AUDIO
Ultra-low distortion test oscillator

CIRCUITRY
Working with current mode amplifiers

APPLICATIONS
Three-chip pager system

RF ENGINEERING
Microstrip design, instability and mismatch

PC ENGINEERING
Schematic entry: good starting point?

COMPONENTS
SOT23 transistor switches 100W

DESIGNER'S GUIDE TO OPTOELECTRONICS
The PC82 Universal Programmer and Tester is a PC-based development tool designed to program and test more than 1500 ICs. The latest version of the PC82 is based on the experience gained after a 7 year production run of over 100,000 units.

The PC82 is the US version of the Sunshine Expro 60, and therefore can be offered at a very competitive price for a product of such high quality. The PC82 has undergone extensive testing and inspection by various major IC manufacturers and has won their professional approval and support. Many do in fact use the PC82 for their own use!

The PC82 can program E/EPROM, Serial PROM, BPROM, MPU, DSP, PLD, EPLD, PEEL, GAL, FPL, MACH, MAX, and many more. It comes with a 40 pin DIP socket capable of programming devices with 8 to 40 pins. Adding special adaptors, the PC82 can program devices up to 84 pins in DIP, PLCC, LCC, QFP, SOP and PGA packages.

The unit can also test digital ICs such as the TTL 74/54 series, CMOS 40/45 series, DRAM (even SIMM/SIP modules) and SRAM. The PC82 can even check and identify unmarked devices.

Customers can write their own test vectors to program non standard devices. Furthermore it can perform functional vector testing of PLDs using the JEDEC standard test vectors created by PLD compilers such as PALASM, OPALjr, ABLE, CUPL etc. or by the user.

The PC82's hardware circuits are composed of 40 set pin-driver circuits each with TTL I/O control, D/A voltage output control, ground control, noise filter circuit control, and OSC crystal frequency control. The PC82 shares all the PC's resources such as CPU, memory, I/O hard disk, keyboard, display and power supply.

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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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CIRCLE NO. 104 ON REPLY CARD

354 ELECTRONICS WORLD+WIRELESS WORLD May 1994
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.
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 1ns - is repetitively transmitted into the subsurface 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 1048ms, 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.

Flexible battery charged to plastic

Researchers at Bellcore say they have created a revolutionary new type of battery that is as flexible as 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 it is not uniform."

Another advantage the searchers had came from the way a buried body decays. The missing flesh is not necessarily replaced by the ground. Voids filled with soft materials, such as rotting flesh, allow the initial transmitted impulse to be 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 2 Gbyte 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 the way a buried body decays. The vision gets disturbed so it is not uniform."

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 43 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.
The BT group’s amplifier uses 500mW pump power to amplify the information signal by 30dB.

One problem BT found with using Pr3+ 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 100ps 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 550Mbit/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*

---

**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 wide 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$_{13}$Y$^+_{13}$ 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.

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 25,000mph - 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...
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.

The piston is propelled by a methane/air charge. This rapidly heats the hydrogen in the tube, compressing it to 7000lb/in² 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.

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 magnetoreistant 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.
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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.

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 $R \times I_o$, 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 sup-
ply 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 1000Ix effectively compressing the useful luminance range to a convenient scale. A microprocessor fitted with a suitable A/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 to 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 1 lx then the 1N4148 diode should be substituted with a low leakage type. The instrument is calibrated by the injection of a luminance 

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 GaAIAs diode and the 950nm GaAs diode. The emitted spectrum (Fig. 7) indicates a good match between a GaAIAs 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/m² 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/m². This equates to an irradiance of 16W/m² 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-
The earth receives approximately 1 kW/m² from the sun at the surface of the earth. Photodiode conversion efficiency is approximately 0.6 A/W at wavelengths approaching the maximum sensitivity. If the photodiode has an active area of 1 mm², the received power would equal 1 mW, generating 0.6 mA of current.

A BPX43 photo transistor with a glass lens generated 2 mA in the collector-base diode (and 1 mA in the emitter-base diode) when subjected to direct sunshine. With an active area of 0.675 mm² 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 1100 nm, useful to the photodiode, while the remaining two thirds are wasted as heat.

The radiated power from a domestic 40 W 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 4πR², creating an irradiation of,

\[ E_e = \frac{P}{4\pi R^2} = 3.2 \text{ W/m}^2 \]

at a radius of 1 m. At wavelengths visible to the photodiode the useful irradiation is approximately one third of the total or 1 W/m².

The illumination from the 40 W bulb was measured to 400 lx at 0.3 m, indicating a correlation between illumination and irradiation of 40 lx to 1 W/m². This correlation is valid only for filament lamps operating at the same filament temperature, approximately 2500 K.

A higher filament temperature will produce more visible light. At 3000 K the correlation is 60 lx to 1 W/m², at 4000 K 100 lx to 1 W/m². Photo components are usually specified at a filament temperature (colour temperature) of 2856 K (standard light A) and the above correlation enables one to perform an approximate conversion between lx and W/m². 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 100 Hz. 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 40 kHz from electronic ballasts. Interference from fluorescent lamps may be reduced by optical filters.

**Power, radiation and calibration**

<table>
<thead>
<tr>
<th>Wavelength (nm)</th>
<th>Power Output (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>400</td>
<td>0.3</td>
</tr>
<tr>
<td>800</td>
<td>0.6</td>
</tr>
<tr>
<td>1200</td>
<td>1.0</td>
</tr>
<tr>
<td>1600</td>
<td>1.3</td>
</tr>
<tr>
<td>2000</td>
<td>1.6</td>
</tr>
</tbody>
</table>
device switching time is 1\mu s.

The led has a half power angle of \(\pm 10^\circ\) but the main intensity is concentrated in a narrow beam within \(\pm 4^\circ\). This often causes alignment problems at extreme range.

Input gating switches the transmitter on or off. The 4.7pf 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 500\mu s.

### 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 sources or have outputs which flicker at low frequencies, a load which is off 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 100\Omega 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.

In the first, photocurrent from the BPW41 raises the base voltage of the low noise ZTX384 via the 330k\Omega resistor until the transistor’s collector voltage reaches about 0.8V by acting as a current generator. For low level photocurrent, 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 this 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 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 10,300 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 behavior dramatically. The voltage gain of the circuit now approximates to:

\[
\text{Voltage gain} = \frac{R_c}{r_e}
\]

With low background light levels. For instance, in a dimly lit room yielding 5\mu A 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 10,300 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 behavior dramatically. The voltage gain of the circuit now approximates to:

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\]

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/m² and 870nm and is reduced to 0.35mA at 950nm. To switch the transistor requires an irradiation of 3W/m². 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 Cth is amplified by the transistor current amplification factor B. The resulting time constant equals Bt/Cth x Rc, where Rc 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, Rc. Sensitivity too is proportional to Rc, 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 100kΩ to reduce the transmitter diode current to a safe level. The Siemens 660nm transmitter diode SFH750V and photo transistor receiver SFH3501 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 100nm detection range. The circuit could be used for data transmission with fibre optics.**
tion is the photo transistor collector-base capacitor. For high speed operation a photo-
diode 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 (150 kΩ). The transistor acts as an impedance transformer reducing the output impedance below 1 kΩ.

