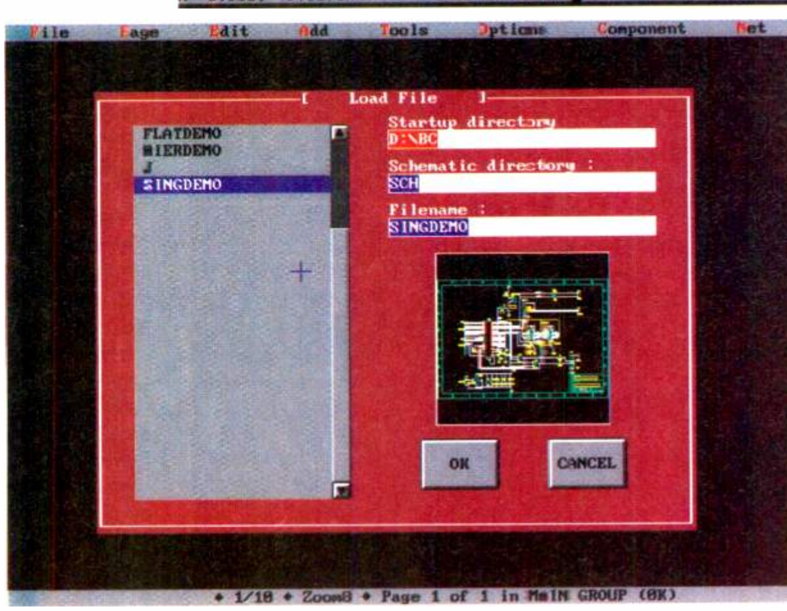
Board Capture is the new schematic capture program from UK company Tsien. It is designed to work with *Boardmaker* 2.51 the latest version of the pcb design product. Last reviewed for *Electronics World* in 1991, the conclusions were that Boardmaker was slow and compromised by embedding the functionality of library editing, schematic capture and pcb edit all in notionally the same operating environment.

The requirements of the three functions are rather different and the package did not take good account of this, particularly in respect of the schematic capture. Now it has its own schematic capture program, have things improved?

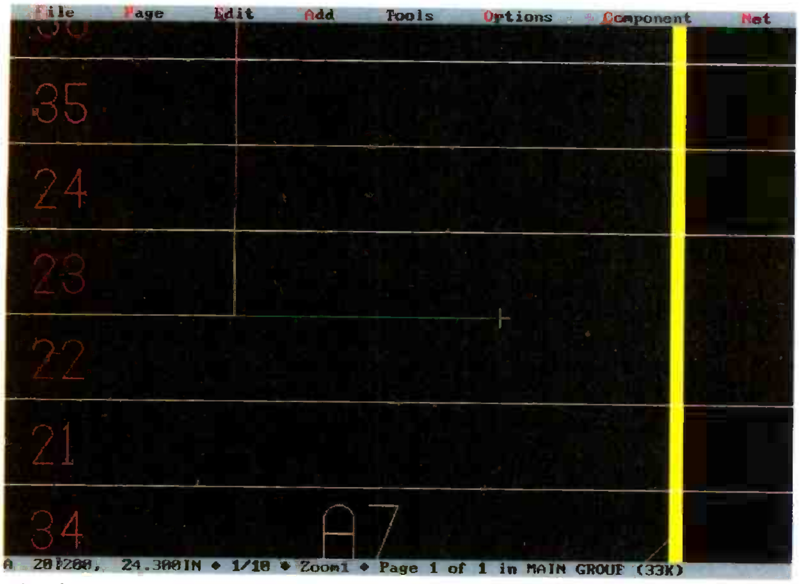
What you get

The software is provided on both 5.25 and 3.5in media together with two professionally printed manuals and a parallel port dongle. Product support for Boardmaker and Board Capture is free for three months. Additional support is available for a further year at a cost of £60 plus VAT per product.

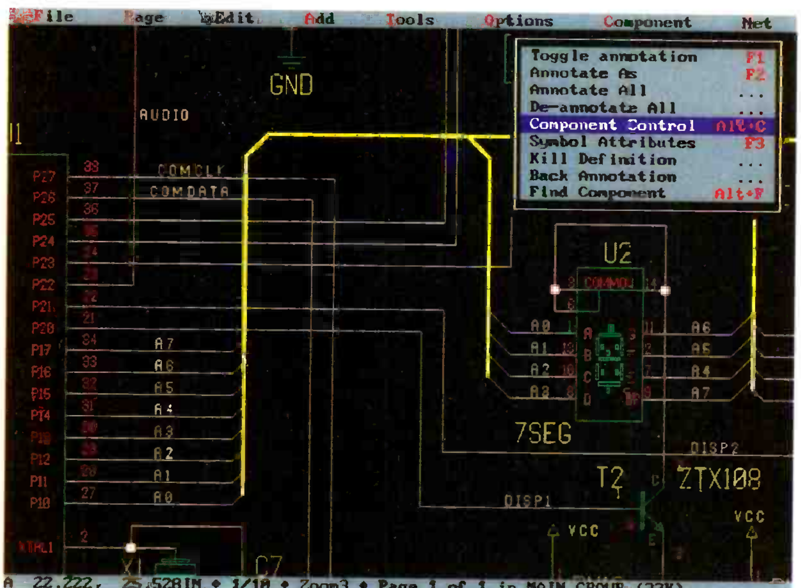
The installation process is straightforward with an install program moving the files onto hard disk and exhuming



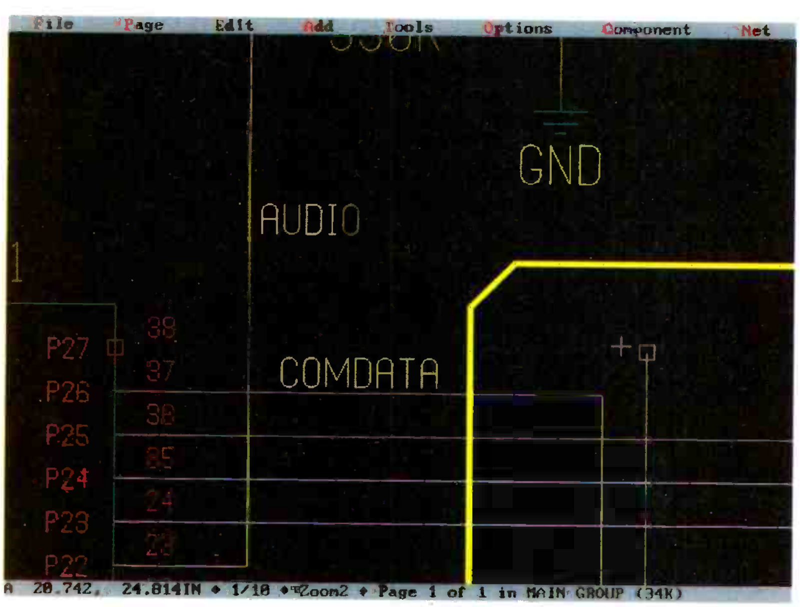
File load screen – note the miniature schematic shown for the selected file.



Line drawing – the left mouse button starts drawing from the nearest grid position.



Loaded file and user interface with pull down menus.



Schematic close-up. Note how unconnected nodes and wires are clearly marked with a small square.

(ugh!) the compressed files. The files, which include the libraries take about 1.5M bytes of disk space for each product. Board Capture may be configured for either standard VGA or super VGA 800x600. Many other aspects may be configured from the regularity of automatic backups to the way in which the mouse is used to pan the screen.

The Boardmaker manual is really the manual for the original Boardmaker 2 product with a few additional leaves describing the updates for versions 2.4 and 2.5. Even the discs are supplied as version 2.4 with a second disk which overwrites the data with that for version 2.51.

Bearing in mind that neither product has on line help, it was particularly annoying that the Board Capture manual hasn't got an index – so learning how to do something depends on remembering what you have read.

Starting up

On starting Boardmaker, you are presented with an initial menu offering schematic, PCB or library editors. Selection of any one of these leads to an almost identical working environment of a standard banner and pull down windows, and a data line across the bottom of the screen. Access to the functions at the banner is either by clicking with the mouse or by typing control with the key letter of that menu. The schematic capture referred to in this menu is the original "pcb" version and Board Capture has to be executed separately. This software has the same simple banner and menu format.

Board Capture

This is the new product, and although it initially looks like the old pcb version, this software offers much of what a schematic should. The zoom, pan and redraw are fast with panning occurring if the mouse is moved to the edge of the screen. The system supports a multi-page hierarchical design with a very effective simple page-up or page-down to jump between pages. There is a further level of abstraction with a tree of multiple nodes, each node may have multiple pages. It is not clear whether this system is memory or disk size limited. Thus the software could potentially handle very large designs.

The file load command menu provides a conventional file selector, but with the addition of a useful preview window which gives a (somewhat unclear) representation of the schematic.

Editing follows the normal scheme for any CAD system, with items being selected and added as required. Symbols are loaded by selecting this feature and then selecting either from the existing symbol list or a symbol from one of the libraries.

One particularly useful feature of the editor is that components are initially loaded with a small square on each unconnected node. As the connections are made, the square disappears making any misalignment of line and node obvious. Deletion is fast using the delete key, but as this is a single key press without a query delete, users must take care.

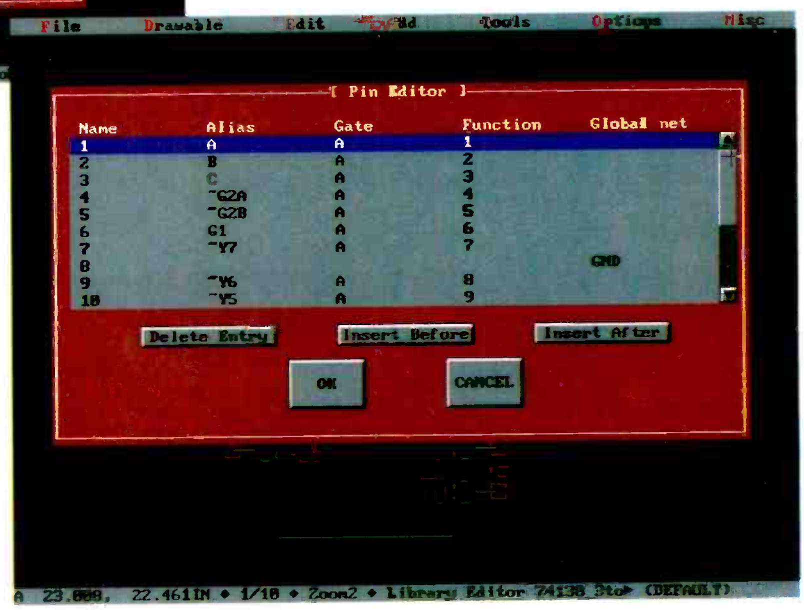
Annotation

Annotation is rather different to any other schematic capture system. When symbols are placed on the drawing they are not annotated (i.e. R_1 or U_9, etc.). Annotation can be added individually as the component is added. There are annotation commands which can annotate and de-annotate all the symbols on the drawing.

Boardmaker provides a component renumber function based on layout position, to aid fast component location. During this process, it creates a changes file for each component whose designator has been changed. This file may be read by Board Capture and used to back-annotate the schematic(s).

Loading a TTL symbol.

The symbol pin editor.



SYSTEM REQUIREMENTS

Requires: A PC, XT or AT compatible computer
640KB RAM, running DOS 3.0 or later.
A parallel printer port (for the dongle)
Graphics screen supported are: CGA, EGA, VGA and Hercules
Options: Epson compatible printer
Mouse
HP LaserJet II or compatible
HPGL plotter
Gerber photoplotter and Excellon NC drilling supported

PRICE

Prices (ex VAT)
Boardmaker 2.51 £295
Board Router 1.08 £200
Board Capture 1.10 £395
Extended support £60 for Boardmaker and £60 for Board Capture
Carr. £5.00

SUPPLIER DETAILS

Product: Board Capture 1.10 and BoardMaker 2.51
Supplier: Tsien (UK) Ltd.
Aylesby House, Wenny Road, Chatteris, Cambs. PE16 6UT
Phone 0354-695959

Library editor

A wide range of standard components are provided in a series of function divided libraries. New components may be added and new libraries created using the library editor. Symbols are created within the Board Capture program by accessing the Library Editor accessed from the file menu. The format of this editor is very similar to the Board Capture editor itself. In the same way as before, when loading a symbol, a miniature version of the component is shown in the preview window. During the symbol editing process, there is access to a separate pin editor. This allows editing of separate records describing the pins. The pin attributes include the pin number, an alias, the gate to which it belongs, its function, global use (e.g. power supplies) and others.

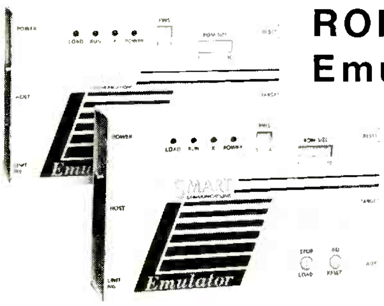
Conclusions

So have things improved for the Tsien products with the advent of the new board Capture program? I must give this a definite yes, because the system now offers a properly integrated solution from circuit to hardware. However the product still shows some problems. The Boardmaker pcb product still has vestiges of the old design methodology, which for newcomers at least will cause confusion. The annotation method is both powerful and confusing and rather different in approach to any other pcb cad systems. The speed of routing remains very slow, but in these days of 486 and beyond perhaps this is not so important.

Users these days expect on-line help and for newcomers the omission of this will slow the learning process. The Board Capture program is rich in features and has an excellent user interface. This will eventually result in the programme providing a high level of productivity.

At £395 Board Capture is not cheap, but it does represent one of the best dos based schematic capture programs. Its companion PCB editor, Boardmaker, remains in need of updating. It is however competitively priced so the combination of the two products does represent excellent value for money. ■

PROMulators



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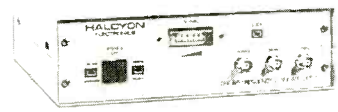

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CIRCLE NO. 121 ON REPLY CARD

Maths functionality for maths-phobics

Mathematical analysis can give valuable insight at the initial circuit design stage. Lionel Snell reports on how Mathematica and Nodal can give engineers the power of maths without the pain.

Mathematica from Wolfram Research and Macallan Consulting's *Nodal* could be just what is needed to put mathematical analysis at the fingertips of every engineer.

Wolfram describes *Mathematica* as "a system for doing mathematics". That makes it very much more than a system for doing calculations, because it can manipulate equations and formulas containing variables like x and y even before specific values have been assigned to those variables.

As with most computer applications, the package does require a special language to be learnt for input and output. But every effort has been made to ensure the language follows familiar mathematical notation. It has simply been designed to get round the problems of inputting mathematical symbols via a standard computer keyboard which does not allow for superscripts, subscripts and other maths symbols.

For example, $\int x^2 \sin x^2 dx$ would be input as:

```
Integrate[x^2 Sin[x]^2,x]
```

and would result in the output:

```
4x³-6xCos[2x]+3Sin[2x]-6x²Sin[2x]
---------------------------------
               24
```

The straightforward language soon becomes second nature and can be used to find roots, solve simultaneous equations, manipulate polynomials and matrices, invert functions and perform many other operations – as well as process numerical values. What is more, new rules or notations can be defined by the user and consistently applied – a flexibility that has allowed third parties to provide specialist packages, such as *Nodal*, to extend the power of *Mathematica* into practical applications.

Graphical output can be particularly useful for engineering applications. Compare, for example, the transfer function in **Fig. 1** with its 3-d graph which shows how the function (`tf`) blows up for certain values of s and ω. The graph gives an immediate visual image of performance and offers far greater insight.

Extensive plotting functions, in two or three dimensions, also include specific engineering forms such as the Smith chart for relating reflection to impedance.

Nodal electronic Interface

Mathematica is a mathematics tool: input equations and

Do engineers need maths?

A senior engineer I knew in the early 70s used to throw scorn on academic qualifications, insisting that no amount of mathematical training could replace years of experience. He had a reputation for being able to "read" a complex circuit diagram the way experienced musicians can read a score: he could predict its performance and see weak points in design that would be confirmed when it was built and tested.

Yet anyone with a couple of years of A-level maths and physics could have written down the relevant equations and would eventually have reached many of the same conclusions. They would have demonstrated at least half as much insight as all the engineer's years of experience, but would also have actual figures to quantity their conclusions.

His attitude was not untypical in the years when computers were first appearing in R&D departments, and apprenticeship schemes were giving way to more academic training.

But even now, people who opt for engineering as a career are attracted by practical work with tangible end products. Such engineers often have far less interest in the abstract processes of mathematics.

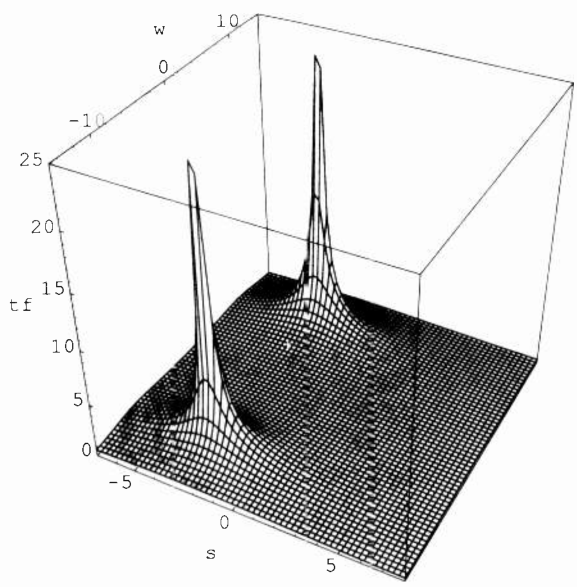

Fig 1. 3-D plot of the s-domain transfer function tf = (1 + (s/w0)2 + s/qw0)-1 for values of s from –5 to +5 and w from 0 to 15. Just one of many different ways that Mathematica can create immediate visual images of system performance – in this case showing the poles where the function 'blows up'.

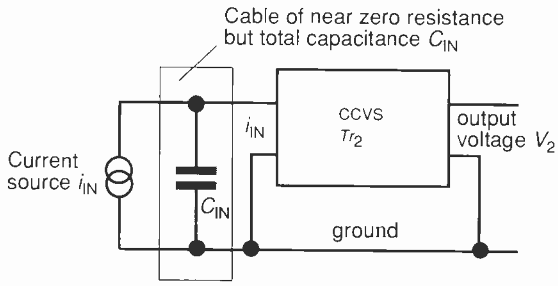

Fig 2. Simple circuit containing a current controlled voltage source. This can be used to model current differencing (Norton) and charge-sensitive amplifiers using Mathematica.

formulas and it returns solutions, graphs or numerical answers. But the user must give it the necessary equations in the first place.

Nodal extends the software into the world of electronic circuits, so that actual circuit descriptions can be entered, causing immediate generation of formulas for manipulation via *Mathematica*.

The two packages are seamlessly integrated so that *Nodal* appears as an extra set of commands within the main program. *Nodal*'s commands amount to a special language formulated to allow two- or three-dimensional circuits to be described through the keyboard. Again, it is logical, based on clearly understandable terminology, and describes any circuit in terms of a number of "nodes" separated by various electrical components.

In **Fig. 2**, for example, there are just three nodes in this sense. All the junctions along the bottom are directly connected and so count as a single node. Between nodes 0 and 1 there are two parallel components, a current source (iIN) and a capacitor (cIN). There is also a "current controlled voltage source" used here to model a charge sensitive amplifier – it has four connections to nodes 1, 2 and twice to 0.

Giving the circuit the name chAmp, it is defined as:

```
chAmp = NodalNetwork[Current
Source[{1,0},iIN],
        Capacitor[{1,0},cIN],
        CCVS[{1,0,0,2}],trz]]
```

Let us say we want to know the output voltage, that is voltage at node 2 (V2 in *Nodal*-speak). The basic command is:

```
NodalAnalyze[chAmp,Result->V2]
```

Remembering that Americans write analyse with a z is the only demanding part. The actual response to this command would be a somewhat untidy first draft formula. But it can be cleaned up with *Mathematica*'s simplify command. In practice the two stages are easily combined as:

```
Simplify[NodalAnalyze[chAmp,Result->V2]]
```

resulting in:

```
iIN trz
```

revealing that output is independent of capacitance.

Practical application

If you need to connect a current source – an avalanche diode detector maybe – to an amplifier, how critical is the connecting cable capacitance?

The last example modelled the situation for a cable of negligible resistance but total capacitance cIN. It showed that output voltage was unaffected by the capacitance.

Replacing the current controlled amplifier by a voltage controlled one defines a new circuit we will call vAmp. The input simply replaces the CCVS with a VVCS as follows

```
vAmp=NodalNetwork[CurrentSource[{1,0},iIN],
        Capacitor[{1,0},cIN],
        VCVS[{1,0,0,2}],gain]]
Simplify[NodalAnalyze[vAmp,Result->V2]]
```

giving

```
-I
--- gain
 2
--------
Pi cIN f
```

Mathematical insight

Circuit modelling software such as *Spice* plays an important role in electronics development, because it allows circuit prototypes to be simulated on a computer. So the performance of various designs can be evaluated quickly and accurately before a single wire has been connected.

The drawback is that this easy trial and error development is at the expense of "insightful" development which takes a more direct route to an optimal solution.

Cad software based on *Mathematica* allows symbolic manipulation of models, leading directly to better design by reinforcing the engineer's intuitive understanding of circuit behaviour.

Mathematica can help engineers at several levels.

Firstly, it performs mathematical manipulation.

What is the point in writing down equations based on Kirchhoff's Laws if you don't trust yourself to solve them or draw a graph of performance without making mistakes? This basic lack of confidence in one's own maths undermines any amount of engineering education. Given simple commands *Mathematica* solves equations, plots graphs, simplifies complex formulas and provides a host of features such as Fourier and Laplace domain analysis. Even the most accomplished mathematician would be glad to hand over such spade-work to a computer not prone to human error. For the engineer it frees the mind to concentrate on actual design principles.

Secondly, *Mathematica* can analyse a real life situation mathematically.

Writing down the initial equations may be just a process of following simple rules, but it becomes a challenge in more complex or unfamiliar cases. Although the software was originally designed purely as a mathematical tool, it has earned such a following that packages are now being created to apply this engine to a host of practical applications. In the case of electronic engineering, *Nodal* allows the engineer to type in a simple circuit description and the relevant equations are instantly generated in *Mathematica* form.

Thirdly, working with *Mathematica* is an education in itself. The creativity required for pure maths research is a rare ability not directly relevant to engineering. All the same, practical experience with *Mathematica* does encourage latent mathematical skills and understanding that can lead in the longer term towards more confident research and development.

where I is the square root of −1, Pi is π and f is frequency.

Here cable capacitance does affect the result. Whether this is serious or not can be readily understood by graphing for appropriate values. Options include a simple plot of output voltage against capacitance at a given frequency; or a 3-d plot of output voltage against frequency and cable capacitance; or a contour plot. For example, the commands:

```
Plot3D[output /. {gain->100, iIN-> pA, f->10^x, cIN->y pF}
       {x,0,9},{y,1,1000},
       AxesLabel->{"log(f)","c(pF)","log(output)"},
       PlotRange->All]
```

define a 3-d graph between chosen values (**Fig. 3**) giving an immediate visual grasp of the situation.

The example is very simple, and a more complex problem would take longer to input. But the saving in manual calculation and mathematics skills would be all the greater – without the risk of human error in calculation.

Graph the result, and the engineer immediately gets the full picture for informed design decisions.

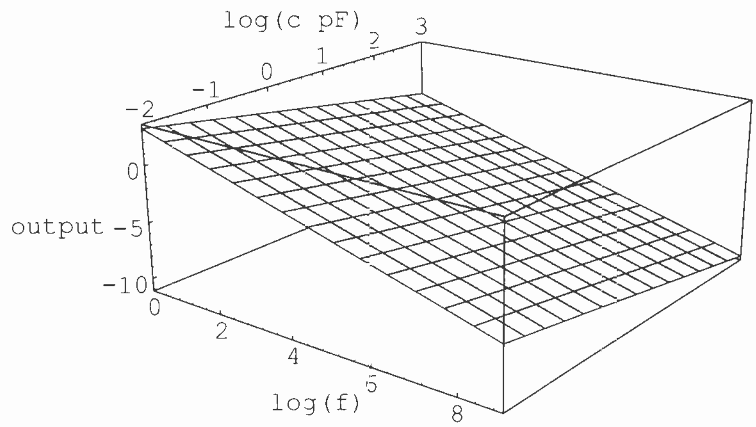

Fig 3. 3-d plot showing effect of replacing the charge sensitive amplifier in Fig. 2 with a voltage amplifier. The output is now seen to be dependent upon the cabie capacitance and the frequency.

Scope of *Mathematica* and *Nodal*

Mathematica itself is a well established mathematical toolkit, and at the time of writing the *Nodal* electronic application package costs around £310 + VAT. Version 2 contains about thirty component descriptions for analogue circuits handling anything from audio to microwave frequencies.

Updates are promised which will include allowance for thermal effects as the temperature of the circuit alters. There is also talk of "mechanical" additions to take into account the mechanical properties of, say, a gramophone pick-up (it is already possible to define certain components in geometric terms, eg capacitance by gap width and plate overlap).

What cannot be handled is non-linear effects or digital circuits. Of course, these are included in the full version of *Spice*, but there is little competition between these two.

Spice is designed to model the most complex circuits at the prototype stage when all the numerical values are known. It can provide accurate performance figures and even quantify the circuit sensitivity around those values.

What *Spice* does not do is provide the symbolic description of overall performance provided by *Mathematica* at a fraction of the cost of the full *Spice* package.

One limitation, is that the output equations are rather clumsy. The input language for *Mathematica* is restricted by the need to enter mathematical symbols on a standard keyboard. For output their is a little more choice of formats. But there is nothing like the clear mathematical output available from a dtp system. As a working tool for design engineering, this is no problem, though it could become tiresome when being used in the preparation of design reports. Proper mathematical typesetting would be welcome.

Unique flexibility

Symbolic analysis of circuits balances the "try it and see" approach by developing insight at the early design stage. Until now this has required mathematical confidence, if not actual skill. But *Mathematica* undoubtedly provides an easy and attractive alternative.

No other system to my knowledge offers anything like the same flexibility for electronic design support – even the *Mathcad* electronics support package is little more than a collection of useful formulas or example circuits.

If there is any other software offering this sort of support to the electronic designer, I would like to find it. ■

Supplier details

Mathematica
Mac: Student version £149; standard version (no coprocessor) £375; Enhanced version (needs 68020 or higher and uses coprocessor) £575.
Windows: Student £149; Standard (386 or higher, no coprocessor) £375; Enhanced (coprocessor) £645.
dos: Student £149; Standard £375; Enhanced £575.

Available from Wolfram Research Europe Ltd, Long Hanborough, Oxon OX8 2LA. Tel: 0993 883400 Tel: 0993 883800

Nodal
£310 + £6 p&p + £55.30 VAT available from Goth, Goth & Chandleri Ltd, 36 Victoria Park, Cambridge CB4 3EL. Tel 0223 321748 Fax: 0223 323804. More information from Macallan Consulting, Tel: 010 1 408 262 3575.

