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FEATURES

Cover: Jamel Akib

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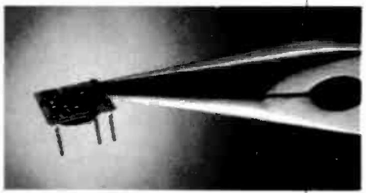
Low Power Ratio Solutions is offering a remote-control transmitter-receiver pair to 1000 readers of *EW+WW* at £11.75 inclusive, instead of the usual price of £19.50. Details on pages 1046 and 1051.



REGULARS

In next month's issue:

Piezoelectric cable – signal conditioning circuits
Audio output stages – reducing distortion
New null filter – simpler alternative to twin-tee featuring low distortion

JANUARY ISSUE ON SALE DECEMBER 29

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Your design ideas are quickly captured using the ULTIcap schematic design Tool. ULTIcap uses REAL-TIME checks to prevent logic errors. Schematic editing is painless; simply click your start and end points and ULTIcap automatically wires them for you. ULTIcap's auto snap to pin and auto junction features ensure your netlist is complete, thereby relieving you of tedious netlist checking.

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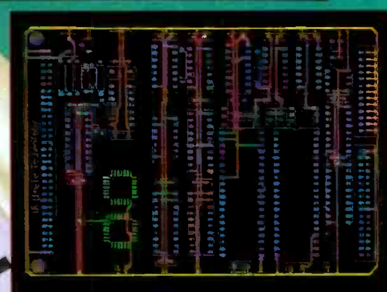

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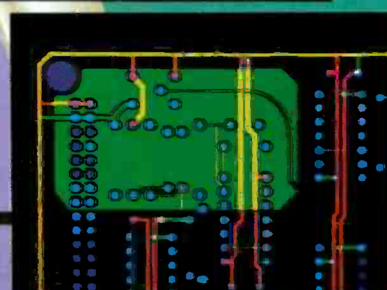

If you need partial ground planes, then with the Dos extended board systems you can automatically create copper polygons simply by drawing the outline. The polygon is then filled with copper of the desired net, all correct pins are connected to the polygon with thermal relief connections and user defined gaps are respected around all other pads and tracks.

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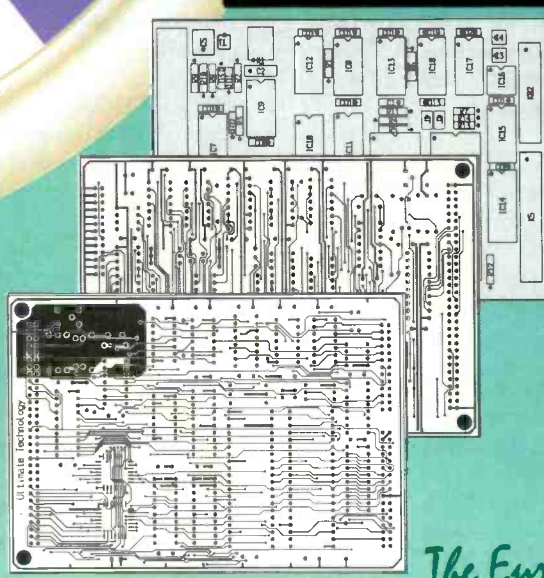

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Thinking ahead

There were two woodcutters, walking into the forest at the start of a day's work. The younger one decided to spice up his day with a small wager, "I bet that I can cut up more logs in a day than you can". The older woodcutter agreed to the challenge and they both started work.

At the end of the day the older woodcutter had cut up far more logs than what the younger one had. "I just don't understand it", said the young man, "every time I looked to see how you were doing, you were having a break".

"Yes," said the older man, "but what you didn't see was that, each time I rested, I also sharpened my axe!".

I do not claim any originality for this story, but the moral behind it is to make sure that your tools are fit for the job. An engineer's best tool is his brain. The basis of the IEE's continuing professional development program is to make sure that engineers do not rest on their laurels. In the future it is possible that, for an engineer to keep his professional status, he will have to prove that he has been keeping up-to-date with new developments. But there is more at stake than status - it is future employment.

Some employers think they are getting more out of their employees by flogging them to death. They think that by employing the 'young woodcutter', working away at a furious rate, they are somehow getting more work done. Wise employers know that, at the end of the day, the young woodcutter will be tired and make mistakes, and before long his tools will be blunt and useless. The wise employer will make sure that their 'woodcutter' regularly keeps his tools sharp, and can therefore do a better job. Although product quality is not important in the simile used in the story, it is important in the world of electronics.

If predictions are correct, more and more people will be working part-time or on short term contracts (see the leader in IEE News 1 Sept 94). Employees will become a commodity, with the most able commanding the highest salaries and greatest future opportunities. In order to become a desirable commodity all engineers should seek opportunities to improve their knowledge. Attending seminars, conferences and tutorials, whether organised by component manufacturers, institutes or others is a valuable source of new ideas. Reading the latest books and technical journals also helps.

If you consider yourself knowledgeable in your specialism, try learning about areas that are not so familiar. Hardware engineers could try to learn a programming language, analogue engineers could learn more about digital techniques. Cross fertilisation can produce novel solutions to old problems. New ideas may help to improve the fortunes of your company. But ideas do not just appear because you think hard, they are formed from many small pieces of knowledge that your mind puts together like a puzzle and, unless you have all the pieces, you cannot find the answer.

Training takes time and money, and a good employer will usually have a budget for training their greatest asset; their employees. An employee's manager is perhaps the best person to identify suitable training, but how many managers have no conflicting pressures? A manager may have the best intentions towards his staff, but fails to train them properly because he cannot spare his time or his employee's time. Often it is up to the employee to help himself, by identifying a suitable training course or seminar. He should also think of a persuasive reason why he should attend: how will it improve performance and save money in the long run. These arguments can then be used by the busy manager to argue for funding from the employer's finance department.

A reasonable training fund should be provided by employers; but this may be impossible if the company is not doing very well. However, training does not have to be expensive; some component manufacturers provide free or low-cost seminars, these are often very useful and full of applications. Employees could help increase their knowledge by reading trade magazines - even magazines aimed at basic hobby electronics can provide valuable information. Text books can be borrowed from libraries or colleagues. Although low-cost training is not an ideal way to learn, it is better than nothing. There is a saying "you learn something new every day", try to make sure that you do. **Steve Winder**

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Electronics World + Wireless World is published monthly. By post, current issue £2.25, back issues (if available) £2.50. Orders, payments and general correspondence to **L333, Electronics World + Wireless World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.** Tlx:892984 REED BP G. Cheques should be made payable to Reed Business Publishing Group.
Newstrade: Distributed by Marketforce (UK) Ltd, 247 Tottenham Court Road London W1P 0AU 071 261-5108.
Subscriptions: Quadrant Subscription Services, Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 0444 445566. Please notify change of address. Subscription rates 1 year (normal rate) £30 UK and £43 outside UK.
USA: $52.00 airmail. Reed Business Publishing (USA), Subscriptions office, 205 E. 42nd Street, NY 10117.

Overseas advertising agents: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine, Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd, 205 E. 42nd Street, NY 10117.Telephone (212) 867-2080. Tlx 23827.
USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. 2nd class postage paid at Rahway NJ Postmaster. Send address changes to above.
Printed by BPCC Magazines (Carlisle) Ltd, Newtown Trading Estate, Carlisle. Cumbria, CA2 7NR
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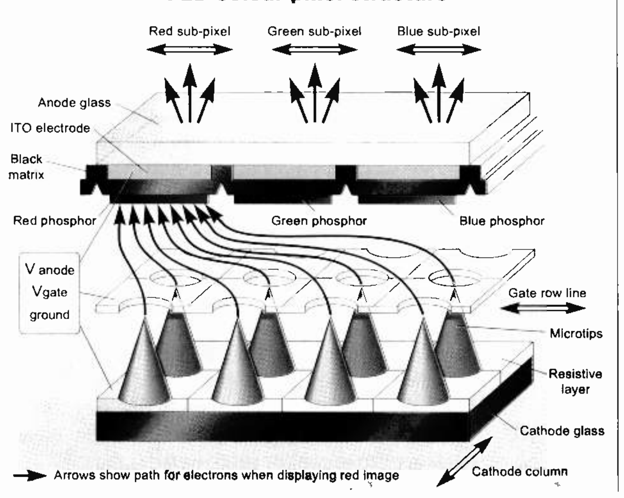
The future for field emission displays

Theoretically, field-emission displays can match the luminosity of conventional crt displays while consuming a fraction of the power.

Last year US technologists believed that the field emission display (FED) was the true flat panel alternative to the lcd for large area displays. Times – and thoughts – have changed. The FED is a relatively immature technology which has yet to be put into any sort of volume production.

FED colour pixel structure



Theoretically, FEDs can match the luminosity of conventional cathode ray tube displays while consuming a fraction of the power and being less than 2.5mm thick. At the Society of Information Display (SID) symposium in the US in June, Raytheon demonstrated a prototype FED which delivered a brightness of 10,000 ft-lamberts, or 34,000 candelas/m^2 and was 50 percent more power-efficient than monochrome crts.

This was only a small 128×128 pixel prototype that measured less than two inches square. Raytheon has set a target date of 1996 to turn that FED technology into a commercially manufactured display. On the other hand, FED flat panel technology pioneer, Pixel Technology of Rousset, Southern France, claims that it will be in volume production of a 6in colour FED next year.

The company was set up in 1992 to develop and market an FED technology developed by the French Atomic Energy Authority (CEA). This technology has been licensed to Texas Instruments and Raytheon in the US and to Futaba, the Japanese fluorescent display maker.

Pixel is also moving ahead with its own product development plans. It demonstrated a 6in colour panel in the US in July 1993 and is due to open its first FED manufacturing plant in Montpellier next month, representing a total investment of \$17m. Pixel plans to be producing its first 6in colour displays in the second quarter of 1995, with 10in by the end of the year and 12in in 1996, which may prove over-ambitious.

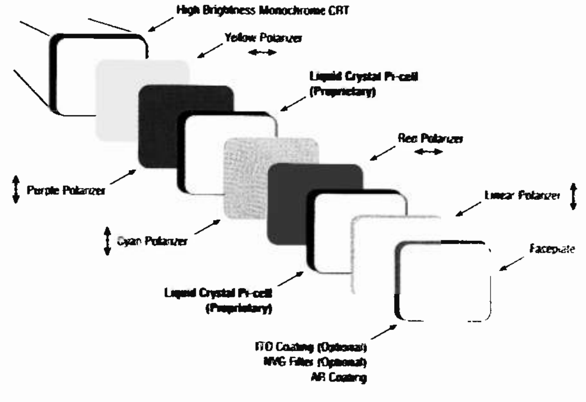
Pixel's FEDs rely on a unique micro-tip cathode (see diagram). Hundreds of 1μm microtips (25,000/mm^2) are required to generate the cold electron emission which illuminates each pixel of the display. A 6in display with 256×256 pixels will have millions of microtips fabricated on a glass substrate.

Even with a level of redundancy built-in this structure will not be easy to manufacture in volume cost effectively. What is suggested to be a more cost-effective alternative has

CRT and liquid-crystal technologies combined

In this new lcd technology, liquid-crystal optical shutters operating at high speeds determine display colour.

Tektronix has reconsidered the design of the earliest electromechanical colour crts and combined it with today's liquid crystal technology. The result is a technique that combines high resolution monochrome crts with an active high quality colour filter.



RGB FULL COLOR, FIELD SEQUENTIAL, LIQUID CRYSTAL SHUTTER STRUCTURE

The technique can be applied to small displays used in instrumentation and workstations just as in the large area displays in aircraft cockpits and air traffic control that are expected to form its initial markets. The potential of the technology, known as the *NuColor* shutter, convinced US display specialist Planar Advance, which earlier this year bought the Tektronix displays group that invented it.

Early electro-mechanical colour crts produced their colour images as a result of the user viewing a black and white screen through a revolving red, green and blue (rgb) colour filter. The new system replaces the mechanical filter with a liquid crystal filter which can be switched at high speed. The rgb filter sub-system consists of two liquid crystal optical switches, known as Pi-cells, which sit in the middle of a sandwich of colour and neutral polarisers. The whole filter sub-system can be fitted .to what is in essence a standard high resolution monochrome crt.

The colour polarisers are used to orthogonally polarise and separate the rgb components of the crt's emission. Under the direction of a drive voltage, the Pi-cells rotate the polarity of the light through 90 degrees. The drive signal to the Pi-cells is derived from the crt's drivers and the filter's switching is synchronised to the display of each of the rgb colour elements on the crt. In this way the colours of the image are built up sequentially, reds followed by greens, followed by blues. The different colour tones, obtained by varying the intensity of been developed by SI Diamond Technology, of Houston, Texas. This uses thin-film diamond cathodes instead of the microtips. The company also claims that the diamond cathode design uses less power, has longer lifetime and is more easily scalable than the microtip design. However, SI's first prototype FEDs were only an inch square with 125×125 pixel resolution and luminous efficiency was only a tenth that of conventional crts.

From Pixel's microtip design, Raytheon has demonstrated a 6in FED with a 500V drive voltage. It has also designed a higher voltage version that can achieve 10,000 ft-lamberts of brightness by increasing the distance between the microtip cathodes and the luminescent phosphors so that an anode grid can be used to focus the electron beam. This design outstrips the brightness efficiencies of crts and the US company claims that it will be easier to manufacture. On the other hand, the anode potentials are increased and at five to 10kV are similar to those of crts.

With Sharp's demonstration of the first 21in active matrix lcd recently, it seems unlikely that FEDs will provide a strong challenge in the flat panel market before the final years of the decade. Enough companies are developing the technology to make it a strong bet for the future, however.

Richard Wilson,
Electronics Weekly

the original monochrome light information, are integrated by the viewer's eyes.

The integration process means that a 180Hz scan rate on the rgb colour fields results in a 60Hz frame rate for the complete colour picture.

Planar Advance has demonstrated the technique on small area 1in square displays right up to 19in diagonal crts with 1280×1024 resolution and a high contrast ratio of 100:1 at 42 ft-candelas. The colour filters use light energy from a single high energy electron beam in the crt, which creates areas of uniformly saturated colour. This is particularly critical in high quality radar screens, but may also find an application in hdtv and virtual reality display systems. ***R.W.***

Government funds cable research

The UK Government is to begin funding research into technology which will allow cable operators to realise the full potential of their growing broadband networks.

An r&d programme jointly funded by the Department of Trade and Industry and the Cable Communications Association will be spearheaded by Eugene Connell, chief executive officer of one of the UK's largest cable operators, Nynex CableComm, who is also chairman of the Cable Communications Association. The plans were revealed at the opening of the European Cable Communications show held in London recently.

Software will be developed to take full advantage of the broadband systems being built by cable operators as part of a high level DTI programme aimed at creating some form of broadband policy similar to those in other countries. The UK government has been criticised for not matching the information super highway plans of the US government but is now talking to companies such as IBM and ICL.

Government broadband policy is to be revealed before Christmas but this is unlikely to lead to a reversal of current tv restrictions on BT. The government is expected to dismiss a recent select committee report that called for the rapid lifting of the restrictions on BT which prevent it from offering tv services on its network until 1998.

US delay in digital radio

Wil Kreuwels of Philips Research and head of the Jessi digital audio broadcast programme has accused the United States of instigating delaying tactics in digital radio broadcasting. The UK, Germany, Canada, France, Scandinavia, Belgium, Australia and India are either using DAB (digital audio broadcasting) or have said that they will. On the other hand, the USA has said that it won't.

According to Kreuwels the US has said that it has a better system but the demonstration of this, first promised two years ago, has still not materialised.

Europe is leading the way in implementing DAB. Over 20 sites are transmitting digital audio broadcasts: 11 in Germany, 5 in France, 2 in Scandinavia, 2 in Belgium and 1 in the UK. Kreuwels says that another eight sites in Europe are in the planning stage.

He adds that Europe's consumer-electronics industry has a unique opportunity to dominate the market for digital broadcast equipment but that highly integrated, low cost receive and transmit chipsets are still lacking. Jessi has produced a chipset for the receiving function which is available now. It has been incorporated in a test receiver system by Philips. This is now available to potential DAB receiver manufacturers says Kreuwels.

David Manners, *Electronics Weekly*

Computer sees complete images

Two US researchers have created an optical computer which can see complete optical images through a pattern recognition device. The computer is designed to compare images and recognise similarities and works on principles similar to those used to create holograms. It consists of a lens, a glass cell filled with caesium gas and a laser.

The 'cpu' of the experimental computer is the glass cell in which "the caesium gas atoms do the actual computing," according to Dr. Randall J. Kluze, a physics professor at the University of Southern California. He has been working on the device for the past three years. The caesium gas is used as an erasable film to store infra-red light patterns.

The researchers shine a single laser through a lens. This splits it into two beams, one of which reflects off the object while the other shines directly into the gas, producing an interference pattern. When a third laser beam is focused at the gas from the opposite direction, it is bent in the areas where the two images are similar, producing bright spots.

Audio entrants to pc market

Two manufacturers of audio equipment – Aiwa, a unit of Sony Corporation, and Teac – are to enter the pc market jointly. Aiwa will use its multiplexing technologies and other expertise in audiovisual equipment, while Teac will provide the motherboards and main circuits for the computer systems.

The first joint products, two pc models on which users can view television programmes, are now available.

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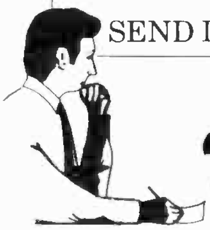

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Integrated baseband processors for DECT

Highly integrated baseband processors have been introduced by Philips Semiconductors and Advanced Micro Devices, indicating that reasonably priced digital cordless telephones for the DECT standard could be less than 12 months away. Philips Semiconductors is further down the road than anyone else to offering a single chip DECT handset design. The first element is a family of single chip baseband processor chips which will be followed shortly by a single device for all the rf down conversion and synthesiser functions.

The *PCD509X* baseband chip integrates all the bulky analogue to digital and digital to analogue converters and an 8bit *80C51* microcontroller on the same piece of silicon as the DECT burst mode logic, the *G.721* ADPCM codec and the *G.711* voice codec. The DECT user interfaces and access protocol, which supports multiple access with collision detection, was developed on the *8051* architecture, so Philips does not believe the on-chip microcontroller will restrict handset designers in their choice of systems software. Four ports are available for interfacing to display, keyboard, interrupt sources and external memory. Interfaces fully support the generic-access protocol (GAP).

Unlike Philips, AMD is keeping the microcontroller external to its *Am79C420* baseband processor, which integrates the burst mode logic, ADPCM, G.711 and analogue to digital converters. On-chip interfaces support d rect connection to Mitsubishi, iAPX, and Motorola compatible microcontrollers.

Philips has also included on-chip emulation of the *80C51* program, two interrupt lines for the burst mode logic and a speech codec to interrupt the *80C51*. There is 64Kbyte of eeprom program memory and 3Kbyte of data memory on chip, and a further lkbyte of data memory is available for the BML and speech dsp. There is a three-channel time-multiplied 8bit a-to-d converter for RSSI measurement and battery voltage measurement. The 8-bit d-to-a converter provides interfacing to the *13824MH* chip crystal oscillator.

This *ABC* family operates over a 2.7V to 5.5V supply voltage range and includes the *PCD5091* handset controller, the *PCD5092* base station baseband controller and the *PCD5093* multi-line base station device which will support up to four 64kbit/s isdn channels.

AMD's is a single programmable device which can be used for handset and base station designs.

UK: Europe's largest semiconductor market...

According to US market analyst, Dataquest, the UK will become Europe's largest semiconductor during 1994, overtaking Germany. This is because the United Kingdom produces half of Europe's pcs and pc compatibles and has the fastest growing automotive semiconductor usage.

At its European Semiconductor Industry Seminar in London last week Dataquest projected that the fastest growing market this year would once again be the UK with growth of 21 per cent, followed by Scandinavia (19 per cent), France (17 per cent) and Germany (15 per cent). The largest growth is seen in computing: up 21 per cent to $9.3bn. The second largest area is communications (at $5bn) with 1994 growth of 14 per cent, followed by industrial ($3.2bn), up 16 per cent; consumer ($2.6bn) up 16 per cent; automotive ($1.9bn) up 19 per cent and military/aerospace($681m), up 14 per cent.

Total European semiconductor sales are $22bn ($20bn for ICs). This breaks down as mos memory $5.3bn, mos micros $5.2bn, mos logic $2.5bn, analogue $3.1bn, and discretes $2bn.

...but third world status for Europe

Europe is heading for a position as the Third World of the electronics industry warns Marco Landi, President of Texas Instruments Europe. The alarm was sounded at the 4th Annual European Microelectronics Forum organised by Future Horizons in Munich in October.

"European semiconductor companies have improved but they are still lagging. Not one is in the top ten" said Landi, "Even in telecomms, Europe's traditional stronghold, we have been losing market share, apart from in GSM."

The old European strategy of supporting National Champions has to end. Landi is insistent that it is a failed concept and must be abandoned. Landi warned that in the most successful high-tech markets, the 'tigers of Asia', they don't write reports about their industrial problems, they tackle them. "We are full of action plans", said Landi, "but where is the action?"

50W psus cleaned up

EC regulations that stipulate equipment power supplies must have a pure sine wave at the socket will now apply to devices down to 50W, not 75W. This will force power supply companies to spend more money between now and January 1996. The lower rating is as a result of German lobbying.

Power factor problems arise mainly in switch mode power supplies because they feed harmonic currents back into the mains as a result of the waveform produced by chopping rectified dc at a high frequency. Another factor is that in uncorrected power supplies, power conversion is as low as 70 per cent.

Malcolm Burchall, a consultant who sits on the relevant BSI Committee, said: "In general, PC manufacturers will meet the new specification by fitting an inductor at a cost of £2 to £3 into the power supply unit".

PowerPC under Acorn's eye

Following the announcement that Acorn Computers is to use IBM *486*-based cards in its current range of Risc-based pcs, it is now believed to be evaluating IBM Microelectronics' *PowerPC* microprocessors for use in its future range of pcs.

Acorn's only comment was that it is committed to ensuring its customers have access to. a whole range of options in the pc arena.

The cards will be made at an IBM plant in Italy and supplied through Blue Micro, IBM Microelectronics' UK representative. ■

Jonathan Campbell

Brighter future for led fibre comms

Light emitting diodes that are five times more emission-efficient than conventional leds in practice – and ten times more in theory – could become valuable sources for optical fibre communications as a result of work being carried out at AT&T Bell Laboratories, New Jersey (EF Schubert *et al*, *Science*, Vol 265, pp. 943-945). In tests, the new resonant-cavity leds (rcleds) significantly out-performed the best conventional devices in photon flux density, a critical measure of a led's usefulness for fibre-optic communications.

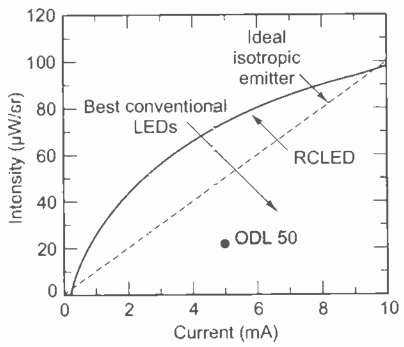
Such a dramatic boost to efficiency has been brought about by integration into the led of a microcavity – an optical resonator with coplanar reflectors separated by a distance of the order of the optical wavelength. Photon energy propagating along the optical axis of the cavity is quantised; placing the photon-emitting active region of a led inside a microcavity allows the scientists to enhance the photon flux density, and so directly influence the optical power coupled into a fibre.

The AT&T rcled is a pn-junction diode integrated with a microcavity whose fundamental mode is in resonance with the 930nm light-emitting active region of the diode. In conventional leds, the spectral characteristics of the devices reflect the thermal distribution of recombining electrons and holes in the conduction and valence bands. But with a microcavity led, on-resonance luminescence is enhanced and off-luminescence is suppressed, so the spectrum actually reflects the properties of the cavity.

Optical sources based on spontaneous and stimulated emission, such as leds and lasers, form the basis of all silica fibre communications. In practice leds are preferred because of their higher reliability, better temperature sensitivity and simpler fabrication costs.

The rcled has been designed for an operating current of 5mA. At this current, its intensity is reported to be 3.3 times that of the best conventional leds, including the state-of-the-art ODL GaAs led.

The researchers say that though the rcled has some similarity with vertical-cavity surface emitting lasers (vcsels) – also candidates for optical interconnect systems – differences in exit mirror losses compared to self-absorption in the active region mean the vcsel has light output intensities in the spontaneous region that are orders of magnitude lower than the rcled.

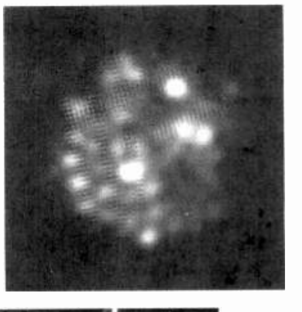


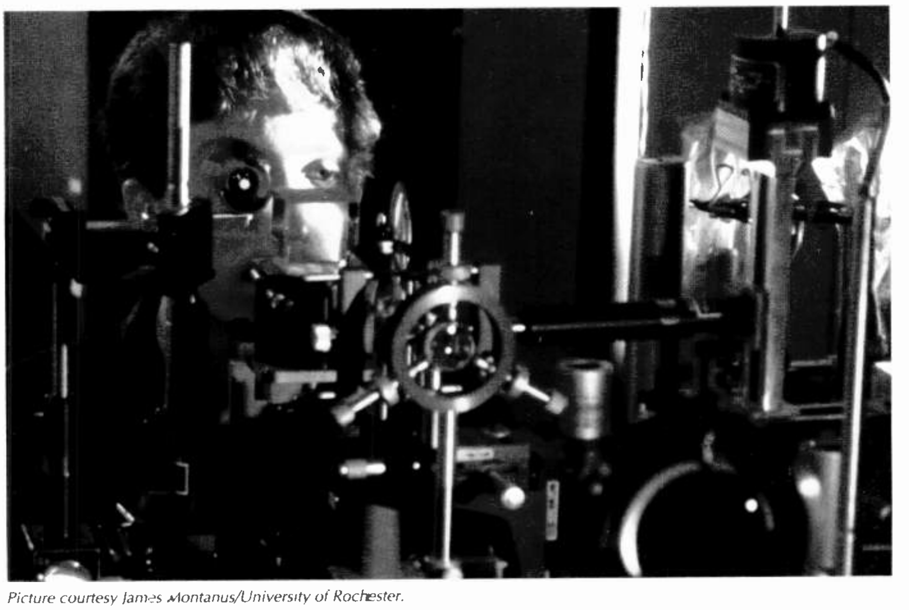
AT&T's rcled shows much better performance than conventional leds (shaded area), and better than a theoretical ideal isotropic emitter

Technical eye-lights: *An electronic camera, many times more sensitive than photographic film, and with a resolution that divides a 0.3mm square into more than 250,000 pixels has helped scientists at the University of Rochester to take the sharpest pictures yet of the inside of the living eye.*

Using the camera – a Photometrics series 200 system with a Thomson TH7895B CCD array – and shooting a low power laser into the eye for a fraction of a second, the Rochester team has recorded individual cones which are about 3μm wide.

Instruments that opthalmologists currently use see structures no smaller than about 10μm. The achievement could help doctors treat age-related ocular degeneration and retinitis pigmentosa which causes gradual deterioration of the eye.

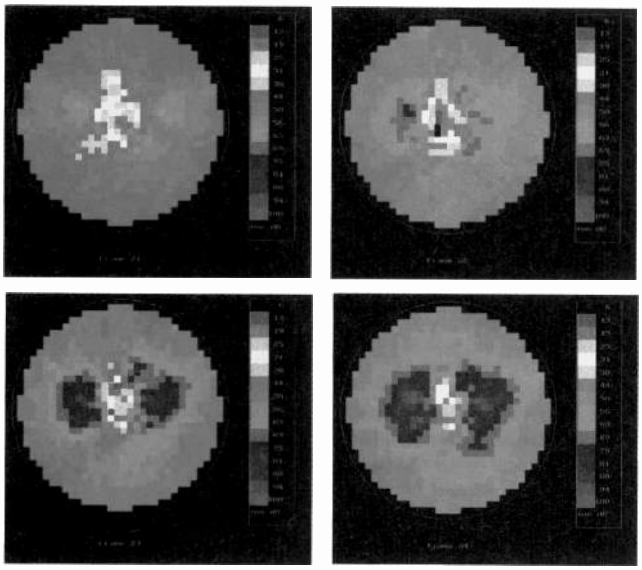




Picture courtesy James Montanus/University of Rochester.

Electrical resistance reveals bodily health

What's your resistance to emphysema? Put yourself in the hands of researchers at the Department of Biomedical Engineering & Medical Physics, University of Keele/North Staffordshire Hospital, and you could find the answer – in pictures.