Using the BPX43 photo transistor the value of Cb is 20 pF at 2 V leading to a bandwidth of 35 kHz if the stray capacitance is assumed to be 10 pF. The bandwidth could be increased if the network resistance is reduced. A similar design operated at 100 kHz with 20 kΩ network resistance and a Honeywell 5D443 photo transistor.

The circuit exhibits a sensitivity of 5 mV/Ix or 330 mV at 1 W/m² with the resistors shown. The photo transistor will accept 5000 lx 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 110 V at 1 m distance.

The collector of a metal can photo transistor is grounded to reduce sensitivity to electromagnetic interference. The photo transistor enclosure 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 5 V 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 600 kΩ providing a fast diode interface. The amplifier limits bandwidth to approximately 80 kHz.

The collector of the first transistor is fixed at 0.6 V 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 npn transistor is fixed at 1.2 V 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 100 lx 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 pnp 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 1 V above the amplifier dc level. The irradiation needed for 1 V output is 0.014 W/m² and with the Fig. 9 transmitter radiant intensity of typically 0.33 W/m², the system should work comfortably at 5 m distance.

The output noise voltage is 0.1 V 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 0.1 V threshold voltage (which corresponds to a range of 15 m) 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 0.2 V at 1.8 m distance.

The circuit of Fig. 13 is a complete proximity sensor with 0.35 m detection range and low power consumption. The 4069 UB 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 100 mA.

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 3 pA at 1 W/m².

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 1 V turning off the output transistor.

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

![Circuit diagram](image)

Fig. 2a Circuit providing a notch at one frequency f. The response at 2f is still 4.77 dB 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.

![Diagram](image)

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

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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 at 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 phase shift in each integrator meant that the loop phase shift 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 phase shift 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 integrator 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 phase shift up to 360°, the circuit will oscillate. 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_{DC}, reducing the loop phase shift 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 centered within its voltage compliance range, and one providing dc-blocked "negative damping" to the HP stage virtual earth. The voltage set by the 10kΩ/2.7kΩ resistors R15 and R16 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, Tpk 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.

Estimating the Q

To see just what Q the filter runs at, Tpk 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 input gives the Q, being approximately 0.5 at second harmonic (Q = (R2/2R1) x Tpk x (Tpk)/2). Thus the operating Q of the filter oscillated over the whole 20Hz-20kHz range with 30MΩ, as shown in Fig. 5b, being fed to the virtual earth of op amp A6a with its gain of 11 was included to ensure the output, as described in Ref. 3, the operating Q is 0.1% at second harmonic, and the third harmonic in the signal from the OTA should be less than 1%. In low distortion measurements, where the harmonics are all much less than 1%, in practical terms (1) and (2) are identical.

Assuming E1 stands for the rms value of the fundamental of a distorted sinewave, E2 for that of the second harmonic component, etc., then the total harmonic distortion is given by:

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

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 E1 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:

\[
\sum E_2^2 + E_3^2 + E_4^2 \ldots \ldots 2
\]

but if all components E2, E3, E4 etc. are less than 10% of the amplitude of E1, then the difference due to each is less than 1%. In low distortion measurements, where the harmonics are all much less than 1%, in practical terms (1) and (2) are identical.

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 was used to drive an output section as shown in Fig. 5b, being fed to the virtual earth of op amp A6a. It can be seen that there is also a contribution from the HP oscillator output, via R31 + R35. This was included so that R31 could be adjusted to suppress any third harmonic component in the output, as described in Ref. 3. The output of A6a drives A3b, 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 S4, and by the 0 or 50dB pads. S4. The odd value resistors R44...R48 and R51...R57 were made up by paralleling standard values to get within 1%.

S3 provides a choice of a low 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, S5, and the main circuit board.

Performance

When set up, the oscillator provided the following performance, all measurements being taken with S3 set to low output impedance. S4 and S5 to 0dB, and using the notch box.

<table>
<thead>
<tr>
<th>Frequency</th>
<th>Total THD</th>
</tr>
</thead>
<tbody>
<tr>
<td>20Hz</td>
<td>0.00062%</td>
</tr>
<tr>
<td>600Hz</td>
<td>0.00034%</td>
</tr>
<tr>
<td>600Hz</td>
<td>0.00022%</td>
</tr>
<tr>
<td>10kHz</td>
<td>0.00092%</td>
</tr>
<tr>
<td>10kHz</td>
<td>0.00062%</td>
</tr>
</tbody>
</table>

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

b. Part circuit of the SVF-filter based low distortion oscillator: output section.
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 let 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 R14 and R25 set to ground), R20 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 R31, which as already stated was included as a way of out-plussing any third harmonic at the output. However, as noted in an earlier design brief the main distortion mechanism in the op-amps used was second order, and there was no visible third harmonic to cancel R31 was therefore set to minimize the second harmonic content at 10kHz. This just leaves the circuitry associated with R24 and R25, which has not been mentioned up till now.

Op-amp A3b 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 R13, to the virtual earth at the input of A3b. With suitable settings of R24 and R25, 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 D2 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 D2. At 50°C the distortion was unchanged from the room temperature value. The particular settings of R24 and R25 used resulted in 1.5V pk-pk at the output of A3a, the waveform exhibiting a percent or so of second harmonic distortion. This is diluted by a factor of about 300 by the ratio of R33 to R34 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%.
90° lag, the BP waveform actually leads that by 90° - note that although inverting stage first of these enables the unit to be used as a transformer’s stray magnetic field.

Another 56Ω 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 R1, which may be used in two different ways. Firstly, with the oscillator running, its frequency can be locked to a low level signal injected via R1, simply by tuning it to the frequency of the injected signal.

Secondly, the oscillator can be disabled by setting S2 to the "filter” position. This has the effect of increasing the op-amp’s negative rail, disabling the positive feedback via the OTA and stopping the oscillation at the same time. At the same time, damping is applied from inverted BP via R37 and R38, defining the Q of the resonant tank defined by C7, R7 and C6.

References.
1. Phase-shifting oscillator Roger Rosens Wireless World Feb, 1982 pp. 38-41
3. Design Brief: Low distortion audio frequency oscillators Ian Hickman EuW 4+4 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.

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 infininte 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 FMFT618 is useful in battery-powered applications. It can carry up...
In battery-powered applications, it is vital that the transistors can absorb inductive transients. The Miller effects—i.e., capacitance mosfets adequately, a high speed buffer control ICs are rarely able to drive larger capacitance mosfets efficiently. High frequency converters, it is vital that the logic output avoids signal inversion. Where 5V logic provides a pulse-width modulation, a buffer can be necessary to trans- fer the h-bridge pairs can be charged to 12V in under 30ns—a feat requiring a peak current of around 4A. Included in 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. 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 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 50N FMMT619 and four npn 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 npn types and 1% for the pnp.