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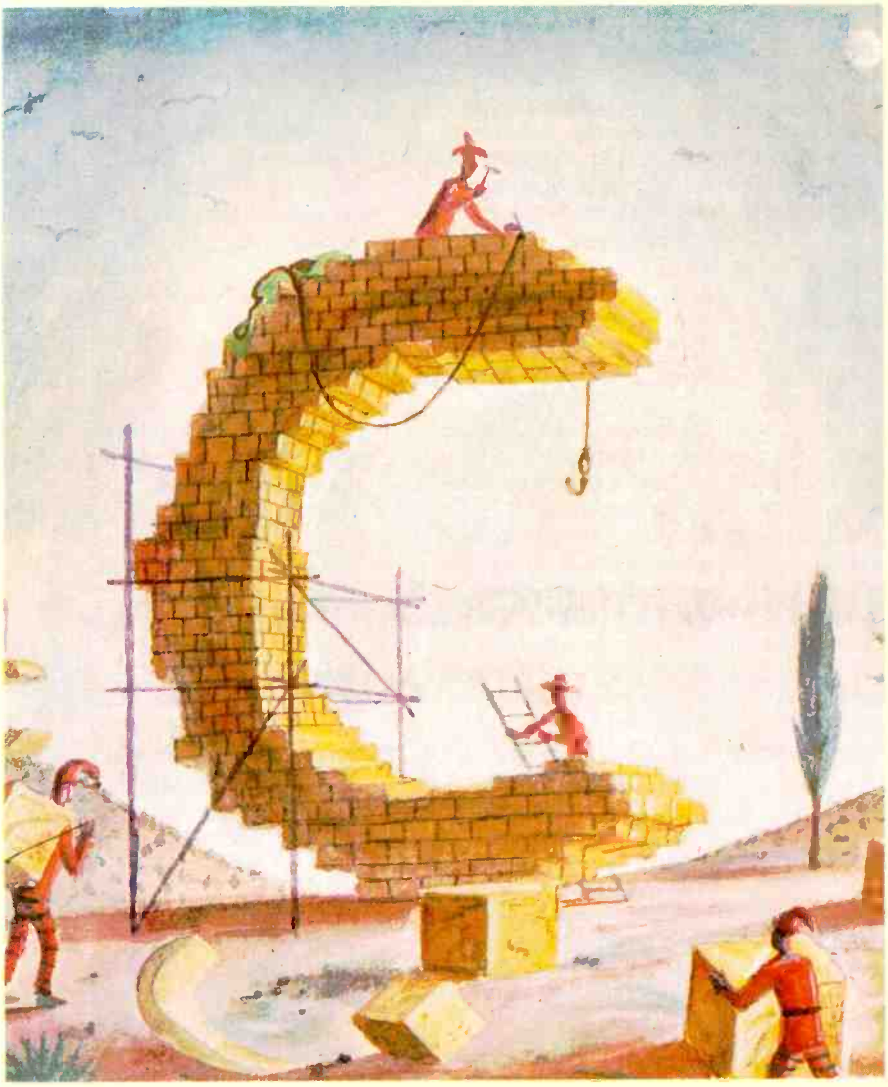

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by

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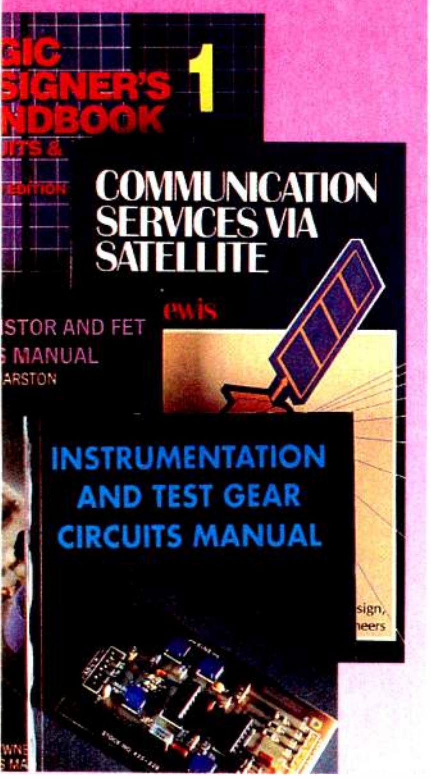

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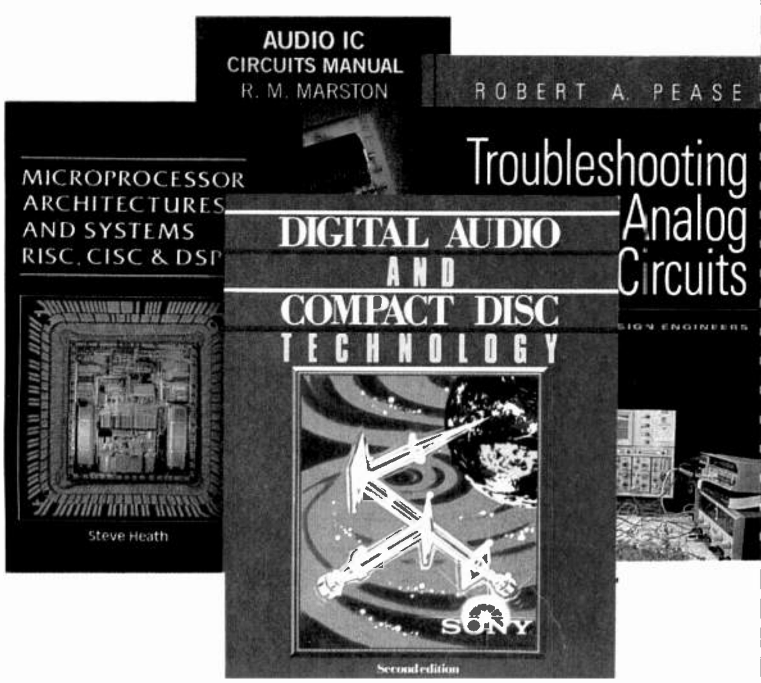

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Current conveyor ICs – a new building block

There are unique benefits to current conveyor ICs in applications from lf to rf. Take a precision rectifier needing only two resistors and two diodes for example. Ian Hickman describes his experience with this new type of component.

At first glance, the simplified circuit in **Fig. 1a** looks like an op-amp. An input to bias the tail current has to be provided externally, but the circuit has the usual npn input long tail pair, feeding current mirrors connected to the positive supply rail.

With complementary emitter followers to provide a low output impedance unity gain buffer at the output, it would simply be an op-amp with facility for trading off supply current against speed. In fact, the buffer is missing. Output is taken directly from two complementary transistors arranged as current sources fighting each other. This circuit is a common operational transconductance amplifier, or OTA.

Figure 1b shows a close relative of the transconductance amplifier. With a unity gain buffer at its output, this circuit becomes a current feedback op-amp. Since the buffer is missing, the device has a high impedance current source output as does an OTA. But instead of inverting and non-inverting inputs both of high impedance, this device has a high impedance non-inverting input and a low impedance inverting input.

Known as a current conveyor, the device is similar to a current feedback op-amp. Unlike the traditional voltage feedback op-amp, the transconductance amplifier is usually used without feedback. Similarly, unlike a current feedback op-amp, so is the current conveyor.

Absence of a feedback path avoids the stability problems that can plague voltage or current-feedback op-amp designs. This is a welcome feature of the current conveyor, providing complete stability when driving reactive loads of either sign. But the device's current output requires a different approach to circuit applications, which cover the frequency range from dc up to 100MHz.

Figure 2a shows a simplified a current conveyor, the relation between the terminal currents and voltages, and pin-out of the *CCII01* integrated circuit. As with a current feedback op-amp, the current conveyor's non-inverting *Y* input is high impedance – 80kΩ at 1kHz – while its inverting *X* input is low impedance, Fig. **2b**.

Packaged in eight-pin DIL, the *CCII01* contains two current conveyors. Conveniently, these can be combined to produce an enhanced composite conveyor. This configuration provides an input impedance of less than 200mΩ at the *X* input up to 1MHz or so, Fig. **2c**.

With its low input impedance, the *X* node of either a standard or enhanced current conveyor can be used as a current summing junction for two or more signals. Its high *Z*-port output impedance – typically 1MΩ at a frequency of 1kHz – means that signals from several current conveyors can be combined simply by hard wiring their outputs together.

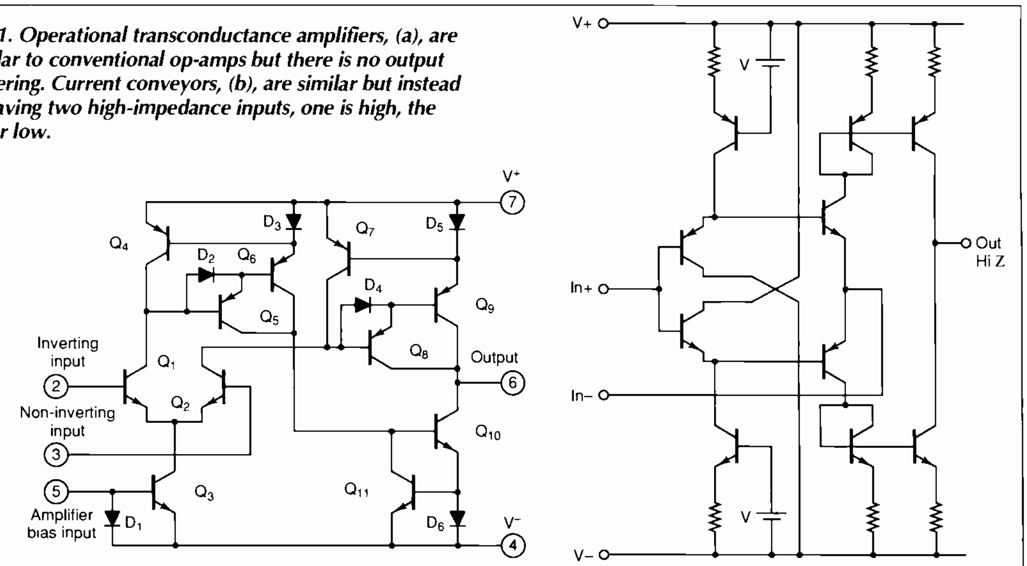
Numerous other applications, including differentiators and integrators, voltage and current-controlled negative-impedance converters, precision half and full-wave rectifiers, double terminated amplifiers are presented in the device data sheet.

Filter

One application is as a biquad filter. Results obtained with the *CCII01* configured as a 5MHz bandpass filter[1] are shown in the data sheet[2]. Equations defining performance of the filter are presented in the panel.

I experimented with the bandpass circuit at audio

Fig. 1. Operational transconductance amplifiers, (a), are similar to conventional op-amps but there is no output buffering. Current conveyors, (b), are similar but instead of having two high-impedance inputs, one is high, the other low.



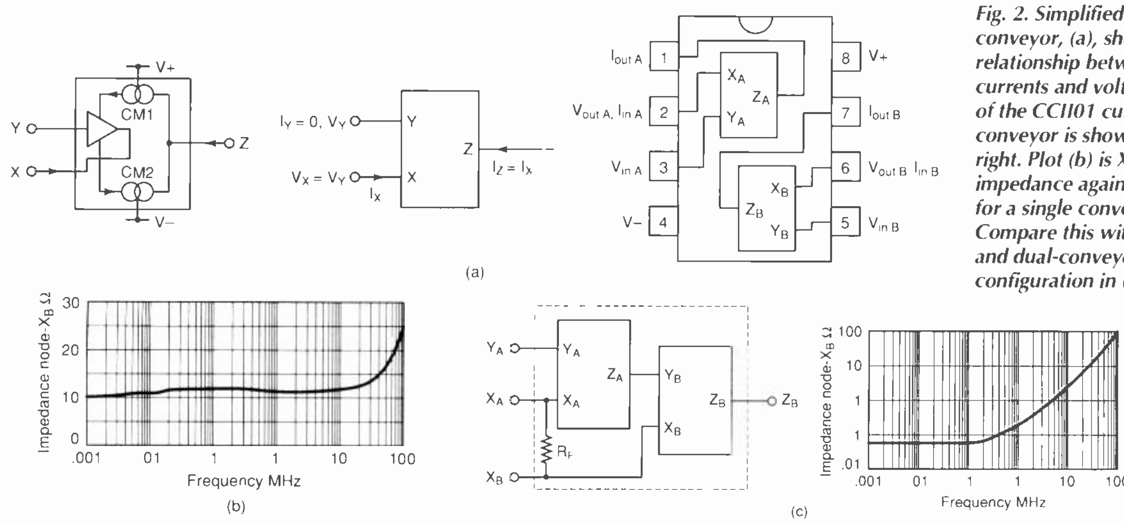


Fig. 2. Simplified current conveyor, (a), shows the relationship between terminal currents and voltages. Pin-out of the CCII01 current conveyor is shown on the right. Plot (b) is X-input impedance against frequency for a single conveyor. Compare this with the plot and dual-conveyor configuration in (c).

frequency, deliberately using extreme values of components to verify the equations and performance. **Figure 3a** shows this version. Component Y_2 is 0.1S, i.e. R_2 is non existent but there is 10Ω of input impedance from X input of the first current conveyor.

From the equations in the panel, the expected frequency peak was at 2250Hz while the measured peak occurred at 2267Hz. As the circuit had 20% tolerance capacitors, the close agreement was more luck than good engineering. No such close agreement was found for Q. The theoretical value was 7.03 whereas the –3dB points at 2052Hz and 2495Hz indicated Q=5.1.

I used the filter to pick out the seventh harmonic of a 324Hz squarewave, Fig. **3b**. Derived from the panel equations, the value of the filter's centre-frequency gain is 9950 or nearly 80dB. This is reflected in the large difference in Y deflection factors for the two traces. Given that amplitude of the seventh harmonic of the 4mV pk-pk squarewave is about 0.73mV pk-pk, the 7V pk-pk average value of the filter output is as expected.

In the time-domain, the damped wavetrain from the filter decays logarithmically after excitation by each

Current conveyor filter performance

A lowpass/bandpass filter using current conveyors is shown here. Interchanging the resistors and capacitors produces a highpass version. In terms of the admittances of the passive components, the transfer function of the circuit is,

$$\frac{V_{BPF}}{V_{IN}}=\frac{-Y_2Y_3}{Y_5(Y_2+Y_3+Y_4)+Y_3Y_4}$$

where Y_2 is the conductance $1/R_2$, Y_3 is the susceptance sC_3, etc, and s is the complex frequency variable. For analysing the steady state, s may be replaced by $j\omega$. Rewriting the equations in terms of C and R gives,

$$\frac{V_{BPF}}{V_{IN}}=\frac{-sC_3R_5}{s^2R_2R_5C_3C_4+sR_2(C_3+C_4)+1}$$

Comparing this with the archetypal form for a bandpass filter,

$$\frac{V_{BPF}}{V_{IN}}=\frac{As}{s^2+Ds+1}$$

peak response occurs when phase shift is zero. In this case, phase shift can also be 180° due to the minus sign in the numerator. The circuit is an inverting filter. This is so when the outer terms in the denominator add to zero, leaving just a $j\omega$ term to cancel top and bottom, i.e. a real negative number.

Since $s^2=(j\omega)^2=-\omega^2$, this occurs when;

$$\omega^2R_2R_5C_3C_4=1$$

Feeding in component values from Fig. 3a produces $\omega=2\pi f_r$=14142 in which resonant frequency f_r is 2250Hz. At this frequency gain is,

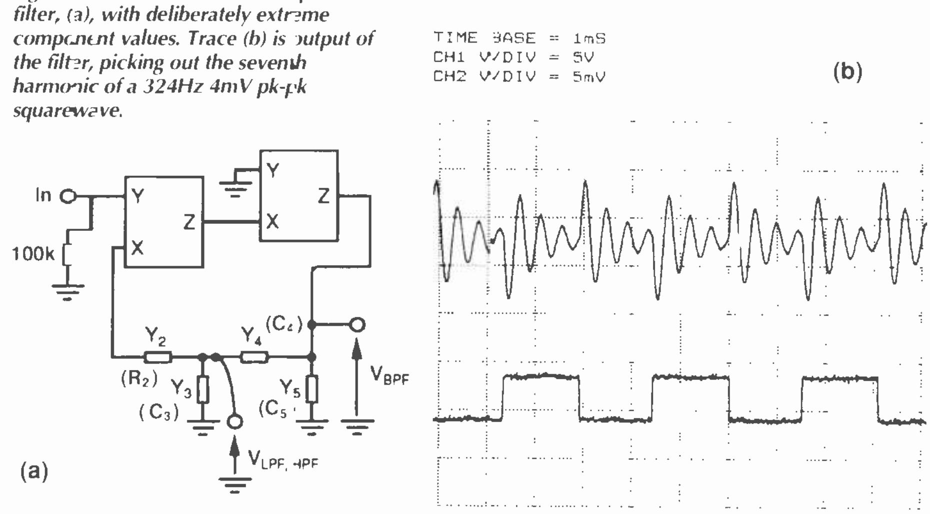
$$\frac{-C_3R_5}{R_2(C_3+C_4)}=-9950$$

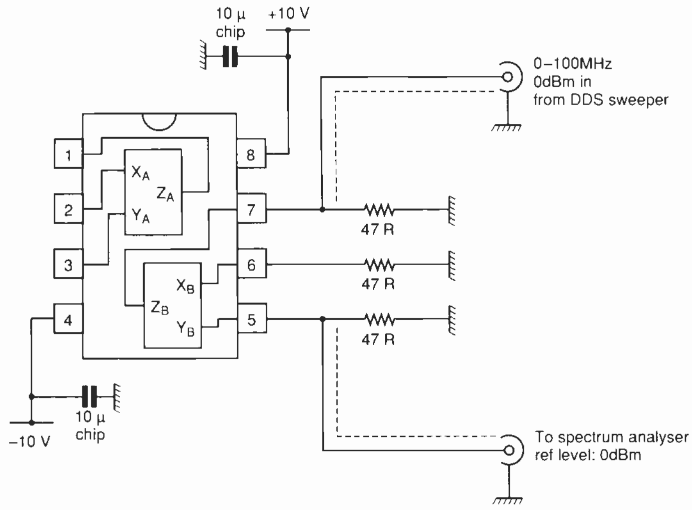
and Q is given by

$$\frac{1}{Q}=D=sR_2(C_3+C_4)$$

In this instance, $1/Q$ is 0.142 and Q is 7.03

Fig. 3. Circuit of a 2267Hz bandpass filter, (a), with deliberately extreme component values. Trace (b) is output of the filter, picking out the seventh harmonic of a 324Hz 4mV pk-pk squarewave.





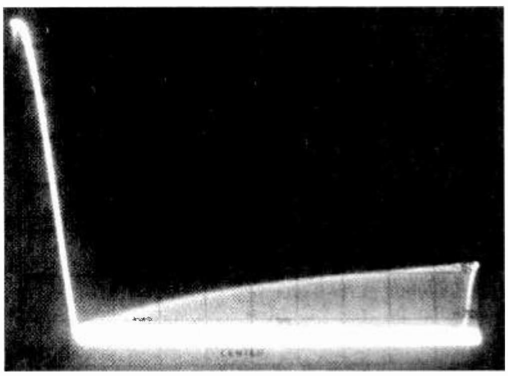


Fig. 4. Measurement set-up used to test reverse isolation of the CCII01 as a function of frequency (a). Sweeper output level was 0dBm. Scan (b) is results of the measurement over 0-100 MHz. Analyser reference level, top of screen, is 0dBm. Scaling is 10dB/div vertical, 10MHz/div horizontal and centre of screen is 50MHz. IF bandwidth is 3MHz.

edge of the squarewave. From this, applying the useful relationship between Q and energy stored over energy lost per radian you can calculate Q.

Careful measurements of successive peaks of the filter output waveform in Fig. 3b show that amplitude falls to 50% over each successive cycle. Energy stored falls to 25% per cycle, or to 80% per radian. Using an approximation only valid for high values, Q is about 100%/20%. This agrees with the measured value, mentioned earlier. I am not clear why centre frequency and gain should agree with the theoretical values but Q not.

Figure 3a is curious. At 0Hz, reactance of each capacitor is infinite, so they effectively no longer appear in the circuit. Gain at the lowpass output is then unity. This is because it is taken via R_2 from the X port of the first current conveyor, the output of the device's unity gain input buffer.

At dc, the circuit is completely open loop, so some output offset might be expected at the bandpass output, given the high value of R_5. In fact, the output offset was zero, even with the 50Ω source disconnected leaving the Y input of the first current conveyor grounded via 100kΩ.

This says something for the accuracy of the device's fabrication, which is carried out by a manufacturer of advanced linear ICs, Elantec. With its 2000V/µs slew rate and 700MHz equivalent gain-bandwidth product, the device is specified for use up to 100MHz.

For any active device intended to be used at high frequencies, an important parameter is reverse isolation. For stable operation – especially in tuned circuits – this should comfortably exceed the forward gain expected from any amplifier using the device.

I measured reverse isolation as a function of frequency using the test set-up of **Fig. 4a**. Output of the DDS-based sweeper was 0dBm at low frequencies, falling linearly by no more than 1dB up to 100MHz.

RF amplifer

Figure **4b** shows reverse isolation over that frequency range. Reference level of the spectrum analyser was 0dBm and IF bandwidth set wide to 3MHz. This was done to produce a bright enough trace to register at each step of the sweep. Reverse isolation exceeds 70dB below 15MHz, greater than 60dB below 70MHz and still 56dB at 100MHz.

With this sort of performance, the *CCII01* should enable stable, high gain rf amplifier stages to be readily realised. I tested this hypothesis via **Fig. 5a**. Here, a series tuned circuit at a current conveyor's X port and a parallel tuned circuit, also 10.7MHz, at its Z output is used to provide a two-pole response.

Appropriate values of L/C ratio were selected for the two tuned circuits. The second current conveyor in the package formed a high input impedance buffer to avoid loading the parallel tuned circuit.

Attenuated –50dBm output of the sweeper was applied to the circuit, output from which is shown in Fig. **5b**. This scan indicates a gain of 30dB and a rather useful but unintentional bandpass response.

With a part having a response extending to the best part of a gigahertz, I used compact layout and a groundplane. As a result there was some coupling between the unshielded coils. Symmetry of the response indicates a lack of internal feedback. This leads to the question - can the circuit be pushed for even more gain and selectivity, without incurring instability?

The 47Ω resistor at the X input of the second current conveyor in the package was replaced by another 10.7MHz series tuned circuit. All coils were screened. I noticed some occasional oscillation at around 400MHz which was cured by inserting a 10Ω resistor between the X port pin 6 and the second series tuned circuit as shown in Fig. **5c**.

Response, shown in Fig. **5d**, indicates that the screened coils prevented unintentional coupling. Output of the circuit is now –7dBm and gain around 43dB. As shown in Fig. 5a however, the circuit does not provide 50Ω input termination.

Slight asymmetry in the response, with a faster fall-off on the high frequency side, indicates some internal feedback. This is not surprising, since although the reverse isolation of the current conveyor at pins 5, 6 and 7 exceeded 70dB at 10MHz the two current conveyors are physically very close.

The remarkable aspect of Fig. 5c is the almost total absence of discrete components, apart from the tuned circuits. Presumably additional tuned circuits, providing 50Ω interfaces to the chip, at input and output, would produce a five-pole response.

In practice, you would probably never need to design a 10.7MHz amplifier using individual tuned circuits as in Fig. 5c. A wide range of block filters covering almost any conceivable requirement is available from numerous suppliers. But if an IF amplifier operating at a non-standard frequency is needed, then current conveyor ICs provide a convenient way of producing the gain and selectivity necessary with a minimum of components.

Figure 6 demonstrates two further current-conveyor applications. Configuration **6a** is synonymous with

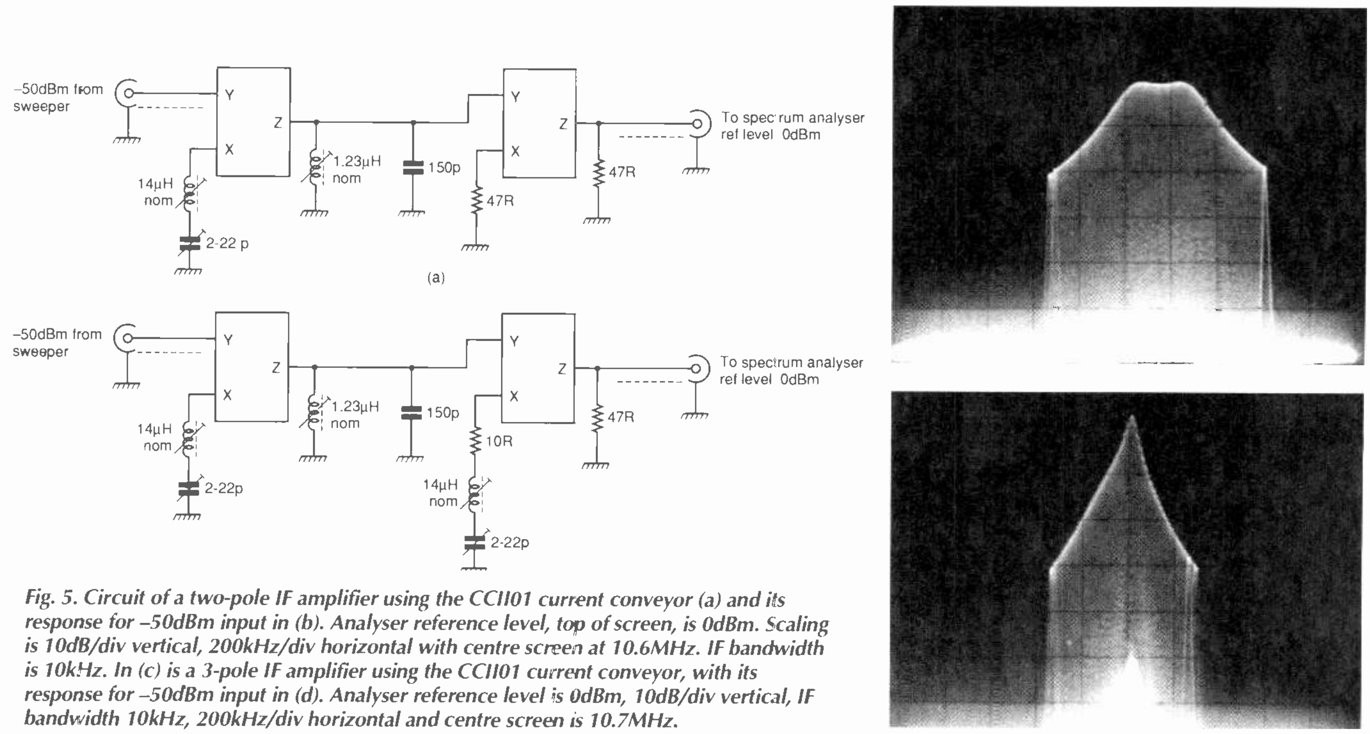


Fig. 5. Circuit of a two-pole IF amplifier using the CCII01 current conveyor (a) and its response for –50dBm input in (b). Analyser reference level, top of screen, is 0dBm. Scaling is 10dB/div vertical, 200kHz/div horizontal with centre screen at 10.6MHz. IF bandwidth is 10kHz. In (c) is a 3-pole IF amplifier using the CCII01 current conveyor, with its response for –50dBm input in (d). Analyser reference level is 0dBm, 10dB/div vertical, IF bandwidth 10kHz, 200kHz/div horizontal and centre screen is 10.7MHz.

the voltage-controlled or short-circuit stable negative-impedance converter. Being a current-controlled impedance converter it is open-circuit stable.

To see how the converter works, imagine Z is a 1kΩ resistor with its lower end is grounded. Voltages at ports X, Y and Z are all initially zero and relationships between the port voltages and currents are defined in Fig. 2a. Raise X to +1V. Since $V_x=V_y$, port Y must also be at +1V so there must be +1V at the top end of the 1kΩ resistor. A 1mA current flows out of port Z but also out of X since $I_Z=I_X$. As a result, port X exhibits a resistance of –1kΩ.

The lower end of Z was assumed grounded solely to simplify the explanation: in view of the device's high common mode rejection – greater than 53dB up to 1MHz – neither end of Z need be grounded. The circuit offers a floating negative impedance.

High output resistance of the Z port, typically 1MΩ at 1kHz, makes it behave like an almost perfect current source. Current passing through the load is not related to the voltage-drop across it. This situation is ideal for biasing diodes in a precision rectifier circuit.

Figure **6b** shows a half-wave rectifier. If R_1 equals R_2, then output is identical to the input for positive voltages and is zero for negative voltages, regardless of waveform or mark/space ratio. Results obtained for a sinewave input are shown in Fig. **6c**. Full-wave precision rectification is also covered in the data sheet.

If you experiment with the *CCII01*, you should be wary of the simulated grounded inductance shown in the data sheet and covered in the first reference. As expected, this set up draws a lagging current from a zero resistance source. Since both current conveyors in the loop are non-inverting, however, the circuit looks like a *negative* resistor at dc – unlike a real inductor. In other words, the circuit is not open-circuit stable and will lock up at one or other supply rail. ■

References

1 *Current conveyor circuits, Electronics World and Wireless World*, Nov., 1993 pp 962-963.

2 *CCII01 data sheet*, LTP Electronics Ltd, 2 Quarry Road, Headington, Oxford OX3 8NU. Tel. 0865 744232.