Electrical impedance tomograms of an adult male. Top left hand picture is end expiration and sequence progresses top right, bottom left, bottom right, to end inspiration. The purple/blue regions are believed to be air in the lungs. Orientation is left/right (as seen) and top is anterior, bottom posterior. The scale indicates 16 intensity levels ranging from no change, black, to 100% change, purple. Resistance readings were made at 21kHz and each image took 100ms to collect.

Because there Paul Record and his team are developing electrical impedance tomography (eit), a technique that uses changes in electrical resistance to create images of our insides.

In theoretical terms, electrical impedance tomography is a technique which, from knowledge of voltage and current conditions on the boundary of a region can be used to determine the spatial distribution of conductivity inside that region. The principle is similar to that long used by geologists who have injected electric currents into the ground and measured voltages appearing on the earth's surface, so determining the conductivity of different layers. They were looking for minerals and oil.

But the Staffordshire researchers have put the technology to use in monitoring the human body, and are working to see eit take its place alongside the more familiar imaging techniques such as x-ray computer tomography, nuclear magnetic resonance imaging and ultrasound.

A patient's first contact with the technology is when 16 electrodes are strapped counter-clockwise around their chest. In effect, they are (almost) circular conductors with electrodes spaced equidistant around the periphery. A current is applied to the first pair of electrodes and the resulting potentials generated between every other pair around the body are measured. By taking successive readings resulting from applying a current to each electrode pair in turn, data on the differing electrical resistances within the body can be built up.

As part of the hardware, a multiplexer both directs current to the electrodes and allows the electrodes to be selected from which the potentials are to be measured. Received signals are amplified, demodulated – the signal being measured is an ac voltage and must be converted to dc for image reconstruction – and digitised. Design of the various hardware components is complicated by the fact that a very large range of potentials has to be handled, setting problems in terms of analogue noise and digitisation accuracy.

Acceptable hardware has now been developed and by taking two complete sets of measurements, separated by time, a picture of the change in resistance in the body can be built up. It is this change-in-resistance map that the Staffordshire team is using to give valuable information about the health and operation of a body.

For example, lung tissue has a resistivity which changes considerably with ventilation, so EIT could be used to image changes in lung air volume. Clinicians need merely to ask the patient to breathe in, take a reading, then breathe out, and take another reading. The resulting image of a resistance change map would indicate any problems. Record says the technique has been well-validated against other lung volume measurements.

Displaying the distribution of air in the lungs could allow conditions such as emphysema and pneumonia to be investigated.The method could also be applied to continuous monitoring in sports medicine.

In general, eit offers a non-invasive, low cost, high speed and highly sensitive method. It does suffer from poor spatial resolution against procedures such as x-ray ct. Unlike ct it could be used for continuous monitoring of a patient on an intensive care unit for example.

At present eit is in pre-clinical trials, being used to monitor lung conditions such as differences in the inflation of each of the lungs.

Other possible applications are in monitoring gastric function, change in tissue temperature during hypothermia treatment of tumours, and changes in lung water. It may also be possible to image heart conditions through tracking movement of blood from the heart to the lungs and back during the cardiac cycle. But complexity of the heart geometry makes calibration for this sort of application much more difficult.

Computer buffs get emotional about user interfaces

It won't be long, according to computer scientists, before we are all getting pally with our "interface agents", software personal assistants that are animated characters which learn our behaviour (and our shortcomings) and accordingly adjust the help they offer us.

Development of agents is involving all the usual tools and complex algorithms so beloved of the computer fraternity. But Joseph Bates, both associate professor at the School of Computer Science and also a fellow of the College of Fine Arts at Carnegie Mellon University, has been researching a less familiar road as part of the project: how can we make those agents believable? The Oz project, an interdisciplinary effort led by professor Bates, is currently working in this area. Writing in *Communications of the ACM* (Vol 37, No 7, pp.122-125) Bates explains how it is emotion that is the key "...because it helps us know that characters really care about what happens in the world".

As such it is felt to be at the heart of creating sympathetic creatures that could serve as components in new user interfaces for the non-specialist.

Endowing computer software with a personality certainly sounds like a wild idea. But remembering hours of work lost through computer failures over the years, perhaps endowing it with a kickable bottom and a thick skin would be the real break-through.

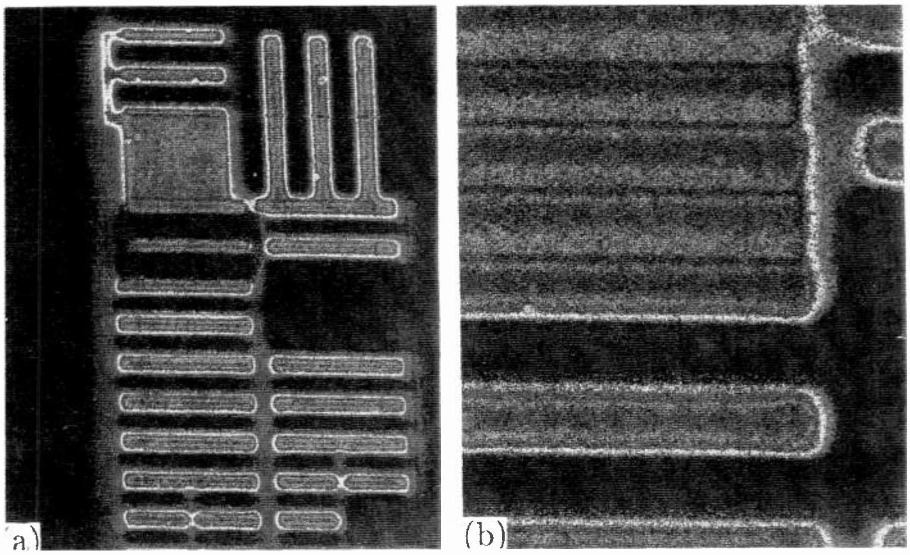
Ion beams reveal device infrastructure

Depletion regions contained within the layered structure of electronic devices could be much easier to analyse at high resolution than previously, with development of an ion beam microscopy method by Mark Breese and coworkers at the SPM unit, Oxford University.

Irradiating chips with ionising beams to generate charge carriers and so monitor electrical and physical properties is an accepted method of identifying manufacturing flaws. Fewer carriers are measured at defects and dislocations because they act as trapping and recombination sites.

But multilayer devices cause difficulties for conventional keV electron beam examination or laser-induced microscopy, as upper insulating and metallised layers can block examination of the structure. Previously surface layers have had to be stripped away or the device cleaved to image the underlying layers.

Breese's technique of ion beam induced charge (ibic), reported in *Physics World*, (October, 1994, pp.26-27), fires MeV ions at the chip to produce very large numbers of carriers which can be detected. Focusing the beam of alpha particles or protons is more difficult than with other approaches but the heavy MeV beams do not suffer so much lateral scattering and so information can be gleaned much deeper down into the structure.



Ibic imaging of an eprom, measured with an incident beam of 2MeV alpha particles (a, 200μm² and b, 60μm²). The technique reveals information about the inner structure of a device without the need for removing upper layers. Different colours represent differing amounts of measured charge, showing up depletion areas for example.

Practical digital amplifier has DSP built in

Digital power amplification, using dsp technology to convert a digital signal directly into a corresponding high power analogue waveform, could move from the possible into the practical as a result of work being carried out by two researchers at Kings College, London, and Universitat Politècnica de Catalunya, Barcelona.

Mark Sandler and JM Goldberg have developed a high-accuracy pulse-width modulation d-to-a converter suitable for high power and low power use and are currently building prototypes.

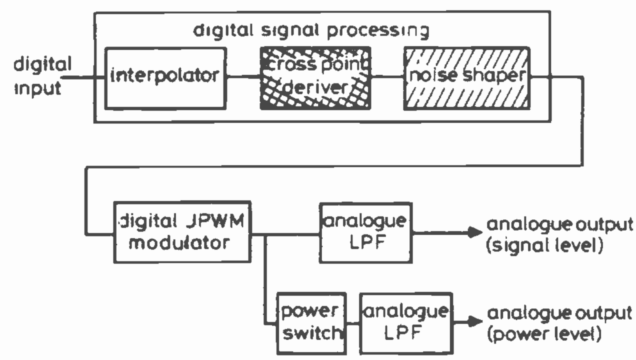
Main difference between their converter and those previously proposed is that use of dsp has made the device much easier to realise in hardware, and distortion looks to be considerably reduced (*IEE Proc – Circuits Devices Syst*, Vol 141, No 4, 1994). DSP is used prior to modulation and seems to greatly enhance the potential for realising a linear 16-bit quality pwm-based d-to-a converter in practice.

Key to the approach is inclusion of a premodulation algorithm – a "cross-point deriver" – designed to digitally emulate the harmonic distortion-free natural sampling process. This "pseudonatural pwm", or pnpwm, has demonstrated dramatic improvements in d-to-a performance, often completely eliminating the harmonic distortion associated with more conventional uniform sampling pwms. In addition, dsp techniques using an oversampled noise shaping network (ons) that reduces input signal wordlength but loses little in base-band signal quality, allow modulator clock speed to be decreased to a rate suitable for hardware implementation. The researchers also say their specially designed ons noise transfer function eliminates undesirable effects associated with more popular designs.

In general terms, the ons/pwm-based d-to-a converter is similar to a one-bit sigma-delta modulation based converter: both use a combination of oversampling, coarse quantisation and error feedback to achieve high performance with reduced hardware requirements. But Sandler and Goldberg explain that the *combination* of ons with pwm reduces stability problems and hardware difficulties associated with single-bit sigma-delta modulation systems.

At the moment, interpolation, cross point derivation, noise shaping and digital pwm sections are being built, and the researchers believe a third-order cross-point deriver could be implemented on a general purpose floating point dsp chip or in about 400mm² of a 1.2μm cmos process (including a fifth-order noise shaper).

The power switching stage is yet to be built.



Using dsp to produce a practical high-accuracy pwm d-to-a converter.

Bendable chip heralds cheaper electronics

Development of an all-organic flexible device by researchers at Laboratoire des Matériaux Moléculaires, France, could herald the first steps in opening up manufacture of large-area, low cost electronics. Not only does the manufacturing process rely completely on printing methods, but the resulting device is so pliable that it can be rolled up, bent and twisted through 90°.

Organic fets and leds have been around for the past four years or so. But as Francis Garnier and colleagues point out, "organic" is a bit of a misnomer since the structures still contain a metallic part: the gold source and drain electrodes in fets or the rectifying calcium or magnesium electrodes in leds. Metal deposition calls for costly high vacuum and high temperatures and the resulting metal-organic interfaces could never be considered fully flexible.

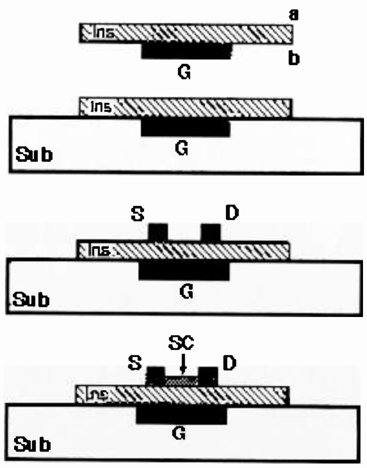
Now the French team has reported (*Science*, Vol 265, pp.1684-1686) a truly all-polymer fet fabricated solely by printing techniques.

The device is constructed by first depositing a 10μm thick layer of conducting graphite-based polymer ink through a 5 x 12mm mask onto a polyester insulating film. This is the gate electrode. Then the device substrate, adhesive tape which provides both flexible mechanical support and self-standing properties of the device, is applied to the gate and insulating layer. Electrical contact for the gate is through the same conducting ink as is used for the electrode. Next the team deposits, through a mask, two 1 x 10mm strips, 10μm thick and 200μm apart, of the same conducting polymer ink onto the opposite side of the insulating film. These are the source and gate electrodes with a channel width of 200μm and length of 10mm.

Finally a 40nm organic semiconducting layer is deposited, by flash evaporation at 350°C, between the source and gate.

The French researchers say that tests show the device demonstrates excellent amplification characteristics in the μA range and their fet demonstrates the p-type character seen in other organic fets with their metallic electrodes.

The current design looks a little over-large – though the team says even conventional printing techniques would allow dimensions to be reduced. There is also potential for macrosized chips needed in the display area.



Constructing an all-polymer fet, starting with the insulator, ins, then gate, G, substrate, Sub, source, S, and drain, D, electrodes and the organic semiconductor, SC.

Atomic electroplating that captures the sun

University of Georgia electrochemist John Stickney has developed an electrodeposition technique that could open up a new way of manufacturing electronic components, including semiconductors and optical electronic devices.

Scientists have long been depositing materials electrochemically, but the structures have been very polycrystalline – composed of many small crystals – and there has been little control over the resulting structure.

What is different about Stickney's recently-patented technique is its atomic scale.

For the past five years, the associate professor of chemistry has experimented with ways to electrodeposit, alternately, one-atom-thick layers of two or more elements, such as cadmium and tellurium, to form a compound. His method, electrochemical atomic layer epitaxy (ecale) produces compounds that could prove useful for many kinds of semiconductors.

For example, one application of cadmium telluride is in the formation of photovoltaics. Stickney points out that cadmium telluride is perfectly matched to adsorb sun light, and so has the potential to produce very efficient photovoltaic devices.

Ecale also might have applications in making phosphor materials which could be essential in developing new high-definition display systems.

To deposit one-atom-thick layers, Stickney combines atomic layer epitaxy – a chemical deposition process – with a surface-limited reaction that stops when the electrode surface is covered. Stickney's process is environmentally friendly in that he uses very small amounts of dilute solutions that are easily recycled. Theoretically it should cost less too since it uses much smaller amounts of chemicals and has simple hardware.

Technophobia – the unrecognised epidemic

Techno-sophisticated cyber-punk *EW + WW* readers apart, we must accept that there are a few old fogeys in this world who are less than at ease with computers and related technology. But a recent analysis in the US suggests that those people are neither few, old or fogeys. They're just as liable to be Nintendo-playing game whizzes as women of a certain age, and they could number half the population!

These surprising findings have come out of ten years of study by Larry Rosen of California State University and Michelle Weil of Chapman University ("What have we learned from a decade of research on the psychological impact of technology?"), *Computers and Society*, 3, 1994). They estimate that between 25% and a half the US population is technophobic, suffering anxiety about computers and technology. Women and older people, say the researchers, are no more technophobic than men and teenagers. The only characteristic that links them all is the avoidance of technology at all costs.

Part of the problem identified by the analysts is that technophobes often had bad early experiences of technology, introduced to it by people who were themselves uncomfortable – one study showed that 45% of school teachers were technophobic.

Fortunately technophobia can be cured – the researchers claim a 92% success rate – by only a few hours of sensitive introduction to computer technology. How long it takes to get a technophobe to sit through a complete Spielberg film is not recorded. ■

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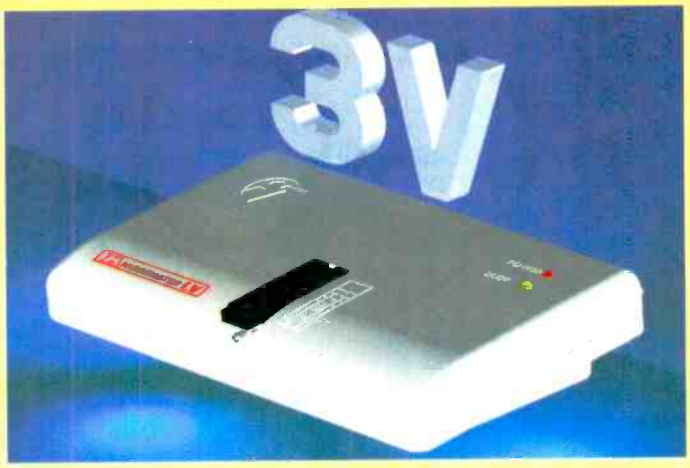

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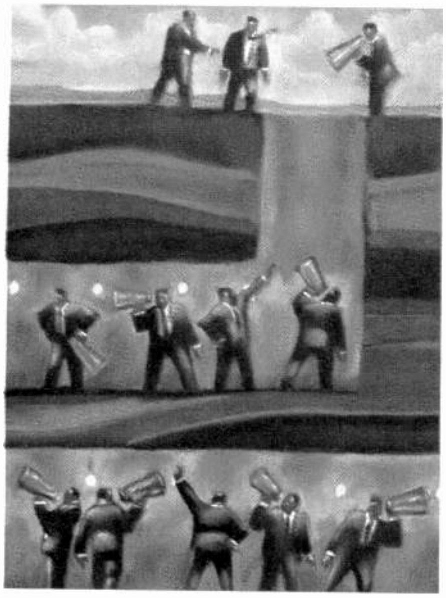


Traditional radio-frequency transmitters and receivers are of little use underground. But there are other communication options available to cavers and potholers, as Mike Bedford reveals.

GOING underground

In the summer of 1953, Geoffrey Workman achieved his moment of fame. In fact, such was the demand for his story that James Lovelock of the *News Chronicle* secured his 'exclusive' by interviewing him in a deserted cave passage hundreds of metres below the ground.

For the previous two weeks Geoff had remained alone in Sand Cavern, part of Yorkshire's famous Gaping Gill pothole, and in so doing had broken the underground endurance record. In the build-up to this event, speculation had been rife regarding the likely outcome of this apparently foolhardy venture. Of prime interest were the effects of his total isolation from the outside world. How would the human body react to being cut off from the daily cycles of the sun? Would the lack of external stimuli render Geoff insane? Or, for that matter, wasn't he surely somewhat deranged in the first place to even contemplate such a feat?

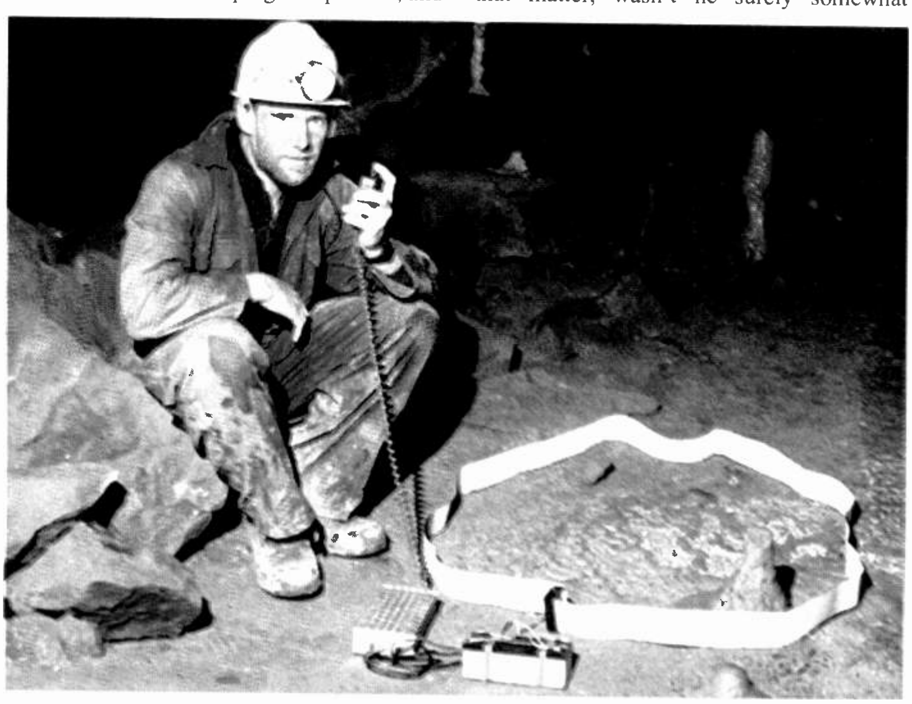
In the event, when Geoffrey Workman emerged to an incredulous world, except for having to wear sunglasses to protect his eyes from the sunlight, he appeared quite unscathed.

What place – you may wonder – has this introductory tale in a magazine dedicated to the application of electronics and radio? Quite simply, it reminds us of how totally isolated the subterranean realm is from the rest of the world. Not only is sound very effectively excluded, but sunlight and a large proportion of the electromagnetic spectrum are totally incapable of penetrating the ground. It also hints at the problems associated with communicating between caves and the surface.

Applications of cave communication

During 1993, the member organisations of the British Cave Rescue Council were called out to 53 incidents involving 128 individuals – all but two of whom were brought safely to the surface.

The use of inductive cave radios played no small part in ensuring the success of many of these rescues. In particular, they have been used to co-ordinate rescue efforts, and to summon medical assistance to the cave entrance to meet casualties. Of course, much of what I say here about cave rescues also applies to the



Caver Stuart France with his experimental 27kHz fm cave radio in Ogof Ffynnon Ddu, South Wales.

Steve Laugher, G7LYN, locates an underground beacon using an 874Hz receiver.

mining industry. During the 70s, the Federal Bureau of Mines, in the USA, sponsored a development program for underground radios in response to a spate of serious accidents. Ironically, it was a spectacular improvement in the safety record in the early 80s which made this development redundant, and so curtailed further work.

As continued media attention illustrates, the same cannot be said of potholing, especially where novices fail to observe the basic safety precautions. Inevitably, therefore, cave rescue groups will continue to have need of effective communication both within caves, and between caves and the surface.

On the afternoon of Tuesday 16 January 1979, Geoff Yeadon and Oliver Stratham secured themselves a place in the annals of caving history by making a 6000 metre through dive from Kingsdale Master Cave to Keld Head in North Yorkshire. Not only was this expedition a major feat of skill and stamina for the divers, it was also one the first success stories for cave radio.

During their record breaking trip, a waterproofed cave radio attached to one of the divers was used to communicate with the surface. The support team located the divers' position, placing flashing road lights on the surface to trace out their path. This episode was an impressive illustration of another application of cave radio – namely radio location.

Since then, the use of cave radios to provide fixed surface locations corresponding to underground survey points has greatly improved the accuracy of cave surveying. This in turn has resulted in a number of spectacular new discoveries. Of particular note was the link-up of Ingleborough Cave and Gaping Gill, involving a 10 hour trip through a series of tortuous passages which had eluded explorers for 150 years.

Cave communications are also used during expeditions to assist the exploration of new cave systems. Additionally they have been used for data logging by cave scientists. Normally, having installed monitoring equipment in a cave, taking readings involves making trips into the cave at regular intervals. With a microprocessor controlled data logger, interfaced to suitable communication equipment, however, regular readings can be stored in memory and transmitted to the surface via the communication link on receipt of the appropriate command.

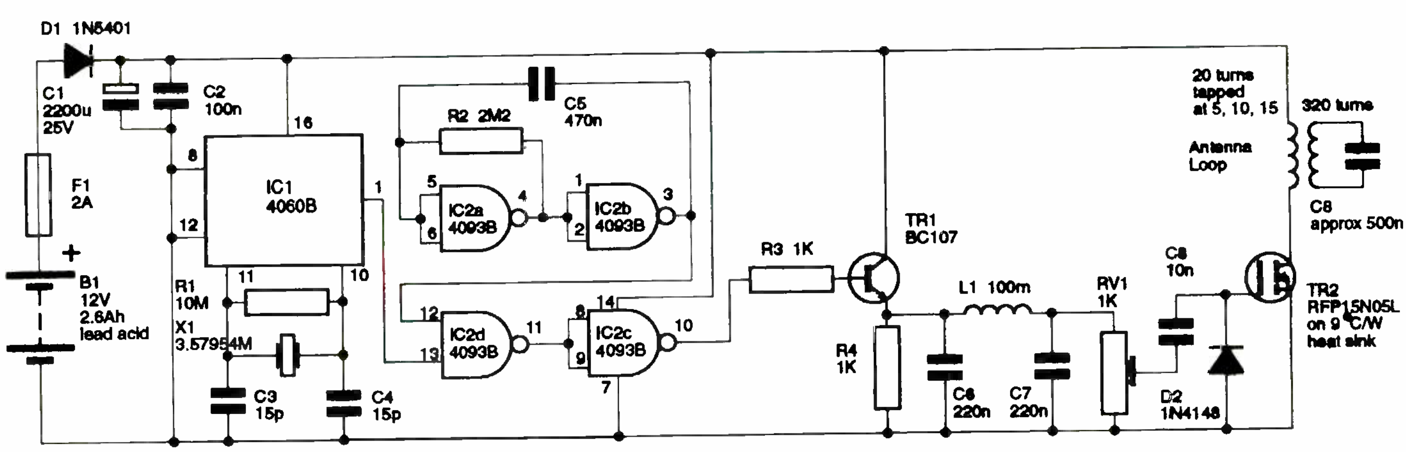
The challenges

It will be no surprise to you that most radio waves are heavily attenuated by solid rock. What is probably not as universally know is the exact nature of the relationship between the attenuation of a radio signal, its frequency, and the characteristics of the rock. Rather than get into heavy mathematics, let me simply state that the attenuation increases with the frequency of the signal and with the conductivity of the rock.

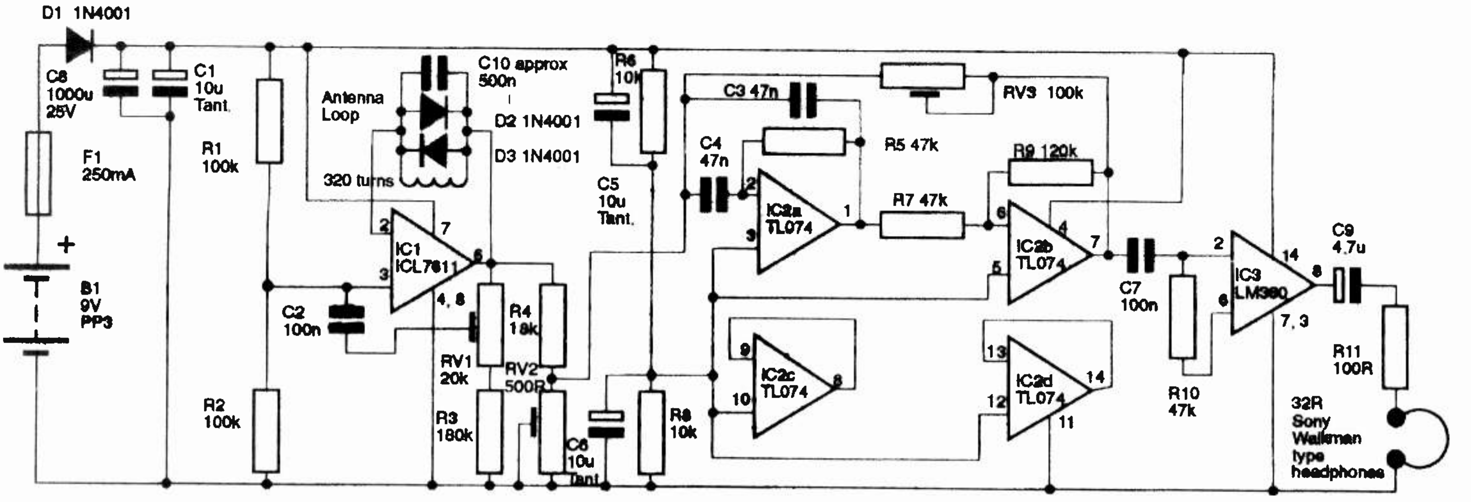
In limestone – the rock in which caves and potholes occur – attenuation at 100MHz varies from 6dB/m to 60dB/m. Even at 1MHz, the range is 0.6dB/m to 6dB/m. Only when we get down to the long wave broadcast band do we find frequencies which are even remotely feasible for transmission to a significant depth. In practice lower frequencies still would be far more beneficial in most instances.

A low frequency equates to a long wavelength and a long wavelength implies a large antenna. However a large antenna is incompatible with small cave passages. At the sort of frequencies which look promising for through rock communication, half-wave dipoles would range from about 750m to 150km! This, in a nutshell is the problem with cave radio.

For the moment, ignore the theoretical difficulties and assume that suitable means are available. Even then, further challenges present themselves. The cave environment doesn't exactly replicate the workshop; nor are potholers renowned for their careful handling of equipment. You can reasonably expect that



In this transmitter circuit for underground communications, on/off keying is used to conserve battery power. This feature also makes the signal easier to detect. More details in the panel on page 986.



Receiver for underground communication at 874Hz. The ICL7611 is configured as a Q multiplier, simulating negative resistance to compensate for the imperfect inductance of the antenna. Diodes in the feedback loop provide protection against strong signals.

any equipment taken into a cave will be dragged in tackle bags and dropped – possibly even into water. Mechanical construction, especially with regard to ruggedness and waterproofing, is a major consideration in the design of any cave communication equipment.

Communication options

As I have said, conventional radio is unsuitable for underground communication. I guess that the most obvious alternative to radio is the telephone.

In mines, telephones are common. One of the attractions of caving, however, is seeing nature unspoiled by human advances. The presence of cables down each and every passage is incompatible with this. In addition, cabling would be expensive to install and maintain.

In practice, rescuers lay a line each time they enter a cave. To keep down costs, cable weight and complexity, single-wire telephone lines can be used, with a return path through the earth.