Saturation losses at 1.5A total only 0.3V and can be easily 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 Ciss, 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 cir- cuits 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 hfe are used. Combined, the FMMT618 and FMMT717 have these characteristics. The 10nF effective capacitance of two I Robots mosfets in parallel can be charged to 12V in under 30ns—a feat requiring a peak current of around 4A. Included in 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 FMM72369 switching transistor, the circuit shown in Fig. 7 converts 5V logic drive to a 12V gate drive signal. Driving the emitter of the 2569 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 2nF-input mosfet.
By using large TO126 or TO220-packaged transistors. Using FMMT transistors not only reduces cost and size but also eliminates the need for protection diodes.

**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 protection diodes. These are needed to divert destructive reverse transients generated by the inductive load.

In Fig. 8, FMMT718 and FMMT720 transistors in SOT23 packaging replace these bulky and expensive leaded transistors, giving other savings too. High reverse 

**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 50mΩ2 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 \( V_{OUT} \). The converter will operate from a supply of \( V_{IN} \) up to 16V. Its output voltage can be set to 5V or 3.3V by altering the value of \( R_3 \). 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 in, say, computer backlighting, need an efficient, compact converter. This design originally had SOT223 BCP56 transistors. Replacing them with smaller FMMT types reduced saturation voltage by half.

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**Components**

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**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 BCP56 transistors. Replacing them with smaller FMMT types reduced saturation voltage by half.**
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CIRCLE NO. 110 ON REPLY CARD
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 matching.

From the book RF Transistors: principles and practical applications.
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.

A common practice is to wind the chokes ($L_1$ and $L_3$) over low valued (10-50Q) 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 0.1nF 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 it losses start producing excessive dc voltage drop.

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

The $L_2\ C_1\ C_5$ 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 $Q$s, and high $Q$s 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 $Q$s 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 these frequencies. The effect is known as varactor instability, and implies that there are $2f_0$ and $3f_0$ 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 $0.5 \times f_0$ since system power gain is much lower at these higher frequencies.

Instead, a stronger $0.5 \times f_0$ spur is generated since in most cases the bandwidth extends to those frequencies and sufficient power gain is available (Fig. 3). The $0.5 \times f_0$ 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 $0.5 \times f_0$ phenomenon does not usually occur.

The ideal approach is to use a dipler 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 50\( \Omega \)) is provided for the amplifier at all frequencies.

For 1dB attenuation, the resistor would have a value of 440\( \Omega \), while for 2dB the value would become 220\( \Omega \).

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.

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 #30 solid copper wire is approximately 20nH. Component tolerances have no meaning at uhf 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 uhf 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:

\[ C_{\text{nom}} = C(1 + (2\pi f)^2 L)^{-1/2} \]

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:

\[ \begin{align*}
C_{\text{actual}} &= 100/(1 + (2\pi 400)^2 * 10^{-9}) \\
&= 61.3 \text{pF}
\end{align*} \]

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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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°, 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 λ/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 \( V_{\text{rms}}/R_L \) 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 usable 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 current appearing at the forward port of the sample line will be 100mA and the voltage will be \( (PRL)^{0.5} = (0.1 \times 50)^{0.5} = 2.24V \).

Coupling coefficient \( C_p \) can be figured as \( C_p = 20\log[1 + (1/2N^2)]/\log(1/N) \), \( N \) 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 \( (C_1/C_2, \text{Fig. 5}) \). 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 (\( \mu = 100 \) 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 \( \mu = 15 \) 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 attenuator.

![Fig. 6. RF envelope and amplifier output of vswr sensor. Horizontal scale is 2μs/div. Vertical scale is 5V/div.](image)

![Fig. 7. Sensitivity vs frequency response of the VSWR sensor at 5:1 load mismatch.](image)
in some applications a fold-back may be developed for testing a system. For example, 386 specific amounts of load mismatch must be.

Circuit testing

Attenuators, such as a pin diode (Fig. 8), are the only way to control the power gain of bipolar transistor amplifiers since the mosfet 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 R1 and R2 (Fig. 5). For this, and the circuit given in Fig. 10, D2 must be shorted to the output of IC1 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.

For example, if \( f \) is 100MHz, \( C_{\text{min}} \) is 7pF and \( C_{\text{max}} \) is 40pF; then: \( L_2 = \frac{1}{4(628) x 7} = 362nH \) and \( L_1 = \frac{1}{4(628) x 40} = 63nH \). 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 0dB 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

\[
\text{Attenuation} \quad \text{in dB} \quad \text{to produce a specific vswr} = \quad \text{loss} \\
\]

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, \( \frac{1}{4}(6/0.67) \) and \( RL = 10\log_{10}\left(\frac{1}{0.67}\right)^2 \) are needed to simulate loads in laser drivers, plasma generators, communications equipment, and certain medical instrumentation.
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 unbreakable 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 easily 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 hank job via electronic mail makes sure that all its messages are so encrypted. 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?

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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ELECTRONICS WORLD+WIRELESS WORLD May 1994
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.

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 Boardmaker PCB edit screen.

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.

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Symbol Attributes
Kill Definition
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occ
vCC
Schematic close-up. Note how unconnected nodes and wire 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 2.5.

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 or U, 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).
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
Car. £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.
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<td>Marconi 222OV 22MHz power supply</td>
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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, \( \int x^2 \sin x \, dx \) would be input as:

\[
\text{Integrate}\left[\, x^2 \sin(x) \, x, x \right]
\]

and would result in the output:

\[
4x^3 - 6x \cos(2x) + 3 \sin(2x) - 6x^2 \sin(2x)
\]

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 \( t (\pm \epsilon) \) blows up for certain values of \( s \) and \( \alpha \). 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 solved them to reach 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 quantify 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.
Fourier and Laplace domain analysis. Even the most accomplished engineer's intuitive understanding of circuit behaviour can be simulated on a computer. So the performance of various designs can be evaluated quickly and accurately before a single wire has been connected.

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 (IN) 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.

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

\[
\text{NodalAnalyze[chAmp, Result->V2]}
\]

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.

\[
\text{Simplify[NodalAnalyze[chAmp, Result->V2]]}
\]

resulting in:

\[
i\text{IN trz}
\]

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

\[
vAmp=\text{NodalNetwork[CurrentSource[1,0],IN]}, \\ 
\quad \text{Capacitor[1,0,IN]}, \\ 
\quad \text{VVCS[1,0,0,2],gain]}
\]

\[
\text{Simplify[NodalAnalyze[vAmp, Result->V2]]}
\]

giving

\[
-1 \quad \text{--- gain} \quad 2 \quad \text{-------}
\]

\[
\text{Pi CIN ²}
\]

Let us now substitute the cable capacitance: 

\[
\text{Pi CIN ²}
\]

\[
vAmp=\text{NodalNetwork[CurrentSource[1,0],IN]}, \\ 
\quad \text{Capacitor[1,0,IN]}, \\ 
\quad \text{VVCS[1,0,0,2],gain]}
\]

\[
\text{Simplify[NodalAnalyze[vAmp, Result->V2]]}
\]

giving

\[
-1 \quad \text{--- gain} \quad 2 \quad \text{-------}
\]

\[
\text{Pi CIN ²}
\]

Mathematica 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.