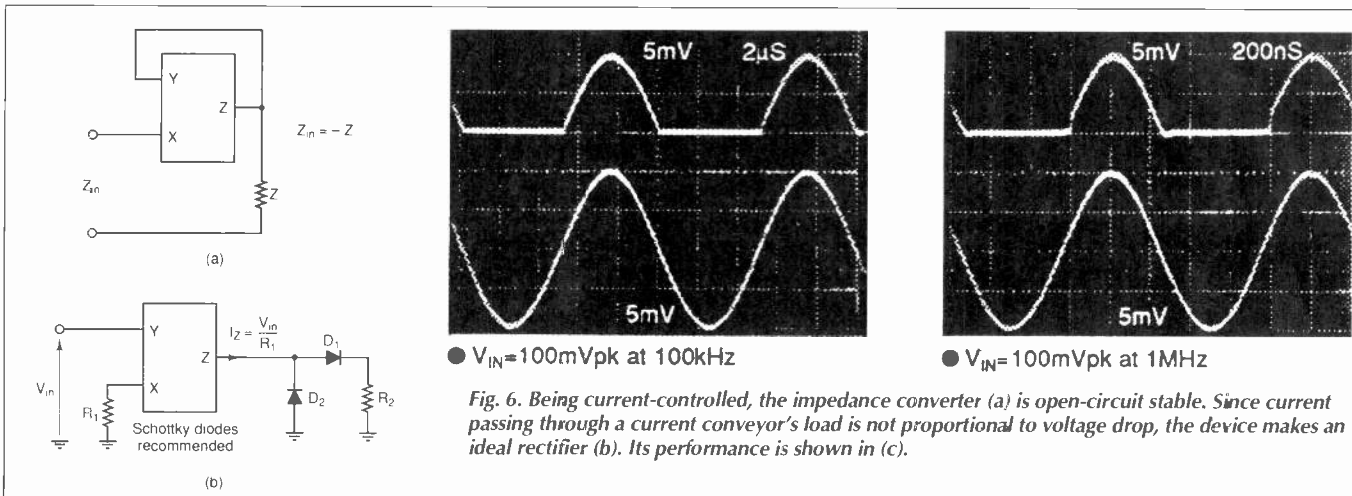
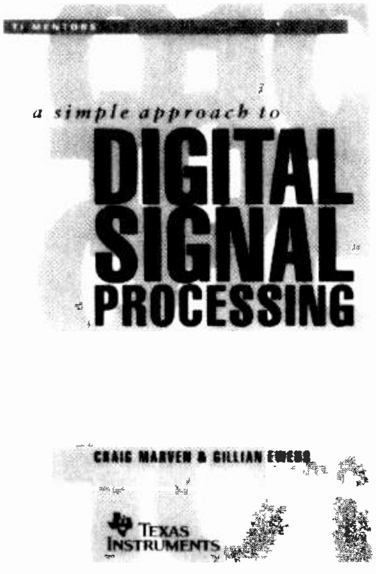


Fig. 6. Being current-controlled, the impedance converter (a) is open-circuit stable. Since current passing through a current conveyor's load is not proportional to voltage drop, the device makes an ideal rectifier (b). Its performance is shown in (c).

BOOK REVIEWS

A simple approach to digital signal processing



Over the past ten years, digital signal processing has attracted a lot of attention. This is reflected by the wealth of text books on the topic.

With subjects such as digital filters and FFTs so well worn it would be difficult to find an area which has not been adequately covered in depth – several times over. However Marven and Ewers have identified a deficiency in the standard book range. There is little to cater for the first time reader who is just starting out on the demanding subject of DSP.

A Simple Approach to DSP begins from ground zero and covers the majority of DSP subjects in sufficient depth to enable the new reader to gain a flavour of what the subject is about without getting bogged down in the maths and the algorithms. In this respect the book succeeds. It is very readable and the page layout is pleasant – a lesson that many other British publishers have yet to learn.

Maths is kept to a minimum and the diagrams are informative. The style is approachable and should maintain the interest of the new reader. Many introductory texts on DSP fail so often to sustain the reader's interest. Mainly, this is because they delve into the complexity of the subject too early with little regard for the struggling reader whose maths may be a little shaky.

This is clearly not the case with this book. On establishing the reader's confidence in the first few chapters, the book succeeds on building on that confidence to introduce the more demanding aspects of DSP – digital filtering and spectral analysis. Each chapter is supported by a reasonable list of references for further reading. However when discussing processor hardware the authors have not presented a balanced overview of commercial products. Since the authors are with Texas Instruments, one would naturally assume a discussion on TI's processors, but not to the exclusion of their competitors. Although there is a brief mention at the start of Chapter 7, a discussion of competitors' products would have provided a useful prospective of hardware performance.

On the whole the book should provide a basic understanding of DSP and should serve as a useful introduction to the subject. If its cost could be reduced to less than £10 it would probably find a wider readership among students as an auxiliary text.

Allen Brown

Craig Marven & Gillian Ewers, Texas Instruments 1993. ISBN 0-904 047-00-8, price £19.

The multiplexer reference manual

Written by Gilbert Held, *The multiplexer reference manual*, is concerned with the multiplexer in communications, rather than the more usual connotation in our field of input multiplexers for signal processing.

The first chapter sets the scene, describing the reasons for the existence of the devices. It proceeds to recount the early uses of multiplexing in telephony and to specify the evolutionary development of the subject from frequency division to optical fibre time division. Each of the methods of multiplexing is given a chapter to itself: frequency and time division, statistical multiplexers, packet assemblers/disassemblers, T-carrier multiplexers and optical fibres. In all cases the alternatives are compared.

In the last chapter, Held looks at currently evolving methods, including fast packet multiplexing and low bit-rate voice digitisation. Networks are, of course, an important aspect of multiplexing and are accorded their rightful share of attention. Perhaps the most important sections are those on statistical and T1 multiplexers, which are given extensive treatment. Engineers concerned in any way – however peripherally – with communications and networking will find the book a valuable reference, although the style is rather unattractive. A more extensive use of the active voice would have helped to make it a little more appealing, but this is a personal preference and in no way detracts from the value of the work.

Philip Darrington

John Wiley, 188 pages, hardback, £24.95.

The Early History of Radio, from Faraday to Marconi

This work, by GRM Garratt is the outcome of a career spent mostly at the Science Museum, much of that time in charge of the Communications Collection. After early work at Metrovick and RAE Farnborough on blind landing, Gerald Garratt joined the Museum in 1934 and, except for a wartime period as Senior Engineer Staff Officer with the Royal Indian Air Force, remained there and made the history of radio his special interest.

Each of the six chapters after the introductory piece, on invention in general and electromagnetic theory in particular, is each concerned with one major figure in this field. The last one, on Marconi, was written by his daughter Susan, after Garratt's death in 1989, using lecture notes and other published work by her father.

As the author reiterates throughout, no single person could be called the inventor of radio communication. Each of the scientists mentioned here prepared the way for Marconi's work by discovering or clarifying a part of the puzzle. On one point, Garratt is adamant: Popov merely used the circuit described by Lodge to detect distant lightning flashes and expressed the 'hope' that such equipment might eventually be used for communication. He made no significant contribution to the subject and his subsequent elevation to the ranks of those who did was engineered by the Soviet authorities for propaganda reasons. They did him a disservice, because he was a good physicist and needed no such false reputation. One is compelled to wonder why, if Popov was a negligible contributor, it was thought necessary to include him in the book.

It seems that leaders in scientific thought were not, even a couple of hundred years ago, automatically able to command respect. Faraday, who was later to become one of the most brilliant scientists of his time, was the son of a blacksmith, understood very little of mathematics and was regarded with extreme caution, being considered a "green-fingered" experimenter with an empirical approach to science, even though his contribution to field theory was seminal. Even Clerk Maxwell's contemporaries were sceptical until he was able to prove mathematically that both the "action-at-a-distance" fraternity and those who worked from Faraday's lines of force had the right answer.

Some of these people were untaught and felt their way in the face of disbelief and often ridicule from the establishment – a familiar story to the present day. Others approached the subject in a rigorously scientific manner and knew exactly what they were doing, but even they were in virgin territory with no one to consult when difficulties appeared; many of them discovered important processes without recognising what they had done. Their courage and determination to go on when the very cornerstones of the work were being questioned by their peers seem almost unbelievable.

This is a valuable little book (96 pages) in that it pulls together all the important names of the period up to and including Marconi: Faraday, Maxwell, Hertz, Lodge, Popov and Marconi. It is not a dull "history book", but is nonetheless complete with all the relevant scientific detail and a little light relief in the form of personal background. It is No 20 in the IEE History of Technology series, published in hard back at £19.00 in association with the Science Museum. ISBN 0 85296 845 0.

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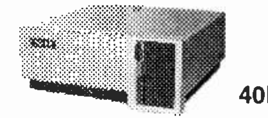

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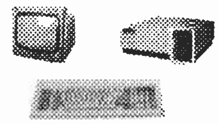

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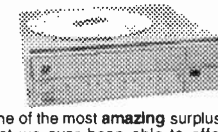

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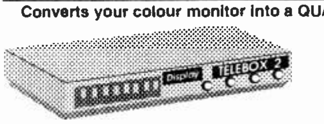

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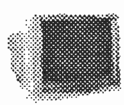

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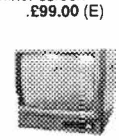

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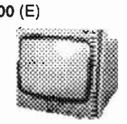

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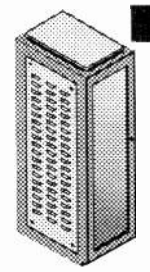

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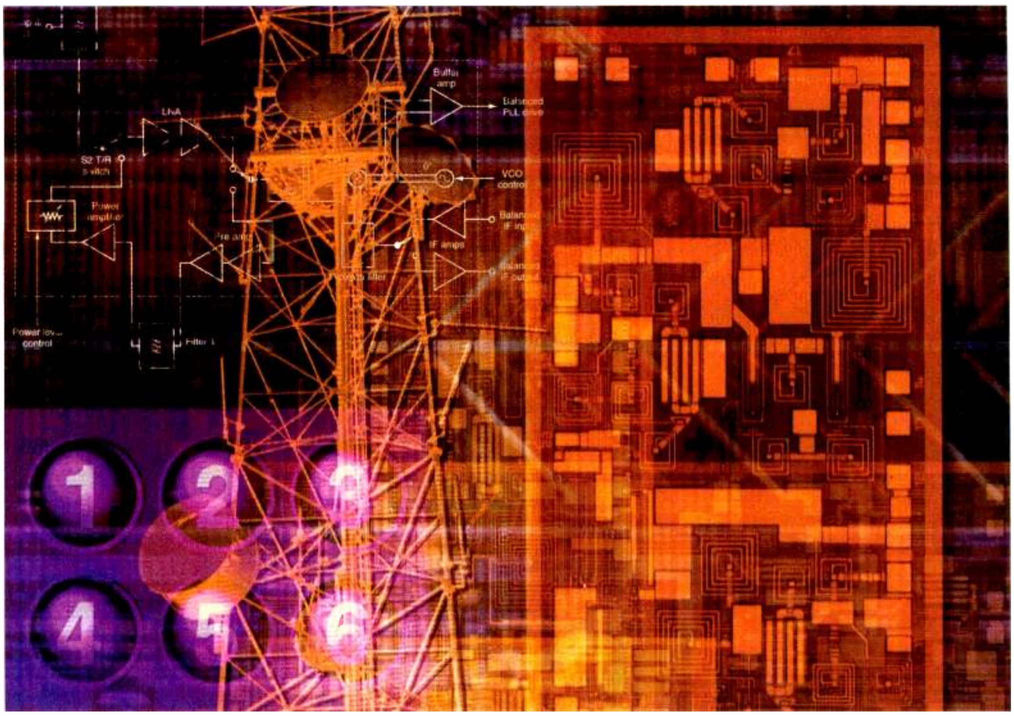


MICROWAVES

2: The laws of microstrip

Microstrip engineering, more than any other technique, provides the foundation to modern microwave design. Mike Hosking* spells out the design rules which turn unwanted reactance into a positive asset.

Most currently designed microwave circuitry is based upon microstrip transmission line. This consists of a conducting track separated from a ground plane by a dielectric layer. The electromagnetic fields propagate and interact but are confined to the region in the vicinity of conducting track and ground plane. One cannot, of course, confine 100% of these fields within the dielectric and some portion will extend into the air above the circuit. Besides affecting the propagation characteristics, such fringing fields will also couple to any adjacent portions of circuit. This latter effect is not always a disadvantage and, as we shall see later, enables a variety of passive components to be realised. **Fig. 1** shows the general structure of a transmission line.

The microwave circuit designer initially needs to know two main circuit parameters: the wavelength within the microstrip circuit (or, equivalently, the dielectric constant or the phase velocity) and the characteristic impedance of the transmission line. Unfortunately, although the configuration of the transmission line appears simple, the answers are deceptively difficult to obtain accurately.

Unlike waveguide or coaxial line, the conducting boundaries are not easily or uniquely definable and an exact synthesis can only be made using, typically, conformal mapping or finite element techniques. The results of such analyses are not convenient, closed-form design equations for engineering use and, in fact, much of the design effort over the past two decades has been spent on developing and improving such closed-form solutions derived from the larger models. The result has been a steady improvement in the accuracy to which circuit designs may be made and the evolution of comprehensive, internationally used CAD tools such as the *Eesof* (recently merged with H-P) and *Super Compact* suites.

The wavelength within the circuit itself (the guide wavelength) must be known, because all designs are based on the electrical dimensions of small sections of circuit. The essence of high frequency circuit design lies in the determination of the inductive, capacitive, or impedance transforming effects produced when an electromagnetic field 'sees' dimen-

Mike Hosking is a lecturer in telecommunications and microwaves at the University of Portsmouth.

sional changes in its transmission line. If all of the field was confined to the dielectric medium then the velocity of propagation, would simply be the velocity of light c divided by $\sqrt{\varepsilon_r}$ where ε_r is the intrinsic dielectric constant (relative permittivity) of the substrate material. However, due to the fact that some portion of the field travels in air (for which $\varepsilon_r = 1$) an effective dielectric constant ε_e must be used and ε_e depends upon several circuit parameters, including frequency. All transmission circuits including free space, present an impedance to the electromagnetic wave, the characteristic impedance, Z_0. Its value must be known in order to design circuits with optimum power transfer and minimised internal reflections.

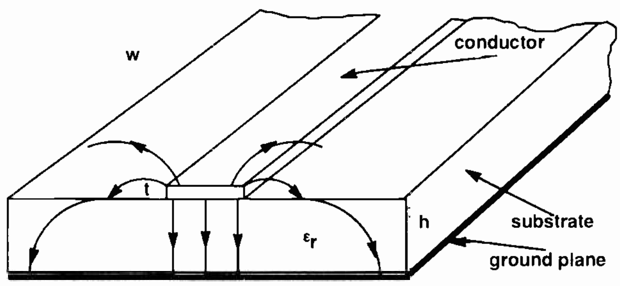
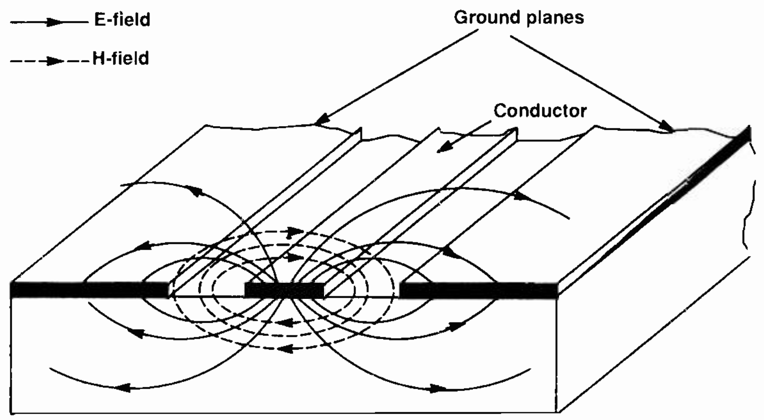


Fig. 1a. General form of microstrip transmission line, showing the electric field pattern with substrate and ground plane.



b. Coplanar waveguide, where conductor and groundplanes are all on the same surface, showing typical field configuration.

Microstrip design equations

The velocity of propagation, hence wavelength, within the microstrip circuit depends upon an effective dielectric constant ε_e. In turn, ε_e depends upon the proportion of field within the substrate to that travelling outside. Both of these depend upon the actual dimensions of the line, together with the substrate material and frequency. For example, we may intuitively appreciate that as the top conducting track becomes wider and wider, then more of the fields can be contained within the substrate. Conversely, for narrow lines, more of the field extends to the air. Thus, the effective dielectric constant must lie somewhere between the extremes of unity and the intrinsic value ε_r.

In practice, the effects of frequency on ε_e can often be neglected below 1 or 2 GHz and, in this case, we have a 'quasi-static' solution to the design (see further reading) where, for example:

$$\varepsilon_e = \frac{\varepsilon_r+1}{2} + \frac{\varepsilon_r-1}{2}\left[1+\frac{10h}{w} - \frac{t/h}{\sqrt{w/h}}\right]^{-0.555}$$

within a certain range of w/h and ε_r.

Thus, the guide wavelength is a function of track width and substrate thickness. For thin metal films, their thickness t can be neglected. Similarly, the characteristic impedance is also a function of w and h, shown graphically in **Fig. 2**.

As a design example, take the two cases of a 50Ω characteristic impedance microstrip line, one on a 1.25mm thick plastic substrate having ε_r=2.2 and the other on 0.125 mm thick GaAs having ε_r=13.1. The former circuit would have ε_e=1.9 and a track width of 3.85mm, while the respective values for the latter would be ε_e=8.4 and w=0.09mm. At frequencies where dispersion must be taken into account (accuracies of <0.5% are usually required) the effective dielectric constant can usually be described by an expression of the form:

$$\varepsilon_e(f) = \varepsilon_r - \frac{\varepsilon_r - \varepsilon_e}{1+P(f)}$$

where $P(f)$ is a semi-empirical modelled term which varies with frequency.

At frequencies higher than those indicated above, account must be taken of what is termed dispersion. In addition to the wavelength changing with frequency, *the actual propagation velocity of the fields changes as well.*

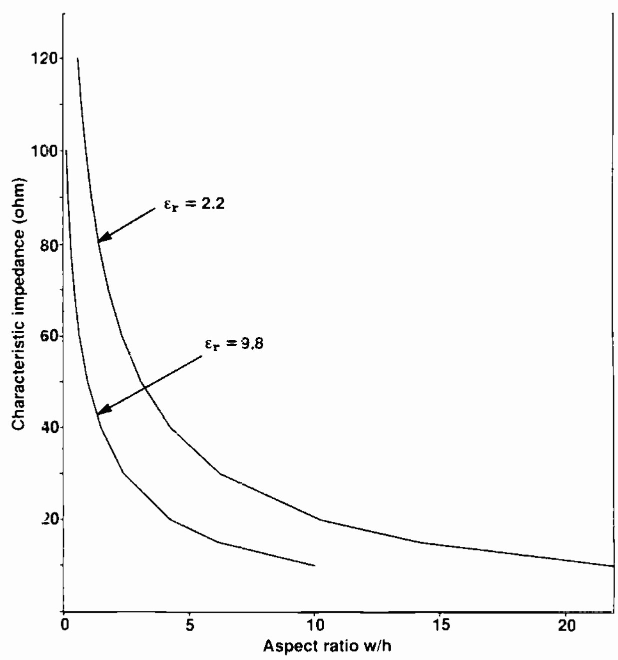


Fig. 2. Graphs of the characteristic impedance of a microstrip line as a function of the line width to substrate thickness ratio for two substrate dielectric constants. It shows that the line width decreases as ε_r increases and that higher values of impedance require narrower lines.

Other important design parameters are the signal losses which occur within a circuit (caused by dissipative losses within the conductors), dielectric loss (caused by the substrate) and radiation loss. There is also an upper frequency limit for microstrip, dependant on thickness and ε_r beyond which a higher mode occurs; this would typically be above 100GHz. Finally, practical circuits must be packaged within some form of enclosure, the sides and top of which can modify their performance by interaction with the fringing fields. Such effects may either be taken into account from analysis of fields within a box, or by ensuring that the enclosing walls are spaced far enough away from the circuit so as to avoid interaction.

Substrate materials

Virtuosity in circuit design accuracy would be of little account if not matched by a corresponding quality of manufacture, particularly in the area of substrate technology. Not only must a choice of materials be available, but their dimensional accuracy, stability and homogeneity must lie within tight tolerances. **Table 1** lists the dielectric constant and loss tangent for some of the more prevalent substrate materials. A wide range of the plastic materials is available. Many of these are based on PTFE, filled with either fine glass or ceramic, although irradiated polyolefin is also used.

Table 1

Material	Tand (10GHz)	e_r
PTFE/glass	10^{-4}	2.1-2.6
PTFE/ceramic	4×10^{-4}	9-11
Alumina (AlO_2, 99.5%)	2×10^{-4}	9.8
Fused quartz	10^{-4}	3.8
Irradiated polyolefin	5×10^{-4}	2.3
Sapphire (anisotropic)	5×10^{-5}	9.4&11.6
Semi-insulating GaAs	6×10^{-4}	13.1

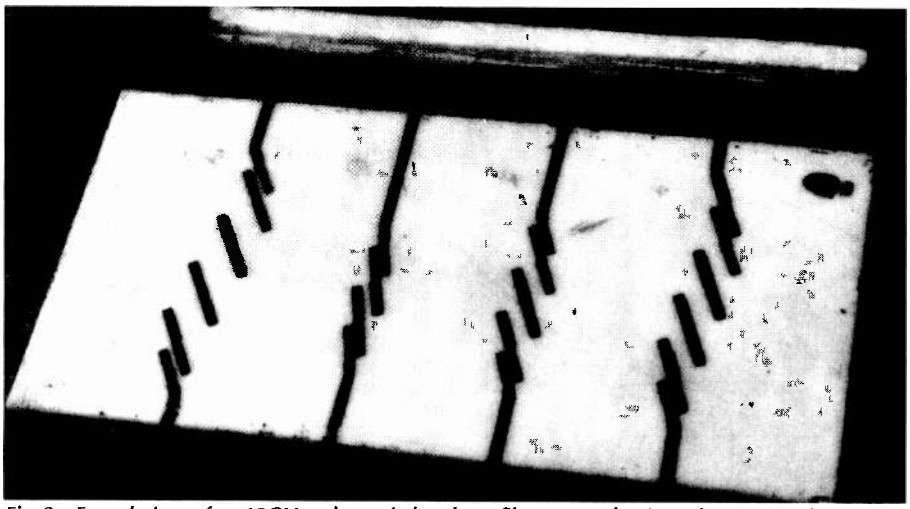
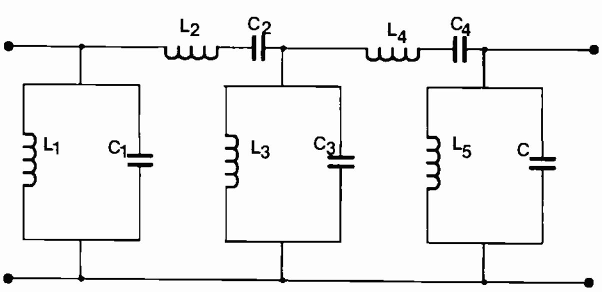


Fig. 3a. Four designs of an 18GHz microstrip bandpass filter on an alumina substrate. The length of each short section of line is approximately $\lambda_g/2$ and the amount of overlap between adjacent lines is $\lambda_g/4$.



b. The bandpass filter equivalent circuit is realised in practice by the parallel resonances due to the $\lambda_g/2$ sections and the series resonances caused by the coupling of the fields between lines.

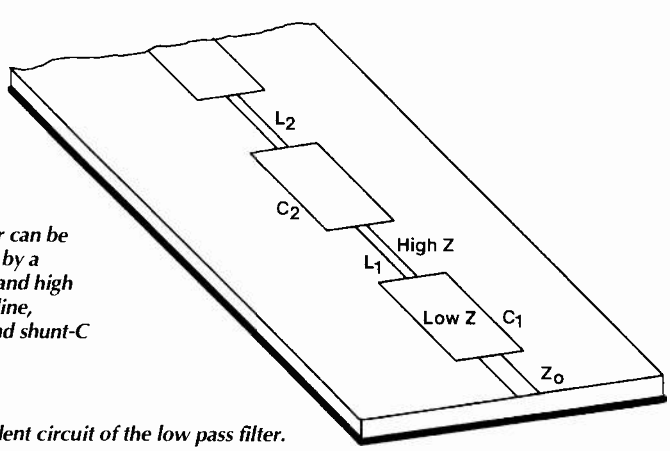
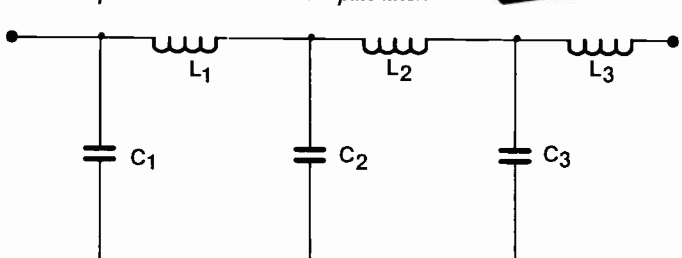


Fig. 4a. A low pass filter can be produced in microstrip by a cascaded series of low and high impedance sections of line, appearing as series-L and shunt-C elements

b. Conventional equivalent circuit of the low pass filter.



Reactance with a difference

If we take a length, l, of transmission line having a characteristic impedance Z_o and we terminate the line with some impedance Z_L, where $Z_L \neq Z_o$, then a portion of any input signal will be reflected, due to the impedance mismatch. We can analyse the circuit in terms of those incident and reflected waves and may eventually express the input impedance (neglecting line losses) as:

$$Z_{in} = Z_o \left[\frac{Z_L + jZ_o \tan\beta l}{Z_o + jZ_L \tan\beta l} \right] \quad \ldots 1$$

where ß is called the phase constant $=2\pi/\lambda_g$ where λ_g is the guide wavelength.

Equation 1 leads to some very interesting results. Take the two extreme cases of making Z_L either a short circuit or an open circuit. In the former case, $Z_L=0$ and equation 1 reduces to:

$Z_{in}=j\ Z_o \tan\beta l$ ...2

This is the expression for a pure reactance, the value of which depends on the length of line l. It can be seen that, for values of l, up to one quarter wavelength ($l_g/4$), Z_{in} is inductive and can have any value between zero and infinity. From $l_g/4$ to $l_g/2$, Z_{in} becomes capacitive and can vary between $-\infty$ and 0. A similar situation exists for an open circuit termination, for which $Z_L=\infty$. In this case $Z_{in}=-j\ Z_o \cot\beta l$ and may again take on any value between $\pm\infty$ as l varies between 0 and $l_g/2$. *Thus, we can make a capacitor or inductor from a simple length of open or short circuited transmission line.* Two special lengths of line are also of interest: when l is a quarter wavelength (or odd multiples thereof) then equation 1 gives:

$$\frac{Z_{in}}{Z_o} = \frac{Z_o}{Z_L} \text{ or } Z_o = \sqrt{Z_{in} Z_L} \quad \ldots 3$$

This is an impedance transforming function, as it has converted the load impedance Z_L into an admittance $1/Z_L$. Alternatively, equation 3 shows that the magnitudes of an input and output impedance may be matched by inserting a quarter wavelength section of transmission line between them, of a characteristic impedance equal to their geometric mean.

The other special length is when $l=\lambda_g/2$; then $Z_{in}=Z_L$. In other words, we could move backwards from the load in half wavelength steps and it would appear as if Z_L was connected at these points. This is a useful aid in designing impedance matching circuitry, a fet amplifier for example, as the circuit elements need not all be crowded together (probably physically impossible anyway) at the actual terminals of the fet.

Actual circuit fabrication is usually a photo-etching process similar to normal printed-circuit board manufacture, the substrates being supplied with a $^1/_2$oz or 1oz copper coating. High purity alumina is a popular substrate and can be obtained with gold metallisation on both surfaces ready for selective etching. Alternatively, screen printing using conductive thick-film inks which are then fired, may be used with alumina. Thick films are ultimately limited by the edge definition, due to the screen, but are successfully used to above 10GHz in frequency. In both of the above types of material, active devices such as diodes and transistors can be bonded to the circuits to produce hybrid components.

Monolithic circuitry with GaAs as the substrate requires a semiconductor manufacturing technology to produce deposited conductor patterns with diffused or implanted regions forming the active devices.

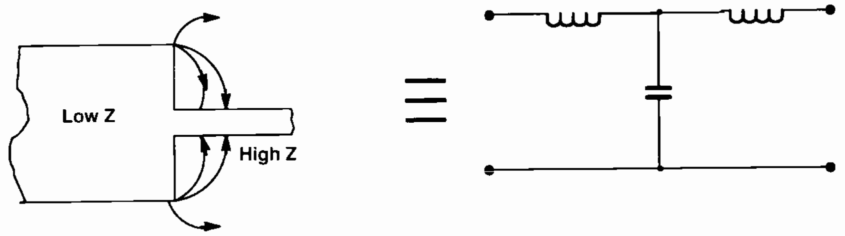


Fig. 5. An abrupt change in conductor width causes a local distortion of the electromagnetic fields to which it appears as an LC circuit in this example. Hence, reactive circuit elements may be produced from physical changes in the line geometry.

Passive component design

The fact that the physical dimensions of microwave transmission lines are comparable in size to the wavelength itself (typically cm or mm) means that the very presence of a section of line, or of a change in dimension, can locally change the phase of the electromagnetic fields and can appear as inductive or capacitive elements. We thus have the concept of "distributed" circuit design. If a conventional type of lower frequency inductor or capacitor were placed into a microwave circuit, it would not behave in accordance with its value. It would appear rather as some complicated discontinuity with a complex equivalent circuit.