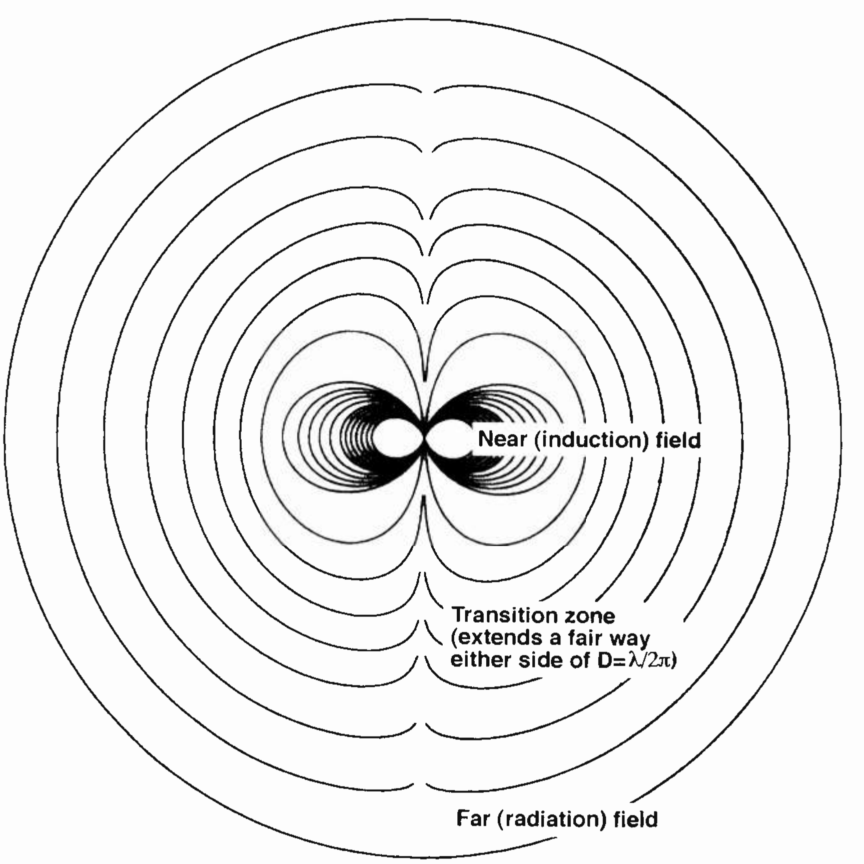
An alternative single-wire technique used in the mining industry is rf guide-wire communication. Here, an rf signal, as opposed to the single-wire telephone's audio-frequency signal, passes along the wire conductor. The main advantage over simple single-wire telephones is that the transmitters and receivers can be coupled to the cable capacitively; a direct connection is not required.

Also, of particular interest to cavers, temporary 'capacitive' repairs to damaged cables can be effected by simply tying together the loose ends. Interestingly, some recent single wire telephone designs, having very high impedance inputs, exhibit similar characteristics.

A third approach, illustrated overleaf, involves a technique used in the first world war and regularly re-invented, albeit not normally for caving applications. If a signal is injected into the ground through a pair of electrodes, that signal can be detected as ground current by another pair of electrodes some distance away. This technique is usually referred to as earth current signalling. It was the subject of experimentation by British radio amateurs during the second world war when amateur licences were revoked.

It was only discovered some time later that these signals penetrate the earth to some considerable depth. Theoretical effectiveness of such a system can be calculated by considering the distance – and hence the earth resistance – between the various electrodes. Hardly surprisingly, the signal-to-noise ratio increases with the separation between the transmitter and receiver electrode pairs, and decreases with the separation between transmitter and receiver.

Something else which was not appreciated by the early experimenters was that the earth current model is not the only one which applies to this form of communication. Experiments have shown that magnetic fields are also set up in the earth and that these can be detected by induction radios introduced in



Induction and magnetic fields in a radiated signal. Within a few wavelengths of the antenna, electric and magnetic fields exist separately. Conveniently, the near-field inductive component can be transmitted and received via a loop very much smaller than a wavelength.

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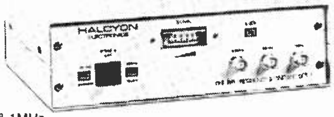

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Typical circuits for radio location

To illustrate the principles of simple radio location cave radios, two circuits are reproduced on pages 983 984 by kind permission of Stuart France and Bob Mackin. You will notice that rf design techniques are not too important at such low frequencies, and in many instances AF circuitry suffices.

The transmitter is a simple four-stage design with an additional 'interrupter'. The first stage is a cmos crystal oscillator/divider based on a *4060B* for generating the 874Hz signal. Two of the *4093B* nand gates generate a 1Hz square wave which modulates the signal using a further nand gate.

The purpose of this on/off keying is to conserve battery drain and to make the received signal more easily recognisable. Based on a BC107, the pre-amplifier filters the 874Hz square wave using a pi-network to provide an approximately sine wave drive to the final stage. The mosfet power amplifier drives the tuned loop, coupling via a tapped link coil.

In the receiver circuit, you will notice that the antenna loop and tuning capacitor are connected in the feedback loop of the first stage amplifier. This device, an *ICL7611*, is configured as a Q-multiplier, simulating negative resistance to compensate for the imperfect nature of the antenna loop as a pure inductor.

Two diodes in the feedback loop provide protection under strong signal conditions. The *TL074* acts as a bandpass filter, and is set up to achieve maximum selectivity without inducing instability. Since the signal is in the audio range, the receiver does not need any sort of detector. Reception is simply a matter of amplifying the signal. Audio output is based on an *LM380*, and drives a pair of personal-stereo headphones, rewired for mono operation.

the next section. In reading the following description, you can envisage earth current signalling and induction radio as one and the same thing.

Low-frequency induction

Earlier on I mentioned that the main problem with using radio for cave communication is that low frequencies are needed, resulting in prohibitively large antennas. For the moment, forget about the size constraints, and consider what happens when an alternating current is fed into a full-sized antenna.

At a distance of a few wavelengths from the antenna you will observe an electromagnetic radio signal – electric and a magnetic fields existing in a fixed ratio to each other.

Much closer to the antenna, a different pattern emerges. You will still find magnetic and electric fields, but instead of the radiated far field signal, you will find the near field – a collective name for the electrostatic (electric) and the induction (magnetic) fields.

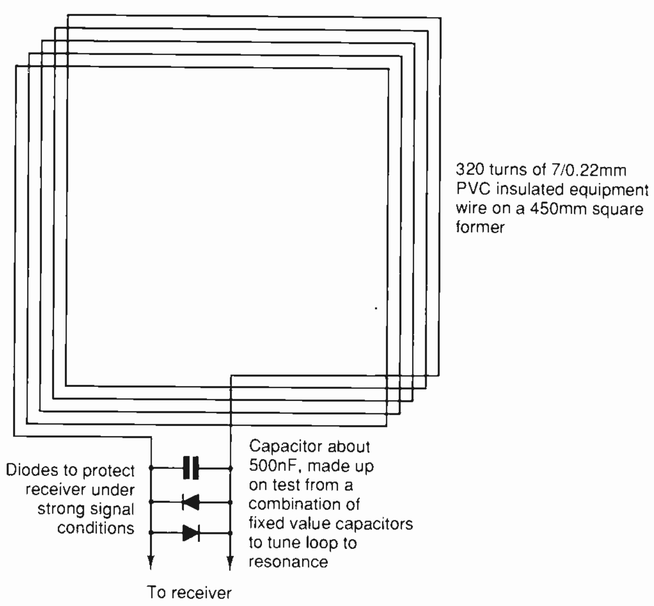
The induction field is particularly interesting since it can be generated not only by a large conventional antenna, but also by a loop very much smaller than a wavelength. At first sight this may seem quite remarkable – low frequency radio from small antennas. But remember that this is not real radio. We are only concerned with the near field which, as its name suggests, does not propagate very far.

In fact the induction field decays with the cube of distance, and so range is severely limited. But since most caves are only of the order of a few hundred metres deep, this is not a problem, and induction is one of the prime means of cave communication.

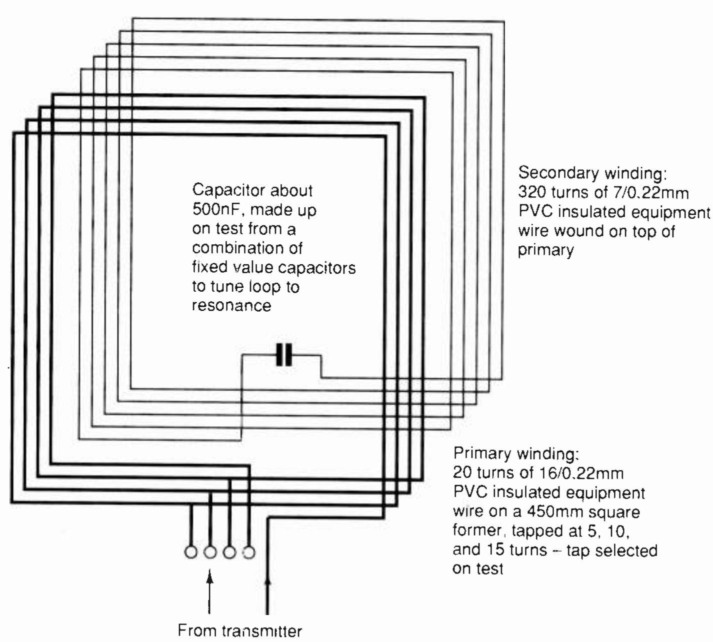
Operation of an inductive communication system can be thought of as being analogous to that of a transformer. The transmitter loop is the primary, and the receiver loop is the secondary. Clearly, the coupling between the two is very loose due to their wide separation. It is imperative to maximise the magnetic field generated by the transmitter, and the signal-to-noise ratio of the receiver.

At this stage, I must introduce the concept of the loop's magnetic moment – a measure of its efficiency in generating an induction field, and hence the communication range achievable. The magnetic moment is proportional to the number of turns, the loop current and the cross sectional area, so you can see in broad terms how to maximise the range. However, if I also refer back to the inverse cube relationship of the induction field, it becomes clear that to double the range, an eight fold increase in either turns, cross-sectional area, or current is required.

Take a brief look at each of these in turn. Considering the turns procuce a rather unexpected result – assume that the range is to be doubled by winding eight times more turns. Infuriatingly, this increases the resistance by a factor of eight and so, assuming that the driver voltage is constant, reduces the current correspondingly. Gain has been cancelled out.



Antenna for the 874Hz receiver. Together with the inductance of the antenna, the protection diodes and capacitor form the Q-multiplier's feedback-loop components.

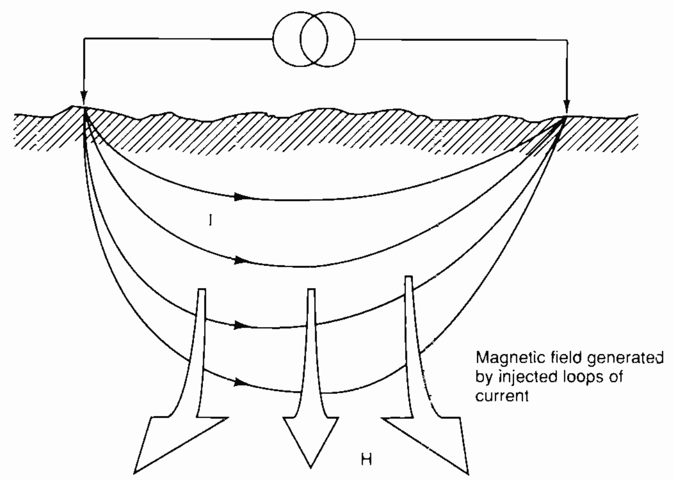


Drive for the transmitting 874Hz antenna is provided by a power mosfet, which is in turn driven by a near sine-wave. Note that both antennas shown here are represented as components on their respective circuit diagrams.

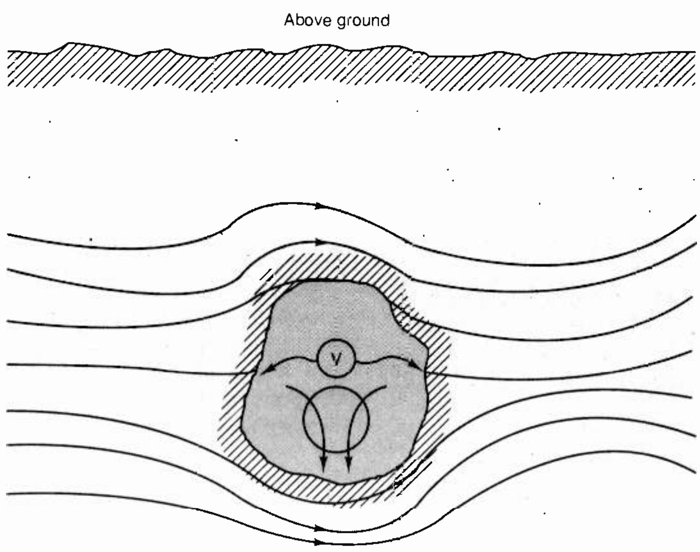
The only way to achieve a gain by increasing the number of turns is to also increase the gauge of the wire. All in all, doubling the range results in a 64-fold increase in weight! In fact, it can be shown that the number of turns is irrelevant. All that really matters is the mass of copper in the loop.

This argument assumes that the loop's diameter is kept constant. Actually a loop with a large cross sectional area is by far the best way forward, but in tight cave passages, the limitations are obvious.

Of course, another approach to improving performance is to push the power up, but once again you encounter the law of diminishing returns – a modest increase in range would require a very significant increase in current, and hence also of battery weight.



Used in the first-world war and regularly re-invented, this technique for communicating underground involves a magnetic field generated by injected loops of current.



Signals produced by inserting electrodes in the ground can be picked up either by probing current or voltage or by picking up the magnetic head using a vertical induction loop in the cave.

Areas for practical work

As already intimated, the challenges of cave radio are unique. In particular, the would-be designer has to work at unfamiliar frequencies, aim to maximise the magnetic moment of the loop while minimising its weight and dimensions, and in addition, engineer the radio to withstand harsh environmental conditions. On the plus side, few people are currently working in this area, so there's plenty of scope for innovation.

A number of avenues are available for experimentation. Scope for commercial exploitation is limited – cavers don't often have a lot of money. However, there is a possibility of making a significant contribution to cave rescue and potentially help to save lives.

In the panel entitled 'Typical cave radio circuits for radio location', I have shown the circuits for a simple 874Hz transmitter and receiver which designed for radio location. With minor modifications, they could also be used for Morse or data communication.

The circuits are simple, so this would be a suitable first step for those interested in cave radio. Despite their simplicity, however, the constructor would still need to address the ruggedness and ergonomic issues discussed earlier, so their construction would provide some valuable experience.

The circuits described here operate below 9kHz – a grey area as far as licensing is concerned. I have been told that the RA no longer regulates this part of the spectrum. The RSGB on the other hand, has been told that although 9kHz and below is not regulated internationally, the UK authorities still do regulate this band. You can carry out sub-9kHz work under a Test&Development license, but this is expensive.

It is expected that a UK vlf amateur radio allocation may come into being in the near future. A frequency in the vicinity of 87kHz has been mentioned, and this would be eminently suitable for cave radio. In the first instance, this will be an experimental permit for people who have already registered an interest with the RSGB. Later, a general allocation may be made for the whole of the IARU region 1, i.e. Europe.

If you decide to experiment with vlf or elf radio, and decide to put it to the test by taking it underground, remember that caving can be extremely dangerous to the inexperienced, and that no one should go exploring alone.

If you want to meet up with electronically minded cavers, you might like to consider joining the Cave Radio and Electronics Group (CREG) of the British Cave Research Association (BCRA). In addition to publishing a quarterly technical journal containing a mix of theoretical and practical articles, the CREG acts as a clearing house for the dissemination of ideas, and arranges twice yearly field meetings.

For details, please send a stamped, self-addressed envelope to me, Mike Bedford, at 4, Holme House, Oakworth, Keighley, W. Yorkshire, BD22 0QY.

My thanks to Stuart France and Bob Mackin for their permission to duplicate their 874Hz transmitter/receiver circuits ■

Further reading

Full constructional and setting-up details were published in *Making a Simple Radio-Location Device*, Stuart France & Bob Mackin, *Caves & Caving 52* (summer '91), pp 7-11. It was reprinted with corrections in *An Introduction to Radio Location*, Mike Bedford, *CREG Journal* 14, December 1993, pp16-18 & 14. Back issues of this journal are available from the Cave Radio and Electronics Group at the address given later.

BETTER AUDIO from non-complements?

A mosfet amplifier with lower distortion, 30% more output using the same power supply, savings in parts and labour and high output voltage compared to conventional units? It's all due to 'virtual complements' says Bengt Olsson.

Bias system symmetry and thermal stability have led to complementary output transistors still being used in power amplifiers, even after 30 years. In fact the complementary concept has never been seriously challenged. Yet the symmetry is actually only theoretical, and though the schematic looks symmetric, in practice, it is not.

So why use expensive p-channel devices with 1Ω resistance when there are rugged 0.2Ω n-channels at half the cost? I believe my new idea for a symmetric mosfet amplifier could prove a workable alternative to amplifiers with high R_{on} and low efficiency.

Symmetry by 'virtual complement'

The symmetric amplifier uses two identical n-channel mosfet transistors, one working in reverse – like a p-channel transistor in a "virtual complement". In reality the mosfets behave like a symmetric pair with a "super-symmetry".

Except for polarity, the virtual p-channel transistor is identical to the real n-channel, and symmetry is perfect. Like the real complement, the virtual complement has excellent thermal stability, and after heavy loads bias current deviates by a maximum ±30% for a couple of seconds before it settles again. But since there is no optimum bias in a mosfet-amplifier, the changeover is inaudible.

Some of the main advantages of the circuit are its extreme simplicity, high efficiency and low distortion.

In the circuit, the bias network is connected to the output terminal (**Fig. 1**), with its voltage fed to the gates of two mosfets via resistors R_1 and R_2. Input transistor Tr_3 provides push-pull symmetric signals to the gates, independent of the voltage across it – resistor R_1 may be above the positive rail while R_2 could be at the negative rail; the circuit will still function perfectly well.

Symmetry depends on an equal current in the two (equal) resistors. When the voltage is rising on one gate it decreases equally on the other gate, assuring symmetry of gate voltages. Because the transistors are equal, the output current will be symmetric too.

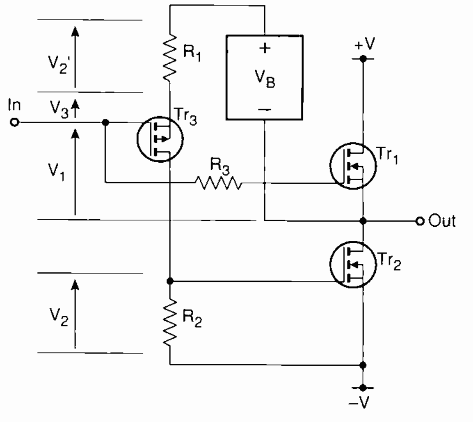


Fig. 1. The basic circuit. $V_B = V_T + V_2' + V_3$ if $R_1 = R_2$. V_B is adjusted to provide the proper bias current in Tr_1, Tr_2. R_3 (= R_2) makes the input impedances of Tr_1 and Tr_2 equal at high frequencies.

Bias current is very stable and works as shown in **Fig. 2**. Quiescent current can be set by a potentiometer. But some kind of temperature compensation is preferable if 'vertical' mosfets are used as these have a higher g_m and so are more sensitive to temperature-dependent bias changes than the more commonly used audio "lateral" channels.

Figure 3 is a variation of Fig 1. It has higher input impedance and lower bias voltage as the gate of Tr_3 is bypassed and is stable with low distortion. **Figure 4** shows an implementation of a "reversed power stage".

Negative side input

Driver transistor Tr_3 may be an n-channel device (**Figs. 5 and 6**): it is only a current path

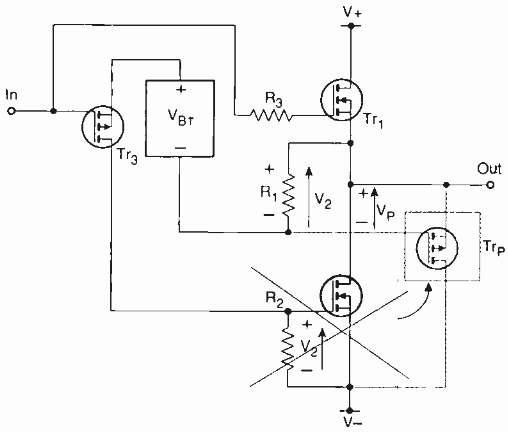


Fig. 2a. Virtual complement, redrawing Fig. 1.

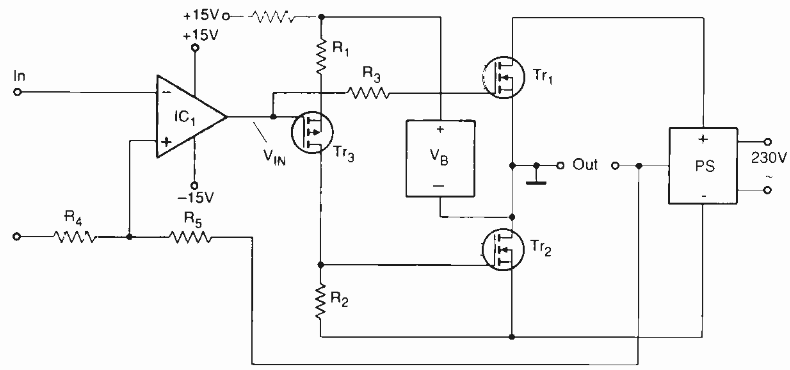
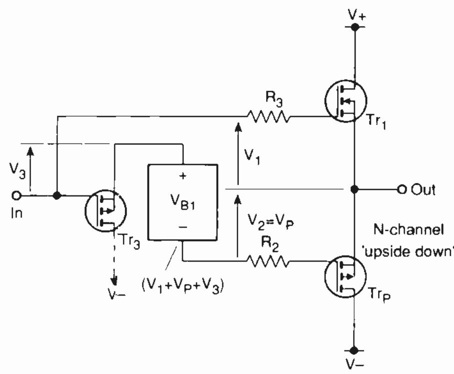
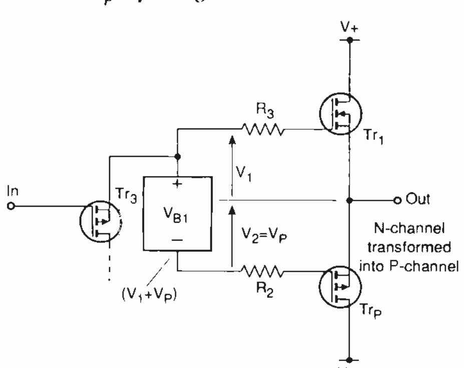


Fig. 4. Block diagram of a reverse power stage amplifier.



2b. Virtual complement, redrawing Fig. 2a with Tr_p replacing deleted area.



2c. *cf Fig. 3.*

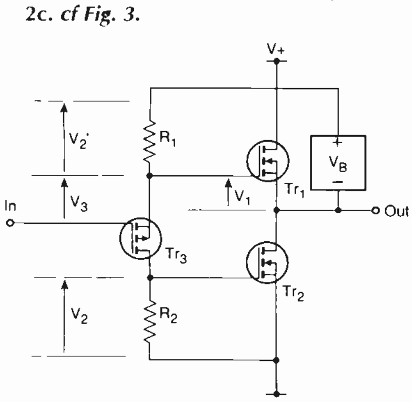


Fig. 3. P-channel, high impedance input. Note: $V_B=V_1+V_2'$

so a change will cause no basic difference. In this case the input to Tr_3 is close to the negative rail and the stage can be driven directly from an input differential pair (**Fig.** 7).

If Tr_2 and Tr_3 gates are connected, another bias-supply, B_2, will be needed. The new bias will compensate for the lower emitter voltage of Tr_3, but may be adjusted by B_1 only (see equations, Fig. 6).

Tr_2 gate will be controlled by the overall feedback, keeping the output terminal at zero by balancing the currents of Tr_1 and Tr_2, and taking no notice of the quiescent current. This circuit needs special attention when the power is turned on or off, as lowering V_{B2} will increase the quiescent current.

Two bias supplies need new bias rules

The existence of B_2 opens new possibilities of bias voltages selection, as the *difference* between B_1 and B_2 controls the gate voltage (No change is involved since B_2 used to be zero.)

Adding the same voltage to B_1 and B_2 will keep the difference constant, leaving more headroom for the bias of transistor Tr_1 and Tr_2: ie, a higher gate voltage at transistor saturation.

The anti-saturation circuit of **Fig. 8** will work the same way and could replace Fig 6. After all, with no negative bias supply needed, what use is B_2 ?

Generally, bias supply B_2 has no function and can easily be avoided, as in Fig. 8.

Any addition or change of bias will affect the quiescent current, and the same is also true of V_s in Fig 8. Increasing V_s has the effect of decreasing the bias current of Tr_1 and Tr_2, even if V_s is only indirectly associated with the bias loop (via V_2).

V_s is shown stabilised, preventing current rise during on and off (when V_s is diminishing). When the power supply voltage drops below 50% , V_{B1} will also be reduced, producing a net reduction in quiescent current.

This result makes it possible to use the main power supply for safe biasing during on and off transients.

Industrial or audio power transistors?

The modern "vertical" mosfet has a high gate bias voltage, typically 3.5-4V at 100mA. But this is not a problem, because it is easy to have a very precise bias supply that is temperature-compensated, using a sensor-transistor mounted on top of one of the power devices.

Unfortunately, it is a sad fact that a high gate voltage is a pre-requisite for this circuit to function, as the gate voltage cannot exceed 2. $V_{g1}=V_{B1}$. The rule is that the mosfet must be able to produce full current at twice the idling gate voltage, the only exception being when a B_2 bias supply is used.

Fortunately, the g_m transconductance of a modern n-channel transistor is very high, typically 5-10A/V, so the current at 7V is often above 20A, which is more than enough.

Hitachi lateral mosfets are often considered as superior to others because they have a lower input capacitance. But it is not the absolute capacitance, rather C_{in}/g_m that determines the capacitive gate current (high g_m means low sr).

High power vertical mosfets are superior in this respect, with a factor of 2-3, and can be used in high quality audio amplifiers.

Bias with adjustable *TC* and voltage

The regulated bias network, V_B, has a temperature sensing transistor Tr_s, mounted in thermal contact with one of the power devices – though not on the heat sink as this may cause thermal run away. The bias can be shunt regulated (**Fig 9a**) or series regulated (**Fig. 9b**). It should match the *TC* of the mosfet channel. Maximum current never exceeds V_{B1}/R_1, and typically it is maximum 25mA for $R_{1,2}$, equal to 330Ω.

TC should be minus 0.1-0.3%/°C and is made up of one constant part (R_3 + R_4 in Fig

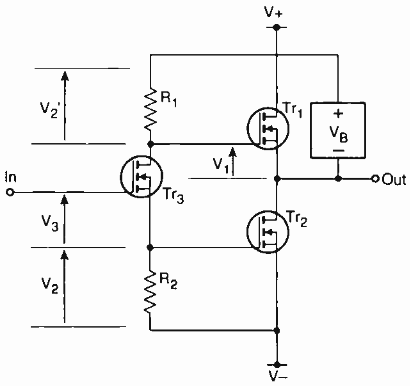


Fig. 5. High input impedance circuit with n-channel drive.

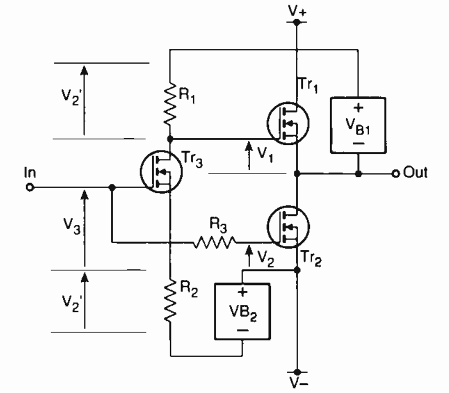


Fig. 6. Anti-saturation circuit (of Fig. 5) with V_{B2}.
Notes. If $R_1 = R_2$
$V_{B1} = V_1 + V_2'$
$V_{B2} = V_3 + V_2' - V_2$
where $V_2' = V_2 + K$
At saturation: $2V_2' = V_{B1} + V_{B2} + V_{sat}$. But V_2 (and V_3) is not limited (isolated gate). V_1 is limited to $V_{B1} = V_1 + V_2 + K$.

9a, R_2 in Fig 9b), and a variable part (–0.33%/°C determined by $R_1 + R_2$ in Fig 9a and $R_3 + R_4$ in Fig. 9b). The current in the sensing device Tr_s is absolutely constant in Fig. 9b so *TC* is determined by the ratio V_E / V_{R2}.