**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 there 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 & Chandler Ltd, 36 Victoria Park, Cambridge CB4 3EJ. Tel 0223 321748 Fax: 0223 323804. More information from Macallan Consulting, Tel: 010 1 408 262 3575.

**System requirements**

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Current conveyor ICs — a new building block

There are unique benefits to current conveyor ICs in applications from If 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 current conveyor, the relation between the terminal currents and voltages, and pin-out of the CCI101 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 CCI101 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 CCI101 configured as a 5MHz bandpass filter2 are shown in the data sheet. 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. Current conveyor ICs have unique benefits to applications from If to rf.
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 9996 or nearly 8000. This is reflected in the large difference in $F$ deflection factors for the two traces. Given that 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

![Figure 2. Simplified current conveyor, (a), shows the relationship between terminal currents and voltages. Pin-out of the CCI101 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).](image)

**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,

$$
\frac{V_{BPFF}}{V_{IN}} = \frac{-Y_2 Y_3}{Y_0 (Y_2 + Y_3 + Y_4) + Y_4 Y_5}
$$

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 $jω$. Rewriting the equations in terms of $C$ and $R$ gives,

$$
\frac{V_{BPFF}}{V_{IN}} = \frac{-sC_3 R_3}{s^2 R_2 R_3 C_3 + sR_2 (C_3 + C_4) + 1}
$$

Comparing this with the archetypal form for a bandpass filter,

$$
\frac{V_{BPFF}}{V_{IN}} = \frac{A_s}{s^2 + D_s s + 1}
$$

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 $jω$ term to cancel top and bottom, i.e. a real negative number.

Since $s^2 (jω)^2 = -ω^2$, this occurs when;

$$
\omega^2 R_1 R_2 C_2 C_3 = 1
$$

Feeding in component values from Fig. 3a produces $ω=2πf=14142$ in which resonant frequency $ω_r$ is 2250Hz. At this frequency gain is,

$$
\frac{-C_3 R_3}{R_2 (C_3 + C_4)} = -9956
$$

and $Q^2$ is given by

$$
\frac{1}{Q^2} = D = sR_2 (C_3 + C_4)
$$

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

![Figure 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.](image)
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 amplifier

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 CCI101 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 10.7Ω resistor between the unshielded coils. Symmetry of the response indicates a lack of internal feedback. This is not surprising, since although the reverse isolation of the current conveyor at pins 5, 6 exceeded 70dB at 10MHz the two current conveyors are physically very close.

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/us 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.

Figure 6 demonstrates two further current-conveyor applications. Configuration 6a is synonymous with...
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 $1\,\Omega$ 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 $+1\,V$. Since $1/=-Y$, port $Y$ must also be at $+1\,V$ so there must be $+1\,V$ at the top end of the $1\,\Omega$ resistor. A $1\,mA$ current flows out of port $Z$ but also out of $X$ since $I_{x}=-I_{z}$. As a result, port $X$ exhibits a resistance of $-1\,\Omega$.

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 $53\,\mathrm{dB}$ up to $1\,\mathrm{MHz}$ – neither end of $Z$ need be grounded. The circuit offers a floating negative impedance.

High output resistance of the $Z$ port, typically $1\,M\Omega$ at $1\,\mathrm{kHz}$, 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 CCI101, you should be wary of the simulated grounded inductance shown in the data sheet and covered in the first reference. As expected, this setup 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
2. CCI101 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 textbooks 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 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

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

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 Metrowick 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 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, even if 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 £10.00 in association with the Science Museum.

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### MOTORS - BATTERY 1-12V

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<th>Current</th>
<th>Power</th>
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<td>9V - 18V</td>
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### POWER SUPPLIES - SWITCH MODE

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<tr>
<td>5V - 12V</td>
<td>1W</td>
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### SELECTIVE BARGAINS

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<tr>
<td>Solar Educational Kit with parts to make solar fan</td>
<td>£8</td>
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<td>3V 200mA solar cell</td>
<td>£2</td>
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<td>400mA solar cell</td>
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<td>1A solar cell</td>
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<tr>
<td>9v dc 250mA battery</td>
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<td>12v 500mA on 13A base</td>
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### LASERS & LIGHT BITS

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<tbody>
<tr>
<td>2MW Laser - Holium neon by Philips, full spec.</td>
<td>£30</td>
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<tr>
<td>Amstrad 3&quot; Disk Drive</td>
<td>£6</td>
</tr>
<tr>
<td>ISOLATORS TRANSFORMERS</td>
<td>£1</td>
</tr>
<tr>
<td>200RPM 60W Motor</td>
<td>£5</td>
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<tr>
<td>500W Motor with gearbox and variable speed selection</td>
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<tr>
<td>1/2HP Motor with 1/2&quot; spindle</td>
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### SOLAR CELLS & PROJECTS

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<td>3V 200mA solar cell</td>
<td>£2</td>
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<tr>
<td>1A solar cell</td>
<td>£3.50</td>
</tr>
<tr>
<td>9v dc 250mA battery</td>
<td>£3.50</td>
</tr>
</tbody>
</table>

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MICROWAVES

2: The laws of microstrip

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.

* Mike Hosking spells out the design rules which turn unwanted reactance into a positive asset.
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 $\varepsilon_r$, where $\varepsilon_r$ 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 $\varepsilon_e$ must be used and $\varepsilon_e$ 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.

**Microstrip design equations**

The velocity of propagation, hence wavelength, within the microstrip circuit depends upon an effective dielectric constant $\varepsilon_e$. In turn, $\varepsilon_e$ 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 $\varepsilon_r$.

In practice, the effects of frequency on $\varepsilon_e$ 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:

$$\varepsilon_e = \frac{\varepsilon_r + 1 + \varepsilon_r - 1}{2} \left[ 1 + \frac{10h}{w} + \frac{t}{h} \right]^{1.655}$$

within a certain range of $w/h$ and $\varepsilon_r$.

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 $\varepsilon_r = 2.2$ and the other on 0.125 mm thick GaAs having $\varepsilon_r = 13.1$. The former circuit would have $\varepsilon_e = 1.9$ and a track width of 3.85mm, while the respective values for the latter would be $\varepsilon_e = 8.4$ 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.