A further design technique makes use of the field coupling which occurs when two sections of microstrip line are brought close together. "Close" typically means between 0.025mm and 5mm depending upon the substrate material and the particular application. With a combination of these design tools, components such as resonators, low pass, high pass, band pass and band stop filters, hybrid directional couplers, power splitters, impedance matching networks and delay lines may be realised. All of these appear as innocuous lines on a substrate but are, in reality, the distributed counterparts of lower frequency discrete-element circuits.

As an example of a passive microwave circuit, **Fig. 3a** shows four microstrip bandpass filters on an alumina substrate operating at about 18GHz.

Each filter has the normal equivalent circuit of cascaded *LC* networks, as in **Fig. 3b**, and indeed, the individual filter elements are calculated in the same way as for lower frequency designs to provide bandpass responses such as Butterworth, Chebyshev, elliptic function; with a trade-off between stop-band attenuation, rate of cutoff, insertion loss and number of poles.

In Fig 3(a), each of the shorter lengths of line is approximately a $^1/_2\lambda$ long and appears, electrically, as a parallel *LC* resonant circuit. Individual lines overlap each other by a nominal $1/_4\lambda$ wavelength and mutually couple, thereby appearing as series *LC* combinations. In practice, an open circuit microstrip line radiates slightly and has fringing fields from the line ends, with the result that it appears slightly capacitive. Thus, a length correction must be made.

A slightly different approach to a filter design, in this case lowpass, is illustrated in **Fig. 4a** with its conventional, electronic equivalent circuit in **Fig. 4b**.

A length of microstrip line much narrower (high impedance) than another section of line loading it, will appear as a series inductance, whilst a wide (low impedance) section will appear largely as a parallel capacitance. Thus, the desired response can be produced by cascading *L* and *C* sections. One typical use of such a filter is in the bias circuit to an active device: dc connections have to be made without allowing the microwave signal to travel into the supply circuit or introducing mismatch.

In practice, this particular type of filter is restricted to frequencies below a few GHz due

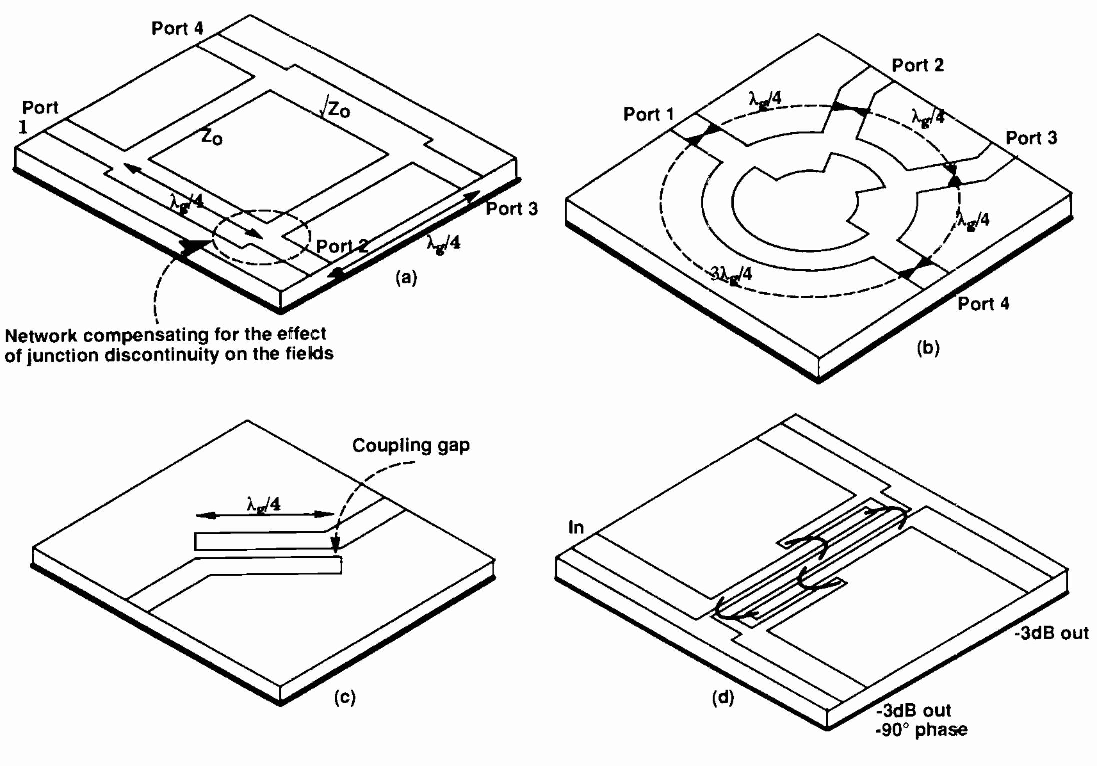


Fig. 6. Four types of microstrip directional coupler in common use: a. A 90° hybrid branch line; b. A 180° hybrid ring ("rat race"); c. Edge-coupled strip having a coupling dependant upon the gap; d. A Lange coupler used on wideband circuits, giving a 3dB power split with a 90° phase difference.

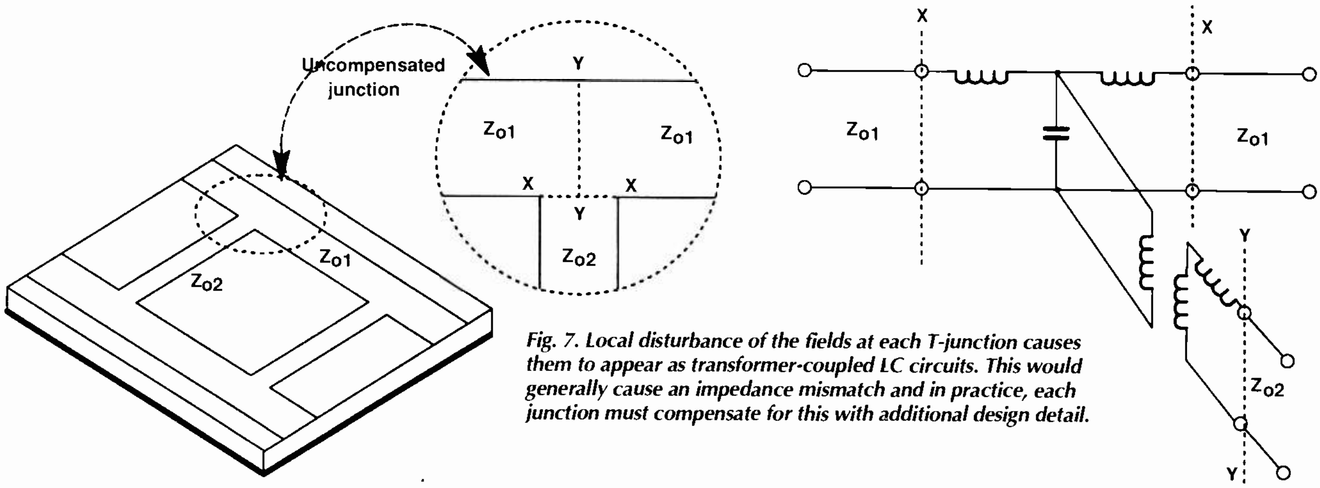


Fig. 7. Local disturbance of the fields at each T-junction causes them to appear as transformer-coupled LC circuits. This would generally cause an impedance mismatch and in practice, each junction must compensate for this with additional design detail.

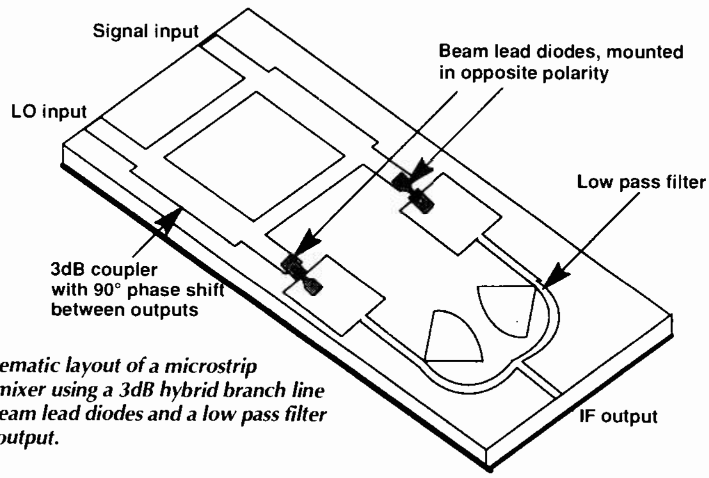


Fig. 8. Schematic layout of a microstrip balanced mixer using a 3dB hybrid branch line coupler, beam lead diodes and a low pass filter for the IF output.

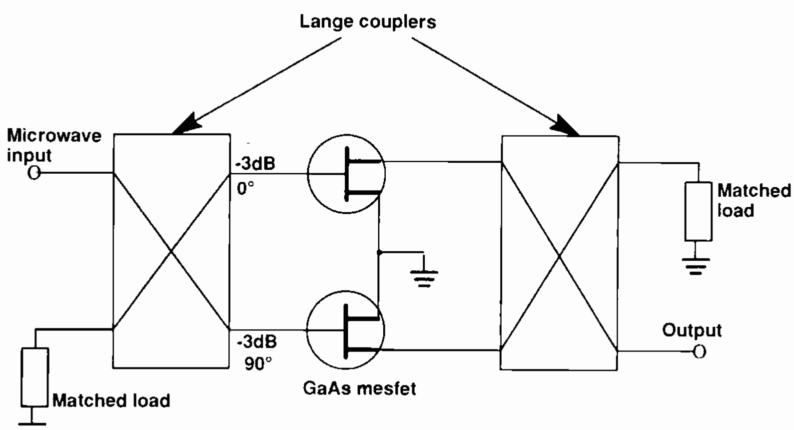


Fig. 9. A balanced amplifier configuration used for wideband circuits reduces the problems of internal mismatch and provides continuous operation should one fet fail.

to fringing fields from the step discontinuity. As shown in **Fig. 5**, these cause the local dimensional change to appear as another form of *LC* network, degrading the filter performance. This latter point serves to illustrate the fine detail which has to be taken into account in the design of high frequency circuits.

Directional couplers

This device is a basic circuit building block used in applications such as balanced, double balanced and image-rejection mixers, balanced amplifiers and certain antenna feed networks. The coupler performs a power splitting function, ie one input and two outputs, but with usually a 90° or 180° phase shift between the two outputs. **Fig. 6** shows four types of commonly used coupler in microstrip form.

Type (**a**) is a 90° hybrid branch line coupler so called because of its ability to provide an equal power split with a 90° phase shift between outputs. Each section of the coupler is normally $^1/_4\lambda$ long and the branched design has the property that a signal input at Port 1, say, will split equally with half going to Port 2 and half to Port 3. However, the output at Port 2 has travelled a distance of $\lambda g/4$, while that at Port 3 has travelled $\lambda g/2$, thereby introducing a 90° phase difference.

In practice, the T-junctions of the lines have an equivalent circuit shown in **Fig. 7** and thus the local detail around the "T" must be altered, in order to introduce additional reactances for matching. Again, this is an example of the design detail required for microwave circuits.

This type of coupler is widely used in balanced mixer design, shown schematically in **Fig. 8**. The two mixer diodes are connected in opposite polarity and their outputs are combined via low pass filters to give the IF, while the overall effect of the phase shift introduced by the coupler is to produce cancellation of the local oscillator AM noise.

The coupler shown in Fig. **6b** is also used in mixer circuits and is called a hybrid ring or rat-race coupler. A phase shift of 180° is introduced between outputs with, usually, an equal power split, although this can be changed by varying the characteristic impedance of the coupler sections.

An input at Port 1 will split equally, say, at the ring junction and the two components will travel in opposite directions around the ring. At Port 2, one component will have travelled $\lambda_g/4$ and the other $5\lambda_g/4$ so they will both be in phase at this port and will thus partially couple out.

At Port 3, one signal will have travelled $\lambda_g/2$ and the other λ_g so they will be 180° out of phase and will cancel, giving no output.

At port 4, the two components will again be in phase and can, therefore, exit but, because they have travelled an extra $\lambda_g/2$ compared with those at Port 2, the coupler outputs have a 180° phase difference.

Instead of forcing the microwave fields to split into two parts at a junction, couplers may be designed just by bringing two tracks into close proximity as in Fig. **6c**. Even and odd mode waves are generated, each having a different impedance and phase velocity. The amount of coupling depends upon the proximity of the lines. However, it is difficult to achieve a coupling of more than about –10dB.

The previous coupler types have the disadvantage that they are inherently narrow band designs, although the bandwidth of the branch line hybrid and side coupler can be increased by adding more sections at the expense of greater substrate area and insertion loss. A design which avoids these problems is the Lange coupler of Fig. **6d**.

Of interdigital construction, it uses bond wires to equalise the even and odd mode phase velocities and gives a 3dB power split with a 90° phase difference between outputs. However, the big difference is that the amplitude and phase balance is maintained over frequency bands in excess of an octave, leading to widespread use in balanced amplifiers, as in **Fig. 9**. See also Fig. 4 in *Part 1*.

A microstrip receiver circuit which brings together several of these elements is shown in **Fig. 10**. The substrate is alumina, about 0.6mm thick and most of the tracks are 50Ω transmission lines. The hybrid ring style of mixer can be seen, with the microwave input signal passing through a coupled line bandpass filter. A single, edge coupled strip is used to sample power from an external source to provide the local oscillator signal. The discs in the foreground are bias magnets for circulators on ferrite substrates. ■

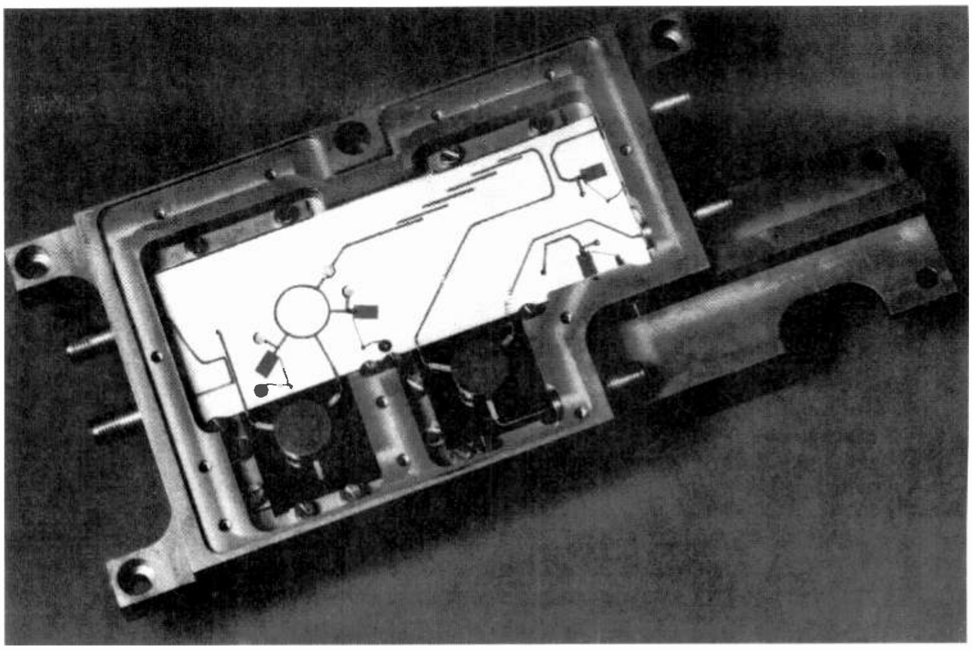


Fig. 10. A complete receiver circuit showing hybrid ring mixer, bandpass filter, edge coupled strips and ferrite circulators.

Further reading

Foundations for Microstrip Circuit Design, 2nd. edition, 1992, T Edwards, Wiley.

Next Month:
lumped components and waveguides.

CIRCUIT IDEAS

SEND YOUR CIRCUIT IDEAS TO THE EDITOR, ELECTRONICS WORLD, QUADRANT HOUSE, THE QUADRANT, SUTTON, SURREY SM2 5AS

DO YOU HAVE A £100 CIRCUIT? EACH MONTH'S TOP CIRCUIT IDEA AUTHOR WILL RECEIVE £100. ALL OTHER PUBLISHED IDEAS WILL BE WORTH £25. WE ARE LOOKING FOR INGENUITY AND ORIGINALITY IN THE USE OF MODERN COMPONENTS

£100 WINNER

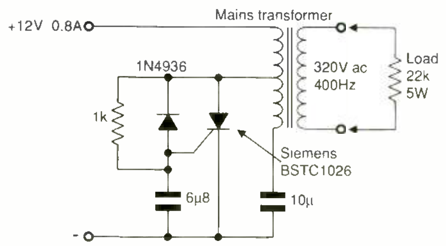
SCR inverter

With one or two drawbacks, the saving grace of this single-SCR inverter is its simplicity. It produces a vaguely sinusoidal waveform of around 320V AC at 400Hz, the frequency being only slightly affected by supply voltage and load. Losses in the RS 209-847 transformer dictate an efficiency of 50% with a resistive load.

At switch on, oscillations must be allowed to build up slowly on a light load, since a heavy load would cause latch-up in the conducting state, necessitating some kind of currrent limiting.

The only variable is the resistor, which will affect the output voltage to some extent and allow lower or higher supply-voltage working. A lower frequency requires a larger resonating capacitor, which should not be an electrolytic type.

D Di Mario
Milan
Italy



Very simple inverter, whose output is virtually unaffected by supply voltage, produces about 320V AC at 15mA.

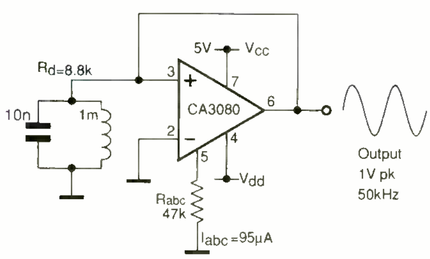
OTA oscillator

In non-linear mode, an operational transconductance amplifier will function as an *LC* oscillator in the circuit originally described by Baxandall in 1959.

The parallel *LC* circuit on the non-inverting input of the *3080* OTA receives feedback from the output. At resonance, the tuned circuit appears as a resistive load $R_d = 2\pi f_0 LQ$. Bias current in the OTA, set by R_{abc}, determines g_m, which is about 2mA/V for an I_{abc} of 1mA. Oscillation at f_0 takes place when $g_m R_d > 1$, increasing in amplitude until the amplifier limits, acting as a switch to drive constant I_{abc} and $-I_{abc}$ into the tuned circuit, the waveform across it being sinusoidal with an amplitude of $4R_d I_{abc}/\pi$ and third-harmonic distortion of $(100/8Q)\%$.

The circuit works to about 100kHz, although distortion rises at higher frequencies.

J Willis
Macclesfield
Cheshire



Simple oscillator for up to 100kHz, using an operational transconductance amplifier to give a 1V pk output at 50kHz.

Voltage-independent time delay

Although this is an *RC* time-delay, no adjustment is needed and supply variations have no effect, since such variations affect both inputs to a voltage comparator equally.

Input pulses cause a '1' at the output of IC_{1b}, this being applied to the *RC* and to $R_{2,3}$, so that

$$v_{C1} = v_1\left(1 - e^{-t_0/R_1C_1}\right)$$

$$v_{R3} = v_1 R_3\left(R_2 - R_3\right)$$

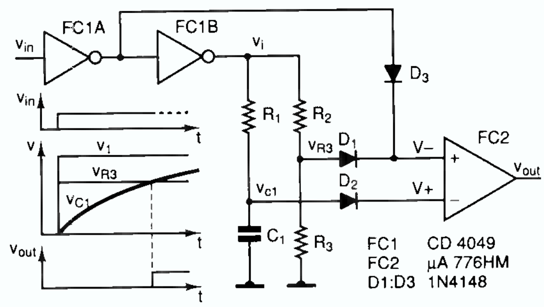
Since the delay is determined by the time needed for v_{C1} to become equal to v_{R3}, solving that equation for t_0 produces

$$t_0 = -R_1C_1 \ln\left(1 - R_3/\left(R_2 + R_3\right)\right)$$

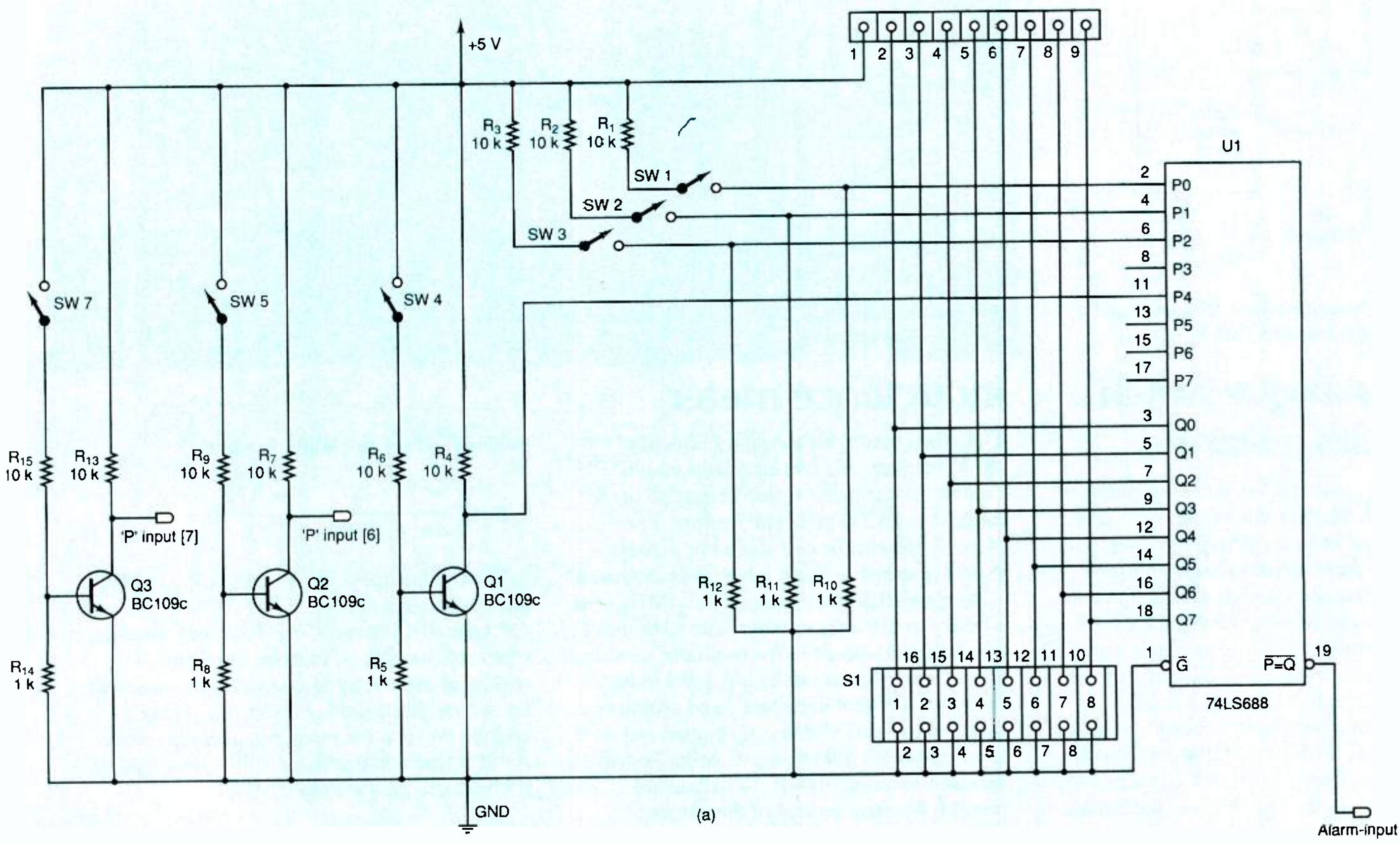
where t_0 depends only on the *RC* combination, and voltage plays no part.

The diodes avoid the possibility of two zeros on the comparator inputs, when the inputs is zero. Inverting input is always high, since either IC_{1a} output or IC_{1b} output is always present.

N I Lavrentiev
Kaliningrad
Moscow Region
Russia



This time delay is independent of supply-voltage variations, even though delay is determined by an RC circuit.



Comparator extends alarm system

Several alarm sensors, presenting either normally open or normally closed contacts, are used as inputs to a digital comparator, which emits an enable to an alarm system when any sensor contact changes state.

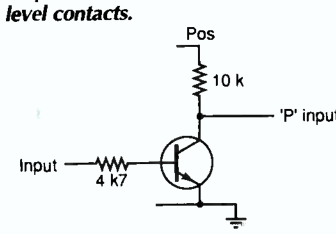
A *74LS688* has two sets of eight inputs, P and Q, and one P = Q output; when all P inputs are equal to all Q inputs, P = Q is low, otherwise it remains high.

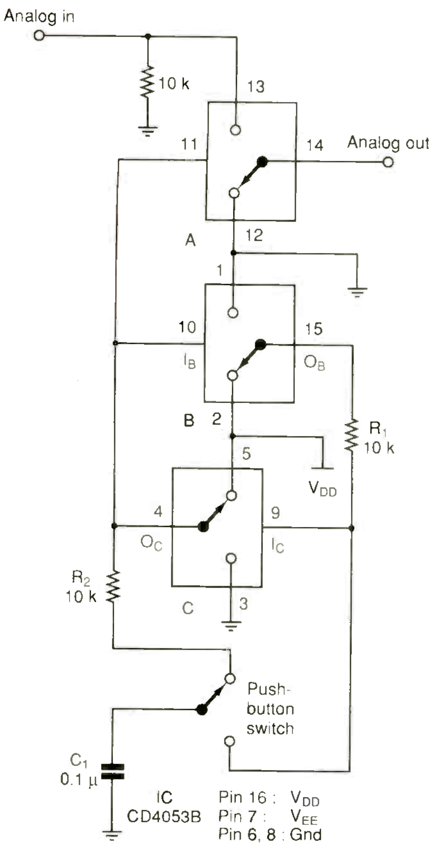
Any Q input may be set at 0 or 1 by the dil switches and the 10kΩ sil resistors. On the P side, normally closed contacts such as that on P0 pull the inputs high, an opened switch taking the input low via its 1kΩ resistor. Normally open contacts allow transistor switches to provide normally high inputs to their P inputs, closing switches again taking the P input low; the use of transistors avoids trouble with varying supply voltages. For low-level sensors, the contact could feed the transistor base directly, as in (b).

Connecting eight *688*s to a summing *688* would extend the number of possible sensors to 64.

M Saunders
Leicester

Up to eight contact sensors connected to one 74LS688 generate an alarm enable when one sensor is actuated, whether the sensors are normally open or closed contacts. At (b) is a simplified transistor feed for low-level contacts.





Push-button alternately opens and closes analogue switch.

Analogue switch with memory

A push-button switch connects or disconnects an analogue signal in successive operations, using one IC, three resistors and a capacitor.

Analogue signals go through one channel of a *4053B* three-channel multiplexer, the other two being connected as a bistable flip-flop so that outputs O_B and O_C are alternately 1 or 0. When O_C is 0, the capacitor is also at 0V; pressing the button connects it to the input I_C and triggers the flip-flop. As the button is released, the capacitor charges again to V_{DD}.

A second operation causes the capacitor to apply V_{CC} to I_C, which retriggers the flip-flop so that the capacitor is once again discharged when the button is released. Since the flip-flop is toggled each time the button is pressed, output O_C opens and closes the top channel to the analogue signal.

Requirements are V_{DD} = 5V-15V, negative supply V_{EE} = −13.5V-0V and (V_{DD}–V_{EE}) = 15V maximum.

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ISRO Satellite Centre
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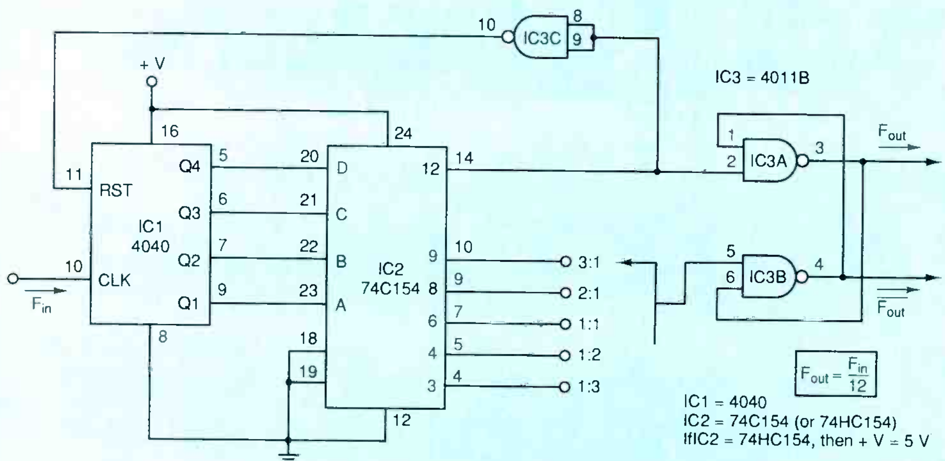
India

Variable-frequency generator has switchable duty cycle

At any frequency up to around 330kHz, the duty cycle of this generator is settable between 3:1 and 1:3.