More power

Measurements indicate that, with the same power supply, output power is increased by typically 30% compared to the old audio transistors with their complementary circuit. Using the new technique with 0.2Ω n-channel transistors (at even lower thd) will also make the new amplifiers run noticeably cooler.

Simplicity of the design is clear (Figs. 4 and **10**), with *grounded* output and floating power supply producing excellent results (Fig. 12).

Negative side input with n-channel driver

The examples in Figs. 7 and 10 include a conventional balanced input stage.

Notice that the commonly used second balanced stage is no longer needed as it is already built into the power stage.

The benefit of a double-balanced stage design – no jolt when turned on or off – is retained in this circuit.

Distortion

Distortion is a dynamic problem in mosfet-amplifiers, and depends on the maximum slewing rate (sr) and gain-bandwidth product (gbw) of the driver amplifier. A high value of gbw, giving a large feedback factor, makes it possible to reduce the static non-linearity to practically zero. There is no hidden or unmeasurable distortion linked to this type of linearisation – except that expected from static characteristics and frequency response, which is constant up to the MHz range.

Very consistent data is a valuable property of n-channel mosfets. Different samples show almost identical gate-characteristics and R_{on}, and the gate-voltages become symmetric, resulting in the lowest possible distortion.

In fact, given a specific bias current, the distortion will typically be 26dB lower than with a conventional amplifier. This is because the driver acts directly on the gate, and not on the superimposed gate-plus-output, as is customary. This type of drive (shown in Figs 4 and 7) can be called "direct drive" (DD).

Distortion proves to be almost unmeasurable below 1kHz in a well designed amplifier. At higher frequencies, the thd will increase in proportion to frequency up to 20kHz, (**Fig 12**).

Symmetry at high frequencies.

Gate input is capacitive (800-2000pF), causing a phase shift at Tr_2 in Fig 1, and cross-conduction at hf unless the generator (drive) impedance is equal for both Tr_1 and Tr_2. According to Thevenin, the drive impedance to Tr_2 is equal to R_2. The drive input to Tr_1 is zero, taking the input as a reference, but it can be made equal to that of Tr_2 if R_3, equal to R_2, is inserted as shown in Fig 4, at the 'source' transistor.

Now the phase difference will be minimal up to several hundred kilohertz, with negligible cross-conduction.

Anti-saturation circuits.

Use of low R_{on} transistors increases the output power, given a specific power supply. But there is one problem: the gate voltage must be sufficient to saturate the transistor.

The gate voltage of Tr_1 is floating and will exceed the positive rail voltage. Conventional mosfet amplifiers usually have an extra positive power supply, but this is not needed in this amplifier.

However, for Tr_2 the situation is different. Turned on hard, Tr_3 will saturate (the voltage is typically only 50mV across Tr_3).

Adding voltages from the negative rail (**Fig. 3**) will accent the problem:

$V_{sat} + V_B = V_2 + V_2'$

$V_{2sat} = {}^1/_2(V_B + V_{sat})$

Normal bias is ${}^1/_2(V_B)$. Thus, the gate voltage of Tr_2 increases only ${}^1/_2(V_{sat})$ at saturation, which is insufficient, bearing in mind the square-law gate characteristic.

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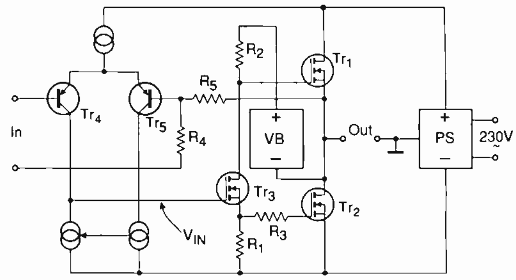


Fig. 7. Block diagram, conventional amplifier.

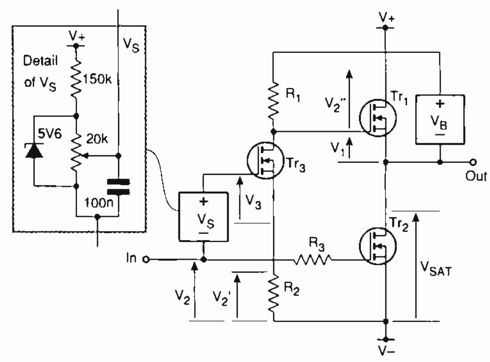


Fig. 8. Anti-saturation circuit (of Fig. 6).]
Notes. $V_B = V_1 + V_2''$
$V_2' \approx V_2''$
$V_2 + V_s = V_2' + V_3$
$\approx V_B - V_1 + V_3$
Make $V_s \approx V_3$. Adjust bias with V_B.

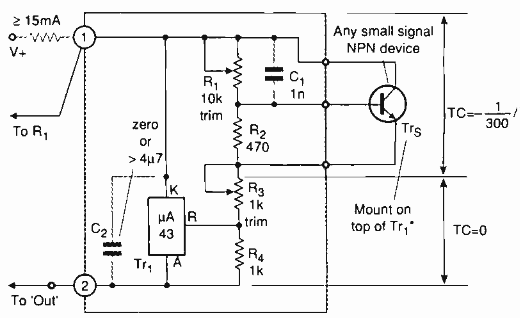
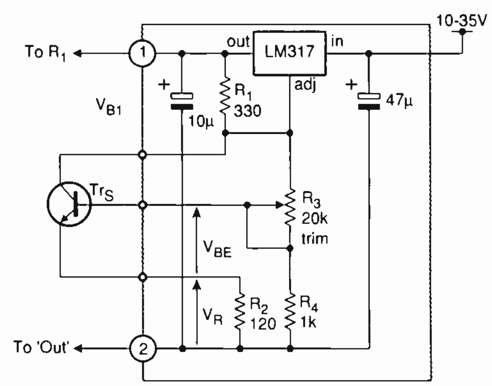


Fig. 9a. Bias with adjustable TC and voltage: shunt regulator.



9b. Series regulator.
Note. $V_R = 0.4724V$ and $V_{BE} = 0.5306V$, so variable part is 52.9% and fixed part is 47.1% which will match (V_{B1} –0.125V) perfectly.

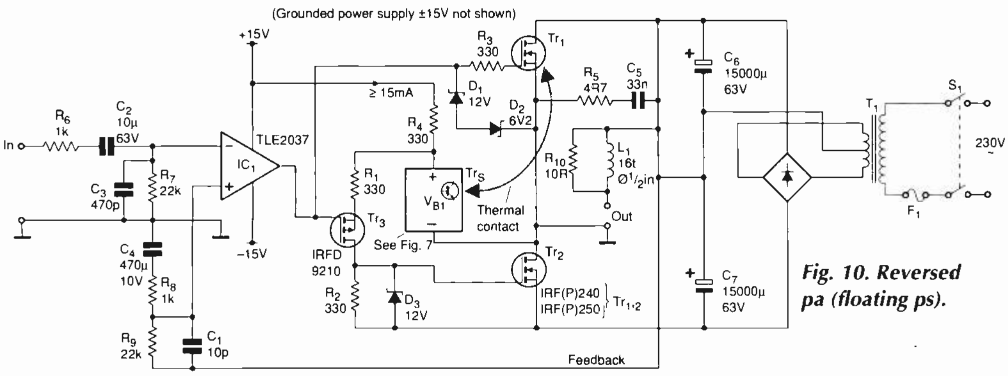


Fig. 10. Reversed pa (floating ps).

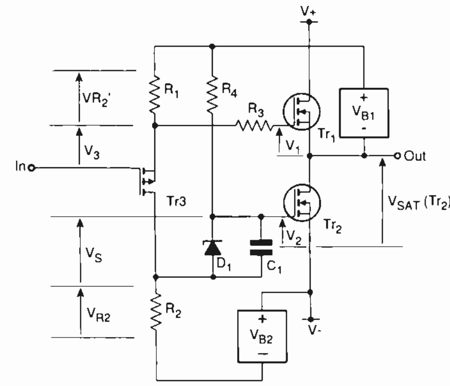
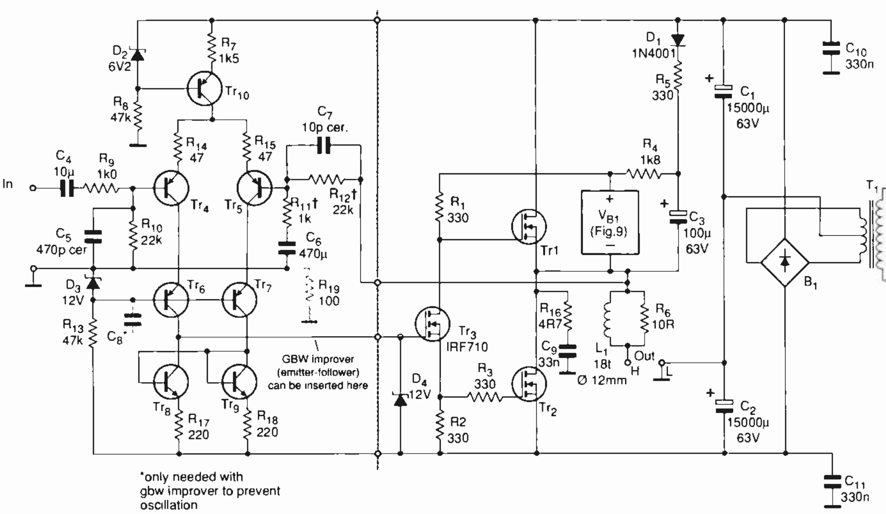


Fig. 13. Anti-saturation circuit (of Fig. 3). Note Tr_3 may be n-channel. $V_{B1} - V_{B2} = V_1 + V_2 - V_s$ (if $R_1 = R_2$).



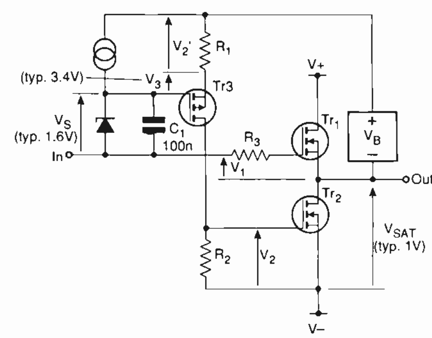


Fig. 14. Anti-saturation circuit of Fig. 1.
Notes. Assume $R_1 = R_2$
$V_B = V_1 + V_s + V_3 + V_2'$
Saturated:
$V_2 = {}^1/_2(V_{sat} + V_B) = V_2 + {}^1/_2(V_{sat} + V_3 + V_s)$

Figure 11. Conventional power amplifier.
Parts list.

R_1, R_2, R_3 330Ω
R_4 1.8k 1W
R_5 330Ω 0.5W
R_6 10Ω 2W
R_7 1,5k
R_8 47k
R_9 1k
R_{10} 22k
R_{11} 1k met f 1%
R_{12} 22k met f 1%
R_{13} 47k
R_{14}, R_{15}, 47Ω
R_{16} 4.7Ω NI 5W
R_{17}, R_{18} 220Ω
R_{19} 100Ω
C_1, C_2 15000μF 63V
C_3 100μF 63V
C_4 10μF
C_5 470pF cer
C_6 470μF 10V
C_7 10pF cer
C_8 N/A unless gbw improver is used (Oscillation suppressor)
C_9 0.033μF 250V polypropylene
C_{10}, C_{11} 0.33μF polypropylene
Tr_1, Tr_2 2, 4, 6 or 8 units. IRF(P) 240 or IRF(P) 250.
Tr_3 IRF710
Tr_4, Tr_5 any low noise pnp
Tr_6, Tr_7 MPSA 92
Tr_8, Tr_9 any high gain npn
Tr_{10} MPSA 92
D_1 IN4001
D_2 BZX55 6V
D_3, D_4 BZX55 6V
L_1 18T dia 12mm.

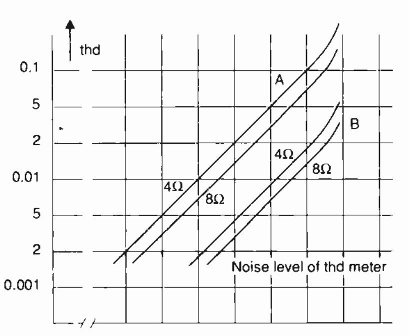


Fig. 12. Distortion at 20Vrms output, single pair of IRFP 240s.
A. For the schematic of Fig. 10.
B. For Fig. 11 with an additional buffer inserted before gate Tr_3 to put less load on the input amplifier and to achieve higher loop gain and more feedback.

Feedback will try to maximise the output voltage, limiting the gate voltage of Tr_2 and raising V_{sat} – but not higher than is consistent with maximum output current and voltage.

This "deficiency", more pronounced at high output current, is indicated by a horizontal clipping which can be observed on an oscilloscope. Approximately 2V is lost in comparison to the situation with natural R_{on} saturation, producing a 10% loss of efficiency at 40V peak output. No distortion is caused below the clipping level.

There are several ways around this problem. **Figure 14** shows a modification of the basic circuit, with additional bias V_s. The bias of B_1 now has to be increased in line with V_s too. Gate voltage on Tr_2 is increased by ${}^1/_2(V_3 + V_s + V_{sat})$, say 3V, when Tr_3 is saturated. That may be enough to saturate Tr_2 at the prevailing output current.

Other ways are possible, but they all involve an extra bias around Tr_3.

Design advantages

The new mosfet-power amplifiers deliver several useful advantages: low distortion (due to "direct drive"), high efficiency (30% more output using the same power supply), savings in parts and labour (typically 25%) and high output voltage (up to 350Vpk-pk in a single output triplet).

Reliability and cost benefits are possible too, as a result of using proven industrial n-channel transistors.

Apart from these improvements, the amplifier is also easy to work with, being very "forgiving" and easy to stabilise. Extremely low levels of distortion are easily achieved and tests show that the gate wave-form is perfectly symmetric.

Maybe the days of the complementary stage are numbered. The troubles with p-channels could be something of the past. ■

I^2C via the PC

There is now a wide variety of I^2C chips performing functions ranging from simple i/o switching to frequency synthesis. This two-transistor interface provides a simple means of communicating with these devices via the pc's parallel printer port, as John Davies explains.

Previous articles in *EW+WW* have looked at the concept of the I^2C bus and also the use of the pc parallel port as a more general purpose interface. This article expands on both topics, presenting a simple way of implementing an I^2C port using the pc parallel port.

Only the master-slave part of the I^2C standard is implemented, with the pc acting as the master. Implementing a full master-master interface using the parallel port would be too processor intensive, leaving little time for the pc to do anything else.

Interfacing entails some very simple hardware together with general-purpose software drivers. These are written in assembler but could easily be translated into higher level languages such as C. No real benefit would be obtained from using interrupts in this application, so the software is simplified even further.

Hardware and software are described by using an example of a simple event counter using the readily available Philips *PCF8583* chip which is a clock/calendar/256byte ram device

Following sections describe both the software and the hardware. I have assumed that you are reasonably familiar with the I^2C bus concept, and with the pc parallel port. Some additional information is presented in the references at the end of the article.

Hardware

Figure 1, on page 996, illustrates the interface's simplicity. Conversion to and from the parallel port ttl level interface to the I^2C bus is done using the two transistors $Tr_{1,2}$. These provide the open collector outputs needed for the data and clock lines on the I^2C bus.

Input to the parallel port for reading the I^2C bus is taken directly from the bus as the 5V level used on this version of the I^2C bus is ttl compatible. Obviously if any other voltage was used on the I^2C bus some level shifting would be necessary.

Discrete transistors are used rather than '74 series open-collector gates partly because they are cheaper and partly because the '74 series device would need a 5V power supply. This is not a problem where the I^2C bus is at 5V level, but it would be at other voltage levels.

Any general purpose npn transistor is suitable, as long as it is capable of switching at around 100kHz. Values for $R_{1,2}$ depend on the physical characteristics of the I^2C bus, but 2.7kΩ is reasonable.

Some of the parallel-port inputs and outputs are inverted in the internal hardware and the transistor stages introduce an inversion. These are

Continued over page

Inside the PCF8583

Within the 8-pin *PCF8583* are 256 bytes of ram, an oscillator, a frequency divider, the I²C interface and a power on reset circuit. The first eight bytes of ram are used to control the modes of operation of the device, and to store counter information. The next eight bytes can be used either as free ram or as alarm registers.

Three different modes of operation can be selected by programming the control and status register at address 0. Format of this register is:

Bit Function
0 Timer flag
1 Alarm flag
2 Alarm-enable bit (1=enable)
3 Mask flag (affects addresses 5, 6)
4,5 Mode: 00=32768kHz oscillator
10=50Hz oscillator
01=event counter mode
11=test mode
6 Hold last count (1=hold)
7 Stop counting flag (0=count)

A memory map of the other registers is shown in **Fig. 2**. The first two modes are both real-time clocks, one running from a 32768kHz clock, the other from a 50Hz input. In these modes, time of day is stored in bcd format. Up to six digits of data ($D_{0\text{-}5}$) are stored in the event counter.

Setting the alarm enable bit of the control/status register activates the alarm function. Via the alarm control register, a dated alarm, daily alarm, weekday alarm or timer alarm may be programmed. Format of the alarm control register is as follows:

Bit Function
2-0 Timer function:
000 – no timer
001 – 1/100s
010 – seconds
011 – minutes
100 – hours
101 – days
110 – not used
111 – test mode
3 Timer interrupt enable
5,4 Clock alarm function:
00 – no clock alarm
01 – daily alarm
10 – weekday alarm
11 – dated alarm
6 Timer alarm enable (1=enable)
7 Alarm interrupt enable

Whenever an alarm occurs, the alarm flag of the control/status register is set. A timer alarm event sets the alarm flag and an overflow condition of the timer will set the timer flag. The open drain interrupt output is switched on when the alarm or timer flag is set.

In clock mode more detail is available in the hours, months and years bytes as follows:

Hours (address 04)
Bit Function
3-0 Unit hours in bcd
5,4 Tens hours in bcd
6 am/pm flag
7 24/12 hour format clock (0=24 hour)

Year/date (address 05)
Bit Function
3-0 Unit days in bcd
5,4 Tens days in bcd
7,6 Years 0-3 (from leap year)

Weekdays/months (address 06)
Bit Function
3-0 Unit months in bcd
5,4 Tens months in bcd
7-5 Weekdays (0 to 6 binary)

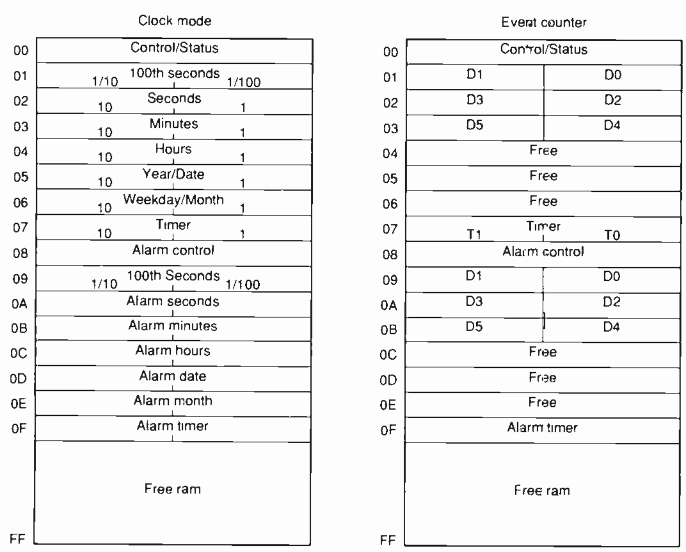
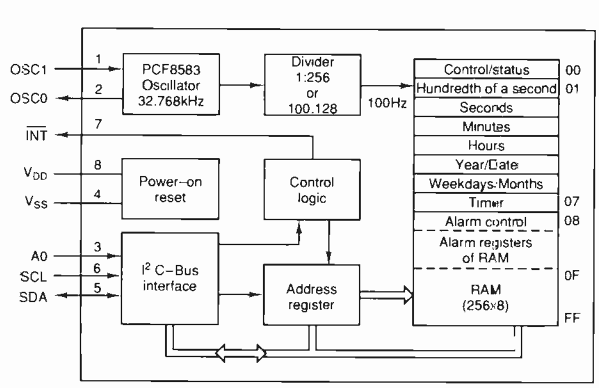
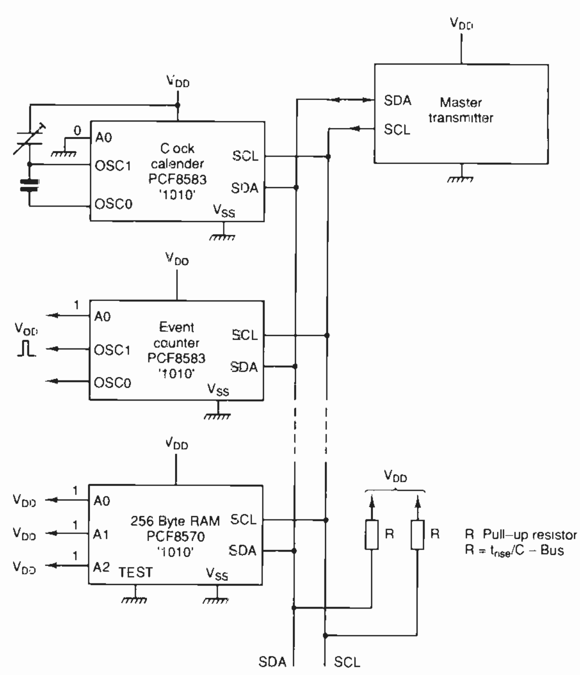


Fig. 2. Control and status registers, left, allow programming of the PCF8583 as a real-time clock or event counter. When acting as an rtc, it can be driven from a 50Hz or 32.768kHz clock. Acting as an event counter, the oscillator input allows pulses to be counted. Below left is a functional diagram of the device accompanied below right by an application circuit showing an rtc configured PCF8583 together with two event counters.





compensated for in the software drivers.

There is an address pin on the *PCF8583*, namely A_0, which can be used to differentiate between two devices on the same bus. This pin is tied low in this application.

Software

There are two sections to the software. Firstly there are the general purpose I²C drivers, and secondly, routines specific to this application.

The application software simply reads how many events have occurred on the *PCF8583* input since the last read and prints the result on the screen. Drivers for the I²C bus are modular, which makes debugging and modification of the code much easier. **Figures 3, 4** show the structure of these drivers.

Application software interfaces to the I²C drivers also divides into two main parts – one for transmission of an I²C message to a specified address and one for the reception of an I²C message from a specified address. Data passes to and from these interfaces in the form of a message giving the number of bytes to be written or read, the I²C address and the actual data to be transmitted, or where the read data is to be stored.

Consider the transmission software first, Fig. 3. The top level interface is a routine called 'tx_mess'. This routine takes the data passed to it. Using the three routines called 'i2c_start', 'i2c_stop' and 'tx_byte', it transmits the message in the correct format.

A further a routine, 'read_ack', is called by 'tx_byte'. This routine checks that the receiver has acknowledged each of the bytes. If the acknowledgement is not received then the sequence is terminated and an error returned.

All of the above call the three lowest level routines. These are 'tx_bit', 'set_clock' and 'rx_bit' and they output the data onto, or read the data from, the i²C bus. Bit inversions due to the hardware are also taken care of here.

Interfacing for receiving, Fig. 4, is a little more complicated. To receive data, some of the transmit routines must also be used for initially addressing the *PCF8583*. Once this has been done, the receive message routine 'rx_mess' calls the receive byte routine 'rx_byte' the required number of times.

The routine 'send_ack' indicates to the slave device that each byte has been received. Note that there is a slight quirk in that the last byte received does not require an acknowledgement. This indicates to the slave that all bytes have been transmitted.

Finally the 'I2c_stop' routine is used to complete and terminate the transmission sequence. Again the low level interfaces to read or write to the parallel port pins are used.

The software initialises the *PCF8583* as a counter device, zeros its count and waits for the pc operator to strike a key. It then reads the *PCF8583* count buffers and displays the result on the screen. An example of the message sequences which are used by the routine 'tx_mess' is as follows

Reset counters:

5 Number of bytes
A1 *PCF8583* address / write command
00 Start address to write
20 Command to start counter
(see explanation of *PCF8583*)
00 |
00 | Zero counters
00 |

Final thoughts

Although the circuit and software described here are simple, they could form the basis of a more sophisticated system. A description of how to extend the range of the I²C bus, or how to isolate the power supplies of two different systems, is given in the I²C articles mentioned below.

On the software side the two message interfaces (tx_mess and rx_mess) could be extended to interface to a high level language such as C. For the really ambitious the software could be extended to include master to master communications but as mentioned previously doing this via the printer port could result in the p.c. spending most of its time servicing the I²C bus rather than doing anything else. ■

Further reading

Busman's guide to i²c – EW+WW Jan/Jun 1994.
Philips I²C bus specification.
Philips *PCF8583* data sheet.

I²C software on disk

An annotated assembly-language listing for transmitting and receiving I²C via the PC is available. To obtain it on 3.5in disk, together with the listing for Mike Button's article in the June 1994 issue, send a postal order or cheque for £12.50 inclusive, payable to Reed Business Publishing, to *EW+WW* Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Please mark 'I²C via LPT' clearly on your envelope.

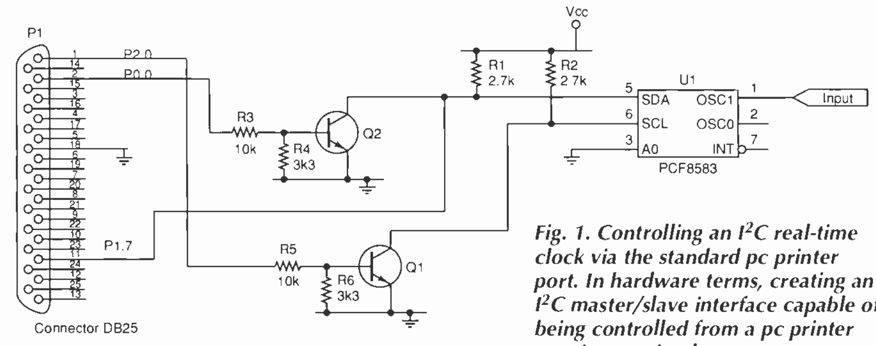


Fig. 1. Controlling an I²C real-time clock via the standard pc printer port. In hardware terms, creating an I²C master/slave interface capable of being controlled from a pc printer port is very simple.

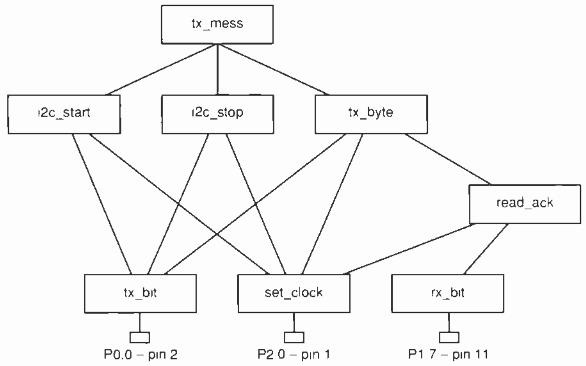


Fig. 3. Software flow for controlling I²C bus transmission from a pc via the printer port.

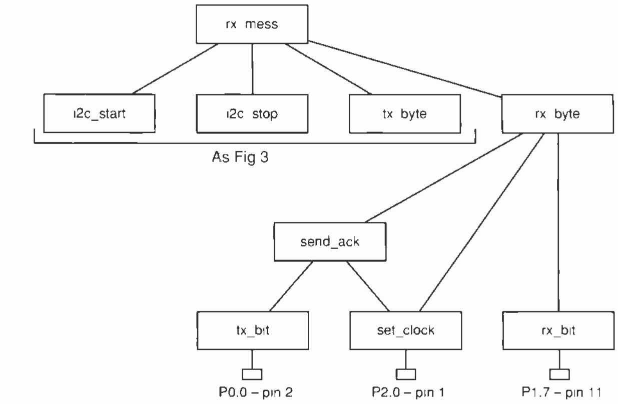


Fig. 4. Receiving I²C data via the PC printer port is a little more complex than transmitting it.

Harmonising theory with practice

Textbooks rarely explain the frequent gap between learned theory and practical result viewed on an oscilloscope. Ian Hickman explores the relation between time and frequency-domain representations of common waveforms to provide an answer.

Take a sine wave of angular frequency ω rad/s, say, or ω/2π cycles/s and compare it with its third harmonic at one third of the amplitude, its fifth at one fifth and its seventh at one seventh of the amplitude. Relative to the fundamental, the amplitude of each harmonic is inversely proportional to its order. All the components start at the left hand end of the plot at time $t = 0$, so that the angle ωt is also zero. As they are all sine waves, they all start from zero, positive-going.