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, dependent on thickness and $\varepsilon_r$ 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>
<thead>
<tr>
<th>Material</th>
<th>Tan (10GHz)</th>
<th>εr</th>
</tr>
</thead>
<tbody>
<tr>
<td>PTFE/glass</td>
<td>1 x 10^-4</td>
<td>2.1-2.6</td>
</tr>
<tr>
<td>PTFE/ceramic</td>
<td>2 x 10^-4</td>
<td>9.8</td>
</tr>
<tr>
<td>Alumina (Al₂O₃ 99.5%)</td>
<td>2 x 10^-4</td>
<td>3.8</td>
</tr>
<tr>
<td>Fused quartz</td>
<td>2 x 10^-4</td>
<td>2.3</td>
</tr>
<tr>
<td>Irradiated polyolefin</td>
<td>2 x 10^-4</td>
<td>9.4&amp;11.6</td>
</tr>
<tr>
<td>Sapphire (anisotropic)</td>
<td>5 x 10^-5</td>
<td>13.1</td>
</tr>
<tr>
<td>Semi-insulating GaAs</td>
<td>6 x 10^-4</td>
<td>13.1</td>
</tr>
</tbody>
</table>

**Reactance with a difference**

If we take a length, l, of transmission line having a characteristic impedance $Z_0$ and we terminate the line with some impedance $Z_L$ where $Z_L/Z_0$, 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:

$$Z_{in} = Z_L + Z_0 \tan \beta l$$

where $\beta$ is called the phase constant $=2\pi/\lambda_g$ where $\lambda_g$ 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:

$$Z_{in} = Z_0 \tan \beta l$$

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 ($l/4$), $Z_{in}$ is inductive and can have any value between zero and infinity. From $l/4$ to $l/2$, $Z_{in}$ becomes capacitive and can vary between $\pm \infty$. A similar situation exists for an open circuit termination, for which $Z_L=\infty$. In this case $Z_{in}=\pm Z_0 \cot \beta l$ and may again take on any value between $\pm \infty$ as $l$ varies between $0$ and $l/2$. 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:

$$Z_{in} = Z_L + Z_0 \tan \beta l$$

or $Z_{in} = \sqrt{Z_L Z_0} \tan \beta l$ ...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 $Z_L=Z_0$. 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 1/2oz 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.

**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 \( \frac{1}{2} \lambda \) long and appears, electrically, as a parallel \( LC \) resonant circuit. Individual lines overlap each other by a nominal \( \frac{1}{4} \lambda \) 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
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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, i.e., 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 \( \frac{1}{4} \lambda \) 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 \( \frac{3X}{4} \), while that at Port 3 has travelled \( \frac{5X}{4} \), 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 \( \frac{X}{4} \) and the other \( \frac{5X}{4} \) so they will both be in phase at this port and will thus partially couple out.

At Port 3, one signal will have travelled \( \frac{X}{2} \) and the other \( \frac{X}{2} \) 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 \( \frac{X}{2} \) 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 \(-10\) dB.

The previous coupler types have the disadvantage that they are inherently narrow band devices, 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 3 dB 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 I.

A microstrip receiver circuit which brings together several of these elements is shown in Fig. 10. The substrate is alumina, about 0.3 mm thick and most of the tracks are 50 Ω 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.

Further reading


Next Month:

- lumped components and waveguides.

---

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

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- Twin Twin PPM Rack and Box Units.
- PPM5 hybrid, PPM9 microprocessor and PPM8 IEC/DIN ~50/+6dB drives and movements.
- Broadcast Stereo Coders.
- Stereo Disc Amplifiers.
**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 current 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.

**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 \( R_d = \frac{2\pi f_0 L}{Q} \). Bias current in the OTA, set by \( R_{abc} \), determines \( g_{mn} \), which is about 2mA/V for an \( I_{abc} \) of 1mA. Oscillation at \( f_0 \) takes place when \( g_{mn} R_d > 1 \), 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 \( 4R_d/abc/1t \) and third-harmonic distortion of \((100/8Q)\%\)

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

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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

$$\frac{v_{C1}}{v_{R3}} = \frac{1 - e^{-t_0/RC}}{R_2 - R_1}$$

$$v_{R3} = v_1 R_3 \left( R_2 - R_1 \right)$$

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

$$t_0 = -RC \ln \left( 1 - \frac{R_3}{R_2 + R_3} \right)$$

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 input 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 diil 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 688s 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.
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

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 1.5, 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. 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

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

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

Analog 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 triggers 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.

M S Nagaraj
ISRO Satellite Centre
Bangalore
India

Electronics World + Wireless World May 1994
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(son) 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
Croiles
France
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 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.

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 the TJES40 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 demonstrate its production with 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 other deficiencies such as feedback-coupled distortion. Ben says he accepts the effect must be at a very low level as it is invisible at 0.0006%. I remember that this is the level of a THD reading that is visually pure noise; though there are

10Hz AP out | Amp out | 1kHz AP out | Amp out
---|---|---|---
Fundamental | 0.0011% | 0.0031% | 0.0012% | 0.0035%
Second | 0.0031% | 0.0092% | 0.0008% | 0.0060%
Third | 0.0035% | 0.0050% | 0.00031% | 0.0024%
Fourth | <0.0002% | <0.0035% | <0.00008% | 0.0048%
Fifth | <0.0025% | <0.0045% | 0.00014% | 0.0024%
Sixth | <0.0006% | <0.0030% | 0.00008% | 0.0021%
Seventh | <0.0006% | <0.0008% | 0.00009% | 0.0009%
Eighth | <0.0003% | <0.0003% | 0.00008% | 0.0016%
Ninth | <0.0004% | <0.0011% | <0.00007% | <0.0008%

AP THD reading: 0.0046% 0.0095% 0.0060% 0.00117%

(80kHz bandwidth)

The same type of fundamental is not perfect, and this is shown as it contributes to the THD figure.

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 off.

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 unsupplied 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 unsupplied regulated output is driven into continuous clipping, but what has this to do with the finer nuances of
termination?

I also wondered when I read that “bass clarity has been shown to be improved by X-exchanging line 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.

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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: the Audio Precision is a brilliant piece of machinery, and that the amplifier is really quite linear too. However, there is nothing revealing 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 Vbe multiplier transistor.

The size of these crossover spikes gives in effect a continuous readout of driver temperature, and the slow variations show some time constant measures 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. This is a strictly a slow (seconds) 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, p.822, EW + WW, October 1992), by an order of magnitude.

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 audio audio, I recent 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 amplifiers 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 30kHz, and not a wholly unrealistic one with some topologies and real-world PSR-reducing compromises found in FA and up-market hi-fi amplifiers.

Douglas Self has shown supply regulation may not be needed to get quite so low static THD+N figures when driving an XLR with his own ideally 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, capacitive 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. 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?

Convincing improvements in bass clarity, the MLS-Rauch 6000 is a 600W/4Ω/channel amplifier with an 8kHz switching supply, but otherwise its audio path replicates Rauch Precision's older DGT 2515s for which there is a conventional supply. Both amplifiers have had an enduring presence in the UK and European live performance, identical and recording studio circuit, and over the past seven years, a number of experienced sound engineers have noted the improved soundscapes 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 dumb to even notice.

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 judged 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 a result 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 hard bit more carefully, he might gather that thermal distortion is the industry’s best working hypothesis with regard to certain A-B tests between nationally 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 формы and concise, diverse and all. Informed knowledge and a sense of the fundamental simplicity, and not using the famous seasickness capacitor location to compensate. His series is much respected in the power floating world.