Four stages of ripple counter *4040* drive the *74C154* 16-line decoder, clock input to the 4MHz counter being divided by 12. When output 12 of the decoder goes low, it sets the output flip-flop, resetting the counter, which starts to count again from zero. As the count reaches that corresponding to the decoder output selected by the switch, the flip-flop is reset. Maximum input frequency is about 4MHz.

Alberto R Marino
Madrid
Spain



At one-twelfth of the clock frequency, this rectangular-wave generator has a duty cycle selected between 3:1 and 1:3.

Inductance meter

When used with a digital frequency counter, this two-transistor circuit indicates inductance within about 5% in the range 0.1µH-2000µH. The reading is not direct, but a simple calculation or perhaps even a graph or look-up table gives the result.

An rf oscillator operating at 3.5-4MHz feeds a buffer to drive the counter. The terminals L_x, when shorted, result in the oscillator's natural frequency; connecting the unknown inductor increases the total inductance and produces a lower frequency, which is displayed and used to calculate the inductance. *C* being known. A receiver covering 300kHz-4MHz could possibly be used instead of the counter.

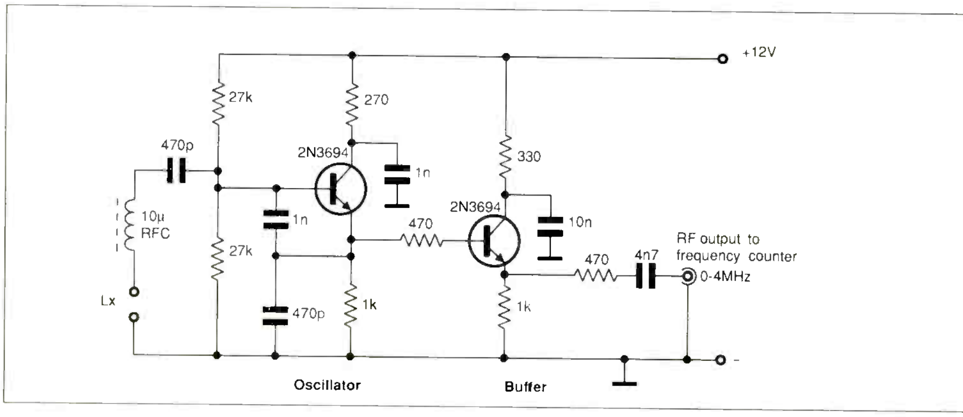
The value of the rf choke must be subtracted from the calculation, which, since the total value of the two 470pF capacitors and the 0.001µF comes to 190pF, becomes

$$L_x = \left(\frac{1}{4\pi^2 f^2 \times 190 \times 10^{-6}}\right) - 10$$

where inductance is in µH, frequency in MHz and capacitance in µF.

Practically, the usual rf precautions must be observed, namely rigid metal enclosure, regulated supply, good-quality capacitors, and the use of short, rigid connections. These include those to the unknown inductor. Any small-signal npn transistor will suffice and the rf choke can be a standard type.

Peter Parker
Bentley
Australia



Easy to use mains timer

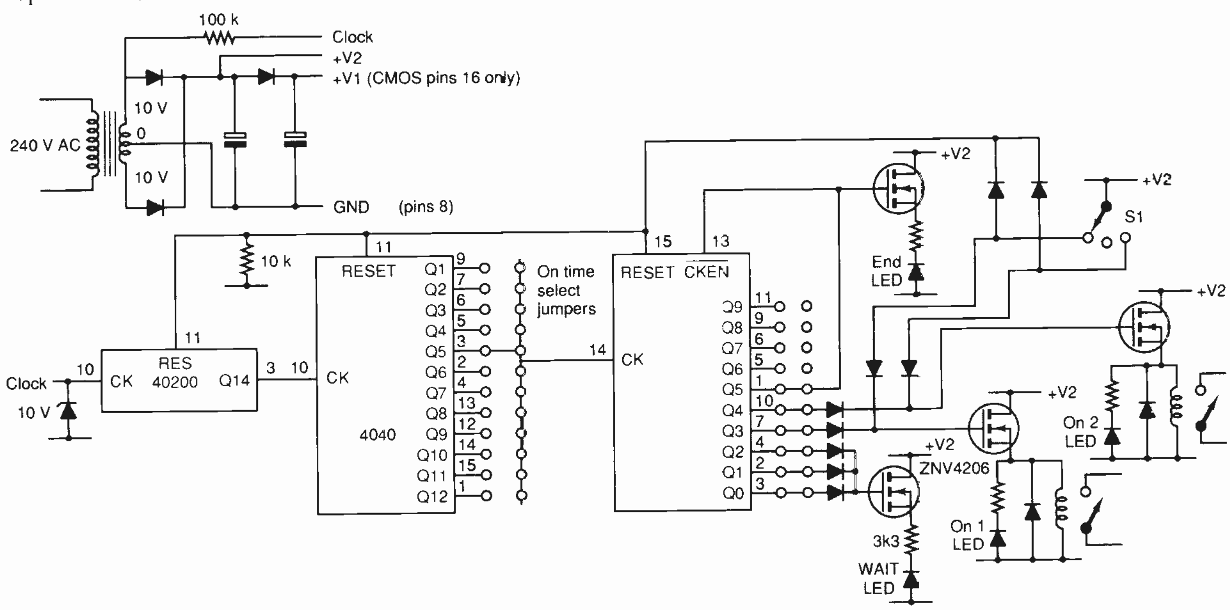
Together with a thousand or so others, our household is currently part of an electricity tariff experiment. During the day, electricity is much more expensive than standard rate, at 12.6p a unit but in the evenings it drops to 5.6p and between 12:30 and 07:30 at night it is only 2.6p. This easy to use timer was designed to run the washing machine and dryer after 12:30 at night.

Two outlets are needed because the dryer and washer cannot be run together from one 13A socket. There is only one control, a switch, which allows manual selection of either of the two sockets. Switching to either resets the timer while switching back to the centre position initiates the timing sequence.

A red LED signals the timer's wait period while a green one signals the end of the cycle. Set up as shown, the first socket turns on about 3h 40m after reset and stays on for just under an hour and a half, after which the second socket turns on for the same period. These periods mean that the timer can be initiated at any time between 8:53 and 12:55 in the evening. Using jumpers allows the on and delay periods to be altered.

Reset must be initiated in both switch positions to eliminate the possibility of having both sockets on at once, overloading the outlet. How long the capacitor can supply the cmos ICs during mains failure is difficult to determine since connecting a voltmeter to the supply increases the discharge rate. It should be at least a few minutes.

James Stevenson
Newcastle-u-Lyme
Staffordshire



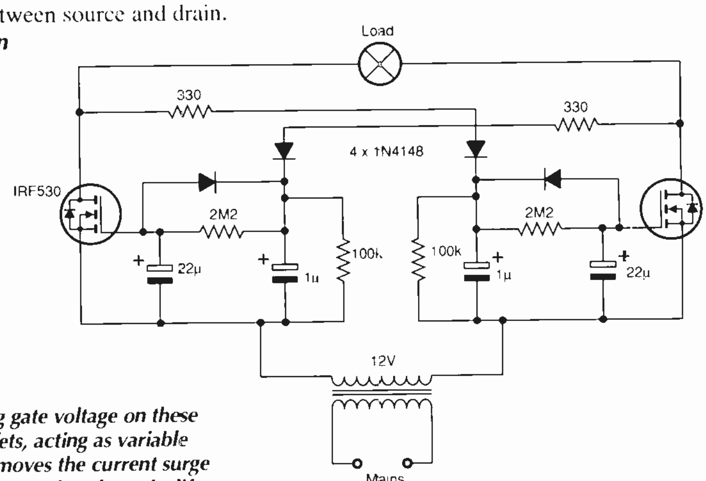
Soft-start filament driver

Two power mosfets with a ramped gate voltage switch on slowly and eliminate a current surge into a cold filament.

The mosfets act as variable resistors, two charge pumps applying a slow-rising DC bias to the gates. Initially, the mosfets are off and the 22µF and 1µF capacitors charge through the 330Ω resistors and the internal drain/source diodes of the opposite mosfets. During this time, the load receives drive from one mosfet and the substrate diode of the other during each half cycle. When the capacitors are fully charged, both mosfets are on, voltage drop through the circuit being determined by the mosfet $R_{DS(on)}$ of about 0.2Ω. The gate drivers give a slow charge, fast discharge characteristic.

Mosfets without the substrate diode can be used, but must have a diode such as the *1N4002* between source and drain.

Joel Setton
Crolles
France



Slowly rising gate voltage on these power mosfets, acting as variable resistors, removes the current surge into a filament and prolongs its life.

LETTERS

Self challenges Duncan over hot audio

Having read Ben Duncan's article *Spirit of Bass* (*EW + WW*, February) I felt there were a few too many unsupported assertions to let them all go.

The alleged benefits of regulated supplies for power amplifiers may have been covered in some depth in previous articles, but this is different from proving that these benefits exist. Ben's previous contribution (*EW + WW*, October) concludes that distortion introduced by ripple and signal on supply lines is a serious problem. But to reach this conclusion he found it necessary to assume that the PSRR of the amplifier in question was about 25dB, which is unrealistically awful even for an unsophisticated circuit.

I question the statement that some sort of "envelope modulation" results between the 100/120Hz reservoir charging frequency and bass frequencies. This obviously occurs when an amplifier with an unregulated supply is driven into continuous clipping, but what has this to do with the finer nuances of reproduction? I can think of no other way for such modulation to happen: perhaps Ben can enlighten me.

I also winced when I read that "bass clarity has been shown to be improved by... exchanging linear and 80kHz switching PSUs". Has it? When and by whom? And what definition of "bass clarity" are we using here? Is it intermodulation distortion, or something so magically non-measurable that only subjectivist hi-fi reviewers can perceive it?

However, my main concern is the reference to "thermal distortion" at the end of the article. Ben seems to take it as given that such a distortion mechanism exists in power amplifiers, but having studied the subject in some depth I have yet to see the effect, and frankly I don't think it exists.

I do agree that it happens in op amps, because of having output and input devices on the same chip, so there is very close thermal coupling between them. But this has nothing to do with amplifiers constructed from discrete devices. I note that his reference 13 deals with op amps only, and even then it seems necessary to measure the thermal distortion effect at 1Hz, well outside the audio band.

	10Hz AP out	Amp out	1kHz AP out	Amp out
Fundamental	.00013%	.00031%	.00012%	.00035%
Second	.00033%	.00092%	.00008%	.00060%
Third	.00035%	.00050%	.000013%	.00024%
Fourth	<.000002%	.00035%	<.000008%	.00048%
Fifth	<.00025%	<.00045%	.000014%	.00024%
Sixth	<.000006%	.00030%	.000008%	.00021%
Seventh	<.000006%	<.00008%	.000009%	.00009%
Eighth	<.000003%	.00003%	.000008%	.00016%
Ninth	<.000004%	.00011%	.000007%	<.00008%
AP THD reading (80kHz bandwidth)	.00046%	.00095%	.00060%	.00117%

NB: The rejection of the fundamental is not perfect, and this is shown as it contributes to the THD figure.

While there have been odd mentions of thermal distortion in power amps in some of the hi-fi press, I have never seen any explanation of how it might work, any estimate of the magnitude of the effect, and a circuit that will demonstrate its production.

In the usual absence of specifics, I can only assume that the alleged mechanism induces parameter changes in semiconductors whose power dissipation varies over a cycle. If this happens, it would presumably manifest itself as a rise in second or third harmonic distortion at very low frequencies, but this simply does not happen. The largest effects would be expected in Class B output stages where dissipation varies wildly over a cycle; the effect is still wholly absent.

One reason for this may be that drivers and output devices have relatively large junctions with high thermal inertia. A few minutes work with hammer and chisel revealed that an *MJE340* driver has a chip with four times the total area of a *TL072*. Given this thermal mass, parameters presumably cannot change much even at 10Hz. Low frequencies are also where the global NFB factor is at its maximum; it is perfectly possible to design an amplifier with 100dB of feedback at 10Hz, though much more modest figures are sufficient to make distortion unmeasurably low up to 1kHz or so. Using my design methodology, a blameless amplifier can be straightforwardly designed to produce less than 0.0006% THD at 10Hz (150W/8Ω) without even thinking about thermal distortion. I think this suggests that we have here a non-problem.

I accept that it is not uncommon to see amplifier TH plots that rise at low frequencies; but whenever I could investigate this, the LF rise could be eliminated by attending to either defective decoupling or feedback-capacitor distortion. Ben says he accepts the effect must be at a very low level as it is invisible at 0.0006%; remember that this is the level of a THD reading that is visually pure noise, though there are

Virtual travel

Compared to previous booms in brown goods, the relatively highly priced camcorder may have been less exciting than video recorders or colour televisions. But the industry must look for a new product, and the buzz is about virtual reality.

A number of new technologies have appeared recently that can be put together to make a remarkable family of devices, which will probably be as near as we will ever get to Star Trek's transporter.

The initial or basic offering will be relatively crude and consist of a transmitter and receiver that can be plugged into telephone sockets anywhere. The receiver will be a virtual reality helmet, with stereo headphones and two video displays with arrangements to focus them as a stereo pair to the viewer. It will be connected to a box that decodes the video and audio data from the fractal compression system used to send it down the telephone system.

The transmitter will be a head sized object with binaural microphones and a pair of television cameras, and the data will be fed via a fractal compression unit to the telephone line. It will probably be sold at high prices to business users so that, for example, someone can be shown round an office or factory to decide whether they are sufficiently interested in buying it before visiting it in person.

Being shown round with this set up would be a bit like a quadriplegic being wheeled around in a wheelchair. Although you would have the feel of being there, you would have to look where your head was pointed.

The next development would be to arrange for the receiver to send to the transmitter signals representing the azimuth and bearing of the head, so that the viewer could turn his or her head around and look where he or she wants. But the head's location would still be moved by the people the other end. This would be more like a paraplegic being wheeled around in a wheelchair.

The stage on from this would be to mount the head on a small airship-type flying machine, and set its computer to keep it at head height. Then directional signals could also be sent to it from the receiving person. I suggest that a flying machine would be easier and cheaper to make than a walking machine. The buoyancy could still be obtained by a cylinder, held vertically by its flight control system, equal in volume to a body and legs and it only need float a few inches above the ground. It may look a bit like an Dalek or animated dustbin, but it would do the job and should be easy enough to mass produce.

However, as it stands people would not be able to use this product to visit each other, as the person receiving the visit would still see the animated dustbin rather than the person.

But a simple trick could be used to get over this. A camera at the terminal of the subscriber making the virtual visit would record his or her appearance, and the person being visited could also wear a virtual reality helmet. However, this helmet would display the home image, except instead of the animated dustbin it would display the image of the person making the visit. It could not be impossible to edit out the VR helmets the two people would be wearing, so each would see the other as if they were a real person.

Of course there will be those who will suggest that this can never replace travel and it would be bad for the soul and so on. However, I can see no technical reason why it would not happen, and in reality it could well have uses to enhance travel rather than suppress it. Never again could the travel agent suppress the fact that the hotel is between the crematorium and the abattoir if customers expect a quick VR tour before laying out their money for the tickets!

John de Rivas
Truro,
Cornwall

real amplifier distortion products buried in it.

I have therefore done some deeper investigation by spectrum analysis of the residual, which enables the harmonics to be extracted from the noise. The test amplifier was an optimally biased Class B machine very similar to that in part 7, except with a CFP output. The Audio Precision oscillator is very clean but this amplifier tests it to its limits, and so the table shows harmonics in a before-and-after-amplifier comparison. The spectrum analyser bandwidth was 1Hz for 10Hz tests, and 4.5Hz for 1kHz, to discriminate against wideband noise.

This further peeling of the distortion onion shows several things: that the Audio Precision is a brilliant piece of machinery, and that the amplifier is really quite linear too. However, there is nothing resembling evidence for thermal distortion effects.

As a final argument, consider the distortion residual of a slightly underbiased power amp, using a CFP output configuration so output device junction temperatures do not affect the quiescent current. It therefore depends only on the driver temperatures. When the amplifier is switched on and begins to apply sine wave power to a load, the crossover spikes (generated by the deliberate underbiasing) will be seen to slowly shrink in height over a couple of minutes as the drivers warm up. This occurs even with the usual temperature compensation system, because of the delays and losses in heating up the V_{be} multiplier transistor.

The size of these crossover spikes gives in effect a continuous readout of driver temperature, and the slow variations seen imply time constants measured in tens of seconds or more. This must mean a negligible response at 10Hz.

There is no doubt that long-term thermal effects can alter Class B amplifier distortion, because as I have written elsewhere, the quiescent current setting is critical for the lowest possible high-frequency THD. However, this is strictly a slow (several minutes) phenomenon, whereas Duncan's comments about "even-harmonic residues" show that he is thinking of the usual sort of per-cycle distortion.

The above arguments lead me to conclude that thermal distortion as usually described does not exist at a detectable level. Nonetheless, if anyone has any hard evidence in rebuttal, I would be interested to hear it. Subjectivist impressions, however, are not required.

Douglas Self.
Forest Gate,
London

Duncan replies

Douglas Self is being economical with the facts. Regulation improved the %THD+N performance of the amplifier for which I developed the regulator (see Fig. 2, p822, *EW + WW*, October 1992), by up to tenfold above 1kHz.

His guidance on PSRR, while correct in the circumstances, has no relevance to the topology I used (the driver stage had a 100% separate and ground isolated regulated supply) and having many years' experience in power audio, I resent any implication that my 0V noding or NFB takeoff were anything other than blameless, long before I even considered improving performance with a regulator.

At 650W/4Ω the amplifier in question also had considerably larger currents flying about and regulation dramatically cleaned up residue on the supply wires, which makes the black art of lead dress and twist less of an issue, opening new possibilities for ultra-compact packaging of high power without runaway noise induction.

More generally, Douglas is attempting to show that supply regulation is just not needed – ever, without even trying it, and based on idealised experience with just one classic topology. Simultaneously, he implies it has never been needed. The latter appears to be an unsupported assertion of his own, as he shows no measurements of PSR vs frequency or the supply residue of commercial products.

My –25dB PSRR figure was clearly intended as a worst case baseline at 20kHz, and not a wholly unrealistic one with some topologies and real-world PSR-reducing compromises found in PA and up-market hi-fi amplifiers.

Douglas Self has shown supply regulation may not be needed to get quite low static %THD+N figures when driving an 8Ω resistor with his own idealistically laid out 50W development of the Lin topology (1956). But where are those measurements demonstrating the extent to which the blameless performance won't be disrupted by RFI on the mains supply or perturbations caused by the transient demands of a hungry, capacitative loudspeaker?

Self's own words are forked, if, as he says in part 1, a power amplifier really is just an op-amp with boots on. So if all power amplifiers can work blamelessly with a raw supply, then why does he and everyone else bother to use regulated supplies for IC op amps in, say, mixing consoles?

Concerning improvements in bass clarity, the MSL-Rauch *P600* is a 600W/4Ω/channel amplifier with an 80kHz switching supply, but otherwise its audio path replicates Rauch Precision's older *DVT 250s* model, which has a conventional supply.

Both amplifiers have had an enduring presence in the UK and European live performance, rehearsal and recording studio circuit, and over the past seven years, a number of experienced sound engineers have noted the improved sonics when changing between the two models.

Subjectivists yes, but also top professionals, and here Douglas ignores reality if he cannot accept that humans necessarily have the final say about equipment intended to communicate with most people who have a right brain, capable of measuring music – something machines are still too daft to even recognise.

As the improvement occurs across a wide range of units, artistes, microphones, venues and speaker systems, it has been robustly demonstrated. The relative absence of 50Hz harmonics and intermod products goes some way to accounting for differences. These and others can doubtless be identified and validated, if someone wishes (is Douglas volunteering?) to sponsor the research effort.

As to looking for thermal distortion in monolithic ICs, the JG Graeme test I cited naturally uses 1Hz to create a fairly gross (say > 3%) error for visibility's sake on a curve tracer, and to overcome display hysteresis. As most low to mid market domestic power amplifiers are nowadays monolithic ICs, the very kind of thermal distortion that even Douglas Self admits is more rife than ever!

Inside his own discrete circuitry, my adversary is wasting his time looking for thermal errors by minutely analysing harmonics with a uniform low crest test signal – and in sluggish output devices of all places.

His figures merely demonstrate that any thermal distortion is not contributing much to %THD with a specific topology and specific devices. This is not the same as proving it doesn't exist, ever, anywhere. Showing that 10Hz %THD did or did not change when the device's thermal inertia is say quadrupled, would be more daring.

If Douglas scrutinised the hated hi-fi press more carefully, he might gather that thermal distortion is the industry's best working hypothesis with regard to certain A-B tests between notionally equivalent components.

Superb as it is, the Audio Precision *SYS-22* is completely the wrong tool for this kind of detective assignment. Measuring a heat wiggle will happen one day; wait until the professionals get there.

Turning over fully to Douglas Self's own series, it on the one hand deserves praise for writing style and content, for engineering pragmatism and clarity of thought, for organising formerly sporadic, diverse and oft half-baked information into simplicity, and not using the infamous seasickness capacitor location to compensate. His series is indeed excellent stuff for teaching.

But anyone bothered about musical accuracy should be cautioned that Self's entire presentation is limited to honing the original transistor amplifier topology for minimum static %THD+N. His reductionist shears have cut too far. I understand the profound satisfaction he gets from steady linearity, and it may have been smart electronics two decades ago when the Space Shuttle was on the drawing board, but today, the real audio design process hasn't even begun.

By the time a truly modern design (likely employing one of the other, more recent, principal topologies with very different rules) has been optimised according to listening tests and %IMD, %DIM and the various AP, Belcher, NMR and other SOTA DSP-based measurements, and after real world requirements like packaging and making it abuse proof have been attended to, then the %THD+N residue will likely be adjusted away from its minima.

Ben Duncan
Lincoln

Diagram as a language

I agree with your editorial remarks in the March issue. It is indeed a pity that the circuit diagram is so rarely accorded its true status as a powerful language.

It can be used at many different yet compatible levels from the briefest block diagram representing a whole system at one extreme to a complete schedule of parts at the other.

Even the final board layout is no more than a topological transformation of the same set of relationships.

Circuit sketches enable instant communication of ideas in a manner that has no equal.

Attempts to replace the concept are to be welcomed if they offer at least the same virtues, but so far have been no more successful than those intended to replace the similarly highly developed and subtle representation of music by a printed score.

RH Pearson
Bourne, Lincs

Sagnac sensors

It may be relevant to mention that the Sagnac effect (1913), referred to by Gerardus Bouw (*EW + WW*, March), is the basis of the laser gyro and other rotation sensing devices.

A laser beam is split into two parts, which are sent in opposite directions round a common optical transmission circuit. Rotation of the structure causes an apparent lengthening of the transmission path in one direction relative to the other (actually a transmission time difference).

This changes the interference pattern between the two sets of coherent light waves and the result is sensed by a photoelectric detector.

Tom Ivall
Staines, Middlesex

c the light

Martin Berner's question (*EW + WW*, March) on relative speeds versus the light barrier can be best understood by considering the velocity of a particle in three dimensions.

If the particle travels at *c* in each direction the classical result would be to add the squares and take the square root, resulting in a speed faster than *c*.

By contrast the correct relativistic calculations result in the overall speed being *c* – effectively one can travel at *c* in each direction and still only be travelling at *c*. Also there is nothing wrong with having two observers whose relative velocity exceeds *c*, the result will only be that they cannot communicate with any device limited to *c*, like a photon.

Changing subject, AJ Quinton (*EW + WW*, March) is on the right lines with his ideas about linking permanent magnets and superconductors, but misses the point in connecting with room-temperature effects.

Magnetism is caused by charges in motion, and to have a north and south requires an area to be traversed by such charges. If one takes the simplest area – that of the circle – and the simplest particle with a magnetic field – the electron, then I would argue that the electron consists of a number of smaller particles, all chasing each other around the circumference of a circle. This would be, in effect, a permanent magnet and a superconductor at any temperature.

Mike Lawrence
Montgomeryshire

O frabjous days

I am surprised by the neo-rejectionism of modern physicists put forth in your magazine.

Anyone who listens to Lewis Carroll, Monty Python and Douglas Hofstadter will most likely contend that everything is screwed up nowadays and that there must be something wrong with a scientific theory that doesn't contain some seemingly paradoxical feature.

This reaction coincides with the closure of all Israeli universities until the end of this semester because the new so-called socialist government tried to reduce the salary of professors, who of course were too intelligent to accept this offer.

Some loony Chabadnik named Amnon Goldberg (I hope he's not a relative of Barukh Goldstein's because he might kill me for this) insinuates that many attempted proofs of the theory of relativity have been shown to be based on fudged evidence. Well I suppose that if physicists keep eating fudge while making measurements, the chewing activity and high sugar content of the fudge may distract their attention from the correct adjustment of their instruments.

Fortunately for physics, no experiment is taken for granted by the scientific community unless it is repeatable in all laboratories.

You are certainly prejudiced if you think that Newton's theory of gravitation has been proven to be universally valid, or even that this validity isn't inflicted with paradoxy.

I was once told by a nuclear physicist that Newton's formula conflicts with the hypothesis that the universe is uniformly filled with matter. If a stone is dropped, the gravitational attraction from stars can certainly be neglected by approximation, and the more distant they are the less their influence.

But near space is not typical of the universe in general. If spherical shells are considered, the mass contained in those shells grows quadratically with distance while gravitation grows inverse quadratically. The net effect of these shells when integrated over the entire mass contained in the universe is to counteract the interaction of the system of the stone and the Earth almost entirely so that the observed constant of gravity must be less than the real by a factor of 50, as was already calculated by Laplace (or was that Poincarré).

He made the same argument re the dispersion of sunlight in the universe. I have been unable to find these statements anywhere in university textbooks on physics, but as it probably is correct, Einstein may very well have had this in mind when he devised his general theory of relativity.

Michael Williams
Beth Shemesh, Isreal

Appalling editor

I was appalled when reading Frank Ogden's comments (*EW + WW*, March) on the use of VHDL and other logic design languages.

Contrary to Ogden, I believe electronics design is about solving real problems for real people, and not about whether one uses schematic entry or wire diagrams.

What Ogden fails to understand is that it is exactly the high level of abstraction possible with modern logic design tools that facilitate creativity in the design process, by allowing designers to concentrate on the problems they are trying to solve, rather than attempting to trace out wires between numerous gates and flip-flops culled from the manufacturers' data books.

He then continues to take a snipe at the using of pre-defined modules in a design, rather than being creative and designing it yourself. I then have to ask what the difference is between using an op-amp or a TTL flip-flop in a design or using an algorithm for implementing the same function on a piece of silicon.

The third point, about circuit complexity, is not very well thought out. The whole reason for moving away from schematic entry is because most humans (even real designers) can only comprehend about five to eight functional blocks at the same time. By implementing functional blocks of high complexity as high-level statements, one reduces the amount of information necessary to interpret the function of the design. Wires and gates convey almost no information about what the design actually does, as anyone getting stuck with a broken state machine of medium complexity will testify.

Regardless of what Ogden claims, the way electronics designers think and work is changing. We will continue to use the best tools available to us, simply because it allows us to design better, faster and more reliable systems.

Frithiof Jensen
Abingdon, Oxfordshire

Short agreement

Regarding your editorial comment on VHDL in March issue: Here! Here!

Richard Ashwell
Newbury, Berkshire

Life is easier with Dos

Jason Ross (*EW + WW*, April) claims naive users have less to remember with Windows and will find it much easier to use than Dos.

In fact a well designed Dos system is far easier to use. It also avoids the confusing and cluttered Windows screen and the difficulties of memorising the meanings of small differences between numerous icons. Icons may be fine for work involving a narrow range of software but are hardly ideal for users with wide and complex interests.

My systems use a single small control file M.bat – I find 8K sufficient. This automatically displays a full screen descriptive menu of 36 subject groups – on booting or at any time from any point in the system by typing M.

This method potentially eliminates the need to remember or type any file names or Dos commands.