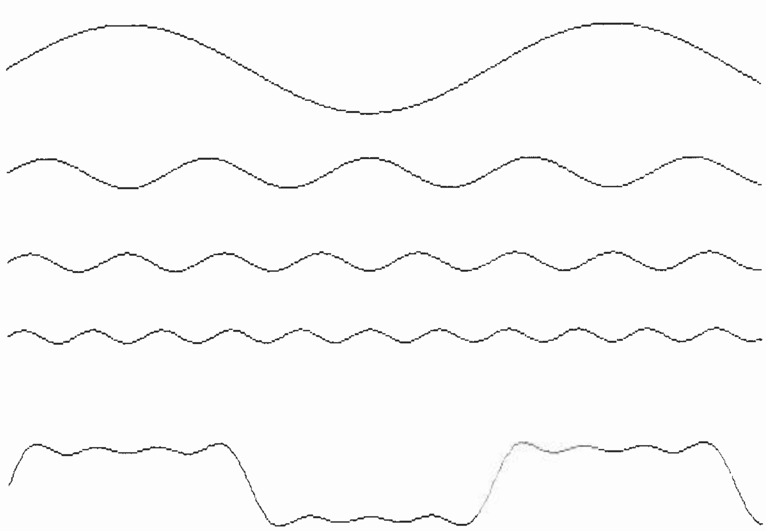
Examining their sum (**Fig. 1**, bottom trace) shows an already passable approximation to a square wave is beginning to emerge.

(a)

The 'flat' top of the square wave has three dips and four bumps, indicating that only harmonics up to the seventh are present. A good few more would be needed to make the top sensibly flat and, in particular, to make the rising and falling edges vertical, at ωt=0 and ωt=π radians. Note that the sine wave positive peak coincides with the negative peaks of the third and seventh harmonics (and 11th, 15th, etc.) while it coincides with the positive peak of the fifth harmonic (and 9th, 13th, etc).

On the other hand, at t=0 all are positive going, resulting in an infinitely steep rising edge, if odd harmonics are included all the way up to infinity.

Fig. 1. Top to bottom, fundamental waveform, with its third, fifth and seventh harmonics, at the appropriate amplitude and phasing. Their sum is the bottom trace. As more harmonics are added, the sum becomes ever closer to an ideal square wave.



The four waveforms shown are of course simply those predicted by Fourier analysis of a square wave. For a 'triwave' or triangular wave, this time the positive peak of the sine wave coincides with the positive peaks of all the harmonics (first four again shown in **Fig. 2**), resulting in the peak of their sum (bottom trace) becoming ever sharper as more harmonics are added. In addition to the altered phasing of the 3rd, 7th, 11th... harmonics, this time the amplitudes of the 3rd, 5th, 7th... harmonics are $^1/_9$th , $^1/_{25}$th and $^1/_{49}$th... that of the fundamental; ie, inversely proportional to the square of their order. So in the triwave (**Fig. 2a**) the fundamental has been plotted at a larger amplitude than in the sine wave, enabling the harmonics, which are now of very low amplitude – especially the 7th and higher – to be seen.

Taking just the fundamental and even a few harmonics, the triangular wave is now very convincing. It needs only to be a bit sharper at its tips to be perfect.

Inverted question

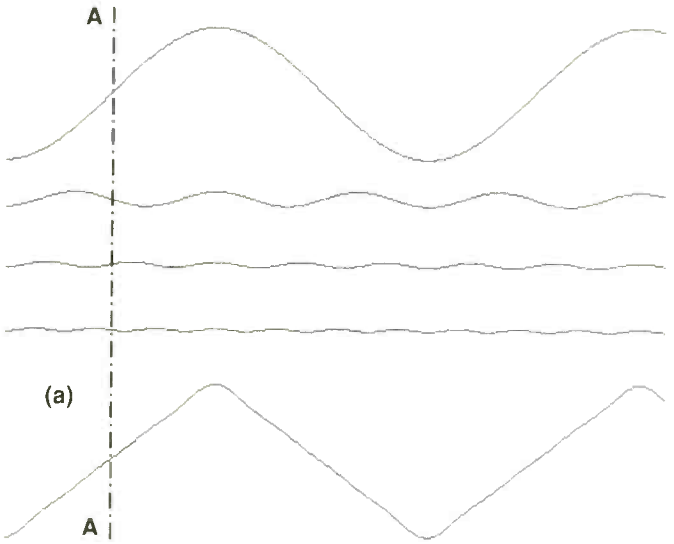
So how is it that, in the triwave, the 3rd and 7th, etc harmonics have become inverted in phase, while the 5th, 9th... have not?

To clarify this, we must change the measurement domain in order to show how the square wave can be converted to a triwave in practice.

Figure 2b shows an op amp integrator and **Fig. 2c** illustrates its frequency and phase response. Its gain is unity at that frequency where the reactance of the capacitor in ohms equals the value of the input resistor R, falling at 6dB per octave above this frequency and rising at 6dB/octave below it (in both cases, for ever more, if the op amp is perfect).

In addition, an input sine wave, of any frequency, suffers a 90° phase lag in passing through the circuit. When the input sine wave is at its positive peak, the output is zero and increasing at its maximum rate.

The more usual sort of op amp integrator is in fact an inverting device, so that the 90° phase lag looks at its output like a 90° lead. So **Fig. 2b** is actually a non-inverting integrator circuit – sometimes known as a de Boo integrator – which behaves exactly like an implementation of the mathematical operation of integration.



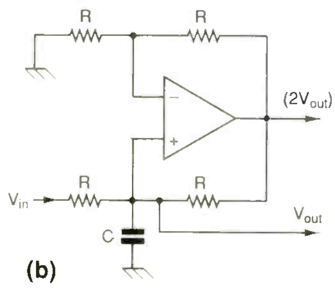
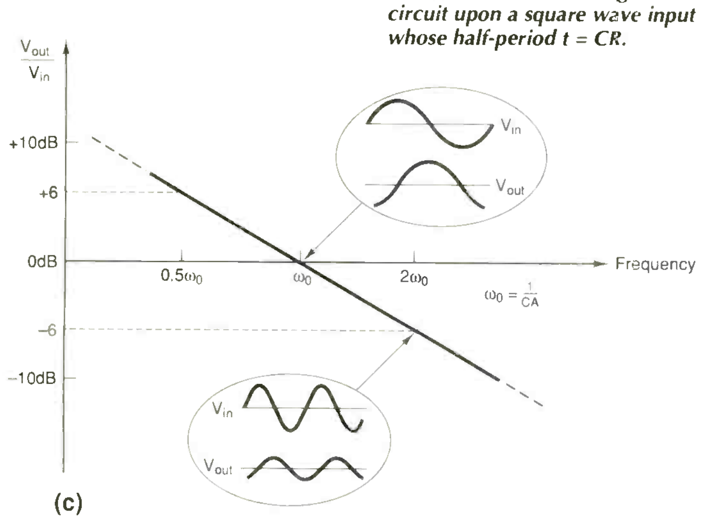
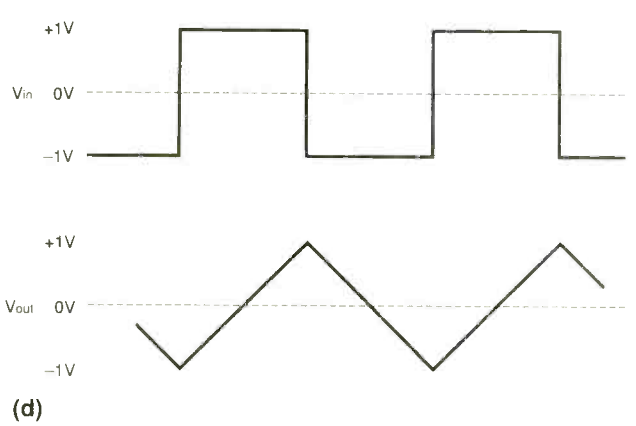


Fig. 2a). As Fig. 1, except that the phasing and amplitude of the harmonics are those appropriate to generate a triwave.
2b). Op amp integrator circuit that behaves like a true integrator. Its output voltage increases when the input is positive, unlike the usual op amp integrator which is actually an inverting integrator.
2c). Integrator's frequency and phase response.
2d). The effect of an integrator circuit upon a square wave input whose half-period t = CR.





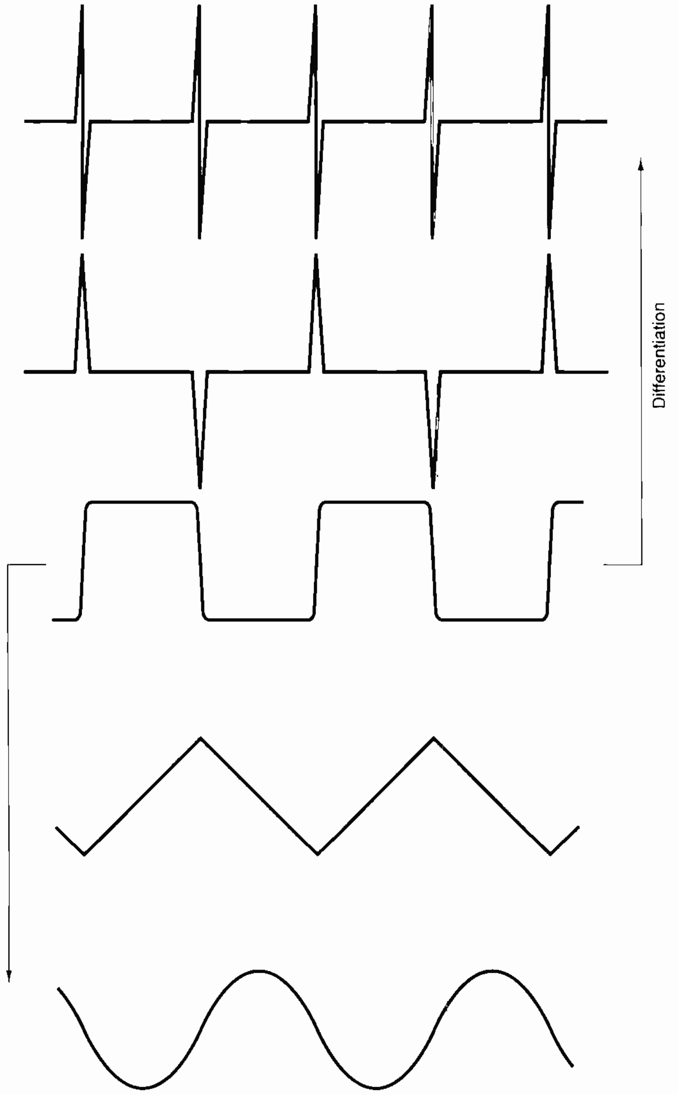


Fig. 3. Effect of repeated integration and differentiation on a (nearly) square wave.

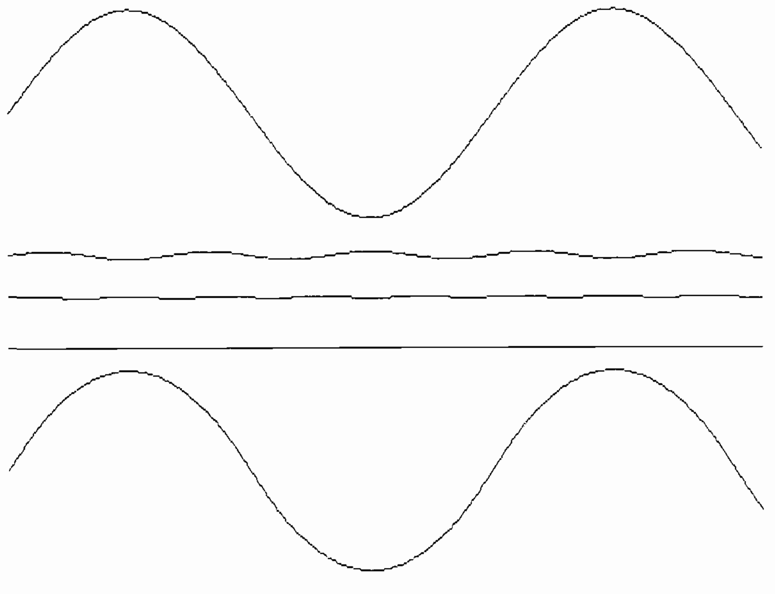


Fig. 4. Fundamental plus 3rd, 5th and 7th harmonics of a triwave that has been integrated, and (bottom) the resultant.

Examining the response of the integrator to a square wave input (**Fig. 2d**) shows that all the while the input is positive, the output increases. Likewise, when it is negative the output decreases. But to understand the relation between the time- and frequency-domain representations, we must see how the effect of the integrator on the individual harmonics in Fig. 1 results in the corresponding harmonic components in Fig. 2a.

Compared to their phasing in Fig. 1, the integrator has delayed all the frequency components in Fig. 2a by 90° – at the left hand side where $t = 0$ they are all at their negative peaks and do not pass through zero going positive until 90° later.

Now, for the fundamental, this corresponds to the time indicated at A–A (Fig. 2a). But at this point the third harmonic has moved through 270°, being at three times the frequency. Discounting the 90° phase shift suffered by the third harmonic itself, this means that, net, it has moved forward 180° relative to the phase relation with the fundamental at the input to the integrator. This is clearly shown in Fig. 2a, where at point A–A (corresponding to the left hand side of Fig. 1) the fundamental is passing through zero positive-going, and the third harmonic is passing through zero negative-going. So its positive peak now coincides with that of the fundamental. But the 90° delay of the fundamental at A–A corresponds to 450° at the fifth harmonic, or 360° discounting the 90° phase lag suffered by the fifth harmonic itself. In fact its phasing relative to the fundamental is unchanged. Now, the positive peaks of all odd harmonics, not just those of the 5th, 9th..., etc., coincide with that of the fundamental, resulting in the sharp point of Fig. 2a.

Integration and differentiation

We should now consider the effect of repeated integration on a square wave, and of repeated differentiation, **Fig. 3**, assuming a finite rise-time to avoid infinite amplitudes. Taking integration first, **Fig. 4** shows the effect of a second integration of a square wave – ie integrating the triangular wave of Fig. 2a. Again, the phase of the third and seventh harmonics, but not the fifth, has inverted. (The 7th might look like a straight line, but its amplitude is in fact $1/7^3$ times that of the fundamental, just –51dB or a more or less negligible 0.29%).

At 1/27th of the fundamental, the third harmonic is responsible for 3.7% distortion, the total harmonic distortion being under 4%. So, for uncritical applications, a twice-integrated square wave could stand in for a sine wave, although as can be seen by comparing the bottom trace in Fig. 4 with the sine wave top trace, it is visibly just a little too rounded at the peak.

In differentiation (**Fig. 5**), the differentiator has a frequency response which rises at 6dB/octave – the very reverse of the –6dB/octave of the integrator illustrated in Fig. 2c. The harmonics are emphasised so that they are now the same amplitude as the fundamental.

Compared with the square wave of Fig. 1, in differentiation all the frequency components are advanced in phase by 90°. Like the triangular wave, but unlike the square wave, their positive peaks all line up with that of the fundamental. This is greatly accentuated into alternate positive and negative spikes (bottom trace) which, as more and more harmonics of appropriate phase and amplitude are added, would turn into infinitely high positive and negative spikes or 'delta functions'.

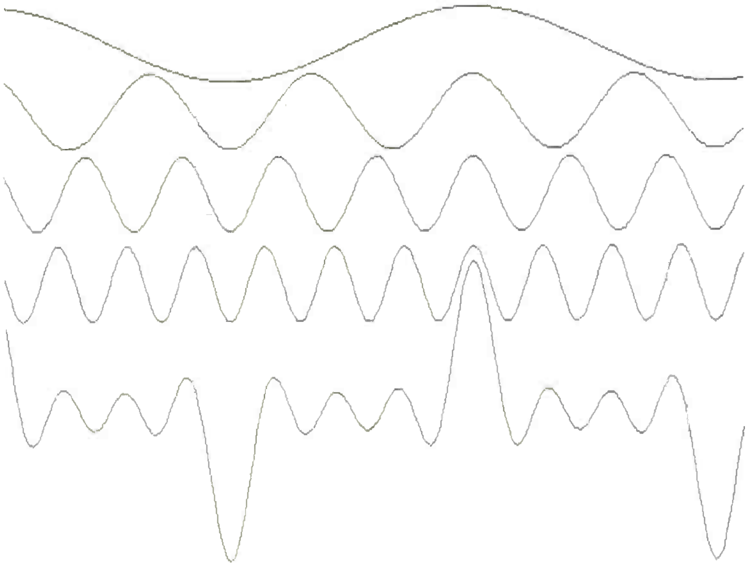


Fig. 5. When a square wave is differentiated, relative amplitudes and phases of the fundamental and the first three odd harmonics add to produce alternate positive- and negative-going spikes (bottom trace). Since, after differentiation, the amplitude of each odd harmonic of a square wave is the same as that of the fundamental, many more harmonics would need to be shown to get very near the ideal waveform.

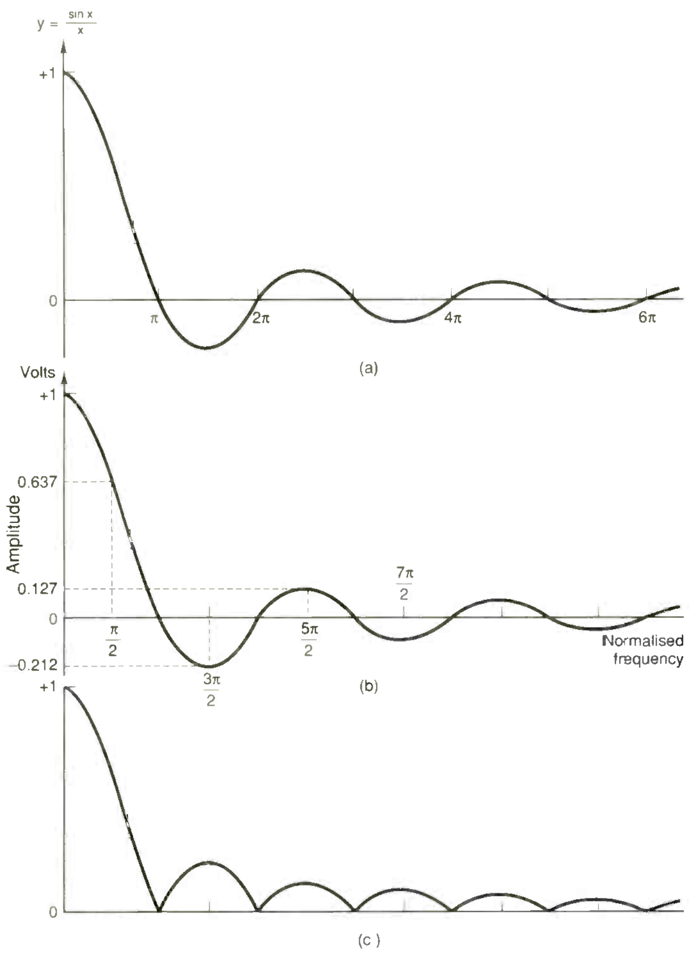


Fig. 6a). The "sin x upon x" function, plotted for positive values of x only.
6b). How the value of y gives the amplitude and phase of the various frequency components of a square wave when the frequency of the fundamental is represented by the value π/2.
6c). Modulus of 6a), showing only the numerical value of the function at each point, regardless of its sign.

Mathematical treatment

Returning to the square wave, the relative amplitudes and phases of the fundamental and its harmonics are described exactly by an important mathematical function which, like π and exponential e, turns up all over the place:

$$y = [\sin(x)]/x.$$

When plotted, clearly the function is going to look like a sine wave, but getting smaller on successive cycles, due to the x in the denominator. But although $\sin(x)$, and hence y, will generally be zero each time x is an integral multiple of π – ie at 180°, 360°, 540°, etc, –0° is a special case. Here, $y = 0/0$. The standard way to evaluate a function at a point where its value is the quotient of two noughts is De l'Hospital's rule. But in this case it can be done by mental arithmetic, remembering that for $x << 1$rad, $\sin(x)$ approximately equals x. The smaller x is, the more nearly exact is the approximation, so that as x tends to zero, y tends to 1.

Traditionally, x is used as the variable in this function, rather than, say, θ, with good reason. The latter suggests an angle, whereas the variable may often be something different. In the case of square waves of all sorts, a useful representation is when x represents not the instantaneous phase angle of the fundamental, but rather the various frequencies involved. The first zero of the function, when $y = 0$, occurs when $x = 180°$ or π radians (**Fig. 6a**).

Now the positive half cycle of a square wave is identical in shape to the negative half cycle, in the sense that if you flip the positive half down below the horizontal and then slide it along half a cycle, it fits exactly. This indicates that the waveform contains no even harmonic components, a point which you can verify for yourself by adding waveforms graphically as in the illustrations here, or by referring to old college notes on Fourier analysis .

What happens if we let the point $x = \pi$ represent the frequency of the waveform's (missing) second harmonic component? The frequency of the fundamental would then be $\pi/2$; of the third harmonic $3\pi/2$; of the fifth, $5\pi/2$ etc. Corresponding values of y would then be $[\sin(\pi/2)]/\pi/2 = 2/\pi$, one third of this value, one fifth and so on... precisely the relative amplitudes of the fundamental and harmonics of a square wave.

Here x represents the radian frequency, often called Φ, and x or Φ equals $2\pi f$, where f is the frequency in hertz or cycles/s. So a radian frequency $x = \pi/2$ corresponds to $(\pi/2)/2\pi$, or 0.25Hz, but the curve can still represent any frequency square wave, simply by introducing a suitable scaling factor.

Similarly, a scaling factor can be used to adjust the amplitude y to represent the actual amplitude of any particular square wave. In this way the curve can represent the constituent frequency components of any square wave in both frequency and amplitude – and notice that its negative loops show the phases of the 3rd and 7th... harmonics to be opposite to that of the fundamental, 5th and 9th..., **Fig. 6b**. Sometimes, the curve is drawn to represent only the amplitudes of the harmonic components, not their phases, **Fig. 6c**. This is similar to a spectrum analyser display, where only the relative amplitudes of the components of a complex waveform are shown, without information defining their relative phases.

The curve of Fig. 6a also fits the spectrum of asymmetrical square waves and pulse trains of all sorts.

In a subsequent article, Ian is to discuss the implications of waveform representations for d-to-a converters in digital audio.

DETECTION before the diode

In the decade before thermionic diodes became widely used for radio reception, the barretter was one of the most popular devices available for rf detection. Here, George Pickworth discusses his work in replicating some of these early electrolytic detector designs.

In 1903, Reginald Fessenden patented the first practical electrolytic detector. He called his detector a barretter – a name apparently derived from the French word for 'exchanger'. This name implies the exchange of ac for dc – i.e. that the device behaved as a rectifier.

Barretter became the generic name for a range of electrolytic detectors based on Fessenden's design. A typical receiver circuit is shown in **Fig. 1**. In North America, the barretter was the main commercial successor to the coherer It competed with many later detectors, including DeForest's audion valve, the gold-point, fused silicon detector and the steel-point carborundum detector, until about 1913 when the triode valve appeared.

A version of the barretter was made by the British Insulated Wire Company, BIWC, but it was not widely used in the UK, where Marconi's magnetic detector was a popular successor to the coherer. The barretter was however popular with North American experimenters and a number of variations evolved.

During its early years, the barretter was the standard for sensitivity and was generally quoted as being somewhat better than Marconi's magnetic detector. On the other hand, the magnetic detector was rugged, reliable and – unlike the barretter – required no further adjustment after initial setting up. As a result, it was widely adopted by the UK for maritime use.

Fessenden's original barretter had an open-top cell, making it unsuitable for maritime use. To overcome this drawback, Fessenden patented his sealed-cell barretter in 1904.

Like many radio innovations, the electrolytic detector seems to have evolved as the result of empirical experiments, but it was not the only electrolytic device employed by the pioneers. The Wehnelt Interrupter, for example was an attractive alternative to the vibrator interrupter and was frequently used by Fessenden;

Although applied primarily in wave-train telegraphy systems, the barretter seems to have been the first practical continuous wave detector. Fessenden and Ruhmer, among others, used it to demodulate experimental amplitude-modulated telephony transmissions.

Unfortunately, a search through the literature failed to reveal meaningful technical data on

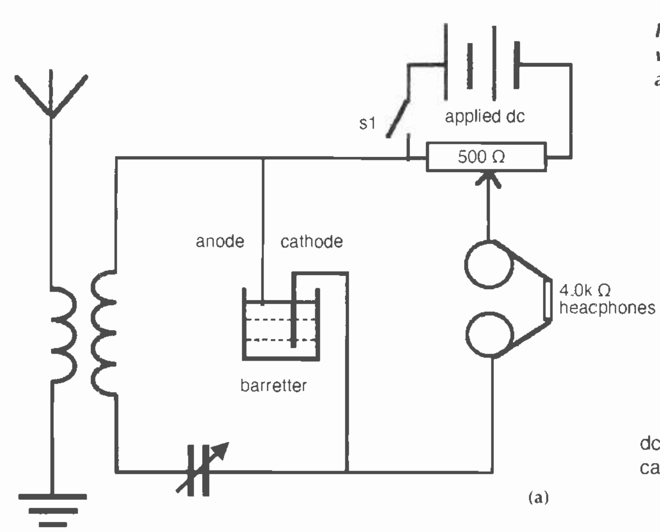
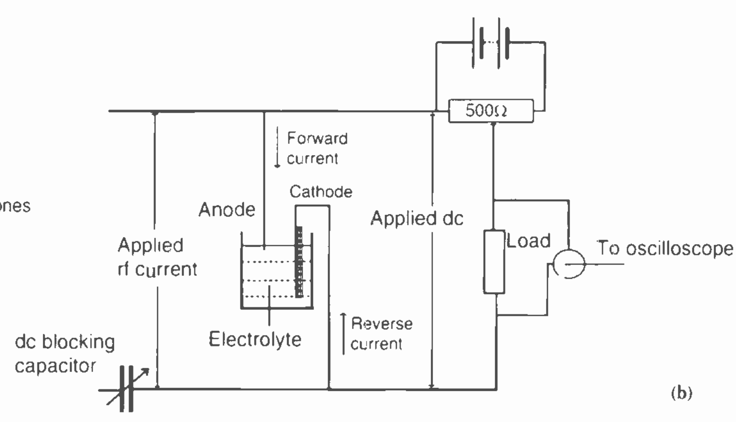


Fig. 1a). Dc bias for the barretter-based receiver, of between 1.9 and 2.5V, was usually provided via a 500Ω potentiometer. Diagram b) details currents and voltages around the barretter electrolytic detector.



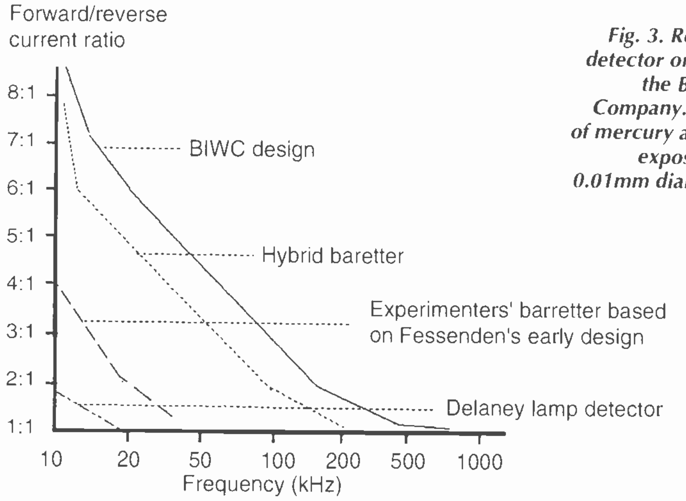
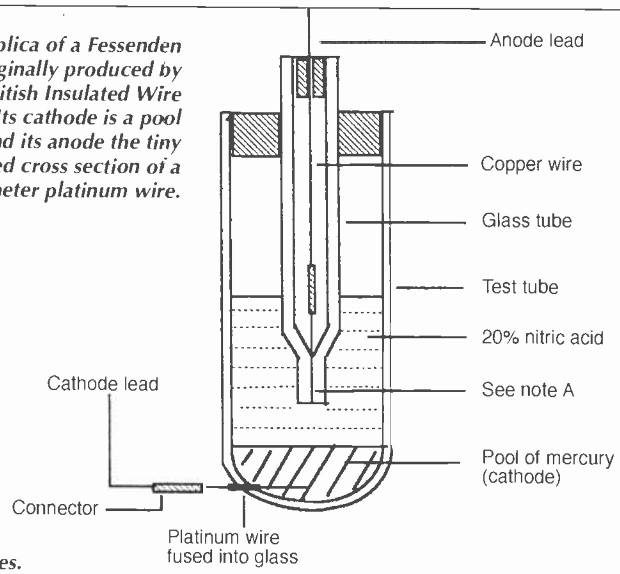


Fig. 2. Electrolytic-detector forward/reverse current ratios. These are derived from relative amplitudes of negative and positive-going half cycles.

Fig. 3. Replica of a Fessenden detector originally produced by the British Insulated Wire Company. Its cathode is a pool of mercury and its anode the tiny exposed cross section of a 0.01mm diameter platinum wire.