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 far too little, I understand the profound satisfaction he gets from reading, diving, and most of all half-baked information into simplicity, and not using the infamous seasickness capacitor location to compensate. His series is relatively unimportant for teaching.

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By the time a truly modern design would even be considered, much of what might 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.

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 representation 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 I 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 Bosw (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 Iall
Staines, Middlesex

LETTERS

(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 Quinlon (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, Monte 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 retrace 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) states that many attempted proofs of the theory of relativity have been shown to be based on false evidence. Well I suppose that if physicists keep eating fudge while making measurements, the fudging 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 infringed with paradox.

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 inversely 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 Poincare).

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, Israel

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 naiive 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 both wide and complex interests.

I use a single small control file M.bat - 1 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 - e.g. 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 the time waiting for completion. At the other extreme gui slowness may mean the expense of extra computers to meet a work load.

RG Silon
Tring, Herts
**NEW PRODUCTS**

**ACTIVE**

**Asics**

3V embedded array. Toshiba’s new 0.5um cmos embedded array combines the functionality and performance of standard cells with the faster turnaround times of gate arrays. With a typical gate propagation delay of 0.25ns and gate power consumption of 2.2pW/MHz, TC180E outperforms 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.

**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 Rs, on values of 10mΩ and 12mΩ. These devices are operating from a 5V logic-level drive as well as the standard 10V. International Rectifier, Tel., 0883 7132; fax, 0883 714254.

Dmos power mosfets in the 14-pam Nexedge family of dmos power mosfets from National Semiconductor are single and dual n-channel and p-channel types and are 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.<, National Semiconductor, Tel., 0104 9141 10330; fax, 0104 9141 10351.

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 33148; fax, 0932 33148.

**A-to-D and D-to-A converters**

Stereo codecs. AD1847 and AD1846 are variable-gain system containing two amplifiers with voltage inputs and a single-ended current output, 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.

**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 LP2951-3.3.5 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 11µJ, 150µJ and 300µJ and initial voltage tolerance ±1%. Gothic Crellon Ltd., Tel., 0734 788878; fax, 0734 776055.

3V RF amplifiers. 1.9GHz, 3V range amplifiers in NEC’s pCP2745T-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-micr measuring 2.9 by 1.5 by 1.1mm and are made in NEC’s 20GHz NESA III silicon bipolar process. NEC Electronics (UK) Ltd., Tel., 0906 691153; fax, 0906 670590.

**Low-power Rx IC**

Temic Telefunken has a family of low-power UHF receiver ICs intended for car security, wireless fobs and appliance control that take a standby current of only 1mA. The U431KB series operate at 433MHz. A modulated carrier is converted by a UHF stage to 10 MHz and amplified by the U 431 X. Its log. amplifier acting as demodulator for AM, FM being handled by a quadrate detector. Digital data is regenerated at the baseband by a clamp mixer and op-amp, the universal output interfacing to a decoder. Temic Telefunken GmbH, Tel., 0104 9171 687274; fax, 0104 9171 99334.

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 CXA1602MN, which is a 100MHz wideband 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 CXA1293MN 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

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Logic building blocks

Fibre-channel transceiver. METL's RCC760 is a cmos transceiver, operating at up to 265Mbaud, which integrates BB10B encoder/decoder, serialiser and deserialiser, a PLL synthesiser, a PLL clock and data recovery and a byte align circuit. PCD evaluation assemblies are available for fibre or coaxial interfaces. Microelectronics Technology Ltd. Tel., 9844 278761; fax, 0844 278746.

Memory chips

Fast srams. Cosmos 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 655mW, 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 Am386e processor in two embedded versions, the Am386DE and Am386ESE. Any 86X PC can be used as development platform, the Fusion680 support program providing software and hardware from more than 60 suppliers, including Microsoft's Windows operating software. Both versions work at either 5V or 3V, with standby modes. Am386DE, the 386MHz type, is available now, the SE following in a few weeks. Advanced Micro Devices (UK) Ltd. Tel., 0483 740440; fax, 0483 66186.

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 HDS2-2 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, red.

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coefficient of 15 ppm/°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 3.3V. Guaranteed input current of National's Femtoamp-input opamp. Polar Electronics. Tel., 0525 377093; fax, 0525 378834.

16-bit i/o-series micro. NEC's V55PS 16-bit, single-chip microcomputer with 16MBbyte 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 V55 and higher-performance peripherals include two UARTs, operating at 780kb/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 A/D converter, a PWM operating up to 24kHz, a watchdog timer, four 16-bit timers and a 16-bit software interrupt timer. Sunrise Electronics Ltd. Tel., 0908 263999; fax, 0908 263500.

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 1ns on a 5V supply or 8ns on 3.3V. Most instructions execute in a single cycle using a 5-stage pipeline to give a 16 ns Dhrystones 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 35μA and 8kbyte of on-chip ram, and the SH7034 with 64kbyte of on-chip ram. Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

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 a 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, videotex and modern functions take the system into telecommunications and picture compression and transmission, computer audio and graphic user interfaces link consumer electronics, telecommunication 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 1150mcd 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 276892.

Two-colour photodiode. The Tandem two-colour photodiode detector combines a 2.5mm diameter silicon diode and a 1.7mm 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 850, 880 and 940nm. Different InGaAs diode diameters are available. A specification is mentioned. Aerotech World Trade Ltd. Tel., 0628 34555; fax, 0628 781070.

**Fast power diode.** Ten members of Harris's Hyperfast RHA series of power diodes exhibit reverse recovery times of 60 to 100ns. These devices have breakdown voltages of 1200V and η, of 75ns at 30A, one of them being offered as a dual device. Two 75A types are 1200V rated and have η, of 100ns; three lower-voltage types recovering in 60ns. A 150A type breaks down at 1000V, with a η, of 25 to 60ns. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 683232.

**Reverse blocking switch.** Replacing two mosfets and their drive circuit, the Silicox Si9718CY is a reverse blocking switch for battery discharge 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 parasitic diodes found in standard mosfets, reducing on resistance to 80mΩ at 3.3V. The device also includes a charge pump and enable circuitry, and under-voltage lockout protects the system. Silicox/Temic Marketing. Tel., 0344 485757; fax, 0344 427371.

**Low-power PWM controller.** Unitrode claims a first for its UC3570 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 1mA 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 100mA. Semtech Ltd. Tel., 0592 773520; fax, 0592 774781.

**Wavelength 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-dash-an in a 10MHz clock and 12-bit resolution, incentives 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 then displayed. Marion Instruments Ltd. Tel., 0494 459200; fax, 0494 335002.

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**Passive components**

**Crossover electronics.** Bipolar electrolytic capacitors in Nichicon's DB GB series are meant for use in audio crossover networks and have values from 1μF to 8μF in 220μF (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance. Working voltage is 50V, leakage current ±1μA and ±5% (DB) and ±10% (GB) tolerance.