A single character – figure or letter – identifies each menu item. Typing the chosen item – for example M B – either runs a program, performs a system operation, lists a full or partial directory tree, lists a directory in alphabetical order or displays a secondary menu with similar facilities. Menu choices may combine these and other possibilities in most combinations.

The method allows the user an efficient structure to describe any requirement simple, specialised or highly complex.

Secondary menus – with two character codes – permit more than 1,000 choices. Tertiary menus – using three characters – expand to 30,000 or more.

Major uses obviously go into the primary directory. The user has no need to remember any details but may at any time go direct to any point in the whole menu tree by typing the appropriate short menu code in full.

Joel Sciamma (*EW + WW*, April) asks why bother about saving a few milliseconds. The greatly increased speed of systems not using a gui can allow real time operation for many program operations, without killing time waiting for completion. At the other extreme gui slowness may mean the expense of extra computers to meet a work load.

RG Silson
Tring, Herts ■

NEW PRODUCTS CLASSIFIED

ACTIVE

Asics

3V embedded array. Toshiba's new 0.5μm cmos embedded array combines the functionality and performance of standard cells with the faster turnround times of gate arrays. With a typical gate propagation delay of 0.25ns and gate power consumption of 2.2μW/MHz, *TC180E* outperforms 0.8μm 5V asics on 3.3V supplies. The device contains up to 340,000 usable gates. Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

Multi-product wafers. A capability known as Multi-Product Wafer (MPW) Train in 0.8, 1, 1.2 and 2μm cmos and 1.2μm BiCMOS is being used by Austria Mikro Systeme to allow the processing of several different devices on one wafer. Development charges are thereby reduced by up to 50% and AMS points out that, at a small extra cost, customers can have several versions of a design at once, saving waiting time for evaluation. AMS takes a tape from the customer and delivers finished, packaged parts. Austria Mikro Systeme International. Tel., 0276 23399; fax, 0276 29353.

A-to-D and D-to-A converters

Stereo codecs. *AD1847* and *AD1846* single-chip, sigma-delta, stereo, digital audio codecs from Analog both support the Windows Sound System 2.0 and Compaq Business Audio. The 1847's serial port allows direct interface to a DSP or system i/o chip, needing fewer pins than a parallel port and costing less. AD1846 is a reduced-cost version of the AD 1848K industry standard Soundport and is pin and register-compatible, offering up to 70dB of dynamic range. Both devices provide CD-quality audio on an ISA or EISA add-in card. Analog Devices Ltd. Tel., 0932 253320; fax, 0932 247401.

Discrete active devices

Low-resistance power fets. Two new TO-220 Hexfet power mosfets by International Rectifier, the 30V *IRL2203* and the 50V *IRL3705* offer $R_{ds(on)}$ values of 10mΩ and 12mΩ respectively, both devices operating from a 5V logic-level drive as well as the standard 10V. International Rectifier. Tel., 0883 713215; fax, 0883 714234.

Dmos power mosfets In the 14-member *NDS9XXX* family of dmos power mosfets from National Semiconductor are single and dual n-channel and p-channel types and complementary n-p dual mosfets. All are in SO-8 surface-mounted packages and the devices are pin-to-pin compatible with other SM dmos mosfets, but with better performance at 4.5V V_{gs}. National Semiconductor. Tel., 01049 8141 103300, fax, 01049 8141 103515.

Switching diode. A fast silicon epitaxial switching diode from ITT, the *BAL99*, has a leakage current of only 2.5μA at 70V and a recovery time of 6ns. Forward voltage at 1mA is 0.715V and 1.25V at 150mA. Power dissipation is 350mW at up to 25°C ambient. The diodes are designed for automatic insertion. ITT Semiconductors. Tel., 0932 336116; fax, 0932 33148.

Linear integrated circuits

3V low dropout regulators. Four new low-voltage, 100mA micropower, low dropout regulators from National feature 380mV dropout voltage at full load. *LP2950-3.0* and *-3.3* are fixed-voltage, three-terminal types, the *LP2951-3.0* and *-3.3* being adjustable and fitted with 3V and 3.3V taps, which avoid external resistors. Quiescent current at light loads is 75μA and there is a shutdown facility and an error flag to indicate when the regulator falls out of regulation by over 5%. National Semiconductor GmbH. Tel., 0104 981 4110 3382; fax, 0104 981 4110 3515.

Precision references. 1.26V, 2.5V and 5V (Refs 12, 25 and 50) micropower references by GEC Plessey use the bandgap principle and thereby avoid the need for an external shaping capacitor. Knee currents are 40μA and 80μA, battery operating power 113μW, 150μW and 300μW and initial voltage tolerance ±1%. Gothic Crellon Ltd. Tel., 0734 788878; fax, 0734 776095.

3V RF amplifiers. 1.9GHz, 3V amplifiers in NEC's *μPC2745T-2749T* range consume 50% less power than others available, running from a supply as low as 1.8V. They are meant for use in receiver buffer amplifiers and exhibit a noise figure of 2.8dB at a gain of 9 to 21dB. Packaging is a six-pin mini-mould measuring 2.9 by 1.5 by 1.1mm and are made in NEC's 20GHz NESAT III silicon bipolar process. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

Variable-gain amplifier. *CLC522* by Comlinear is a DC-coupled, two-quadrant multiplier with differential voltage inputs and a single-ended voltage output and forms a complete variable-gain system containing two input buffers and an output op-amp. Signal channel and gain-control bandwidth is 165MHz at a gain of 10, maximum gain is set externally over the range 2 to 100, the gain control giving more than 40dB variation. At the smallest maximum gain of 2, bandwidth is 350MHz and gain control non-linearity 0.5%. Comlinear Europe Ltd. Tel., 0203 422958; fax, 0203 422961.

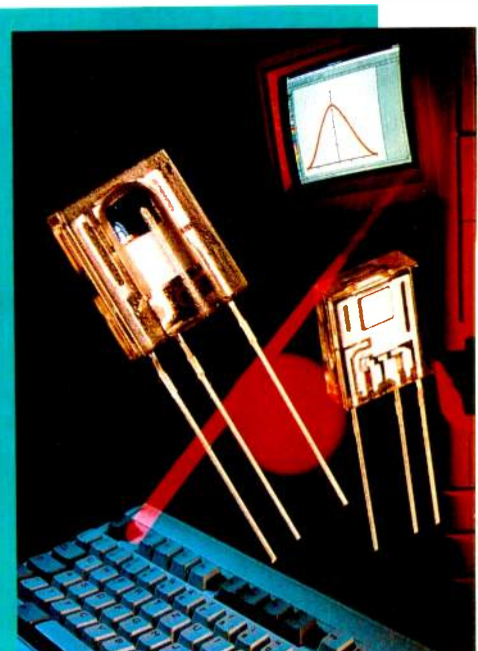
FM IF amplifiers. Sony has a family of low-voltage, low-power FM IF amplifiers for cordless and cellular telephones. The newest model is the *CXA1683M/N*, which is a 100MHz wide bandwidth double balanced mixer design for cordless telephones operating on 1.8-to 6V DC. The chip has a programmable low-pass filter at the detection output stage, a squelch filter operational amplifier and a signal strength indicator with a 70dB dynamic range. Typical current drain is 4mA from 2.3V. The *CXA1293M/N* is similar, but with features for cellular telephones. Sony Semiconductor Europe. Tel., 0256 478771; fax, 0256 818194.

3.3V references. Knee currents down to 15μA, with a temperature

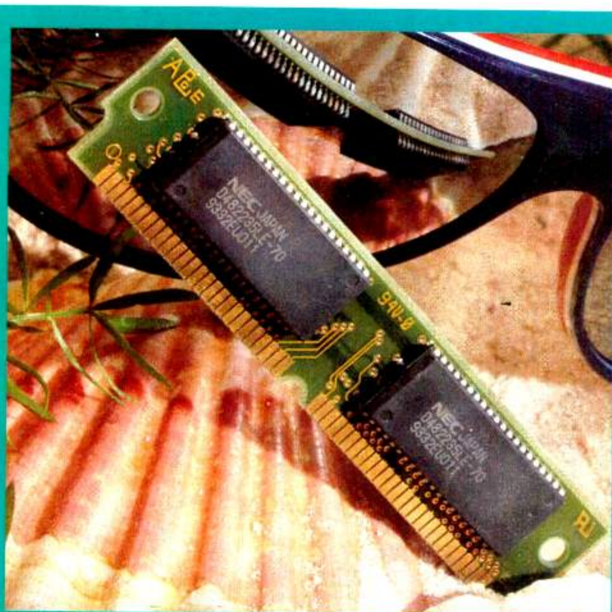
Linear integrated circuits. Two new voltage-feedback op-amps from National, the *LMC6572* and *LMC6574* dual and quad devices, were released early in the year and are intended for use in current-to-voltage conversion or amplification in mobile radio or instrumentation. They offer 2.7V and 3V operation at 40μA for each amplifier, an input current of 20fA and 120dB gain on a 2.7V rail.

Theses amplifiers are therefore well suited to battery power and will cope with either 3.3V digital logic regulated supplies or unregulated rails. Since the output swing is virtually rail-to-rail, within 20mV of the supply with a 5kΩ load, enhanced signal-to-noise ratio and dynamic range make the devices highly suitable for interfacing with A-to-D converters. They will drive heavy loads and their high voltage gain reduces the need for cascaded amplifiers, resulting in better accuracy. Both amplifiers have guaranteed performance over the –40°C to 85°C temperature range and are available in 8-pin and 14-pin plastic dips and 8-pin or 14-pin plastic SOICs. National Semiconductor GmbH. Tel., 01049 814110382; fax, 01049 814103515.

Low-power Rx IC. Temic Telefunken has a family of low-power UHF receiver ICs intended for car security, wireless lans and appliance control that take a standby current of only 1mA. The *U 431X B* series operate at 433MHz. A modulated carrier is converted by a UHF stage to 10.7MHz IF and amplified by the U 431X B, its log. amplifier acting as demodulator for AM, FM being handled by a quadrature detector. Digital data is regenerated in the baseband by a clamping comparator and op-amp, the universal output interfacing to a decoder. Temic Telefunken GmbH. Tel., 01049 7131 672747; fax, 01049 7131 993342.



Please quote "Electronics World + Wireless World" when seeking further information



Video ram simm. Video ram *MC-250* is a 256K by 32bit device made up of four *µPD482235* video rams in a simm format, with two decoupling capacitors for each device, the whole module measuring 82.5mm by 17mm. Each ram has a random-access port and a serial read/write port connected to an internal 16384-bit data register. The device has a split serial buffer, so that data is transferred into one half as serial data is being read from the other half, avoiding flyback noise or flicker in video systems. All i/o is at TTL levels. Sunrise Electronics Ltd. Tel., 0908 263999; fax, 0908 263003.

coefficient of 15ppm/°C are obtained from the Zetex *ZRC330* 3.3V precision references, which use a bandgap technique needing no external capacitor. Operating current is 20µA-5mA and the device tolerates temperatures from –40°C to 85°C. Since the device is based on an asic, metal mask programming allows Zetex to offer any voltage between 2V and 10V and voltage trimming does not affect temperature coefficient trim. Zetex plc. Tel., 061 627 5105; fax, 061 627 5467.

5GHz dual differential amplifier. A dual amplifier for RF and IF down and up converters, Harris's *HFA3102* is a dual long-tailed pair with a transistor in each tail, giving a power gain-bandwidth product of 5GHz. Current gain is 70 and is matched between the halves to within 10%. Noise figure is 3.5dB, collector cutoff current is 10nA and collector leakage current 0.01nA. The IC is pin-compatible with other available devices. Spice models are available. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

Infrared receivers. Infrared communication inside offices, say between computers and printers, is the province of Temic Telefunken's new receiver modules, which are required to work in conditions of high visual and electrical interference. *TFMS 5..0* and *TFMT 5..0* operate on carriers between 30 and 56kHz, special versions being made for 20 to 60kHz. A pin diode detects the carrier and is followed by signal processing and a Schmitt output. Shielding by the housing reduces the effects of both visual and electrical interference on paths of up to 40m. Temic Telefunken GmbH. Tel., 01049 7131 672747; fax, 01049 7131 993342.

Battery charger/monitor. The Microchip *TrueGauge MTA11200* monitors the state of health of rechargeable batteries and recharges them. The device provides information on remaining capacity measured during discharge only, total capacity, voltage, current and temperature, a single-wire to the host microcontroller allowing a module with a *Truegauge* to stay in the battery pack. There are 35 programmable system parameters from battery warning levels to end-of-discharge voltage, and the device is optimised for different battery types by specifying negative delta peak detection and protection against thermal overcharge, time-out overcharge and peak voltage. Development software is available. Polar Electronics. Tel., 0525 377093; fax, 0525 378367.

Femtoamp-input op-amp. Guaranteed input current of National's *LMC6001* op-amp is 25×10^{-15}A, other characteristics including an offset voltage of 350µV, 10µV/°C maximum drift and input-referred noise of 22nV/ Hz, which allows a better signal-to-noise ratio than jfet electrometer amplifiers. Versions are available with input currents of 25fA and 100fA. Thame Components Ltd. Tel., 0844 261188; fax, 0844 261681.

Logic building blocks

Fibre-channel transceiver. METL's RCC700 is a cmos transceiver, operating at up to 265Mbaud, which integrates 8B/10B encoder/decoder, serialiser and deserialiser, a PLL synthesiser, a PLL clock and data recovery and a byte alignment circuit. PCB evaluation assemblies are available for fibre or coaxial interfaces. Microelectronics Technology Ltd. Tel., 9844 278781; fax, 0844 278746.

Memory chips

Fast srams. Cmos srams in the Alliance Semiconductor *7C256* 256K series are fast – 12 to 25ns – and sparing of power, maximum active power in the 15ns type being 605mW. Standby power levels as low as 1.1mW are achieved and data is retained down to 2V. Supply is 5V or 3.3V. Hunter Electronic Components Ltd. Tel., 0628 75911; fax. 0628 75611.

Microprocessors and controllers

Embedded 386. AMD is offering the *Am386* processor in two embedded versions, the *Am386DE* and *Am386SE*. Any X86 PC can be used as a development platform, the *FusionE86* support program providing software and hardware from more than 60 suppliers, including Microsoft's Microsoft At Work operating software. Both versions work at either 3V or 5V, with standby modes. *Am386DE*, the 33MHz type, is available now, the SE following in a few weeks. Advanced Micro Devices (UK) Ltd. Tel., 0483 740440; fax, 0483 756196.

16-bit V-series micro. NEC's *V55PI* is a 16-bit, single-chip microcomputer with 16Mbyte of address space and 64Kbyte i/o space; the CPU is 8086-compatible, so that existing software is usable. The CPU executes instruction about twice as quickly as the V35 and higher-performance peripherals include two uarts, operating at 780kbit/s asynchronous and up to 3.125Mbit/s clock synchronous, and an 8-bit parallel interface configures as either a general-purpose 8-bit i/o port or as a Centronics type giving high-speed bidirectional parallel communications as a slave or as a driver. There are also a four-channel DMA controller, a 4-channel 8-bit A-to-D converter, a PWM operating up to 24.4kHz, a watchdog timer, four 16-bit timers and a 16-bit software interval timer. Sunrise Electronics Ltd. Tel., 0908 263999; fax, 0908 263003.

32-bit risc microcontrollers. The *SH* series of microcontrollers by Hitachi use a 32-bit risc core, the first to be available operating with cycle times of 50ns on a 5V supply or 83ns on 3.3V. Most instructions execute in a single cycle using a 5-stage pipeline to give a 16 Dhrystone Mips performance. Other features include 16, 32-bit wide general-purpose registers and a hardware multiplier performing 16 by 16-bit plus 42-bit multiply and accumulate operations in 100 to 150ns. Current consumption is 100mA at 20MHz and 40mA at 12.5MHz for the 3.3V type, several sleep modes being provided. First models to appear are the *SH7032*, which is romless with 8Kbyte of on-chip ram, and the *SH7034* with 64Kbyte of rom/prom and 4Kbyte of ram. Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

Mixed-signal ICs

MPEG2 decoder. Toshiba's new single-chip video decoder meets the MPEG2 standard and decodes compressed 1153 by 1024 line, 30 frames/s digital signals with HDTV-level resolution. Parallel decoding in the variable-length decoder and the risc microprocessor allows both to decode data bit streams at the same time to provide the necessary speed. Toshiba Electronics UK Ltd. Tel., 0276 694600; fax, 0276 691583.

Optical devices

Miniature leds. Seven-segment leds in HP's new *HDSP-U* series have grey or black surfaces for better contrast and provide an 8mm character height in a 11mm by 7.1mm by 5mm package. Red, orange,

Mixed-signal ICs

Consumer chips. ITT's *Digital Systems 3000* is effectively the next generation of digital consumer electronics, building on the earlier, self-contained, television-oriented *DIGIT2000* system of ICs to become a structure for virtually unlimited application in multimedia systems. Most *Digital Systems 3000* ICs are self-contained and combine analogue and digital techniques.
The system copes with functions beyond the current consumer area: services such as telefax, videotext and modem functions take the system into telecommunications and picture compression and transmission, computer audio and graphic user interfaces link consumer electronics, telecomms and computers in the area of multimedia services.
Since all the modules are small and powerful and are produced by the same technology, tailor-made, one-chip implementations for special applications can be made relatively cheaply.
The company says it is ready to put into practice the idea of single-chip systems for the new, extended area of consumer electronics. For example, ITT is able to integrate several DSP kernels and their ram and rom on one chip, whereas the existing practice is to use an MCU and an interface chip, taking up more space and requiring more extensive interfacing. ITT Semiconductors. Tel., 0932 336116; fax, 0932 33148.

yellow or green character colours are available at luminous intensities of 480mcd for yellow to over 1000mcd for the red and green. Hewlett-Packard Ltd. Tel., 0344 362277; fax, 0344 362269.

Laser diode for optical memory. Intended for use with optical disk memory or memory cards, Mitsubishi's *HS01* high-power HF Super-Imposition laser diode operates from 5V and is in a steel packing-in-shield case for low noise; RIN figure is –125dB/Hz. Maximum output is 35mW continuous and 45mW pulsed. Mitsubishi Electric UK Ltd. Tel., 0707 276100; fax, 0707 278692.

Two-colour photodiode. The *Tandem* two-colour photodiode detector combines a 2.5mm diameter silicon diode and a 1mm diameter InGaAs diode in a single TO-5 package to provide sensitivity over the 400 to 1700nm spectral range, each detector being individually addressed. It can be optimised at 820, 880 and 940nm. Different InGaAs diode diameters can be specified. Aerotech World Trade Ltd. Tel., 0628 34555; fax, 0628 781070.

Oscillators

Clock oscillators. *Raltron C04810* and *C04910* clock oscillators are only 2mm high, with a footprint of 7.5mm by 5mm. Frequency stability is 50ppm and power consumption of the 04810, depending on frequency, is 20 to 35mA, driving 10 LSTTL loads or 15pF mos. The C04910 drives 10 TTL loads or 50pF mos, but still takes only 25 to 60mA. Frequency range is up to 60MHz, a 3.3V version being also available. The *H-13* is a low-profile SM device, with a height of 1.3mm and consuming 0.1mW. Flint Distribution. Tel., 0530 510333; fax, 0530 510275.

Power semiconductors

High-voltage mosfets. Motorola has upgraded its entire 400 to 600V range of power mosfets to 800V and 1000V, using a new technique in which multiple rings provide a better field shaping towards the edge of the mosfet die. This gives improved voltage blocking to protect against surface charges at high voltages. As examples, *MTY10N100E*, a 1000V type, has an $R_{ds(on)}$ of 1.3Ω at 10A and the 800V *MTP4N80E* has a 3Ω $R_{ds(on)}$ at 4A. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

1500W suppressor. Semtech's *SMCJ* transient voltage suppressors are either unidirectional or bidirectional and are in a DO-214AB plastic package. Response time is 1ps and forward surge rating 200A. Having a voltage range of 5 to 170V, the devices have a steady-state power dissipation of 3W (1500W peak). Semtech Ltd. Tel., 0592 773520; fax, 0592 774781.

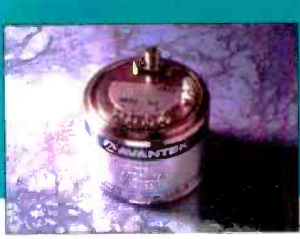
Fast power diode. Ten members of Harris's *Hyperfast RHR* series of power diodes exhibit reverse recovery times of 60 to 100ns. Three of the devices have breakdown voltages of 1200V and t_{rr} of 75ns at 30A, one of them being offered as a dual device. Two 75A types are 1200V rated and have t_{rr} of 100ns; three lower-voltage types recovering in 60ns. A 150A type breaks down at 1000V, with a t_{rr} of 100ns. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

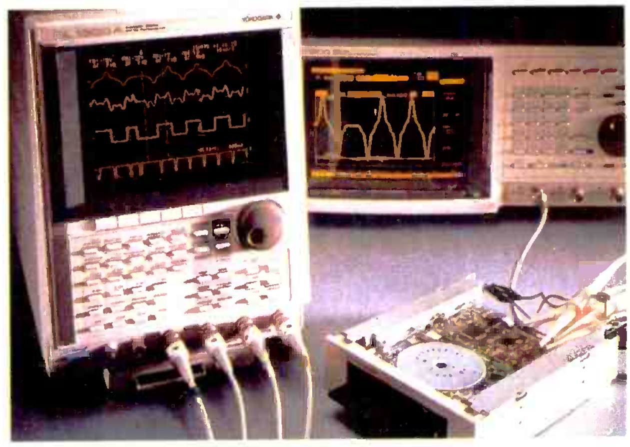
Reverse blocking switch. Replacing two mosfets and their drive circuit, the Siliconix *Si9718CY* is a reverse blocking switch for battery disconnect application in dual-battery notebook computers, allowing the computer to switch from one battery pack to the other before cells completely discharge. A new process eliminates the parasitic diode found in standard mosfets, reducing on resistance to 80mΩ at 3.5A. The device also includes a charge pump and enable circuitry, and undervoltage lockout protects the system. Siliconix/Temic Marketing. Tel., 0344 485757; fax, 0344 427371.

Low-power PWM controller. Unitrode claims a first for its *UCC3570* BiCMOS voltage-mode PWM controller for use in isolated, high-frequency switched-mode power supplies. It features an 85μA start up current, 1mA run current and the ability to drive a 1A mosfet gate at up to 500kHz. Using voltage feedforward, the device responds accurately and within one clock cycle to wide line-voltage variations and is not noise sensitive. Unitrode (UK) Ltd. Tel., 081 318 1431; fax, 081 318 2549.

300W transient suppressors. Semtech's *SM* series of surface-mounted silicon transient voltage suppressors are low-cost devices for data line or supply rail use, having a peak pulse power of 300W, a response time of 1ps and unidirectional or bidirectional operation, depending on type. Voltage range is 5 to 24V, breakdown 5 to 26.7V minimum, clamping voltage at 5A, 11 to 55V and leakage current 1 to 100μA. Semtech Ltd. Tel., 0592 773520; fax, 0592 774781.

40GHz YIG-tuned oscillator. Avantek, an HP subsidiary, offers the *AV-20040*, which it believes to be the first YIG-tuned oscillator to cover 20 to 40GHz without filters, multipliers or amplifiers to produce a minimum of 10dBm, with –60dBc spurious output. It is hermetically sealed and operates from a single 15V supply. Frequency drift is 60MHz maximum. Hewlett-Packard Ltd. Tel., 0344 362277; fax, 0344 362269.





Waveform generator. Yokogawa's *AG1200* arbitrary waveform generator captures and reproduces signals from digital storage oscilloscopes. It interfaces to the DSO via a GPIB port and has a 3.5in floppy drive for waveform storage. Waveforms loaded from the DSO are reproduced, displayed and output directly, or modified and edited on screen. Output is 4-channel with a 10MHz clock and 12-bit resolution, instruments perhaps being combined to give 16 channels or channels combined to provide 32-bit patterns. Waveforms are also produced from functions and/or generated by the Scope Draw technique in which the waveform is sketched and edited on screen. Martron Instruments Ltd. Tel., 0494 459200; fax, 0494 535002.

PASSIVE

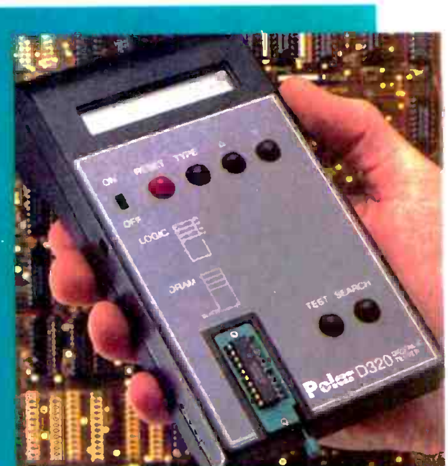
Passive components

Crossover electrolytics. Bipolar electrolytic capacitors in Nichicon's *DB.GB* series are meant for use in audio crossover networks and have values from 1μF to 68μF in ±20% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current 3μA and allowable ripple 205 to 1200mA (DB), 760 to 1120mA (GB). The radial-lead components resist most halogenous cleaning fluids. Nichicon (Europe) Ltd. Tel., 0276 685393; fax, 0276 686531.

High-temperature ceramic capacitors. Vitramon's *H* range of chip capacitors uses the X8R dielectric, which has the stability of X7R but withstands temperatures up to 150°C, making them suitable for unfriendly environments such as under the bonnet of a car. Sizes 0805, 1206, 1210, 1812 and 2225 are available in values from 470pF to 1μF at 50V DC and in tolerances from ±5% to ±20%. Insulation resistance at 25°C is 100GΩ and a minimum of 10GΩ at 150°C. Vitramon Ltd. Tel., 0628 524933; fax, 0628 525435.

Vertical inductors. In values of 0.47μH to 1000μH at ±10% standard tolerance, *microSpire MIVI-PK 0455* series of vertically mounted inductors save board space by offering diameters of 5 to 13mm and a height of 11 to 17mm. Self-resonant

Hand-held IC tester. Capable of testing 74/74LS/74HC TTL and 4000 series cmos logic devices and 41/44 series dynamic rams, the *Polar D320* is hand-held and is powered by a 9V PP3 cell. More than 350 devices are in its built-in library and it can be switched into a search mode to compare an unknown device with available data. Components are placed in a 20-pin zif socket. Polar Instruments Ltd. Tel., 0481 53081; fax, 0481 52476.



Please quote "Electronics World + Wireless World" when seeking further information

frequencies are 2.5MHz for the 1000μH version up to 145MHz for the 0.47μH type. Surtech Interconnection Ltd. Tel., 0256 51221; fax, 0256 471180.

Multilayer ceramics. Multilayer ceramic capacitors in AVX's *CM* series are in 0603, 0805 and 1206 cases sizes at working voltages of 16V. Values up to 100nF for the 0603, 470nF in the 0805 and 1μF for the 1206 types are available, all with leakage currents given by an insulation resistance of over 10GΩ or 500MΩ/μF, whichever is the smallest. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Resistor networks. Made to the customer's specification of values, tolerance and circuit configuration, *Beyschlag* surface-mounted resistor networks are available in quantities down to 100 pieces. The thin-film resistors are the Beyschlag Micro-MELF type, welded to a metal frame which is shaped to form the circuit. Values in the range 10Ω to 2MΩ are used to tolerances of 1% or better and with temperature coefficients of 50ppm/°C or better. The networks are said to possess a high pulse load capability. Flint Distribution. Tel., 0530 510333; fax, 0530 510275.

Connectors and cabling

Adaptable board connector. 65,536 different connection patterns are possible in a housing smaller than a dil package with the *Erg* system. The two-part component is a pin header and matching jumper block which is selectively loaded to obtain the required pattern of contacts. The multi-jumper block fits over the 1-16 three-pin rows of gold-plated header pins, enclosing and protecting them and setting up the interconnection pattern. Standard blocks are fully loaded, but Erg can supply any pattern to order. Blocks are either polarised or reversible. Erg Components. Tel., 0582 662241; fax, 0582 600767.

Filters

Notch filters. A new approach to temperature compensation in cavity filters has been adopted by Wainwright Instruments GmbH in Germany in their range of ultra-stable notch filters for GSM, PCN and other mobile comms systems. Drift is less than ±1ppm/°C. The technique has also been applied to filters for other frequencies and characteristics. Wainwright Instruments GmbH. Tel., 01049 8152 2245; fax, 01049 8152 5174.

Hardware

Conductive keypad. The Grayhill *Series 90* conductive rubber keypad is now available from Highland. The pads are available in 3 by 3 and 4 by 4 forms with matrix circuitry and are rated at 12V DC and 5mA for 0.5s. Contact resistance is 100Ω and bounce time less than 12ms. Highland Electronics Ltd. Tel., 0444 236000; fax, 0444 236641.