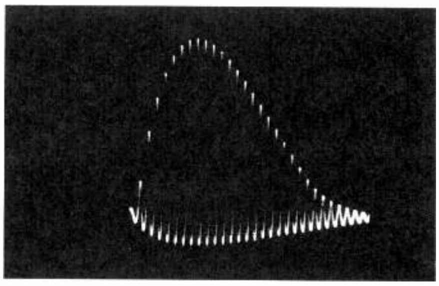
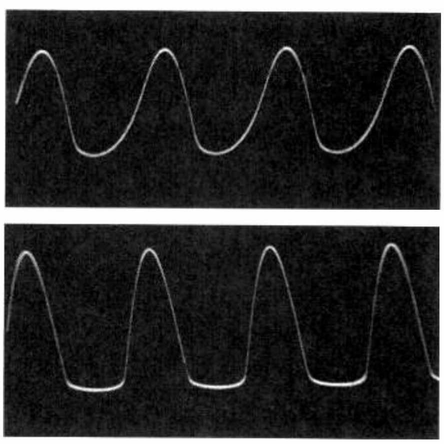
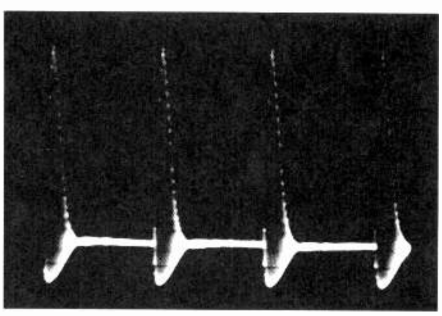


Photo A. When rectifying a 50kHz wave train, the BIWC replica barretter resulted in this half-pear-shape outline.



Photos B,C. Output from the barretter electrolytic detector receiver, top, compared with output from the same receiver but with the barretter replaced by a diode. Both are 50kHz continuous waves.



the barretter, so, from data gleaned from various sources, I constructed a replicas and conducted my own research with a number of variations. These included a simple Fessenden Barretter, a *circa* 1905 experimenter's barretters based on Fessenden's original device, a BIWC type and a hybrid design.

Rectification

This study confirmed that the end result is rectification due to dissipation by the barretter of positive-going rf half cycles; this effect can of course be simulated with a modern diode.

Operation of the barretter is complicated and involves the movement of ions in the electrolyte. A simplistic explanation is that positive-going rf half cycles ride on the steady forward current through the barretter: this is created by the applied dc.

I found the sensitivity of my replica shown in Fig. **1a)** to be poor. With greatly reduced with rf input power, I obtained a tone of similar loudness in a pair of headphones from around 1920 by substituting a diode for the barretter and its applied dc source. Furthermore, rectification efficiency deteriorated as rf frequency was increased.

Rectification efficiency was based on the forward/reverse current ratio, measured by the relative amplitude of negative and positive-going half cycles across the load resistor of headphones, Fig **1b**).

Generally with pear-shaped wave trains, **Photo A**, the best I was attained was about 10:1 forward/reverse current ratio at 10kHz. This deteriorated to about 3:1 at 100kHz and to about 1.2:1 at 500kHz, **Fig. 2**. Rectification efficiency of the barretter is compared with a modern diode in **Photos B** and **C**.

I found that with wave train transmissions, **Photo D**, a forward/reverse current ratio as low as 1.3:1 still gave a clearly audible tone with my 1920 magnetic headphones. But, as

Photo D. Pear-shaped waveforms at 50kHz produced by a receiver based on the hybrid barretter detector.

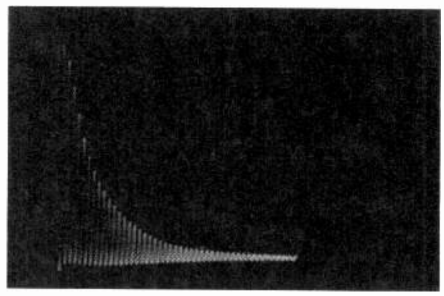


Photo E. Exponentially-declining wave trains at 10kHz produced by a replica of BIWC barretter design.

maritime spark systems operated on frequencies between 500kHz and 1MHz, I assumed that commercial barretters had a better high-frequency response. More about this later.

A better than 5:1 forward/reverse current ratio was found to be necessary to give reasonable speech quality when demodulating am voice transmissions; with my replicas, this was only possible with low frequencies. However, with precise adjustment of rf input and applied dc, almost perfect rectification was attained with exponentially declining waves with a frequency of 10kHz when using the BIWC replica, **Fig. 3**, and **Photo E**.

Recreating Fessenden's experiment

With the kind co-operation of the DTI, I was able to re-create conditions in a disused railway tunnel, similar to those experienced by the pioneers more than 90 years ago. I used my miniature spark transmitter, which, like Fessenden's system, had a spark coil and Wehnelt interrupter. It produced a wave train repetition frequency of approximately a 1kHz. Frequency was 50kHz.

I faithfully replicated equipment used by Fessenden – the only exception being that the transmitter was inductively coupled to the receiver. Hearing a musical note virtually identical to that first heard by Fessenden was a remarkable experience.

Musical note

To produce a pure musical note, the period between wave trains and must be precisely constant Photo D. Transmitters with rotary dischargers gave the purest tone, but as you have seen, a spark coil and Wehnelt interrupter could also produce pleasant tone: the makes/breaks were 'clean' with an almost constant period and much faster than was possible with vibrators type interrrupters; these suffered from contact bounce and imprecise make/break. (See *Spark transmitter technology EW+WW* Nov. 93.)

Experiments showed that the shape of the wave-train envelope was not critical and that a pleasant note could be obtained with both pear-shaped, Photo A, and exponentially declining wave trains, Photos A&E.

Loudness of the tone was set by the amplitude of the positive-going half cycles, as displayed by the oscilloscope. It was not significantly reduced by an increase in the magnitude of the negative going half cycles, provided that the forward/reverse current ratio did not fall to less than 2:1

Rf filtering

The circuit configuration precluded rf current from being filtered from the headphone circuit. A capacitor connected across the headphone terminals would be in series with potentiometer, creating an *R/C* circuit that upsets the tuner.

Despite the high inductance of the headphone, their diaphragms respond to a train of radio-frequency half cycles as if they were a single pulse.

Successive wave trains cause the diaphragm to vibrate at the wave train repetition rate, typically 200 to 1kHz, producing a corresponding tone in the earpieces, Photo D.

Diaphragms generally had a natural resonant frequency of about 1kHz, so were most sensitive to wave trains having a repetition rate of this frequency.

To increase both sensitivity and selectivity, the resonant frequency of the diaphragms was occasionally tailored to a specific wave train repetition frequency. Signalling was by transmitting short or long groups of wave trains corresponding to the dots and dashes of the morse code.

Demodulation

Unfortunately, at Radio 4's frequency of 198kHz, the forward/reverse current ratio of my replica barretters was insufficient to allow the transmissions to be used to evaluate the barretter as an am detector.

Nevertheless, experiments involving modulating the function generator output with speech signals confirmed that the replica BIWC barretter was capable of demodulating amplitude-modulated continuous waves at frequencies below about 50kHz. Speech reproduced was reasonably clear.

Barretter versus coherer

As the barretter was a direct successor to the coherer, it was interesting to compare these two devices. The coherer obviously behaved as a latching relay; it was triggered by individual wave trains, thereafter a local dc source operated the morse register. (see *Coherer-based radio EW+WW* July 94)

The principal drawback to the coherer was that it had to be re-set or 'restored' after responding to each wave train; this limited wave train repetition rate to the order of a 100Hz while making signalling very slow.

When the barretter was found to work in a circuit virtually the same as that used by the coherer – with the exception that the headphones were replaced by the morse inker and restorer – it was understandable that the the pioneers considered it as a relay type detector.

The barretter was therefore perceived as a polarised, very-high-speed, non-latching relay

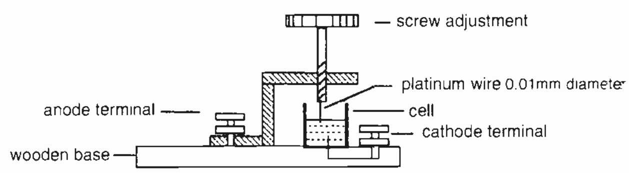


Fig. 4. In this simplified Fessenden barretter, platinum-wire electrodes are used in conjunction with a 20% nitric acid electrolyte solution.

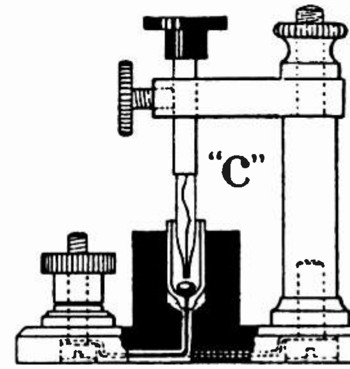


Fig. 5. Diagram of the original British Insulated Wire Company's barretter showing the thumb screw used to wind the anode into the electrolyte. The cathode was a mercury pool.

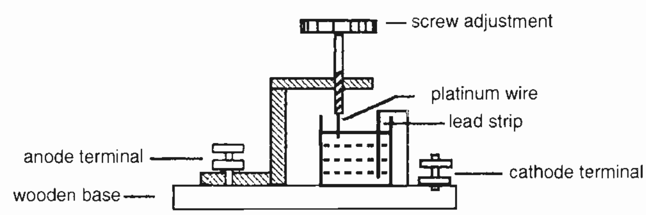


Fig. 6. Experimenter's barretter from around 1905, based on Fessenden's original design. In this case, the electrolyte was 20% sulphuric acid and the cathode a lead strip.

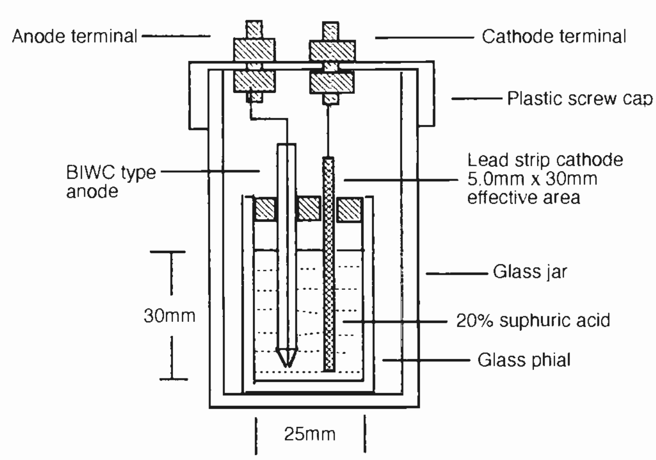


Fig. 7. Hybrid barretter involving an anode similar to that used in the British Insulated Wire Company's design, Fig. 3.

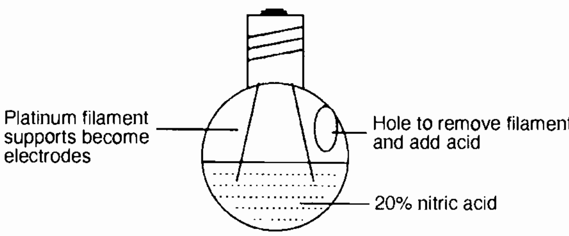


Fig. 8. Delaney lamp detector was unusual in that it had electrodes equal in size. These electrodes were simply the filament supports of a small bulb.

that responded to individual waves within a train. As it did not need to be restored after each operation, it became known as a 'self-restoring' detector.

While the relay analogy of a barretter was found to be faulty, the result was indeed pulses but with a repetition rate beyond the capability of a morse register. As a result, it was possible to hear and read signals via headphones.

Being a two-terminal device, isolating the input and output circuits of a barretter presented problems similar to those with the coherer. Circuit configuration had to avoid shunting the barretter with a dc path via the tuning coil; this was usually achieved by series tuning, as in Fig. 1.

Parallel tuning in conjunction with a dc blocking capacitor was occasionally adopted however. Many early experimenters simply connected the barretter directly to the antenna in the so-called 'untuned' mode originally used with the coherer.

While the coherer was best suited to trains containing very few waves, the barretter, being a rectifier type detector, was suited to continuous wave systems where tuners could be used to give a high degree of syntony. At that time however, wave-train transmissions prevailed. Although the barretter itself was well suited to these transmissions, *LC* tuners were not. For this reason a high degree of syntony could not be achieved.

Electrolysis

Electrolysis is fundamental to the operation of the barretter, but as as the controversy over cold fusion shows, it is still not fully understood. Platinum, which seems to be an essential component of the barretter, is closely related to palladium. So it is not surprising that my experiments raised many questions that remain to be answered.

A typical a barretter comprised a small cell containing 5 to 10ml electrolyte, either 20% nitric or sulphuric acid and a pair of electrodes. It required an applied dc, typically 1.9 to 2.5V, and typically derived from two 1.5V cells via a 500Ω potentiometer.

The barretter generates a potential of about 1.7V that opposes the applied dc. As a result, forward current remains virtually zero until applied dc reaches about 1.7V. Here, forward current increases linearly with applied dc.

This phenomenon was known to the pioneers. They perceived that if applied dc was set at a level slightly below 1.7V, positive-going rf half cycles would add to the applied dc. In turn, this would cause current pulses to flow through the circuit in sympathy with the rf half cycles.

Typical barretters

An essential feature of barretter is that one electrode presents a minute, almost a microscopic surface area to the electrolyte; this was generally the anode but some early literature shows it as the cathode.

Although my replicas would work when the minute electrode was the cathode, efficiency was far inferior. So for the purpose of this study, the minute electrode is the anode. The polarity of the electrodes is of course set by the polarity of the applied dc, Fig. 1b).

The term barretter applies only to devices based on Fessenden's design, ie devices requiring applied dc and having one minute electrode. A large number of pioneers experimented with electrolytic detectors and numerous variations evolved.

Fessenden detector construction

I made a simple Fessenden detector using a small cell with an anode comprising a short length of very fine platinum wire. This wire was attached to the end of a screw so that it could be adjusted to just 'tip' the 20% sulphuric acid electrolyte. The cathode was a small platinum strip. With my replica, I found adjustment critical. It was also upset by vibration, which presumably caused slight variations in the level of the electrolyte, **Fig. 4**.

Incidentally, early literature often referred to platinum wire as Wollaston wire; presumably after Wollaston who developed and patented a technique for producing platinum, thus making this metal more readily available and dramatically reducing its price.

Insulated Wire's design

The British Insulated Wire Company, BIWC, design had a cell with a capacity of only about 2.0ml; the electrolyte was 20% nitric acid. The anode consisted of a short length of platinum wire, said to be only 0.004mm diameter, fused into the end a glass tube; the end of the wire was ground flush with the tube to expose only the wire's cross section. The cathode was a pool of mercury, **Fig. 5**.

For my replica, Fig. 3, I used 0.01mm diameter platinum wire for the anode, as this was the finest I could obtain. As you will see this difference in wire gauge may well account for the better high frequency response of commercial barretters.

Other designs

Most early experimental designs were based on Fessenden's original screw-thread anode. Cathodes varied from a tiny platinum plate, platinum wire, a pool of mercury to lead strips. The electrolyte was either 20% nitric or sulphuric acid, **Fig. 6**.

The hybrid design. I found the BIWC-type anode convenient as the electrode was simply immersed in the electrolyte to the optimum depth. No further adjustment was needed. Moreover, I found the hybrid design of **Fig. 7**, which incorporated a BIWC anode and a lead strip cathode, to be most practical. Depth to which the lead strip cathode was immersed in the electrolyte could easily be adjusted, thereby changing its effective surface area.

The hybrid barretter cell had a capacity of about 20ml and contained about 10ml of 20% sulphuric acid; it was housed inside a screw-cap jar to prevent spills. Although slightly less efficient than the BIWC design it was reliable and gave consistent results.

Delaney lamp detector. Unlike the barretter, the Delaney-lamp electrolytic detector, so named because it was constructed from a small electric lamp, had electrodes of equal size, **Fig. 8**. Experimenters used a blow torch to melt a small hole in the glass envelope and remove the filament; the bulb was then partly filled with 20% nitric acid. The filament support leads formed the electrodes; these were platinum as this was one of the few materials available at that time that could be fused into the glass.

The Delaney lamp detector was reported to be less sensitive than the barretter but tolerant to very strong transmissions which could damage the barretter's minute electrode; this gives a good indication of the enormous energy present in wave train radiated by a spark transmitter.

Voltaic-cell designs. The Shoemaker detector apparently had a minute platinum anode, but its cathode was a small zinc plate. The electrolyte was dilute sulphuric acid; it was in effect a tiny voltaic cell so did not require applied dc.

Other voltaic-cell designs apparently used platinum in conjunction with carbon, zinc or copper electrodes while some designs used copper-sulphate solution as the electrolyte. ■

In a second article, George will be discussing behaviour of barretter in more detail.

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

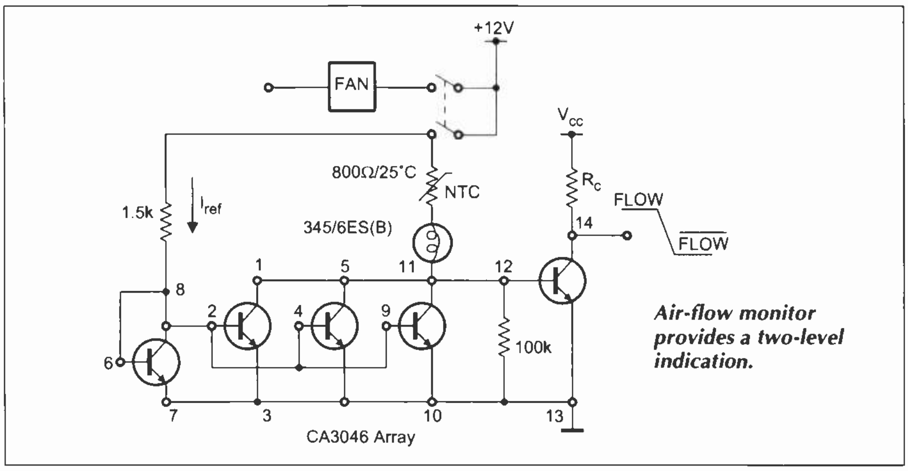
Do you have an original circuit idea for publication? We are giving £100 cash for the month's top design. Other authors will receive £25 cash for each circuit idea published. We are looking for ingenuity in the use of modern components.

Winner of the first spectrum analyser goes to the circuit: One chip air-flow monitor

An 800Ω thermistor with a lamp in series provides combined negative and positive temperature coefficient, and can accept voltage excitation; the positive coefficient of the lamp filament prevents thermal runaway, but allows sensitivity to heat dissipation in the air stream.

With normal flow, the thermistor possesses high resistance and passes a low current to node 11 of the *3046* transistor array. The triple current mirror therefore turns off the output transistor. If air-flow drops, the temperature rises, reference current and current through the monitor increase and the output transistor conducts and saturates.

The supply voltage and load resistor R_c should be chosen to provide the required output levels; limits for the *3046* are 15V and 10mA collector current. Trim reference current to take account of varied ambient temperature.



Air-flow monitor provides a two-level indication.

John A Haase
Fort Collins
Colorado
USA

Capacitive continuity tester helps locate cable breaks

£100 WINNER

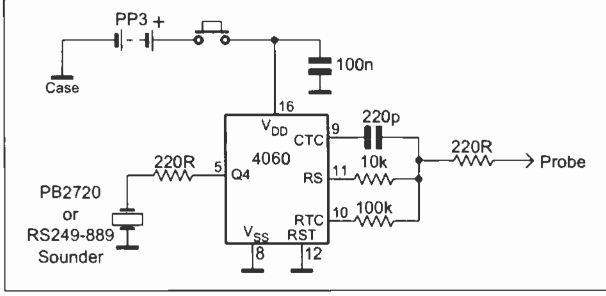
It is common knowledge that breaks in multi-way cables with moulded connectors always occur at the other end to the one you cut off. This device uses a single probe to test continuity and will indicate at which end the break is.

A *4060* oscillator/counter runs its oscillator at 20kHz, additional resistance or reactance connected to the junction of the 100kΩ and 220pF tuning components changing the frequency. An output from Q_3, Q_4 or Q_5, depending on personal preference, drives a piezoelectric sounder to give the indication and the circuit is housed in an unpainted metal case.

Hold the cable in one hand, the tester in the other and touch the rigid probe to each pin in turn; if the pin is connected, stray capacitance decreases the tone frequency. Do the same at the other end of the cable. Breaks are at the end where the tone frequency is least affected by strays. Carry out pin-to-pin checks by touching one pin with the probe pin and the other with a finger; the oscillator stops if there is continuity.

Motors and large inductors can be compared to indicate shorted turns, but this will need a little practice.

Robert Atkinson
Christchurch
Dorset



Using one probe, this tester indicates at which end of a multi-wire cable a break has occurred.

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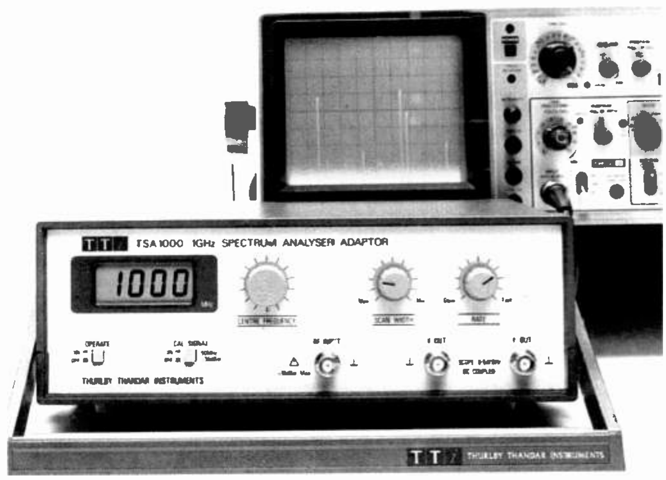
Our judging criteria are ingenuity and originality in the use of modern components with simplicity particularly valued.

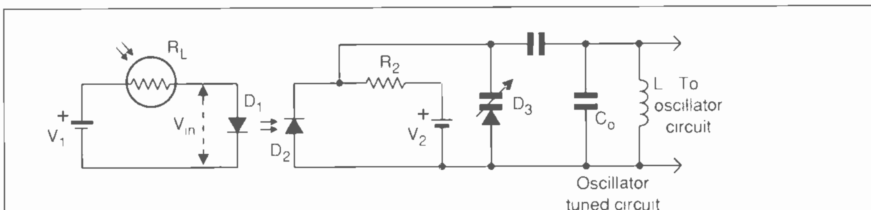


Light-control for oscillators

Varying the amount of light falling on a light-dependent resistor varies the frequency of an oscillator.

Light on R_L, the ldr, determines the light output of the opto-coupler led and therefore the output of its photodiode. Reverse bias on the variable-capacitance diode forming part of the oscillator tuned circuit is now dependent on the ldr illumination, which sets the frequency of oscillation.

K N Sunil Kumar
Visakhapatnam
India



Amount of light falling on the light-dependent resistor R_L determines oscillator frequency,

HF converter for car radios

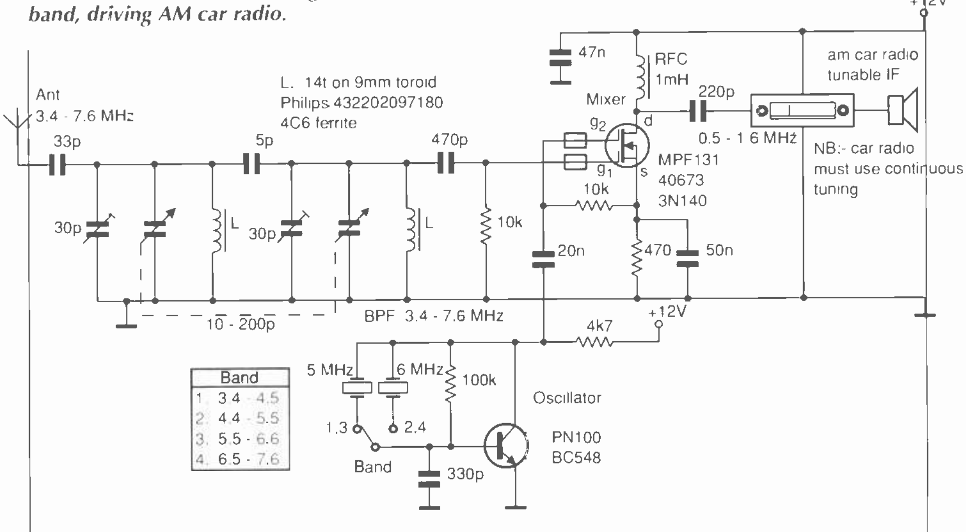
A simple frequency converter in front of an AM car radio offers a convenient means of monitoring hf signals. Frequency stability and easy tuning are among the advantages over more exotic circuitry and the selective front end contributes to good image rejection.

The converter shown is conventional: it is crystal-controlled, each crystal giving two tuning ranges, although one of them is reversed. Two crystals therefore give four 1MHz ranges. 5MHz and 6MHz crystals covering 3.4-7.6MHz continuously. ssb/cw signals can be received if a 455kHz bfo is used.

Inductors *L* are wound on ferrite toroids. House the converter in a metal box to avoid picking up local broadcast signals.

Peter Parker
Bentley
Australia

Tuned front end converts hf signals to broadcast band, driving AM car radio.



Magnetic field detector

External magnetic fields affect the permeability of a magnetised ferrite rod far more than that of a non-magnetised one. Making the ferrite tune an oscillator produces a beat frequency with the output of a similar reference oscillator.

These rods are 8 by 110mm and wound with 110 turns of 0.3mm enamelled copper wire and have a little magnet glued to one end, with reversed polarity one to the other. You may find similar rods and coils in a salvaged mw receiver.

Reference and scanning probes are identical, mechanically and electrically, to confer invulnerability to temperature and voltage variation. Both oscillators tune to 1.25MHz, a small magnet being attached to the outside of the scanning probe case, to vary the beat frequency when it is turned.

Though screening is vital, a metal cae stops oscillation, so a plastic case is used, with metal strips or wires parallel to the rod and earthed at one end. Other circuitry is in one box, with the halves screened from each other to avoid locking one oscillator to the other.

Zero beat is obtainable with the fine frequency control and a very low beat frequency is detected by using the lf oscillator to interfere with the beat and produce an audible sound.

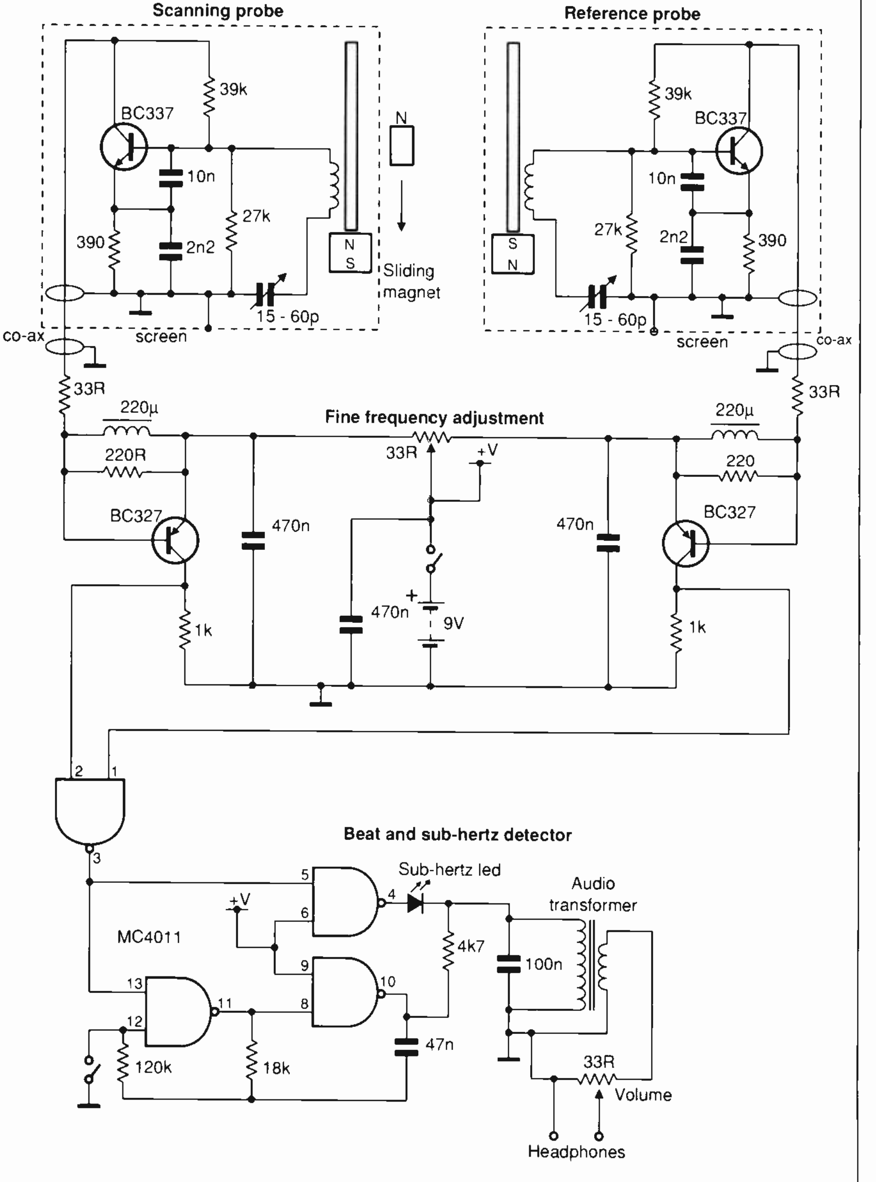
Turning the scanning probe through 180°in the Earth's field produces a 1-12kHz beat frequency change.