**High-temperature ceramic capacitors.** Vitramon's H range of chip capacitors uses the XBR 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 of ±5% to ±20%. Insulation resistance at 25°C is 100GΩ and a minimum of 10GΩ at 150°C. Vitramon Ltd. Tel., 0286 524933; fax, 0286 525435.

**Vertical indentors.** In values of 0.47μm at 1000μm at ±10% standard tolerance, microSpire MIVI-PK 0455 series of vertically mounted indentors save board space by offering diameters of 5 to 13mm and a height of 11 to 17mm. Se-I resonant.
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 1000μF for the 0603, 470μF in the 0805 and 1μF for the 1206 types are available, all with leakage currents given by an insulation resistance of over 1TΩ or 500MΩ/F, 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, Boyschagl surface-mounted resistor networks are available in quantities down to 100 pieces. The thin-film resistors are the Boyschagl Micro-MELF type, welded to a metal frame which is shaped to form the circuit. Values in the range 1μ to 2MO 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 510275.

Connectors and cabling. Adaptable board connector. 65.536 different connection patterns are possible in a housing smaller than a dpa package with through connector. 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 capacitor filters has been adopted by Wainwright Instruments GmbH in Germany in their range of ultra-stable notch filters for GSM, PCN and other mobile comm 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 236600; fax, 0444 236641.

Shielding can. Made in tin-plated steel. West Hyde's/isolator is a board-mounted shielded attach to the board by four solder prongs set on 0.1in pitch. The one-piece unit is fully soldered to the internal faces to aid high level potential. 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 738181; fax, 0453 866667.

Instrumentation. GPB test gear. Thurby Thandar has add-ons to its range of rack-mounted, GPB test equipment, including power supplies, multimeters, generators, logic analysers and oscilloscopes, which are in 2US. All instruments conform to IEEE-488.2 in addition to IEEE-488.1. Thurby 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 or as part of an ATE system, the Gigaga-ronics 60X series of RF signal generators covers the 100kHz-2.1GHz range in three models: 6060/1, the 60602 having fast-rise pulse modulation for transmitter and radar testing. As standard, a 10MHz crystal oscillator gives ±5X/10MHz and warms up in an hour. Alternatively, an optional oven-controlled reference oscillator at ±5X/10MHz and warms up in 30min. Sematron UK Ltd. Tel., 0734 819709; fax, 0734 819786.

Micro-based thermometers. ETI's Microtherm 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 celcius to fahrenheit. Electronic Temperature Instruments Ltd. Tel., 0903 202191; fax, 0903 202445.

Auto timebase oscilloscopes. Hitachi's V0105 and V065, 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 autobrake in the usual modes. Thurby 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 Microvib CM/410 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 links. 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 characteristics, monitors Ethernet traffic and measures inherent noise over time to locate intermittent trouble. An autoroute 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. 0644 787060; fax. 0644 785573.

Literature

Audio products. Crystal's complete range of audio ICs is described in a 102 page book. Information includes all specifications, block diagrams, theory of operation, applications and schematics. Devices detailed are data converters and codecs, DSPs and synthesizers, digital audio transmitters and receivers and volume controls. Crystal Semiconductor Corporation. Tel. 0111 512 4427555; fax. 0111 554 4351.

Power supplies

Rechargeable 996. A direct replacement for the 996 dry battery. Power-Sonic's 6V, 5Ah sealed lead-acid battery type PS600 can be recharged up to 1000 times and includes an overload self-resetting cut-off. The combined 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 memfet or hemt devices and most of them incorporate temperature compensation. Many are in the low-cost quasi-MMIC form, in which all components are integrated and the active devices bonded separately. Custom designs incorporating limiters, special filters, variable gain etc., are undertaken. Connection options include standard worldwide, coaxial and microstrip. Anglia Microwaves Ltd. Tel. 0277 630000; fax. 0277 631111.

Switches and relays

Solid-state relays. Series RA relays from Teledyne handlode handle voltages up to 250V RMS from 49V to 440V AC, with positive latching and pure power devices down to 0.2. Inverse parallel SCLRs, 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 recycle and case temperature return 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 Controls offers a vision 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 50m and a carring factory system in which cans are checked for damage. Pro-Active Control Ltd. Tel. 0223 300801; fax. 0223 300979.

Thermocouple data acquisition. DI-21TC from Datag is a 12-bit, 16-channel portable data acquisition system, optimised for grounded thermocouples but suitable for other analog inputs to the built-in signal input. A sensor on the PCB signal terminal receptacles providing cold junction compensation. The instrument line diagram software automatically in real time with DSP-based, 10th-order polynomial calculation software, voltage inputs being automatically converted to temperature readings in the user's choice of range, either ±1200°C or ±120°C. It plugs into the parallel port of a PC and has a hard disk battery. Keithley Instruments Ltd. Tel. 0734 575666; fax. 0734 566469.

40MHz DSP board. LS1 says its PCM/C-3140 applications board is the first to be based on the Texas Instruments TMS320C31 40MHz digital signal processor. It is a two-thirds-length AT PC card and comes with a complete set of software 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 LS1's Windows-based View31 tool and libraries of C-called DSP and maths functions. Loughborough Sound Images Ltd. Tel. 0509 231843; fax. 0509 238243.

Computer systems

Industrial computers. SealTouch computers by the American Deeco company are high-performance units in panel-mounting configuration and proved against hostile environments. CPUs are 486DX or DX-2 and control 386CX and EX embedded in-circuit emulators for the Intel 80386/CX ICE. Hitex announces full in-circuit emulators for the Intel 80386/CX and EX embedded microprocessors, using the existing PC-based T122/360SX emulator as a base to allow true ICE of the central processor at up to 25MHz in 100G0P and MOPP packages. The H10C360 debugger is now extended to support the 80386's new peripherals and 16-bit code by Microchip, Borland and Intel are supported. Real-time debugging and analysis are enabled. Telek (UK) Ltd. Tel. 0203 692066; fax. 0203 692131.

Development and evaluation

386CX/EX ICE. Hitex announces full in-circuit emulators for the Intel 80386/CX and EX embedded microprocessors, using the existing PC-based T122/360SX emulator as a base to allow true ICE of the central processor at up to 25MHz in 100G0P and MOPP packages. The H10C360 debugger is now extended to support the 80386's new peripherals and 16-bit code by Microchip, Borland and Intel are supported. Real-time debugging and analysis are enabled. Telek (UK) Ltd. Tel. 0203 692066; fax. 0203 692131.

Data communications

Parallel port data acquisition. About the size of a laptop computer, Parallel Pad by Computer Instrumentation acquires data for a PC 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 701156; fax. 0903 702001.

Portable data acquisition. Containing not only analogue interfacing circuitry for on-site data acquisition, but also a 486 25MHz PC, Onsite's Techstation-ct 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 measuring instrument and printer port of a PC and has its own processor board. There is also the TI floating-point C compiler and debugging is assisted by LS1's Windows-based View31 tool and libraries of C-called DSP and maths functions. Loughborough Sound Images Ltd. Tel. 0509 231843; fax. 0509 238243.