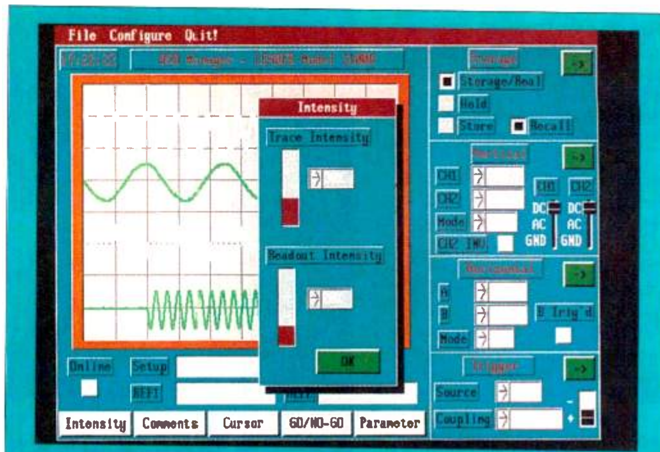
Shielding can. Made in tin-plated steel, West Hyde's *Isolator* is a board-mounted shield attached to the board by four solder prongs set on 0.1in pitch. The one-piece unit is fully soldered on the internal faces to provide efficient sealing. It comes in a range of sizes and can be made to order in seven days. It is also available at lower cost without soldering. West Hyde Enclosures. Tel., 0453 731831; fax, 0453 886637.

Computer peripheral

Contactless smart card. Measuring 43mm by 54mm and 5mm thick, Mitsubishi's new contactless *MelCard* is tough enough to stand wear and will withstand being made into a key fob. The cards operate a read/write device linked to a computer network for data i/o, read and write operations, the device automatically transmitting data at up to 455kHz; the frequency can be customised for specific applications such as operating a ticket barrier or recognising parts in a factory.

Cards are read at a distance up to 800mm in less than 0.2s. They contain a single-chip, 8-bit microcontroller which uses little power, the use of sram for user memory also takes little power and enables high-speed read and write. RF transmission avoids the problems associated with dirty contacts and facilitates volume throughput.

The cards have been thoroughly tested while being bent and twisted at temperatures from –20°C to 60°C and, should the occasion arise, they will survive being immersed in 1m of water for 30 minutes. Mitsubishi Electric UK Ltd. Tel., 0707 276100; fax, 0707 278692.



PC oscilloscope control. *DSO Manager* is a package by FemtoTek for the control of a Leader Instruments 3100D digital storage oscilloscope from a PC, using a GUI and a mouse. It is based on National's *LabWindows* software and comes as the software only or as a complete system including a GPIB interface board. All the oscilloscope functions are settable from the PC and waveforms can be stored on and retrieved from disk, plotted on a printer and sent to Ascii files for later use. National Instruments UK. Tel., 0635 523545; fax, 0635 523154.

Instrumentation

GPIB test gear. Thurlby Thandar has additions to its range of rack-mounted, GPIB test equipment, including power supplies, multimeters, generators, logic analysers and oscilloscopes, which are in 2U cases. All instruments conform to IEEE-488.2 in addition to IEEE-488.1. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Loop-powered process meters. *Models 81 and 82* 3.5 and 4.5 digit process meters are powered by the 4-20mA current loop process signal, offset and scale values being adjustable to allow the use of a range of engineering units. Accuracies are ±0.1% and ±0.02% respectively and the signal input can be set to 4-20mA or 10-50mA. Panel-mounting hardware is included. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 0273 570215.

Clean RF generators. Designed to be used as either bench-top instruments or as part of an ATE system, the Giga-tronics *606X* series of RF signal generators covers the 100kHz-2.1GHz range in three models; *6060/1/2*, the 60602 having fast-rise pulse modulation for transponder and radar testing. As standard, a 10MHz crystal oscillator ages at <±5X10^{-7}/month and warms up in an hour. Alternatively, an optional oven-controlled reference ages at <5X10^{-10} and warms up in 30min. Sematron UK Ltd. Tel., 0734 819970; fax, 0734 819786.

Micro-based thermometers. ETI's *Microtherma* thermometers are lightweight, handheld instruments based on microprocessors to allow continuous and automatic recalibration. Temperature range is –200°C to 1370°C with either 0.1°C or 1°C resolution. Each has a 4-digit display with open-circuit and low battery indication. Type K,J,T, E or N sensors can be selected by the keypad, and there is a conversion facility from celsius to fahrenheit. Electronic Temperature Instruments Ltd. Tel., 0903 202151; fax, 0903 202445.

Auto timebase oscilloscopes. Hitachi's *V1065* and *V665*, 60MHz and 100MHz oscilloscopes, in a new family of compact instruments, provide automatic timebase setting, with manual override, and on-screen display of A and B sweep time, delay time and hold-off. Channel sensitivities are also displayed and the instruments have a cursor readout for voltage, time difference and frequency. Two models in the range feature a 4-digit frequency meter and all models have delayed sweep and autotrigger in the usual modes. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Vibration analyser/data collector. A technique developed by SKF, spectral emitted energy technology (SEE), is used in the *Microlog CMVA10* single-channel vibration analyser to give a true multi-parameter capability. The technique is an acoustic enveloping process that helps to detect rolling element defects and predict bearing failure at an early stage. The instrument has 4Mbyte of on-board ram to allow the use of a multi-plane balancing routine in the unit. It is also compatible with the PRISM computer-aided condition monitoring software. Endevco UK Ltd. Tel., 0763 261311; fax, 0763 261120.

Lan cable tester. For the use of lan installers and maintenance people, the *Fluke 650 Cablemeter* tests either coaxial cables or unshielded twisted pairs in both Ethernet and Token-ring lans. A single function knob and a four-line display, together with display and storage of up to 50 results, make for ease of use. It tests a circuit's characteristic impedance and will monitor Ethernet traffic and measure inherent noise over time to locate intermittent trouble. An autotest feature performs a set of tests . compares the results against standards and gives indication of a problem by a tone and fail message. Jensen Tools. Tel., 0604 787060; fax. 0604 785573.

Literature

Audio products. Crystal's complete range of audio ICs is described in a 1072 page book. Information includes all specifications. block diagrams, theory of operation, applications and schematics. Devices detailed are data converters and codecs. DSPs and synthesiser, digital audio transmitters and receivers and volume controls. Crystal Semiconductor Corporation. Tel., 0101 512 442 7555; fax, 0101 5512 445 7581.

Power supplies

Rechargeable 996. A direct replacement for the 996 dry battery, Power-Sonic's 6V, 5Ah sealed lead-acid battery type *PS650* can be recharged up to 1000 times and includes an overload self-resetting cut-out. The company now has more than 100 models of SLA rechargeables with capacities in the range 0.5Ah to 4800Ah. Power-Sonic Europe Ltd. Tel., 0268 560686; fax, 0268 560902.

Radio communications products

GaAs fet amplifiers. Multi-octave and narrow-band GaAs fet amplifiers by Lucas covering the 0.1-40GHz range are based on a library of modules for rapid development and modification and low cost. They use mesfet or hemt devices and most of them incorporate temperature compensation. Many are in the low-cost quasi-MMIC form, in which all passive components are integrated and the active devices bonded separately. Custom designs incorporating limiters, special filters, variable gain, etc,. are undertaken. Connection options include standard waveguide, coaxial and microstrip. Anglia Microwaves Ltd. Tel., 0277 630000; fax, 0277 631111.

Switches and relays

Solid-state relays. *Series RA* relays from Teledyne handle loads voltages up to 250V RMS from 40Hz to 440Hz, with reactive loads at power factors down to 0.2. Inverse parallel SCRs, configured for zero-voltage turn-on, handle current surges to 250A. If temperature exceeds set limits, the relays shut down and latch off until the input recycles and case temperature returns to normal, the trip being signalled by a status output. Teledyne Electronic Technologies. Tel., 081 571 9596; fax.081 571 9637.

COMPUTER

Computer board-level products

Vision inspection system. *PAC-Scan* by Pro-Active Control is an inspection system quality control, on-line inspection and component identification, using a range of standard cameras and monitors. It is based on a Eurocard and three processors are offered for large or small frame stores, monochrome camera and a toolkit. Resolution is 512 by 576 pixels and there are 256 greyscale levels. Examples of the system's application are a PCB positioner, accurate to within 50µm and a canning factory system in which cans are checked for damage. Pro-Active Control Ltd. Tel., 0223 300801; fax, 0223 300979.

Thermocouple data acquisition. *DI-221TC* from Dataq is a 12-bit, 16-channel portable data acquisition system, optimised for grounded thermocouples but suitable for other analogue inputs to the built-in signal i/o panel, a sensor on the PCB signal terminal receptacles providing cold junction compensation. The instrument linearises thermocouple signals in real time with DSP-based, 10th-order polynomial calculation software, voltage inputs being automatically converted to temperature indications in the user's choice of range, either ±1200°C or ±120°C. It plugs into the printer port of a PC and has its own battery. Keithley Instruments Ltd. Tel., 0734 575666; fax, 0734 596469.

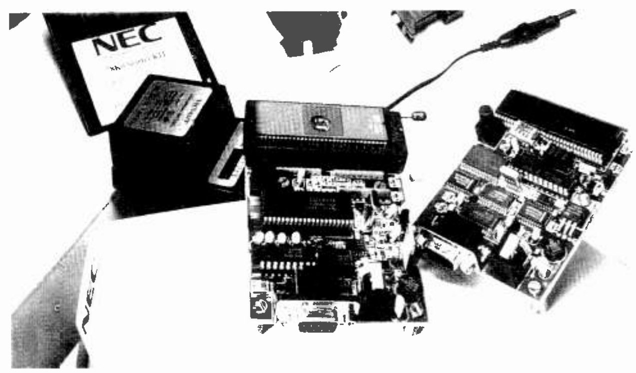
40MHz DSP board. LSI says its *PC/C31-40* applications board is the first to be based on the Texas Instruments TMS320C31 40MHz digital signal processor. It is a two-thirds-length PC AT card and comes with a complete Windows 3 development and integration software package, the Hypersignal Block Diagram software automatically generating ANSI C code and providing real-time support for the board. There is also the TI floating-point C compiler and debugging is assisted by LSI's Windows-based View31 tool and libraries of C-called DSP and maths functions. Loughborough Sound Images Ltd. Tel., 0509 231843; fax, 0509 262433.

Computer systems

Industrial computers. *SealTouch* computers by the American Deeco company are high-performance units in panel-mounting configuration and proof against hostile environments. CPUs are 486DX or DX-2 and control is by an infrared system that "looks" like a mouse to the computer. The system includes a sealed active-matrix 10.4in VGA display, a 3.5in drive, a 120MB hard disk, four expansion slots, on-board dram and dos. Deeco Systems. Tel., 0101 471 4700; fax. 0101 489 3500.

Data communications

Parallel port data acquisition. About the size of a laptop computer, *Parallel Pad* by Computer Instrumentation acquires data for a PC or notebook computer at up to 100kHz. Its 12-bit converter has programmable ranges from ±10mV to ±10V. Additionally, a screw terminal block may be plugged in to provide eight 4-wire inputs for thermocouples, PRTs, accelerometers and other sensors. There are 32 programmable logic i/o channels and a pair of counter timers. A stand-alone logging mode allows the unit to continue collecting data when its notebook computer has gone home. Computer Instrumentation Ltd. Tel., 0903 700755; fax, 0903 700788.



Controller design kit. A starter kit for embedded system design from NEC costs only 5% the price of a full kit, but still contains everything needed to design systems based on the NEC 78K0 range of microcontrollers. It includes an assembler and full-screen debugger for use on a PC, an evaluation board with 32K ram, a programmer for UV and one-time-programmable devices, a UV-erasable 78K0 microcontroller and all necessary hardware. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

Portable data acquisition. Containing not only analogue interfacing circuitry for on-site data acquisition, but also a 486 25MHz PC, Onsite's *Techstation -ci* series are 16-channel, 100kHz analogue signal recorders based on a notebook PC, offering on-line spectrum analysis while logging data to disk. The whole thing measures 15in by 12in by 3.5in and weighs 6kg. Everything works under Windows, so that no programming is needed and the data can be output in common formats for transfer to other software or loading to mainframes. Software supplied as standard is Windows, dos and Onsite's *DasyLab* acquisition software. Laplace Instruments Ltd. Tel., 0692 500777; fax, 0692 406177.

Development and evaluation

386CX/EX ICE. Hitex announces full in-circuit emulators for the Intel *80386CX* and *EX* embedded microprocessors, using the existing PC-based T32/386SX emulator as a base to allow true ICE of the central processor at up to 25MHz in 100PQFP and MQFP packages. The *HiTOP386* debugger is now extended to support the CX's new peripherals and C compilers by Microsoft, Borland and Intel are supported. Real-time debugging and analysis are enabled. Hitek (UK) Ltd. Tel., 0203 692066; fax, 0203 692131

Cheaper Checkmate. *Checkmate's* Intel 80C186EA in-circuit emulator has been reduced to £3500, including the *Paradigm Debug* debugger. The emulator does not interfere with the target hardware's operation and does not use the target's stack or insert wait-states when running from overlay memory and is guaranteed to run in the user's target. Great Western Instruments Ltd. Tel., 0272 860400; fax, 0272 860401.

Programming hardware

Field programmer. *Sprint Plus 48* by Concentrated Programming is a low-cost device programmer supporting more than 3000 devices, including most popular FPGAs, CPLDs, PLDs, microcontrollers, eeproms and eproms. It uses the ram, CPU and hard disk of the user's PC via a parallel port for menu-driven read, blank check, program, verify, sum check and ID check, the use of a notebook or laptop giving portability. Software is updated when required. Concentrated Programming Ltd. Tel., 0279 600313; fax, 0279 600322.

Software

RISCOS software. Steve Hunt offers software applications for the Acorn *RISCOS* computers, with the needs of RF engineers in mind, including a filter designer, a receiver noise figure/intercept point analysis tool, an inductance ready reckoner and a Smith chart design aid. Steve Hunt. 0604 858090 (evenings).

Mathcad updates. Two new versions of *Mathcad* technical calculation software have appeared: *Mathcad 5.0* and *Mathcad PLUS 5.0*. The former has been made easier to use, with easier equation and text editing and pull-down menus, yet offers new functions such as trace and zoom for graphics, print preview and a spell checker for document preparation. Mathcad PLUS has the same ease of use, but also provides a new set of functions, including differential equation solvers and advanced matrix algebra. Mathsoft Europe. Tel., 0344 23491; fax, 0344 873461. ■

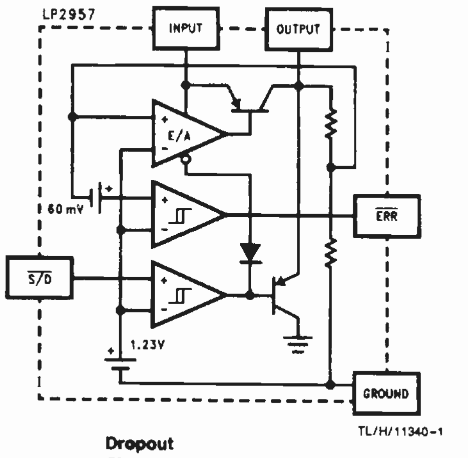
Low drop-out regulators

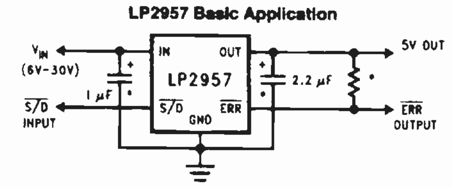
Given a 250mA load, early linear voltage regulating ICs like the *LM78xx* series needed an input supply at least 1.5V higher than their output before their regulators started to function properly.

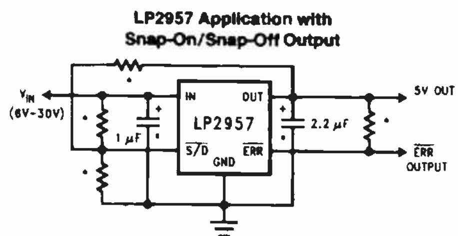
Two recent ICs produced by National Semiconductor need an input-to-output differential of only 470mV to achieve the same output current. One of these devices is specifically for microprocessor applications, the other a more general-purpose type designed to consume very little quiescent power.

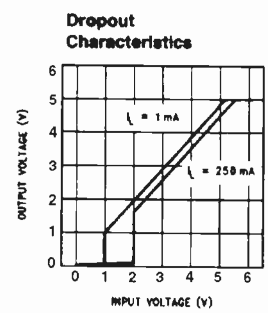
Called *LP2957*, the IC for microprocessor applications has a fixed 5V output. Being designed for minimal energy losses, this IC can compete with switching regulators in battery powered applications. For this reason, reverse battery protection is incorporated. Via its integral power transistor, the IC can deliver up to 250mA yet its quiescent current is only 150μA.

As the first set of diagrams show, the TO220-packaged *LP2957* can act as a basic 5V three-terminal regulator but with shut-



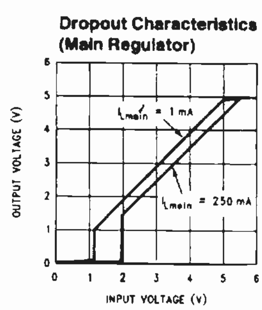


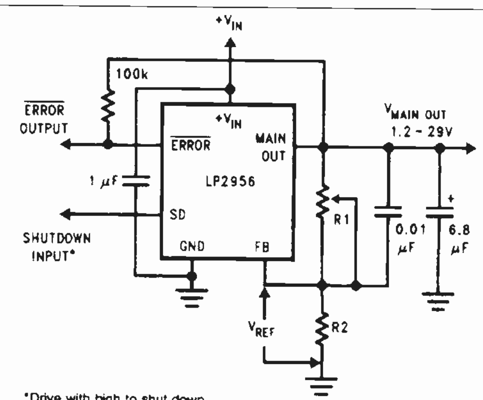


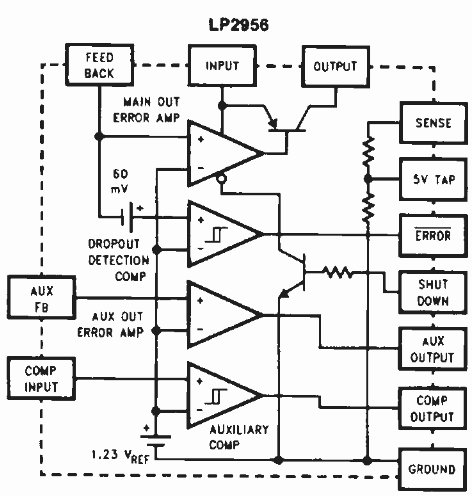


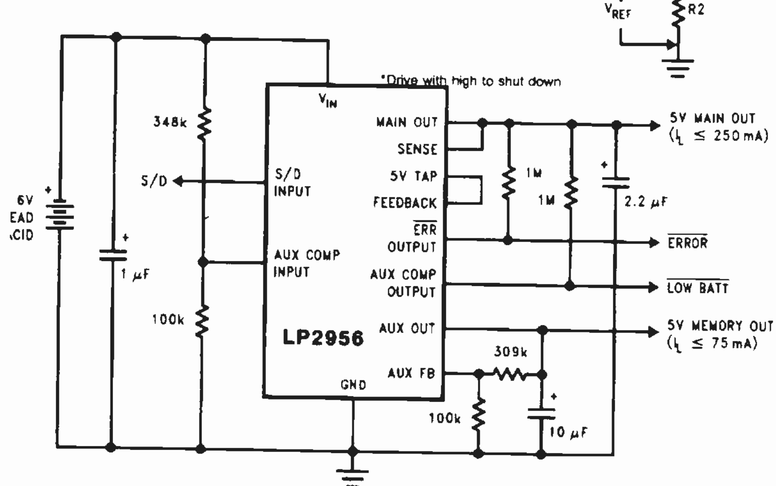
Producing a fixed 5V output with low energy losses, the LP2957 low drop-out regulator is designed specifically for microprocessor and microcontroller applications. Adding a few passive components to the basic regulator circuit provides snap-on/snap-off output. This ensures that, as input voltage falls, supply to the microprocessor cannot linger at a level that might cause it to operate unpredictably

At light loads, quiescent current of the LP2956A low drop-out regulator is 170μA, making it suitable for battery-power applications. Shown here are both 5V fixed and and 1.2 to 29V variable output configurations. Fixed-voltage circuit shows the auxiliary supply which remains on when the main supply is disabled.









down input and error output features added. The error output drives low whenever the output falls out of regulation by more than about 5%.

In the same set of illustrations is a snap-on/snap-off configuration. This arrangement is used to prevent the unpredictable microprocessor operation that occurs when supply input voltage falls below a preset level. If input to the regulator falls below the threshold, supply to the microprocessor 'snaps' off, causing a clean power down.

Housed in a 16-pin DIL or surface-mount package, the general-purpose *LP2956* shown in the second set of diagrams is more flexible than the five-pin *LP2957*. Its output is adjustable from 1.23 to 29V and it has an auxiliary output that can be used, say, for powering memory when the main supply is turned off via its shut-down pin

Again, this device has a comparator that signals when the main output falls more than 5% out of its regulation limits. Output voltage of the 75mA auxiliary regulator – which also features low drop-out capability – is independently adjustable.

Comprehensive performance curves and applications data are presented for both devices in their individual data sheets.

***National Semiconductor**, The Maple, Kembrey Park, Swindon, Wiltshire SN2 6UT. Tel. 0793 614 141, fax 697522.*

Designing with DC-to-DC converters

Possibly the most comprehensive book on using and selecting DC-to-DC converters and power supply systems has been produced by the Energy Systems Division of Ericsson. Comprising over a hundred A4 pages, the *Powerbook* covers topics ranging from circuit needs to electrical and thermal design of decentralised power systems.

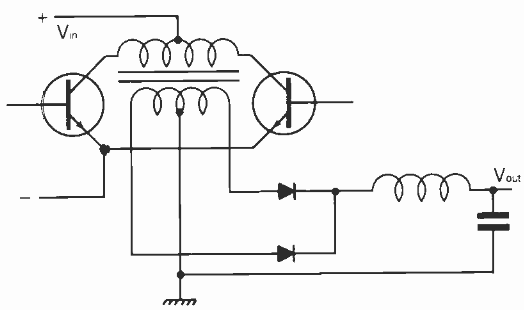
Particularly popular in telecomms, decentralised power systems have a number of advantages over traditional alternatives involving a single central power supply. Heat caused by power supply regulation is evenly distributed throughout the enclosure. Regulation is carried out local to the PCB being served so cable voltage drop is minimised.

Upgrading a 100W centralised PSU to 110W, say, would involve replacing an expensive unit. Upgrading a system with ten 10W converters involves simply adding another 10W converter. Likewise, the failure of a distributed regulator is less drastic than the failure of a central PSU. In some cases, a distributed regulator can even be replaced while the system is powered up.

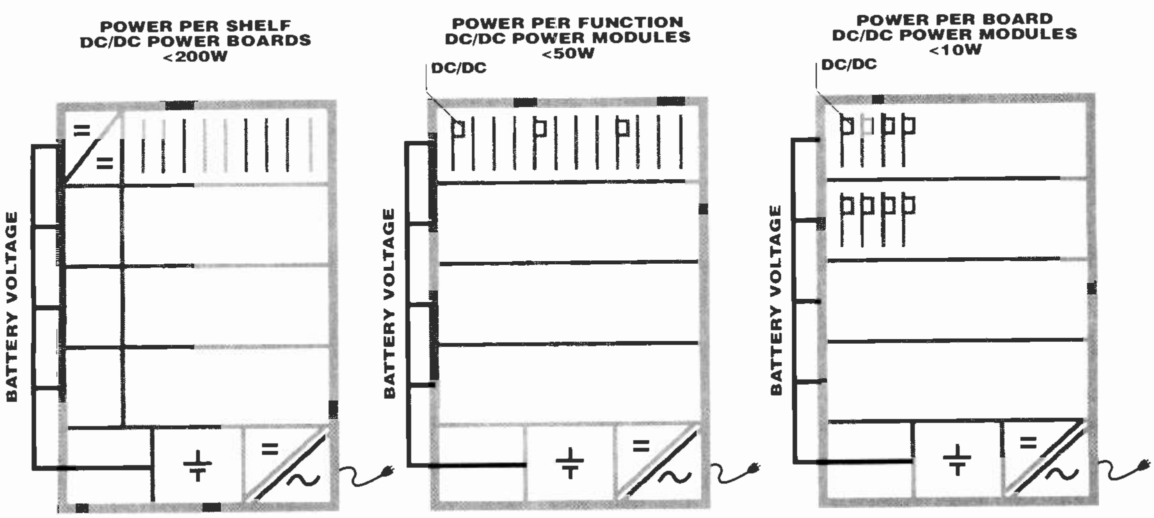
Finally, the cost of holding replacement power supplies for maintenance is also lower and energy consumption can be reduced since converters designed for distributed power can usually be turned on and off under microprocessor control.

In a section on converter topologies, clarification of the various types of converter is provided. Because different names can be used to describe the same topology, and since there are so many different types, describing topologies often causes confusion. According to the book, there are hundreds of different topologies and variations but no universally accepted classification scheme exists; but one option is presented on the next page.

Most widely used in commercial DC-to-DC converters are flyback, push-pull, half-bridge and full bridge topologies. In the flyback converter, not shown here due to its simplicity, transformer voltage 'flies back'

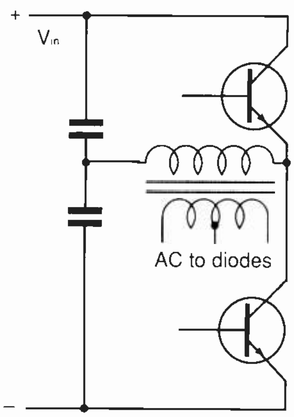


Push-pull DC converters are most useful for lower input voltages since each of the power switches sees a voltage twice that of the supply rail due to the tapped transformer primary.

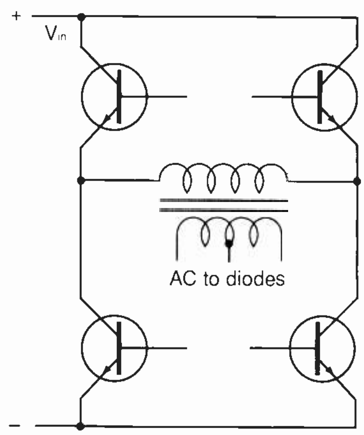


Widely used in telecomms, decentralised power architectures are gradually evolving towards the solution on the right. Dividing up the regulation into one DC converter per PCB has numerous advantages over the original solution comprising one regulated power supply per rack system.

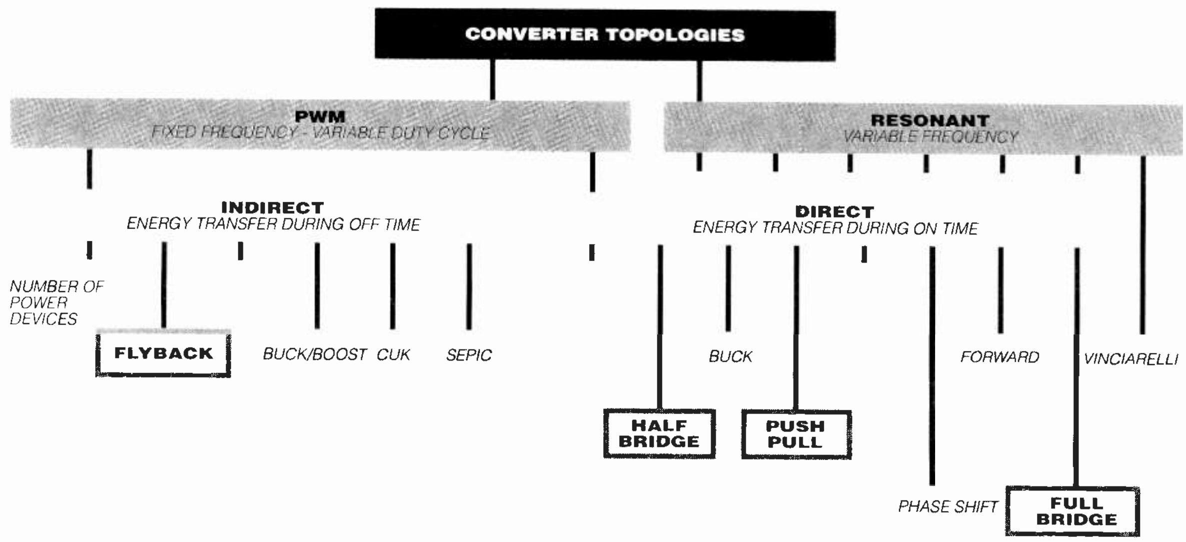
TOPOLOGY	NUMBER OF SWITCHING DEVICES	DEVICE VOLTAGE STRESS (VOLTS)	OPTIMAL POWER LEVEL (WATTS)	COMMENTS
Flyback	1	$V_{IN} + N V_O$	5 to 50	N = Tansformer turns rat Low complexity Low cost
Push-Pull	2	2 V_{IN}	15 to 150	
Half bridge	2	V_{IN}	15 to 200	
Full bridge	4	V_{IN}	150 to 1500	



In the half-bridge topology, two capacitors keep one side of the primary half way between the supply rails so the power switches only experience half the voltage stress of those in the push-pull converter. Component count is the penalty.