It is possible to detect the beat clearly at, for example, 20cm from the front of a tv set, 1m from its side, 25cm from an oven and 60m from a 300kV power line.

D Di Mario
Milan
Italy

Magnetic stray field of appliance and other sources and distance for clearly detectable modulation.

14in tv - front	20cm
14in tv - side	1m
300W halogen lamp	10cm
25W soldering iron	15cm
Kitchen oven	25cm
1100W iron	20cm
Toaster	15cm
Fluorescent light	22cm
Compass needle	5cm
Car – from bonnet	60cm
380kV Power line	60m



Stray magnetic field detector, using magnetised ferrite-cored coils as scanner and reference probes. Indication is beat note in headphones.

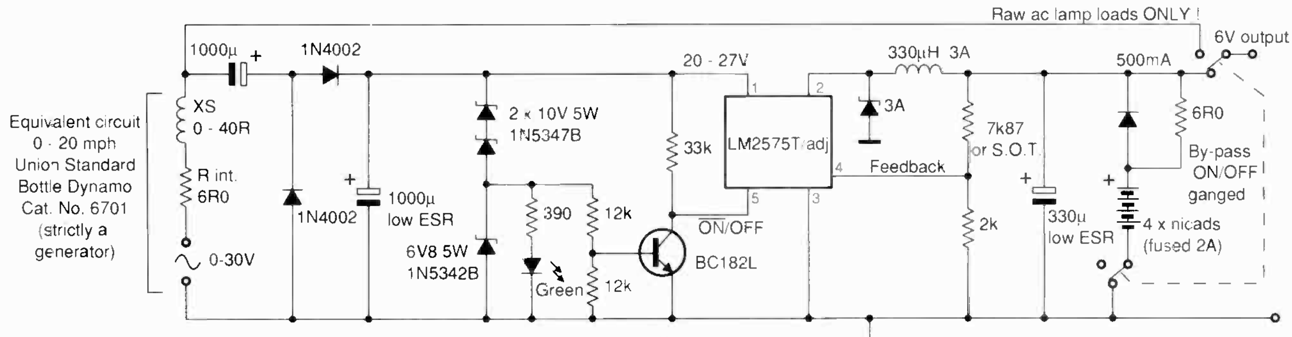
Bike battery charger

This cycle battery charger using the standard lighting-set dynamo or generator has been in use for three years with no problems.

A voltage-doubler, driven by the generator, provides 3.5W of cycle-frame-referred dc – slightly higher in voltage than when driving lamp loads. Three zeners and Tr_1 provide a gentle start-up, protect the circuit when the cycle is travelling fast and when there is no load and the led indicates that normal operating voltage exists. A National Semiconductor *LM2575-ADJ* set to 5V by means of the feedback, converts the rectified voltage to 6V.

Four NiCd batteries in a solder-tagged pack use any power offered by the generator at any speed from walking pace to about 14mile/h, the stabilised voltage having prolonged the life of the 2.4W halogen bulbs. The whole thing fits into a drink bottle.

P W Fry, G4SBF
Southampton



Battery-backed dynamo cycle lighting works from walking speed to about 15mile/h.

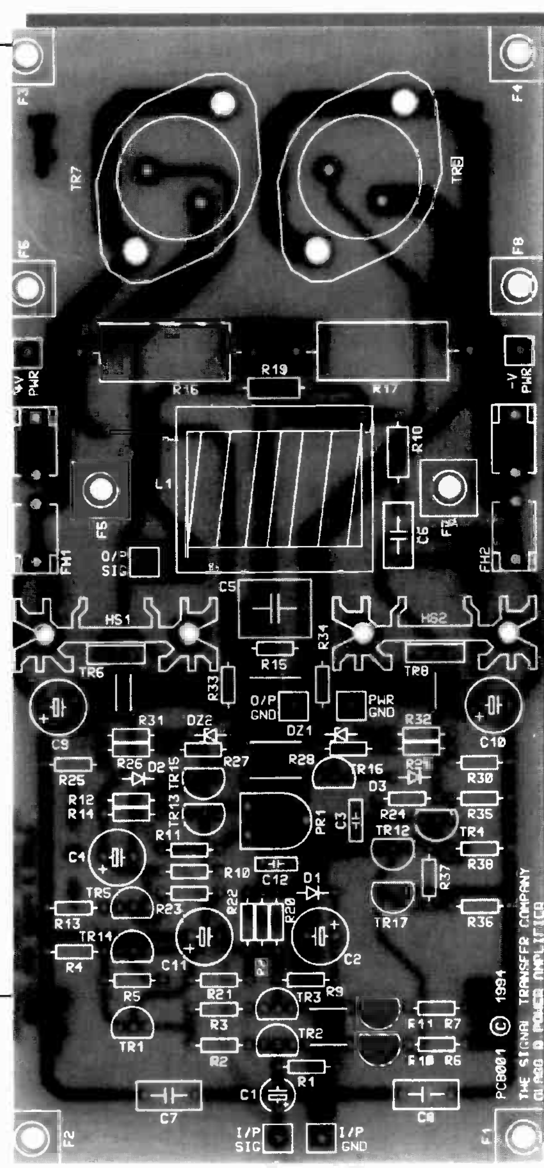
PCBs for Douglas Self's power amplifier series

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 of 0.0015% at 50W and follows a new design methodology.

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, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.



Wide-range, low-power, programmable timer

Drawing only 10µA, this timer can be programmed for delays from 8ms to over a year, to an accuracy determined by the crystal.

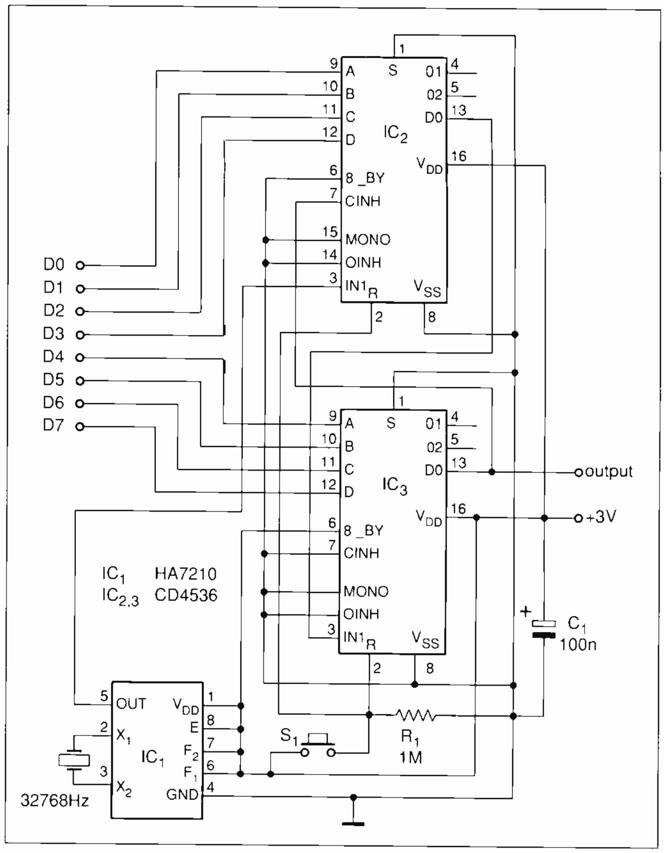
A low-current (4µA) *HA7210* oscillator drives a pair of *CD4536* 24-stage binary ripple counters, whose a-to-d converter and 8-bypass pins select which stage is used for output. If 8-BY is low, the first eight stages are bypassed; if high, one of the first 16 stages feeds the output, determined by the a-to-d converter inputs.

In the diagram, 8-BY of IC_2 is low and that of IC_1 high, which causes the the pair together to form a 24-stage counter with an eight-stage prescaler, in which $D_{0\text{-}7}$ select the output stage.

When the selected output goes high, it inhibits the IC_1 clock, so that the output remains high until S_1 resets the counters.

Since the minimum number of stages selected is eight and the maximum forty, the range of delays is 7.8125ms with $D_{0\text{-}7}$ all low to 33554432 seconds with $D_{0\text{-}7}$ high.

Yongping Xia
Torrance
California
USA



For a current expenditure of 10µA, delay in this programmable timer ranges from milliseconds to more than a year.

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Reverse engineering FOR GRAPHS

Designers frequently use computers to produce graphs, but there are times when it would be useful to have the computer read in data plots from curves on hard copy. Allen Brown looks at software to do just that.

Graphs on paper are excellent tools for conveying trends, but copying a graph's numeric information into a computer in order to make use of it in software tools can be extremely tedious.

Given a calibration curve of a sensor, for example, it can be handy to have the information not as a plot, but as an array of numbers. Digitising plots or graphs using a digitising tablet is one way of achieving this. But there is now another alternative, and that is to use a pc software package running under DOS called *UnGraph*

Published by Biosoft of Cambridge, *UnGraph* can be used manually by clicking the mouse on the plot points. Alternatively you can use the software's automatic line tracer which follows the plot trace and generates the [x,y] coordinates as it progresses – an action referred to as vectorising.

There is sufficient intelligence in *UnGraph* to correct for possible skew in the image. When in automatic mode, the software is usually able to distinguish between the plot curve and its grid lines.

To make effective use of *UnGraph* you will need some means of obtaining an image of the graph that you wish to digitise. This normally needs a scanner – whether hand held or flat-bed – interfaced to your pc. The scanner must be able to produce either .tif or .pcx format line-art image files. Although *UnGraph* has the ability of directly controlling a *ScanMan* hand-held scanner, most scanners these days conform to the Twain standard, which is Windows based. *UnGraph* is only available for dos, and cannot use the Twain driver. Therefore, when run via Windows, the image must be obtained separately before *UnGraph* is evoked as a whole screen dos program.

When evoked, *UnGraph* recognises the graphics standard of the pc and provides the user with a menu bar of options. These include 'Convert' which is required to convert the .tif or .pcx image file into a .scn image file for *Ungraph* to process.

When selecting the vertorisation option, the image appears on the screen, complete with skew. If the image file is larger that 50Kbyte, only a fragment of the graph will be visible at any one time. Resolution of the image appears poor, but seems sufficient. The user enters three coordinate positions in order to define the scaling and skew of the graph.

Graph-tracing options

There are five tracing options available to the user – Manual, Manual with Grid Elimination, Automatic, Automatic with Grid Elimination and Digitise Individual Points. From the manual it is not at all apparent how the Grid Elimination process is performed, and there is no indication in the vectorising mode how this can be performed.

Graphs with dashed-line plots need filling in using *UnGraphs*'s Paint facility before the Automatic Tracing mode can be used.

Automatic-trace mode is an attractive feature of the package, but it does need a degree of user interaction. Sometimes, as it is following a curve passing through a grid line, the Auto Trace function follows the grid line instead. It is relatively easy to rectify this by back tracing and manually digitise a few points past the difficult part of the curve. However it does sometimes fail completely to progress and you have to start again.

It may be necessary to use the Erase facility, via the mouse, to remove parts of the plot which are causing problems. Auto Trace does not perform well on curves with very sharp peaks, such as those found in a spectra plot. The user has to manually digitise the process up and over the peaks.

When in the Auto mode the extrapolation between the digitised points defaults to a linear algorithm. But the user has the choice of a quadratic algorithm which is more appropriate for a sharply changing curve.

When the digitising is complete, the user is asked to confirm the start and end of the x coordinates, and the required separation between the samples. This results in the highlighted data points of collector current as a function of irradiance for a photo-transistor, as illustrated in the screen shot.

When storing the vector as x and y coordinates, the allowed precision is only two decimal places, and scaling may be necessary. If a graph contains several curves, the process can be repeated for each one.

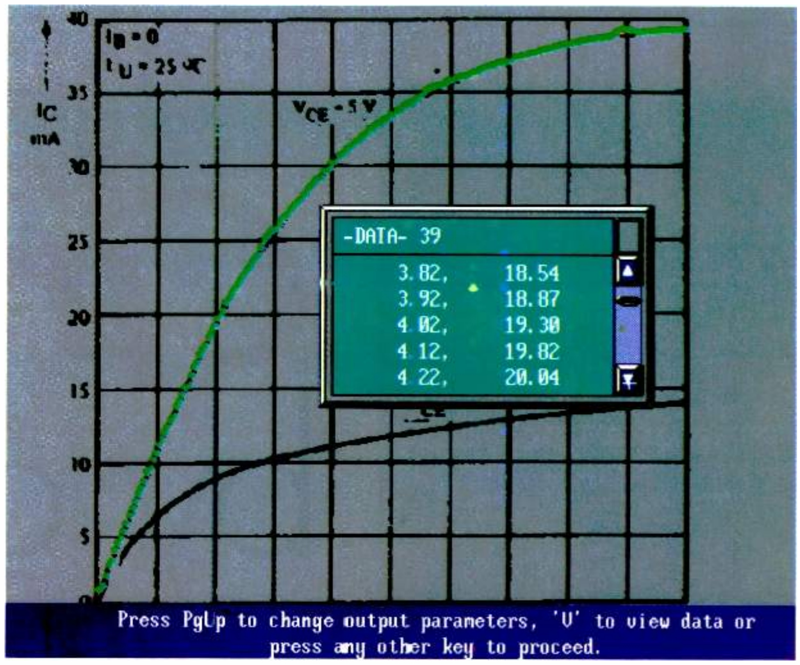
Conclusion

The fact that UnGraph is able to digitise graphical data automatically with comparative ease makes it an attractive tool. It would be preferable to have a Windows version to make its integration into other processes easier. Nevertheless it can be used effectively once its limitations are recognised.

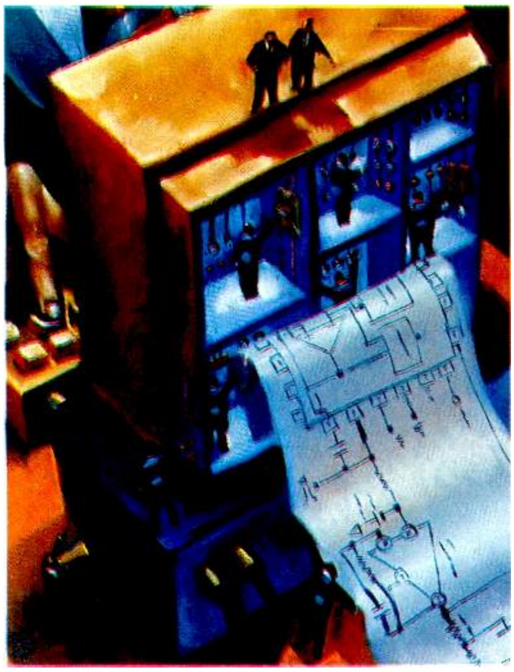
The software would benefit from having some of its rough edges ironed out – in particular the problem relating to the Auto Trace getting lost and remaining lost. On the whole *UnGraph* should prove to very useful to you, especially if you want to use someone else's data that is only available in a graphical format. ■

Software source

Biosoft, 49 Bateman Street, Cambridge CB2 1LR. Tel. 01223 68622, fax 01223 312873. Price £199 fully inclusive.



UnGraph pulling data from a previously scanned graph. Resolution of the scanned image appears poor but is sufficient for the task in hand.



Circuits by design

2: dc modelling

Using a common-emitter amplifier as an example, Owen Bishop illustrates how modern circuit simulation software helps to demonstrate dc circuit performance under various quiescent conditions.

Examining the behaviour of a common-emitter amplifier, **Fig. 1**, under various quiescent conditions gives us the chance to become more familiar with some of the basic procedures available in *SpiceAge for Windows.*

The circuit is minimal, with a constant voltage source providing base bias. Later this will be replaced by resistors, but for the moment, using a voltage source allows the correct biassing conditions to be found more readily.

As usual, the first step in the simulation is to enter the editing window and key in the netlist, **Fig. 2**. The transistor specified by the entry 'qnpn.lib' is a general-purpose npn transistor model, the details of which are stored in a library file. These details are called upon by the program when a simulation is run. I will discuss the details of this model later.

Power supply is represented by the voltage generator, appropriately called V_{cc}. This constant 12V dc source is specified as having no excitation, that is, no superimposed waveform, and an offset of 12V.

Voltage V_{bb} is the input level, which we want to vary from zero up to a suitable level, in order to study how the circuit behaves. A convenient way of doing this is to specify V_{bb} as a voltage generator with a ramp waveform.

The value of such a waveform is the amount, in volts, by which it changes during a second. A look at the netlist shows that we have specified a ramp of $1Vs^{-1}$.

A suitable range of input voltages is obtained by letting V_{bb} ramp for 2s. This is achieved by going to the Sweep Time dialogue box, from the Time menu, setting the start time to 0, the stop time to 2s, and sampling every 5ms, **Fig. 3**. This gives 400 samples of V_{bb} and the node voltages and currents during a sweep.

To begin with, the probes are set to read the input voltage at node V_{bb}, and the output voltage at node C. Selecting Analyse, then Transient produces **Fig. 4**. The scale on the *x*-axis is the time in seconds and, knowing that the lower curve shows V_{bb} ramping up at $1Vs^{-1}$, we read this scale as the voltage V_{bb}.

The other curve shows the collector voltage initially at 12V, then starting to fall as V_{bb} reaches about 0.5V. It has fallen almost to zero – saturation – by the time V_{bb} is around 0.8V. The cross-hairs in Fig. 4 are aimed at the point on the collector voltage curve where it passes down through 6V, which is half the supply voltage. This is a key point for distortion-free operation of the circuit.

At the bottom of the screen, the time indication can be interpreted as representing a value for V_{bb} of 0.6856V at this point. As a check on the simulation, set up the circuit on a breadboard, using a *BC548* transistor. Voltage V_{bb} is provided from a 6V battery with a potentiometer connected across it, tapped at the wiper to produce a variable voltage.

Figures obtained from the analysis are matched by those found from measurements on the circuit. For example, at the operating point, when output voltage is exactly 6V, a meter shows that V_{bb}=0.683V. Allowing for tolerance, this is very good agreement.

Specifying generators

There are two ways of specifying generators, the most straightforward being to key the required characteristics directly into the netlist. Alternatively, first list the generator and its connections in the netlist, then select Time, then Generators to obtain the Signal sources and excitation control dialogue box. A list of generators is displayed there.

To set the V_{bb} generator, for example, highlight the V_{bb} generator by clicking on it in the Generator panel; click the Ramp option; key '1' in the Value box. On returning to the netlist, we find that the option and its value have been added to the netlist.

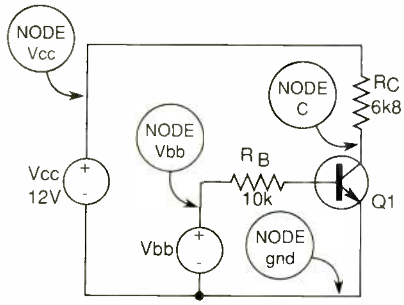


Fig. 1. Simple simulation of a common-emitter amplifier is useful for illustrating how altering the circuit parameters affects dc conditions.

```
* CBD02 – Common-emitter amplifier

V Vcc   –output:gnd+output:vccEx=NoneOf=12.00000
V Vbb   –output:gnd+output:vbb    v=1.000000    Ex=Ramp
R Rc    p1:vcc       p2:c         v=6.80000k
R Rb    p1:vbb       p2:b         v=10.0000k
> Q1    qnpn.lib     collector:c  base:b        emitter:gnd
```

Fig. 2. In the netlist for the common-emitter amplifier of Fig. 1, the transistor qnpn is a general-purpose type specified from within a library.

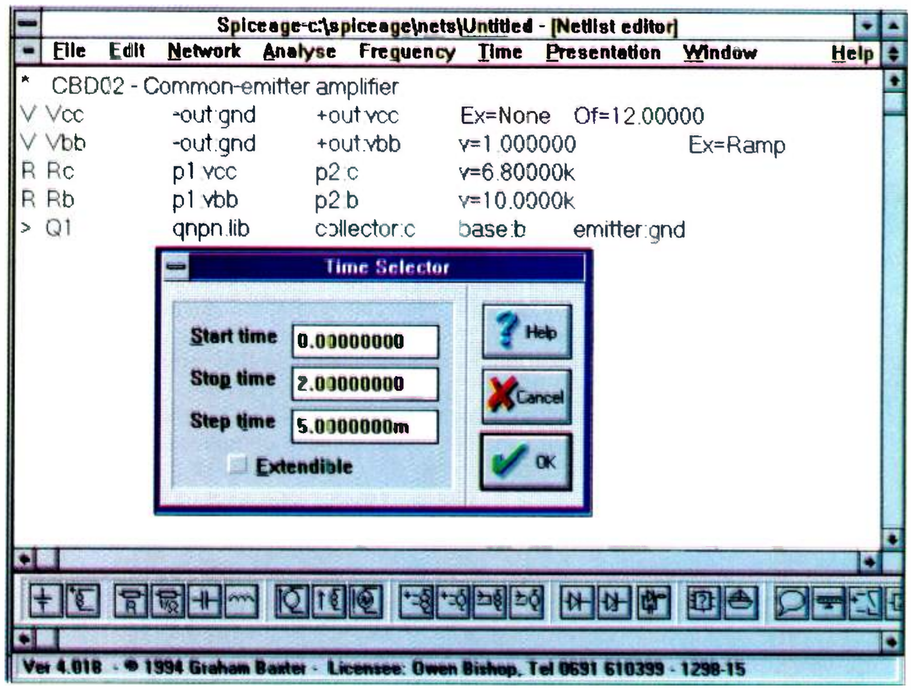


Fig. 3. Defining a sweep. Having 5ms steps in a 2s sweep provides 400 readings of the various node voltages as base voltage gradually rises.

Collector current

Figure 9 plots input and output voltage, as before, and also the current through R_c, with the 'qnpn.lib' model installed. In order to give all three curves sufficient amplitude suitable scale factors are set on the probe control panel **Fig. 10**. With these settings output voltage is plotted on the default (×1) scale, V_{bb} is multiplied by 5, and the current by 5000.

Figure 9 shows current beginning to flow as V_{bb} reaches about 0.5V, rising linearly and saturating at about 8.5/5000=1.7mA. When output is 6V, current is 4.2/5000=0.88mA, a suitable value for such an amplifier. With V_{bb} at 0.7V, output voltage is 6.5V.

These values are found to be true in the breadboarded circuit, too, using a *BC548*. When we substitute a high-gain *BC550* on the breadboard, output voltage falls to only 0.17V when V_{bb}=0.77V. This result is obtained in the simulated circuit if we substitute the *BC109C* model.

To establish quiescent conditions in the circuit when output is 6V, first find the exact value of V_{bb} at this point. Measurements are taken directly from the graph, placing the cross-hairs on selected points and reading the coordinates in the panel below the graph. The voltage graph shows that V_{bb} is 0.6856V when output is exactly 6V.

Next edit the netlist to make V_{bb} a constant dc source at this voltage by changing its excitation to 'none' and its offset to 0.6856V. Then select Analyse followed by DC Quiescent. The quantity analysed, whether voltage, current or power, depends on the selection for Probe 1. Since this probe is already set to measure current through R_c, we obtain a table giving currents through all components.

Current through R_b is 8.88µA and that through R_c is 879µA – the same result as obtained above by reading from Fig. 9 – resulting in a current gain of 99, as might be expected. You could also perform a quiescent analysis for the voltage at each node, or for the power dissipated in each component.

Using Mathematica

It is possible to analyse a resistive network by giving the *Mathematica* portion of *SpiceAge* eight simultaneous equations to solve – a task well within its capabilities. The same approach

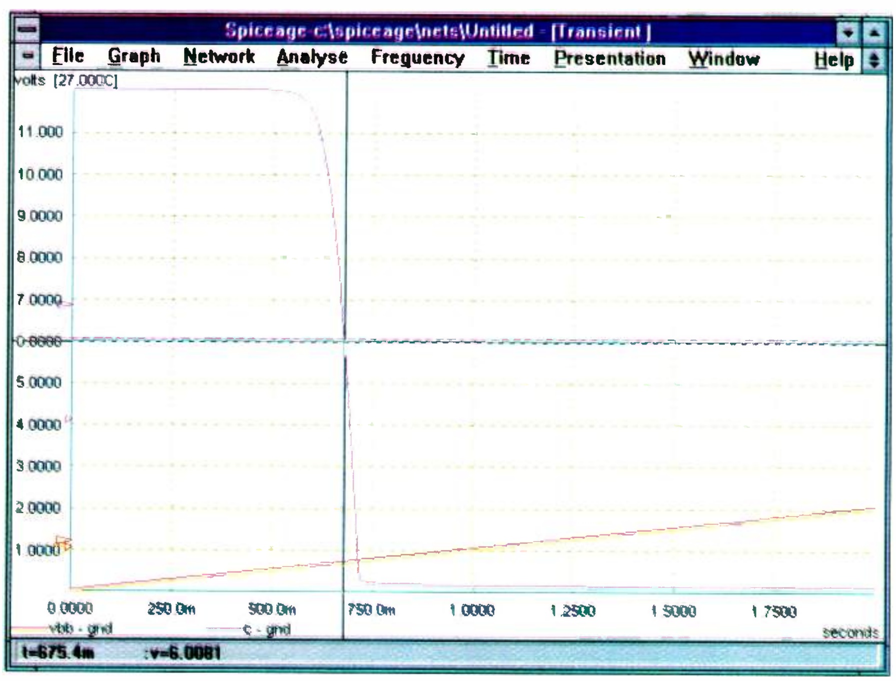


Fig. 4. Analysing transient response of the common-emitter circuit. Since base voltage ramps up proportionally with time, the horizontal scale can also be read as V_{bb}. Collector voltage is on the vertical scale.

Transistor models

In the last line of netlist Fig. 2, the transistor is specified by quoting a file name. The library file of this name is itself a netlist of a sub-circuit which is the equivalent of a transistor.

Selecting Network, then Exploding reveals the netlist. The last statement of the netlist is immediately expanded, Fig. 5, to the full transistor netlist. Fig. 6 shows the equivalent circuit. It consists of two diodes connected back-to-back to simulate the base-collector and base-emitter junctions.

Transistor action is simulated by the voltage-controlled current source, Q1.gm, specified by letter *G* in Spice (also by the code VCVS in SpiceAge). Its control terminals – not shown in Fig. 6 – are 'connected' to nodes 'Q1.ei' and 'gnd'. The effect of this is that it is controlled by the voltage v_r across the emitter resistor. This resistor is included in the model simply to generate the controlling voltage.

The value of the source is 990m, which means that the current (actually, the collector current i_c) in amps is 0.990 times the voltage v_r.

A pair of equations can be written:

$v_r = i_c/0.99$,

from the voltage-current relationship of the VCCS, and,

$$v_r = r(i_c+i_b) = 1\times(i_c+i_b) = (i_c+i_b),$$

incorporating Ohm's Law, and summation of currents.

Substituting for v_r:

$i_c/0.99 = (i_c+i_b)$

$\Rightarrow i_c = 0.99 i_c + 0.99 i_b$

$\Rightarrow i_c = 0.99 i_b/0.01 = 99 i_b$

The result is a device with a current gain of 99. In general, if the value of the voltage-controlled current source is ς (always <1), the current gain is ς/(1−ς).

Instead of this simple model, we can use a more complicated one, based on the characteristics of a particular type of transistor. *SpiceAge* has a comprehensive transistor library, including many of the Zetex transistors.

In Fig. 7, 'qbc109c.lib' is substituted for 'qnpn.lib' and then exploded. The sub-circuit is similar to the previous one but has a base resistor and a second generator, a voltage-controlled voltage source, Fig. 8.

Although the positive terminal of voltage-controlled voltage source appears

```
* CBD02 – Common-emitter amplifier

V Vcc    –output:gnd+output:vccEx=NoneOf=12.00000
V Vbb    –output:gnd+output:vbb       v=1.000000  Ex=Ramp
R Rc     p1:vcc     p2:c      v=6.80000k
R Rb     p1:vbb     p2:b      v=10.0000k
Di       Q1.be      anode:b   cathode:Q1.el
R Q1.e   p1:Q1.el   p2:gnd    v=1.000000
Di       Q1.bc      anode:b   cathode:c
GQ1.gm   –out:c     +out:b    +con:Q1.el     –con:gnd v=990.000m
```

Fig. 5. Exploding results in an expanded netlist with the transistor fully described.