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 850800 (evenings).

Mathcad updates. Two new versions of Mathcad technical calculation software have appeared. Mathcad 50 and Mathcad PLUS 5.0. The former has been made easier to use, with eased eration 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 all 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 837461.

Cable communication systems

Switches and relays

Solid-state relays. Series RA relays from Teledyne handle handle voltages up to 250V RMS from 49V to 440V AC, with positive latching and pure power devices down to 0.2. Inverse parallel SCLRs, 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 recycle and case temperature return to normal, the trip being signalled by a status output. Teledyne Electronic Technologies. Tel. 081 571 9596; fax. 081 571 9637.
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 150uA.

As the first set of diagrams show, the TO220-packaged LP2957 can act as a basic 5V three-terminal regulator but with shut-

At light loads, quiescent current of the LP2956A low drop-out regulator is 170uA, making it suitable for battery-power applications. Shown here are both 5V fixed 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.
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 and MV6639.

Operating at 153MHz, the system is a direct-conversion receiver that converts frequency-shift-keyed FSK into digital data for decoding. Output data is 512 baud tone data. All filtering is carried out on chip and two reference voltages are available for RF amplifier biasing.

Within the MV6639 POCASG* 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, \( L_4, C_{11} \), and \( C_{12} \), is optimised for noise figure and best overall device sensitivity.

In the local oscillator, \( L_4 \) tunes out the crystal capacitance and suppresses oscillation at the fundamental frequency. It may need tuning to suit the crystal.

Circuitry around \( T_{25} \)'s collector is designed for resonance at 153MHz. Output feeds an RC quadrature network comprising \( R_{16}C_{18} \) with \( R_{17}C_{19} \) 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 \( T_{24,5} \) to isolate receiver data output and the POCASG decoder data input. On leaving the buffer, POCASG data feeds the MV6639 decoder, whose reference is a 32kHz watch crystal. Radio identification code, bit rate selection

Previous page indicates.

Ericsson Components AB, Energy Systems Division, 516481 Kista-Stockholm, Sweden. Tel. +468 721 6356, fax 721 7001

*POCASG

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 alternating 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 SW2.

Antenna matching shown provides the necessary 50Ω source impedance. With it is a curve showing the dip it produces at 153MHz. Variable capacitor VC1 is capable of tuning the dip to within at least ±5MHz of the resonant frequency. Coupling capacitor Cy 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.

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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.

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,
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 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 left alone those categorised as a solution looking for a problem.

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 pounce 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 invalidate 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 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 invoking 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 busi-
ness 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 tech-
nical 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 accep-
tance. 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-pro-
duction and marketing costs; does not divert
resource from existing profitable operations,
and that the terms of acquisition are not unac-
ceptable.
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.
A deal is a deal?
Inventors usually become frustrated with the
tedious business of ipr transfer and the appar-
ten 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, particu-
larly 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 advan-
tage 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.
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 trawling around the country listening
to constant refusals. Companies usually move cautiously and a
decision not to proceed after months of nego-
tiation can take the fight out of even the most
enthusiastic of inventors.
The bird in the hand, though not good poli-
cy, is all too common and the burden of per-
sonal involvement and sundry costs can quick-
ly exhaust the patience.
But instead of despair the exercise should be
reconsidered in terms of whether the objec-
tives 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 newspa-
ters 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.
But 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 ama-
teur 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 constraints typical negotiation.
Few people realise just how much resource
is needed when investing in a new idea, par-
ticularly 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, underesti-
nated.
Risks are large, with costs ranging from the
staff needed, to the set-up costs. Time spent by
personnel in establishing the pro-
duction 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 famil-
ilar 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

**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 "share risk" basis, giving a licence to the
recipient and placing a duty on them to get
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 com-
nences. Whatever happens, royalties will be
as 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
conventions. In the case of a non-product
orientated deal (process or cost saving ideas)
the inventor receives a part of the cost saving.
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 instantaneously and confer 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, 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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<th>Price</th>
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<tr>
<td>HP8510B 10MHz-22GHz Microwave Spectrum Analyser</td>
<td>£9550</td>
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**High Performance features include:**
- External mixing to 11GHz
- Internal Preselection
- 1.7-22GHz
- Wide resolution range 100Hz-5MHz
- On-screen readout
- Chipset output
- Digital storage
- IP output: +20dBm, HP8.

### TEST EQUIPMENT

<table>
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<tr>
<td>Tektronix 492 22GHz Spectrum Analyser</td>
<td>£8500</td>
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<tr>
<td>Tektronix 3751 A/B/C 24-channel sync input analyzer (W/A) with 3752B, 3753, 3757 RF downconverters (1.1-1GHz)</td>
<td>£4000</td>
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<tr>
<td>Tektronix 4193A 5GHz Firewire analyzer</td>
<td>£7000</td>
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<td>Tektronix 3796B 8 GHz frequency counter</td>
<td>£5000</td>
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<tr>
<td>Tektronix 4324B 6GHz optical power meter</td>
<td>£6000</td>
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<td>Tektronix 334A-4 microwave power supply</td>
<td>£2000</td>
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<tr>
<td>Tektronix 8505A digital photometer</td>
<td>£1500</td>
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<tr>
<td>Tektronix 8508A automatic modulation meter</td>
<td>£1000</td>
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<tr>
<td>Tektronix 8502B network analyser system including 8503A S-parameter test set and 8406A comb generator</td>
<td>£2000</td>
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<tr>
<td>Tektronix 8500A network analyser system including 8505A S-parameter test set and 8514 A/B waveguide</td>
<td>£3000</td>
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<tr>
<td>Tektronix 8503A network analyser system including 8504A S-parameter test set and 8511 A/B waveguide</td>
<td>£4000</td>
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<td>Tektronix 8502A network analyser system including 8501A S-parameter test set and 8504 A/B waveguide</td>
<td>£5000</td>
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<tr>
<td>Tektronix 8501A network analyser system including 8500A S-parameter test set and 8504 A/B waveguide</td>
<td>£6000</td>
</tr>
<tr>
<td>Tektronix 8500A network analyser system including 8503A S-parameter test set and 8514 A/B waveguide</td>
<td>£7000</td>
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**Please note:** All our equipment is now operationally verified tested before despatch by independent laboratory.

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Optional software support for Microsoft Windows includes a complete Dynamic Link Library (DLL) and two icon driven packages, Signal Centre for signal capture and analysis and TRACS for process monitoring, visualisation and control.

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The board has 48 lines of TTL compatible digital I/O provided as six 8 bit ports four of which can be programmed to be all inputs or all outputs and two which can be split to be 4 inputs and 4 outputs if required. There are also three 16bit counter timers which can be used to generate and measure pulses at up to 5MHz.

An on board 4MHz crystal oscillator is also included on PC 14AT along with LEDs showing the status of a selection of the I/O ports, these can be used as programming and debugging aids.

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