Full-bridge circuits, mainly used for converters delivering more than 150W, have four switches each seeing the full supply voltage but experiencing only half the current of their counterparts in the half bridge.



when the power switch turns off. Of the four types mentioned, this is the only one where the transformer acts as an energy storage device. It comprises a single transistor driving a transformer primary winding and rectification together with smoothing following the isolated secondary winding.

All four topologies are reviewed but the book points out that topology should not be a major criterion when selecting a DC converter for a particular application. No one topology is best, as the table on the previous page indicates.

Ericsson Components AB, Energy Systems Division, S16481 Kista-Stockholm, Sweden. Tel. +468 721 6356, fax 721 7001

Radio pager design

Full circuit details, PCB layout and adjustment procedures for an fsk radio pager are presented in GEC Plessey's note *AN172, Radio pager design using the SL6649-1 and MV6639*.

Operating at 153MHz, the system is a direct-conversion receiver that converts frequency-shift-keyed rf into digital data for decoding. Output data is 512baud tone data. All filtering is carried out on chip and two reference voltages are available for rf amplifier biasing.

Within the *MV6639* POCSAG* decoder, which is capable of operating from a supply as low as 1V, is a voltage doubler intended to drive an LCD.

In the circuit shown, two references within the *SL6649* bias an external cascode rf amplifier. This amplifier then feeds differential inputs of the mixer through a transformer. Input networking to the amplifier, L_4, C_{31} and C_{32}, is optimised for noise figure and best overall device sensitivity.

In the local oscillator, L_1 tunes out the crystal capacitance and suppresses oscillation at the fundamental frequency. It may need tuning to suit the crystal.

Circuitry around Tr_3's collector is designed for resonance at 153MHz. Output feeds an RC quadrature network comprising R_7/C_{18} with R_6/C_{17} and subsequently the mixer local oscillator input ports. Screening is important around the oscillator to prevent sensitivity fall-off due to feed-through to the rf input.

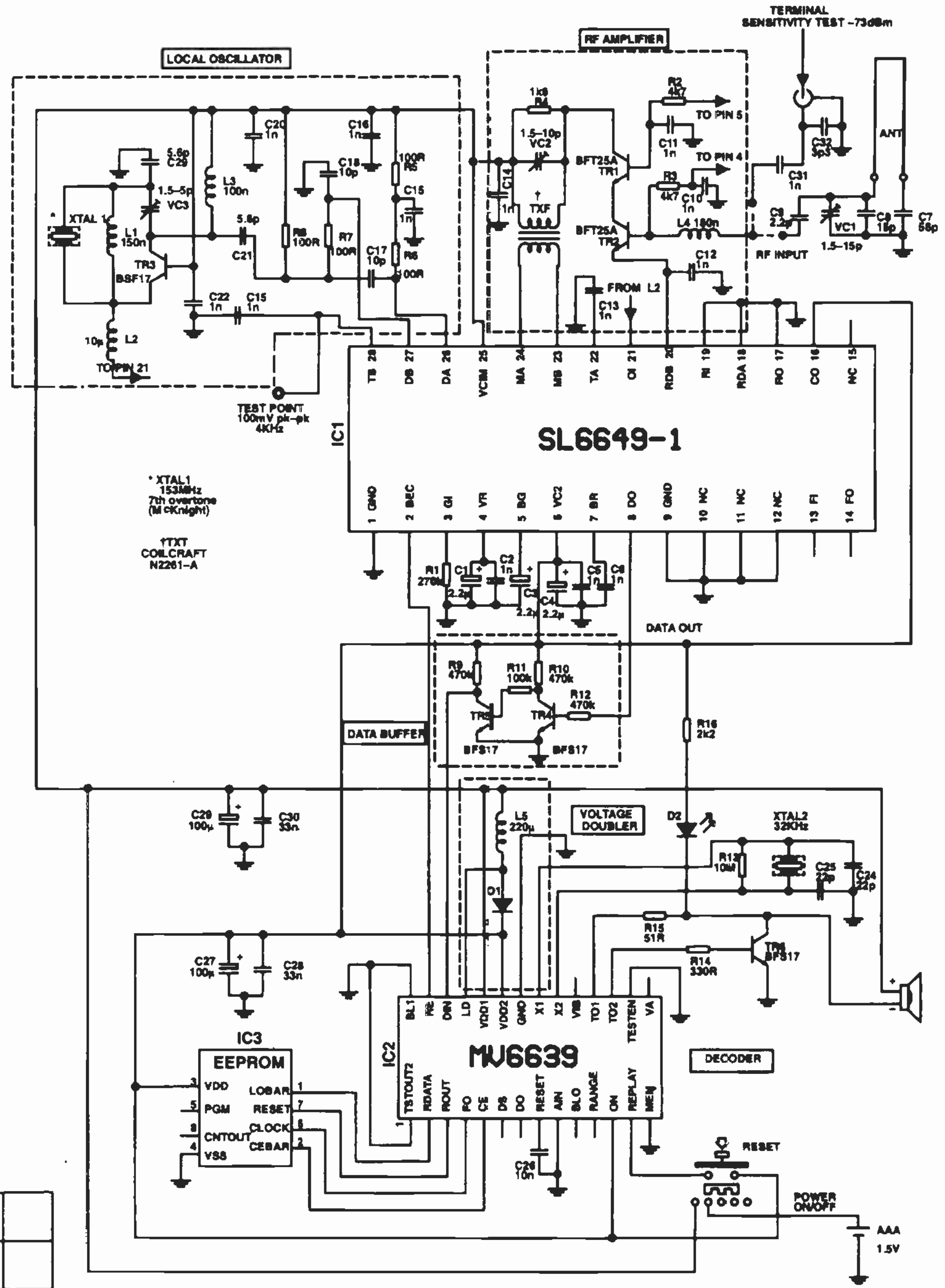
Buffering is provided by $Tr_{4,5}$ to isolate receiver data output and the POCSAG decoder data input. On leaving the buffer, POCSAG data feeds the MV6639 decoder whose reference is a 32kHz watch crystal. Radio identification code, bit rate selection

***POCSAG**

Covered also by CCIR RPC No1, Pocsag is the most widely accepted radio paging standard. It accommodates over two-million pager IDs, two of which can be recognised by each *MV6639* Pocsag decoder.

An acronym for Post-Office Code Standardisation Advisory Group, Pocsag transmission code comprises at least 576 bits of alternate one, zero sequence preamble followed by batches of code words. Each batch starts with a sync string followed by eight frames dual 32bit code words.

In addition to simple bleeping, Pocsag also allows simple messages to be transmitted.



Complete three-chip radio pager for 153MHz uses three ICs. Output is tone only but included in the MV6639 POCSAG decoder is a voltage doubler for powering LCD-readout.

and housekeeping information is held in eeprom. At switch-on, the decoder resets and reads information from this prom.

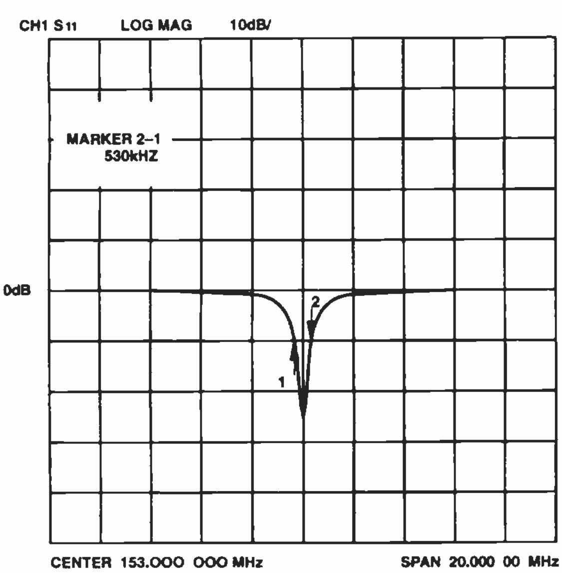
On receiving a correct message, the sounder receives a low level for 4s then a high level for 12s, unless the signal is terminated via SW_2.

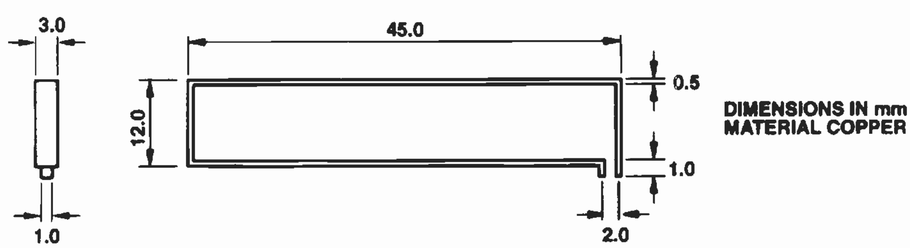
Antenna matching shown provides the necessary 50Ω source impedance. With it is a curve showing the dip it produces at 153MHz. Variable capacitor VC_1 is capable of tuning the dip to within at least ±5MHz of the resonant frequency. Coupling capacitor C_9 also contributes to the tuning but it should be kept small to maximise Q.

Antenna material should have a high conductivity. This is because the radiation resistance of this type of antenna is very low. Any ohmic loss will degrade efficiency.

Within the note is a description of the procedure for finding the pager's terminal sensitivity, which should be −126dBm or lower. ■

GEC Plessey Semiconductors, *Cheney Manor, Swindon, Wiltshire SN2 2QW. Tel. 0793 518 000, fax 518 550.*





Loop antenna for the 153MHz pager. When combined with the matching components shown on the circuit diagram, its terminal impedance is 50Ω. High-conductivity material such as copper must be used to minimise ohmic losses.

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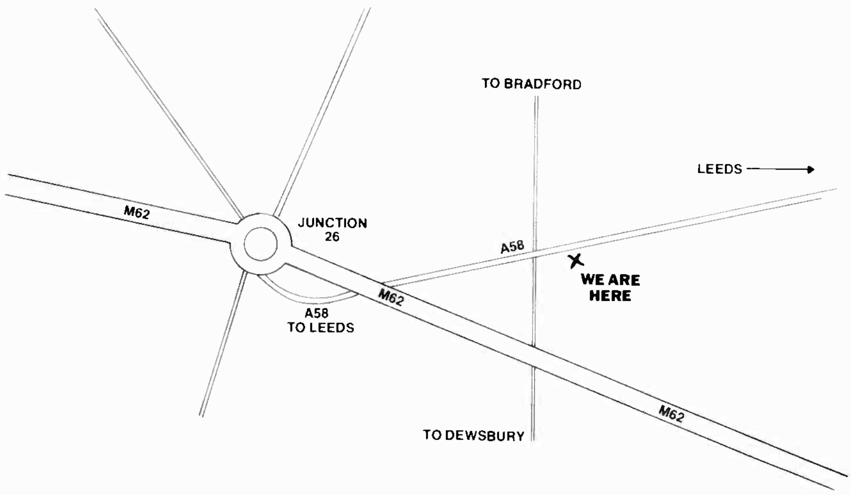

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Illustration Tracy Martin

Patently unclear

Just because it's patented doesn't make it good. And just because it's good doesn't mean it will be successful. Barrie Blake-Coleman* explains how big companies play by the unwritten rules of the patent game.

"Anything that won't sell I don't want to invent – a sale is proof of utility and utility is success". Thomas Alva Edison 1847-1931.

**Barrie Blake-Coleman is Industrial Liaison Officer at CAMR, Porton Down.*

A single patent can often embody, in unemotive terms, years of toil and accumulated genius. Eventually, it may become the only accessible record of a project, representing a huge commitment in time, facilities, money and intellectual effort.

Patents emanating from corporate resource-hungry exercises fall into this category. They are usually the outcome of dogged, but straightforward problem solving by large corporate machines able to throw capital at a particular project.

In contrast, there is the patent based on a moment's revelation by a gifted individual. Like the team effort, the inspirational pathway to invention is just as likely to originate in a corporate environment as from the perseverance of an individual.

Whatever the source, the result can be a patent specification worth a lot of money. A successful patent can gain substantial intrinsic value by virtue of what it cost to attain the information contained in it.

On the other hand, many patents are never tested technically or for their commercial potential and there is no real intention of exploiting their merits. These come about for a number of reasons, not least that of corporate bodies defending current, or future, trading positions.

In the case of the typical gizmo, a device of impressive cleverness stamped all over with "World Patents", we see something that fills a need few of us had spotted, and we stand in admiration at the commercial perception that created it.

But probably greater wealth is buried in the deep undercurrents of industrial technology – and here the patenting game is far more serious. With vast sums committed to r&d, the most important innovations are typically those which reduce production costs. Unfortunately,

Going for brokers

The person least likely to arrive at a realistic assessment of the business difficulties of promoting an invention is the inventor. Some are street wise and become highly skilled at getting their product off the ground. But most find it difficult to sell their idea, lacking the time, experience and negotiating skills.

The alternative to hawking a patent about is to look for professional help. Reputable brokers for ip (intellectual property) tend to operate on their best estimation of the likely returns on the venture. Some require ' a small fee now and a large piece of the action when I make you rich'. Others will waive the initial fee and simply settle for a percentage when the idea generates income. Yet others will not charge an inventor who can not afford the starting fee but will use the shotgun approach: "I won't make a formal, systematic search, but if I happen to pass a target that needs your invention I'll shoot the idea at them".

All are reasonable methods of sharing the load and in most cases have the merit of giving the broker or agent an incentive to make a real effort.

As a rule, a broker will tell an individual inventor the hard truth about the value of a patent. After all, he or she has the none too difficult task of finding outlets for the manifestly marketable ideas let alone those categorised as a solution looking for a problem.

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creative product or process engineering enshrined in patents is seldom without risk.

A strategy for defence

Only a fool would hand a competitor a route to a competing technology, particularly if they haven't paid for it. Even more foolish would be to patent – and so publish – a method showing competitors precisely which technology you have chosen to pursue. This is particularly true if a patent fails to protect procedures that might do the same thing. Thus is born the defensive patent, part of the new policy of strategic patenting.

Organisations committed to aggressive competition may look on strategic patenting as a way of maintaining a product lead. For example, blanket patenting entails patenting a preferred primary technology or process, then filing other patent applications describing procedures which conceivably could do the same thing. No-one needs to establish the validity of these parallel methods, the aim is simply to think up all other possible variations which could bypass the primary patent.

Success in obtaining patents on all your applications means competitors will (supposedly) be technically isolated. Of course, strategic patenting is perfectly legal. It has only one immediate drawback – cost.

Single-nation patenting costs can be modest. But to be truly effective a patent must be filed universally. One "International" patent (Patent Cooperation Treaty), assuming no hiccups, might soak up between £15,000 and £45,000, and servicing/renewal costs are prohibitive. The expense gives perspective to the comparison of patents granted to the leading industrial nations.

The patenting record of the Japanese as compared to the UK, US, France and Germany reflects a willingness to spend money on patents, rather than their technical and commercial value. We should never presume that all patents are viable. Statistics give only the number of patents filed, but give no clue as to commercial fitness. The true value of a patent lies in its ability to be exploited.

Falling numbers of British patents say less about the decay of British inventiveness than about the decline of British industrial solvency. In recent times, not only has government and industry neglected to resource the r&d that generates ideas worth protecting, there has been a marked reluctance to finance exploitation – even when lucrative ideas are on offer.

Unless we use our creative talents to create wealth, the wherewithal to protect our ideas will dwindle. With it will go the incentive to patent at all.

Keeping ahead of the competition

It is naive to believe that competitors will not poach on an originator's "natural" right to monopolise an invention. Patents are designed to protect an idea, but they can invite competition at a time when it could best be done without.

One lesson is that – strategic reasons apart – if you can't afford to do it right with a full inviolate patent portfolio, don't do it at all. As a result, there are probably more technical breakthroughs left unpublished than there are locked up in the patent office.

Indeed, this catch-me-if-you-can policy is an alternative to strategic patenting, though it calls for a lot of confidence in an ability to run faster than the opposition.

An application that fails or is abandoned still constitutes a publication and so can impede a competitor's chances of acquiring a similar patent. It may also cause a collision at some later date with subsequent patents filed by the original applicant.

Nevertheless, the guiding rule should be that if the idea makes commercial sense and has the realistic potential to generate income – then the wise decision is to proceed. Like accident insurance, owning the right patent means you are unlucky if you have to rely on it.

But if the application is for less sound reasons, the maxim remains – a patent is only as strong as the litigation that tests it. A patent grant says only that the examiner is prepared to allow it, not that the examiner is infallible and guarantees the legal invulnerability of the patent. Furthermore, the rights conferred on an inventor tend to focus on preventing others from profiting from the method embodied in the patent and have little influence on manufacturing rights.

Finding an interested party

Let's assume you have found someone reputable to pick up your intellectual property (see box **Going for brokers**) and run with it – beware of cowboy patent shops and never sign over rights to an agent.

To attract, the idea must appeal to the recipient and make commercial sense. But any professional involved with ipr (intellectual property rights) can cite a massive array of negative, as well as positive, reasons why a business will or won't acquire patents.

Seldom is it obvious why a company declines interest in a technical acquisition of apparent value to it. Even some of the positive reasons tend to be defensive. For example, a patent might be acquired by a company to ensure it does not impede parallel "in house " development; or the company might want to keep the idea out of the hands of competitors. Or it may want to be certain that no other factor could change the commercial life expectancy of its products.

There are many subtle technical and manufacturing reasons that can prompt companies to turn down a patent. For example:

- failure to understand the implications of a technical development;
- existing product range viewed as sufficient without inviting undue competition;
- appropriate pre-production development/design facilities are lacking;
- new idea departs from core business;
- idea not seen as technically feasible or production orientated (ie requires too many ex-house skills), or
- client has technical confidence but is unfamiliar with or lacks marketing skills to make new idea successful.

But internal politics is by far the most dangerous area for new ideas. Jobs go with products and a new idea may mean changes of responsibilities and skills. It could signal future security for some, but perhaps not for those linked to an old product.

There are also risks in adopting a new product. Those responsible for deciding an acquisition will want it to succeed, perhaps bleeding resource from other areas or depleting already overstretched budgets.

For these reasons, many new ventures never see the light of day and the inventor receives yet another refusal.

Right decision, right reasons?

Of the positive reasons for acquiring a patent, the most compelling concerns the wish to use it as a major source for income generation or saving money.

The ideal vehicle for a new idea is one with the technical and financial resource quickly to extract the full value of the patent. Willingness to do so depends on whether the patent is properly understood in terms of a marketing and manufacturing strategy.

A good agent will establish a pyramid of prospective takers for an invention. At the top will be those with an industrial interest that aligns itself with the invention. Then there are those who are evidently diverse in their business strategy and will be open to persuasion as to the business merits of taking on the idea.

Below them comes the option of an entirely fresh business venture, requiring complete start-up from scratch. This clearly would be indicated where finance could be acquired to begin a new manufacturing operation.

In many cases, a company's existing technical expertise in a related area bodes well for a new idea – making it easier to evaluate in terms of the potential commercial worth.

A good agent will pick out the appropriate decision makers in an organisation and put up a convincing argument in favour of acceptance. The aim will be to show that the new idea is technically sound; aligns itself with the company's business interests; demonstrates a return on investment that justifies pre-production and marketing costs; does not divert resource from existing profitable operations, and that the terms of acquisition are not unacceptable.

In short, the sale of the ip must be with a full and sympathetic understanding of the factors that need to be satisfied in the business and manufacturing environment.

Patent agreements

Primary risk venture: the recipient agrees direct purchase of ownership and the inventor relinquishes all rights and control for a lump sum.

Royalty: the inventor is paid a substantial royalty (or moderate lump sum + royalty) on a "shared risk" basis, giving a licence to the recipient and placing a duty on them to go into production.

'Split risk': a substantial lump sum is paid up-front for a licence (or ownership) and the inventor takes a chance on when/if use, manufacture and any royalty on sales commences. Whatever happens, royalties will be at a much reduced percentage.

Royalty agreements are generally exclusive to the recipient or pre-defined in terms of the industrial and marketing area covered. Sometimes, where the invention may spawn other novel products, or where the recipient wants absolute market control, the licensing arrangement may include sub-licensing conditions. In the case of a non-product orientated deal (process or cost saving ideas) the inventor receives a part of the cost saving.

A deal is a deal?

Inventors usually become frustrated with the tedious business of ipr transfer and the apparent inability of prospective takers to appreciate the breathtaking revelations detailed in their patent specification. But the more experienced inventor knows that any recipient of an idea is placing a heavy commitment on themselves.

Many factors have to be satisfied, particularly if rights and licensing agreements are coupled to dates when production is to begin.

Typically, a patent is acquired as a primary risk venture, on a royalty basis or as a split risk (see box **Patent agreements**). The advantage to the recipient in buying out the inventor under a primary risk venture is obvious. In accepting a once-only payment, the inventor surrenders control and interest in the patent. As such, the organisation acquiring the invention is not obliged to use it. Similarly, "option" deals pay the inventor a retainer for exclusive use by the recipient at some future, indeterminate, date.

To win the best deals, inventors must realise they are an integral part of the acquisition. Peddling ipr is as much a public relations exercise as a conveyancing of technical information. Few companies are likely to want an unfamiliar technical process without direct access to the originator. The patent is seldom enough and it is not uncommon for inventors to be overly defensive about their ideas, not even trusting the safety of a patent. So, by deliberate omission or oversight, many specifications are deficient in essential technical features, making it very difficult to repeat the work.

Granted, in certain circumstances the mere exposure to an idea is enough for a technically-competent individual to succeed in producing the same end product. Likewise, the merit of a patent may be in its ability to identify a new application for a well established procedure.

But, whatever the situation, the inventor is usually seen as important to the success of the undertaking. An acquisition should be seen as an exchange: the inventor gets the chance to see his or her idea used – with a possible financial benefit – in exchange giving expert knowledge and a patent.

Hard negotiation

Having negotiated a half-decent deal, should we think twice before refusing it? The offer could have been long in coming, following years of traipsing around the country listening to constant refusals.

Companies usually move cautiously and a decision not to proceed after months of negotiation can take the fight out of even the most enthusiastic of inventors.

The bird in the hand, though not good policy, is all too common and the burden of personal involvement and sundry costs can quickly exhaust the patentee.

But instead of despair the exercise should be reconsidered in terms of whether the objectives are sound. Open-minded applicants will learn from the observations made at various meetings. Perhaps opinion agrees that the idea embodied in the patent is heavily flawed. For the inventor, the facts can be hard to swallow but the more quickly they are, the less painful it will be.

The consensus could be that the idea is good, but the application or approach is in error. Perhaps the industrial sector targeted is actually inappropriate.

But the positive side of getting used to "no" is that it can teach how to get a "yes". Edison (1093 patents) knew the game very well: "It is easy enough to invent things and set the newspapers talking, but the trouble comes when you try to perfect your inventions so as to give them a commercial value."

Inventors need to cultivate pragmatism and good business sense to profit by their efforts. Put another way, vendors of ipr need to be able to say no too. If a thing is valuable, it shouldn't be given away. Again, professional help is advisable – if only to suggest where to invest the money.

On what terms?

Corporate r&d now accounts for the greater proportion of patents and the day of the amateur inventor is past its zenith. Nevertheless, patents are still bartered, exchanged and sold and it is as well to consider how terms are agreed and what constrains typical negotiation.

Few people realise just how much resource is needed when investing in a new idea, particularly if it represents a new product. The importance of the intellectual contribution can easily be overestimated and the mechanics of turning the concept into a reality, underestimated.

Risks can be large, with costs ranging from the staff needed, to the set-up costs. Time spent by personnel in establishing the production and marketing operations must also be included.

So when negotiating a licence for ipr or the transfer of rights, it is in the vendor's interests to agree terms which are favourable to the recipient in the first instance. Recipients should not have their chances of success marginalised because of an over-demanding "money now" agreement. This epitomises the reason for the shared risk approach. Better to wait for long-term rewards than risk short term failure, as can happen when small firms take on new patents requiring substantial investment in time and equipment.

It is a salutary lesson that most of the familiar patented processes took a long time to make it. Typically, the period for a concept to materialise as a product is seven years.

Float glass took seven years to perfect; major advances in petroleum cracking needed twelve years; nylon was only perfected after six years and, even with an effective recovery process, large scale production of penicillin took twelve years. The transistor only became reality five years after the first crude demonstration of the effect, and production of crease-resistant fibres required another five years of work following the first patent disclosure of the method.

For the invention to succeed, it has to be possible for the user to succeed. That does not mean concluding a bad deal for the vendor.

Some agreements are immensely beneficial to the vendors immediately. But these are generally where the rights transfer instantly confers some financial advantage to the recipient. Being able to market without hinderance a product whose launch would infringe the patent in question is just one example. The further proportionately-smaller investment needed to acquire a licence frees the company to continue profitable trading and eliminates the need to write off the original manufacturing investment.

Unencumbered exchange of licences, where two competitors have both invested in nearly identical products, is another and precludes costly litigation or mutually destructive price wars.

Rewards of success

As we have seen, the delay before a product based on a patent makes its mark commercially can be protracted. The individual inventor has, on past records at least, tended to produce ideas "before their time" and must face an uphill struggle to convince others. Often, too, inventors see economic advantages as secondary to the pleasure of conceiving and demonstrating a novel concept.

But the problems faced in converting ideas into industrial reality remain essentially the same today as in the inventive heyday of the 19th century – as do the rewards if successful. The Black and Decker *Workmate* made its inventor Ron Hickman very wealthy. But, as is now part of inventor folklore, when he first approached B&D they declined to pursue the idea because they thought that too few would be sold to warrant volume manufacture. He set up his own manufacturing operation and when sales clearly started to pick up he offered the idea to B&D again.

A redesign for cost effective mass manufacture resulted in massive sales. In this case, Hickman needed to prove his idea commercially, beyond its technical merits. In the end, the popularity of his product proved that his marketing perception was better than that of the professionals.

It is possible to win and lose at the same time. At the end of the 19th century Westinghouse, desperate to break the hold that his arch rival, Edison, had in dc electrical generation offered a straight million dollars for Tesla's patents on ac electricity supply. They shook hands on the offer just minutes after their first meeting started. As it turned out, the patents were actually to be worth much more than the fabulous sum agreed. Though Tesla was delighted, Westinghouse later confided that had Tesla demurred, he would have been prepared to go as high as three million dollars.

However, Westinghouse was unsure how his board of directors would react. They were unclear of Westinghouse's intentions and on hearing about the million offered to Tesla thought Westinghouse had taken leave of his senses. Tesla believed he had done extraordinarily well but, ultimately, nearly exhausted all his money on a fine lifestyle and virtually sterile experimentation. Curious, then, that in retrospect Tesla was content with far less than he could have got, Westinghouse achieved more than he expected, while the Westinghouse board thought the whole arrangement was foolhardy as it stood, without realising how bad it could have been. In the end Westinghouse was vindicated but might not have been had he committed the full three million, making it difficult for the company to develop Tesla's technology. A lesson for us all? ■

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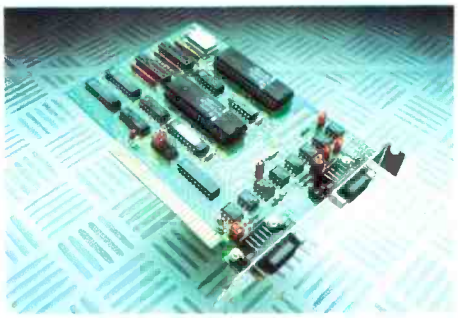

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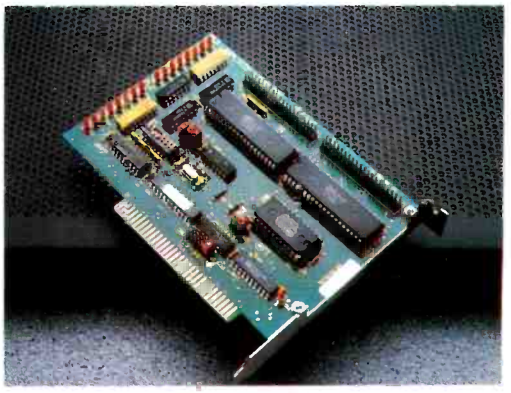

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