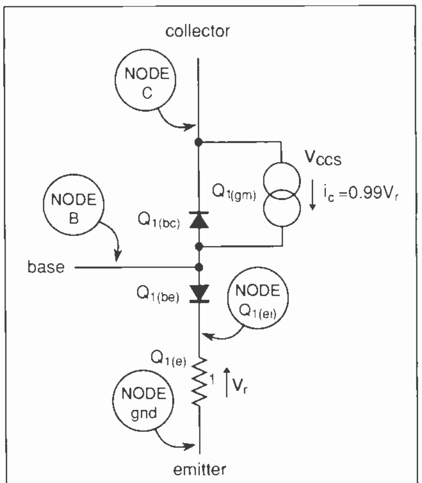


Fig. 6. Equivalent circuit of the general-purpose transistor described in the netlist of Fig. 5 Back-to-back diodes simulate the junctions.

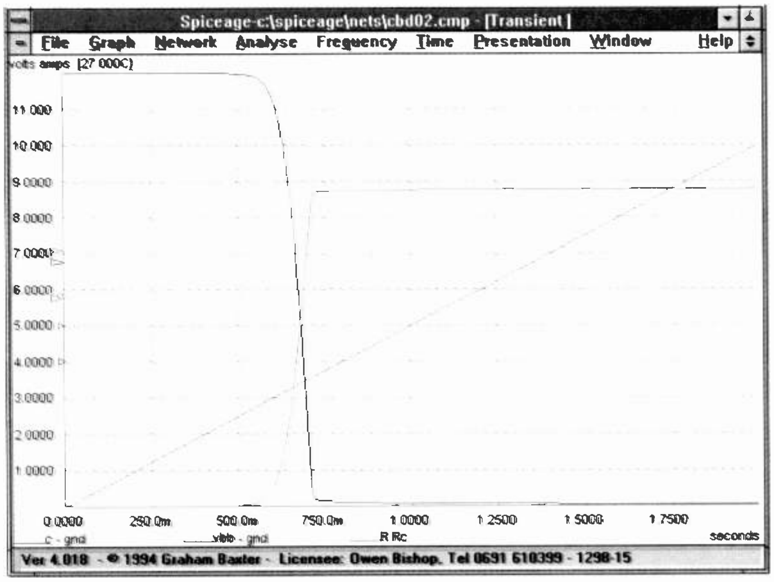


Fig. 9. Input and output voltage of the common-emitter amplifier, together with current through the transistor collector resistor.

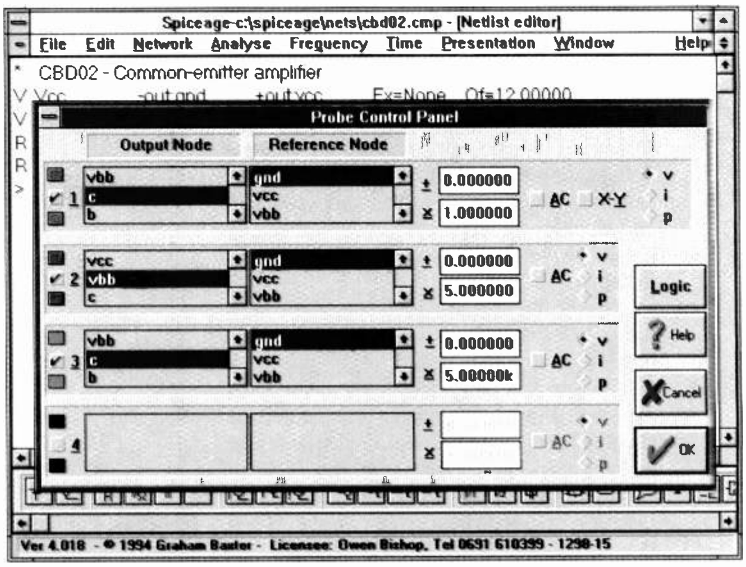


Fig. 10. Setting scaling factors to make sure that the curves of Fig. 9 provide useful amplitude displays.

unconnected, examination of the netlist shows that node Q1.ee is connected to the positive control input of the current source. The voltage source is controlled by the voltage across the emitter resistor and, in turn, the voltage-controlled current source is controlled by the voltage across the voltage source.

The voltage source reproduces the voltage across the resistor, but with a delay of 1ns. The effect of this arrangement is that the current source is controlled as before, but with a delay of 1ns. This simulates the base transit time. Together with the 6pF capacitor this makes the model sensitive to the effects of frequency, particularly the Miller effect at high frequencies.

The current-source value is 0.9976, resulting in a current gain of 0.9976/0.0024, which is 416. As with the other models, the characteristics can be edited. This makes it possible to match the model more precisely to details of a particular type as quoted in the manufacturer's data sheet.

One of the disadvantages of using a more complicated model is that it takes longer for the analysis to be completed.

```
* CBD2 – Common-emitter amplifier

V Vcc      –output:gnd+output:vccEx=NoneOf=12.00000
V Vbb      –output:gnd+output Vbb   v=1.000000          Ex=Ramp
R Rc       p1:vcc        p2:c        v=6.80000k
R Rb       p1:vbb        p2:b        v=10.0000k
* Q1.BC109C hfe=420 Ccb=6p Ft=150M
* Q1.Does not model storage time, delay time, nor inversion mode
* Q1.The Ft modelling holds up to a frequency of approx. twice Ft
* Q1.
Di         Q1.be         anode:Q1.intbase       cathode:Q1.ei
R Q1.e     p1:Q1.ei      p2:gnd      v=1.000000
Di         Q1.bc         anode:Q1.intbase       cathode:c
VCCS       Q1.gm         –out:c      +out:Q1.intbase
   +con:Q1 ee            –con:gnd    v=997.600m
R Q1.bb    p1:Q1.intbase             p2:b       v=50.00000
C Q1.miller p1:c         p2:Q1.intbasev=6.00000p
VCVS       Q1.transit    –out:gnd    +out:Q1.ee
   +con:Q1 ei            –con:gnd    v=1.000000          De=1.00000n
```

Fig. 7. This netlist is similar to the one used to produce Fig. 6 but the general-purpose transistor qnpn is replaced by a BC109.

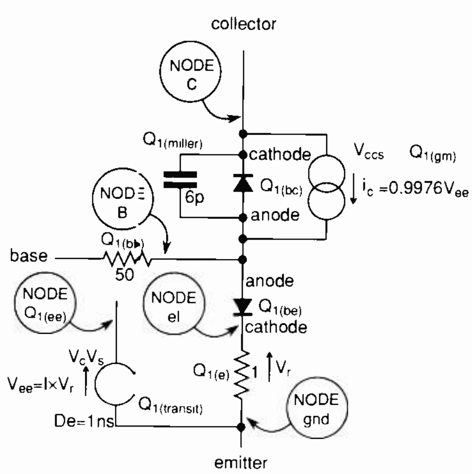


Fig. 8. Equivalent circuit for the BC109 transistor of netlist Fig. 7, showing the additional base resistor and voltage source.

Means

The time-averaged means of plotted quantities are indicated on the y-axis of the graphs by triangular markers. The longer markers indicate root mean square values. The shorter markers indicate arithmetic means. The colours of the markers are same as those in which the curves are plotted. In Fig. 9, reading from the top downward, the rms markers refer to current, output voltage and V_{bb}; the mean markers refer to current, V_{bb} and output voltage.

can be applied to find V_{bb}, writing out all possible equations for the circuit and asking for them to be solved.

In the case of Fig. 1, the equation for base-emitter voltage V_{be} is exponential and the program is unable to produce a solution. It needs to be helped along by having the equations presented to it in the order that will allow it to proceed step-by-step to the solution. With such small steps there is little point in using *Mathematica*. A scientific calculator will do the job just as well. However, I will take this opportunity to look more closely at the syntax and some other features of the program.

To start with, take a very simple example, the calculation of the collector current when the output of the circuit – i.e. voltage at the collector – is exactly half the supply voltage. Given R_c and V_{cc}, the collector current in milliamps is:

$$cc = \frac{V_{cc}}{2R_c} \times 1000$$

Under section (1) in **Fig. 11**, I show how to incorporate this equation into a procedure. The function *cc* is a function of two variables, *rc* and *vcc*. Note that I have not used capital letters because these are reserved for the program's own objects.

The expression following ':=' defines what the function is to do when called. The expression is simply the right-hand side of the equation above, using *Mathematica* syntax. Remember that a space between two terms indicates multiplication. Having typed in this function, or having loaded it from a file, select the Evaluate button on the tool bar and click the mouse once, or press Shift+Enter. Nothing appears to happen, apart from a blue *In[n]:=* indicator appearing before the command, but the function has now been defined. All you need to do subsequently is to type in the function name, but include actual values of the variables.

Clicking on the Evaluate button causes the function to be evaluated for the input variables. The value of the function (0.882353) appears on the screen, as in Fig. 11. One point to note is that the voltage is typed with a decimal point, i.e. '12.', even though no figures follow the decimal point. The reason for this is to obtain a relatively low-precision result, to six significant figures. It does not matter which of the values has the decimal point, as long as at least one of them has.

If all of the input values are typed as integers, with no decimal points, the program evaluates the function exactly. This usually means quoting the result as a rational number, in this case $^{15}/_{17}$. Summing up, *cc* is a 'one-line program' for finding the collector current.

The next function in Fig. 11, called *bev*, calculates V_{be}, and requires the entry of three values, R_c, V_{cc}, and H_{fe}. This is a two-line program, the two parts of the procedure being separated by a semi-colon. The first part takes the previous program a stage further and evaluates the base current bc as the collector current divided by H_{fe}. It then takes the value of bc and uses this to find the base-emitter voltage, V_{be}.

This presents a problem because this variable is in the index of an exponential function. You cannot use the Solve command, which is one reason why *Mathematica* can not simply be given all the relevant equations and asked to solve them simultaneously. Instead, FindRoot is used to search for the solution, using Newton's method. To commence this you need to state – or guess – a starting point for the iteration.

The two values in curly brackets indicate that the root to be found is the value of V_{be} and its initial value is to be 0.5, which is likely to be close to the final value. Having evaluated the function as before, we key it in with a set of values. This time the 100 happens to have the decimal point. When this is evaluated, the function returns the value V_{be}=0.532982, in volts. Because of approximations in the values inserted into the equations, probably only the first two figures are significant, and V_{be}=0.53V.

In general, values like *bc*, calculated within a procedure, are available for use outside the procedure. To find the value of *bc* after running the procedure, type '*bc*' and select Evaluate. This returns 8.82353×10^{-6}, or 8.8μA.

Because FindRoot is a rule defining how to calculate V_{be}, but is not actually a value, V_{be} is not available for further calculation. So we use a third procedure to find the value of V_{bb}, the total of V_{be} and the voltage across the base resistor. Keying in the function *vbb* with the values for V_{be} and rb returns the base bias, 0.621217V, or more realistically, 0.62V.

The text and equations of Fig. 11 are saved as a 'Notebook'. This Notebook, and others included on the disk to be published in connection with this series, can be loaded in *Mathematica* and used as an annotated source of ready-made calculation routines. After loading, place the cursor on the function and then click on Evaluate. Then key in the function with variable values and evaluate.

```
CBD02 - Common-emitter bjt amplifier

1) Collector current (mA), when output is at half
                                the supply voltage.
            rc = collector resistor
            vcc = supply voltage
cc[rc_,vcc_]:=(vcc/(2 rc) 1000)
cc[6800,12.]
0.882353

2) Base-emitter voltage, given
            rc = collector resistor
            vcc = supply voltage
            hfe = small-signal current gain

bev[rc_,vcc_,hfe_]:=
(bc=vcc(2 rc hfe);
FindRoot[bc==10^-14 (Exp[38.647 vbe]-1),
{vbe,0.5}])
bev[6800,12,100.]
{vbe -> 0.532982}

3)base bias voltage (vbb), given
            vbe = base-emitter voltage
            rb = base resistor
  Requires program 2 to have been run.
vbb[vbe_,rb_]:= (vbe+rb bc)
vbb[0.532982,10000]
0.621217
```

Fig. 11. Steps for incorporating an equation into a procedure. In this case, the equation is for collector current when output voltage is at half the supply rail.

Diode voltage drop

The voltage drop V_d across a diode, or the base-emitter junction of a transistor, is defined by:

$$i_D=I_0(e^{Vd/nVt}-1)$$

in which i_D is the current across the junction, I_0 is the reverse saturation current, n is a constant, which may be taken to equal unity unless otherwise specified, and V_T=kT/q. In the latter expression, k is Boltzmann's constant (1.38x10-23JK^{-1}), T is the temperature (taken as 300K in this simulation), and q is the electron charge (1.6×10^{-19}C). Substituting these values in the equation gives:

$$i_D=I_0(e^{38.647Vd}-1)$$

For the base-emitter junction of a *BC548*, $i_0\approx10^{-14}$A, which is the value taken for the Spice model of a diode.

Biasing

Returning to the *SpiceAge* circuit, the final task is to select a pair of fixed resistors to provide operating-point bias from the V_{cc} power rail, **Fig. 12**. An appropriate approach is to decide on a value for R_{b1}, and to sweep R_{b2} over a range of values to find one that puts Node *c* at 6V. Edit the netlist (Fig. 2), deleting the voltage source V_{bb} and replacing it with R_{b1} and R_{b2}. Fix the value of R_{b1} at 1kΩ. Allot a temporary value to R_{b2} (say, 300Ω). Its exact value is not significant as it will be swept from 100Ω to 470Ω, which a rough mental calculation indicates will produce a suitable voltage at Node V_{bb}.

The lower and upper sweep voltages are specified in the netlist statement by quoting 'vs=100' and 'vf=470' for the starting and finishing sweep values respectively. This technique can be used when sweeping values of other components, such as capacitors and inductors. To switch on sweeping, select Analyse then Tolerance, Value and Temperature Sweeping.

In the dialogue box select Value sweep, click on the Linear option and make *N*, the number of sweep steps, equal to 10. A small value of *N* produces a quick sweep. Later, this number can be increased to, say, 50 to bring the results closer to the required operating point.

To set the display of the results, select Time then Probes; turn on Probe 1 as a voltage probe, at Node V_{bb}, with its reference at 'Gnd'. When all is set up, run a Quiescent sweep. The graph displays the value of V_{bb} as R_{b1} is swept from 100Ω to 470Ω.

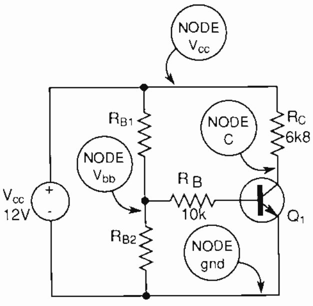


Fig. 12. Complete common-emitter amplifier with base current derived via a potential divider.

At this stage V_{bb} is too high over the whole sweep range. Rather than sweep from a lower value, increase R_{b1} to 3.3kΩ. Repeating the analysis shows V_{bb} ranging between 0.3 and 0.6V, within which range the operating point should lie. Make Probe 2 active, to measure the voltage between Node *c* and ground, with scale factor of 1. The aim is to find the conditions under which Node *c* is at 6V.

Repeating the quiescent sweep shows Node *c* and 12 V, then falling rather rapidly to 0V. Narrow the sweeping range so that it just spans the fall at Node *c*. The scale on the *x*-axis of the analysis is not graduated in ohms or any other physical quantity but represents sweep numbers. If *N* is 50, the scale is graduated from –25 to +25. Any given point on this scale is converted to resistance by using the formula:

$$R=\text{min}+\frac{range(n+N/2)}{N}$$

in which min is the lower end of the sweep range, range is the sweep range, and *n* is the sweep number. For example, at the operating point, sweeping from 100Ω to 225Ω in 50 steps, the sweep number is 6.58. Substituting in the formula:

$$R=100+\frac{125(6.58+25)}{50}=178.95\Omega$$

A standard 180Ω resistor is the obvious choice. Edit the netlist to make R_b equal to 180Ω and delete the sweep specifications. Cancel Value sweep in the dialogue box and click on Nominal. Turn on the voltage probes. Now select Analyse then DC Quiescent (not sweep), to obtain a table of node voltages.

Node *c* is at 5.72V, which is close to the operating point. Change Probe 1 from voltage to current through R_c. Repeat the analysis and obtain a table which shows that the collector current is 923μA, close enough to the required 1mA.

Looking at the currents through other resistors we see that R_{b1} and R_{b2} are both passing more than 3mA, which might be considered wasteful. Edit the netlist to make R_{1a} and R_{1b} ten times bigger, 33kΩ and 1.8kΩ respectively. Repeating the analyses shows that now the collector voltage is 6.1V, and its current is 867μA, which are close enough to the required operating conditions. ■

CIRCLE NO. 126 ON REPLY CARD

Keeping a tag on technology misuse

The mix of rf and digital engineering that is tagging technology can already keep a limited track of people. How far could we – or should we – go asks Nigel Burke?

Electronic tags are coming, ready or not. Society has not yet developed a cultural attitude to wireless tagging, nor worked out its effects on the workplace. But the tagging market will be supply-driven by the neat rf modules that are now becoming available to engineers.

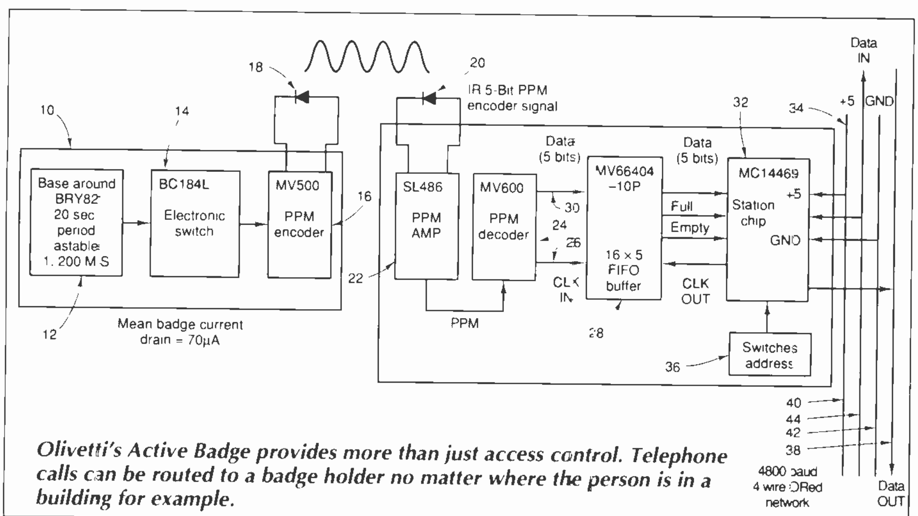
The Criminal Justice Bill provides for fresh trials of electronic prisoner-tagging next year, and the baby stealing headlines of only a few months ago have focused thinking on access control to hospitals, and even the direct radio-location of unsupervised children.

Techniques for tagging people are the same as those for inanimate articles and livestock – except that tags may not be stuck nor rivetted onto humans. It is also natural for people to try to subvert anything which inconveniences them. In 1989 when the Home Office first tried electronic tagging offenders, more than a third of the taggees cut and ran.

Present people-tagging devices deal mainly with locations around a site. Machinery that could send world-wide coordinates back to any controlling point is still unwieldy, though quite feasible, since gps or loran data need no longer be returned to base by shortwave or satellite phone. Channelling through the nearest Internet dial-up host is a possibility. The Home Office expects that its American contractors for the new judicial tag will supervise curfew violations by modem from the USA.

The tag will only monitor proximity to a base station connected to a BT socket. It will not yet give latitude and longitude.

Humbler tagging applications have evolved from the familiar anti-shoplifting tags. Types containing a lumped resonant *LC* circuit are still in use, while library books and flatter articles are protected by printed coils and stripline resonators. Such tags are inert until they approach a 'read field' in which a pulsing rf emitter excites the circuit, and listens for the damped oscillations. There is no way to differentiate between two tags using the same band and range tends not to exceed the 2m needed to cover a doorway. Edinburgh's Simpson Memorial Hospital is evaluating this kind of system as a proposed solution to baby-napping – though it is hard to see why a kidnapper would fail to detach the tag. Simple tags are also in use at Addenbrooke's Hospital to warn staff when elderly patients have wan-



Olivetti's Active Badge provides more than just access control. Telephone calls can be routed to a badge holder no matter where the person is in a building for example.

Supertag security

The *Supertag* long-range rfid device was developed by the South African CSIR organisation, and has been licensed in Britain to ICL through the British Technology Group.It is a battery-less uhf transponder device, intended to replace barcodes for labelling goods in shops.

The circuit is broad band, relying on time division and addressed transmissions to supply data to and from the correct tag. Working energy is extracted from the interrogating rf using a Schottky diode charge pump.

Because goods will be piled together, a good anti-clash technique is needed to differentiate each tag, especially where tags return the same id code.This can be achieved by programming tags to respond at random intervals, 25items/s. Whenever a clear and uninterrupted tag signal is received, it is addressed with a control signal to stop transmitting its 64 bit id, and counting rate changes to fifty items/s.

The British Technology Group, London, sees the Supertag as technology ideal for personnel tagging. If economies of scale do allow it to displace barcode labels, it would certainly be cheap enough for a throw-away visitors pass.

But the necessary simplicity of the digital circuits, having only 300 gates and 64 bits of eeprom memory, prompts security questions concerning the whole technology. A tag could be rewritten for a different price, or a different visitor security clearance. Even a read-only tag is susceptible to a record-and-playback hack.

As with the first generation of radio car locks, any device that transmits the same code every time can be imitated by a programmable transmitter. BTG's plans for the development of the Supertag include tamper-evident features. Unfortunately, the best security lies with a tag that records all reads and writes in an audit trail – a considerable pressure on production cost.

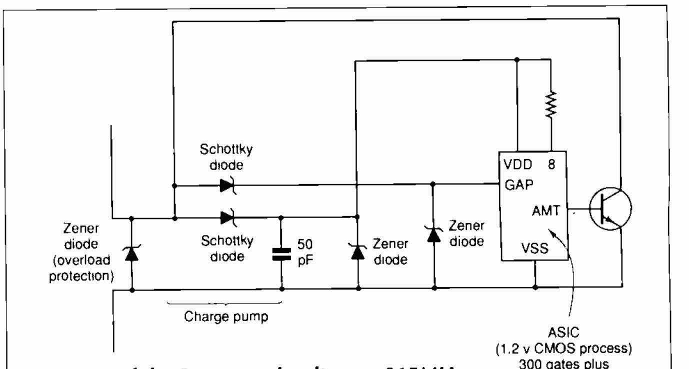
dered off. They can be effective in giving slightly-vague patients semi-autonomy. But they will vary in how far the semi-autonomy represents freedom from dependency, or the loss of full autonomy.

Making tags more useful

Tag usefulness is extended by modulation of the return signal with a unique identification code. Current tags such as Cotag's security badges and stock-control labels do not have their own rf frequency source. Instead, they intercept the reader-equipment's wave and re-radiate it having performed frequency division with a flip-flop, and chopped it with a digital waveform. The battery-less versions rectify some of the incident wave to provide dc power for the digital circuits.

Eagle Tracer in Cheshire manufactures a read/write variant that can return a 64bit id code, and be reprogrammed on-the-fly. RFID tags of this nature, produced to wheat-grain dimensions and made biologically inert, can be implanted in flesh. In the Irish Republic, licensed guard dogs have had to have rfid implants since 1989, and there is no medical reason why human beings could not have implanted tags.

RFID can plainly help with access control. Eagle Tracer tags are meant to be integrated with pir sensors and a computer-governed environment, such that people without valid tags in the form of badges are detected. In the hospital context, security would mean giving tags to workers, nursing mothers and cleared visitors. Tag-reading doorways would be looking for intruders, not smuggled infants.



Prototype of the Supertag circuitry – a 915MHz tag achieving counting rates of 50 items per second within a 3m sphere around the operator.

Over at Olivetti Research Labs, the smart security badge has been transformed into the active badge, a technology that provides services to the user, rather than just the stern function of access control. An active badge is an addressable tag for office workers, that communicates with the office computer network via infra-red transceivers distributed about the site. It transmits its id using an led baseband ppm signal at 9600 baud, then listens for any paging messages from the network.

Badge wearers' locations are mapped room by room – relative to colleagues and badged pieces of shared portable equipment – and can be paged, have phone calls, work sessions and mail diverted to the nearest workstation. Thankfully a do-not-disturb status can also be requested through the system.

A smart governing network can take far more advanced decisions than whether or not to let someone through a door. It can decide whether staff are adequately supervised by more senior workers, or whether movements seem suspicious or unproductive. Built-in security includes authenticating badges by shooting a random number at them and requiring the badges to process it through a cryptographic algorithm and respond with the right answer.

Analysing how workers spend their day using all the available context is potentially more powerful, since businesses lose more to inside cheats than interlopers. But there is also the potential for misuse of active badges or similar technology, having the potential to increase the general serfdom of hireling workers, just as networked keystroke-counting can be used to squeeze the work rate of typists.

Wider context

In the wider world, serious rf is needed to keep tabs on people. Anti-kidnap tags have been available for years for VIPs but have been expensive, relying on matched receivers to track them from great distances.

Classical df (direction finding) techniques have reached the mass market in the form of Tracker, AA's vehicle location product. A vhf transmitter hidden in the car is activated at the owner's request by a signal from the system's administrators, and transmits a signal that suitably equipped police cars can track, using a phased array of roof-mounted rod antennae. Range is said to be six miles, though the tracker does not know where the vehicle is, being just an emitter. As such it is not the future for outdoor people tagging.

The cell phone networks are a more promising prospect for location information. A rough-grained location facility is already built into cellular networks. Users are tracked from cell to cell to allow distribution of frequency slots, and in urban areas, the cell may be barely a hundred yards wide. Spectrum re-use amongst non-adjacent cells makes it possible for far more users to emit locator signals than on a free-range basis, as in Tracker.

Cellular phone networks and data networks such as Paknet could be enhanced with direction sensing antennae, and offer a convenient route back for location data to whatever business or authority is using it. Or gps/loran data could be collected by the tag, and returned by cell phone.

Certainly, cell phones are underdeveloped for personal security. Low tariff emergency-only phones would be useful for children, and ccd camera chips could be incorporated to enable subscribers to threaten aggressive strangers with sending a snapshot straight to the police.

Rf and digital engineering have both come together with a mighty clash to get tagging off the ground. But when it comes to tagging people, even more technologies are converging.

Graphical information systems will allow industrial site operators, rescue services and pizza deliverers to map staff on screens showing not just route data, but warning information too: from chemical hazard zones to credit blackspots.

Neural network pattern analysis will be used to examine working patterns for inefficiency and fraud, while tagging supermarket trolleys could lead to development of new means of consumer anthropology. Video and biometrics will be used to ensure that people tagging is monitoring the specified persons. The problem for engineers is not so much to make it all work as to predict which among all the technological permutations will be commercial winners.

For individuals, the problem is deciding whether to have a subcutaneous Visa card fitted, a cell-phone that can send a distress signal with coordinates down to 10m, or even one that transmits a signal like that all the time, home to Mum – or to Big Brother's bunker.■

Economics of total surveillance

The notion that everyone might one day be tracked by computer is always titillating to fiction writers. But it is unlikely that such tracking will come about in a unitary and planned way.

RFID techniques can certainly be used to police buildings, and perhaps even security zones such as the City of London. *Ad hoc* location data can be culled from cameras scanning car number plates, and the Inland Revenue has already used cashpoint machine records to map the movements of tax exiles, to see whether they are honestly claiming non-resident status.

But what about tracking all our movements to a 10m resolution throughout the whole of our lives?

Let's do the sums.

If you live for 76 years – the UK national average – that is a total of 39945600 minutes. The equatorial circumference of the Earth is 40,689km, giving 4,068,900 reference points of 10m, ie longitudes. We can refer to any longitude by a seven figure denary, or twenty-two bit binary number. Taking a liberty with the actual shape of the Earth, we'll say that the latitude can be expressed in the same way, so we can transmit a person's location to this accuracy in forty-four bits. Doing this once a minute accumulates a lifetime tachometer disk only 219.7Mb in size. That includes a lot of redundant data, because people are stationary for most of the time, and only slight displacements need be recorded. So the data is compressible.

If everyone has to transmit a unique identity number, location and some parity and cellular handshake data, the data rate will turn out to be about 1baud, and the system's bandwidth requirement should be comparable to existing cell phone networks.

MAXIMISING monopole bandwidth

Monopoles on a ground plane are simple and versatile but one of their most attractive features is exceptional bandwidth of up to 50% of the quarter-wave frequency. Richard Formato describes how the monopole's bandwidth behaves and how to maximise it.

In the base-fed monopole on a ground plane, the radiating element is a conducting cylinder of height L and diameter D, **Fig. 1**. The bottom of the element is insulated from the ground plane by a base insulator, and the rf source (transmitter) is connected between the bottom of the antenna and ground[1].

In theory, the source is a 'delta-function' generator, which means that the base insulator is infinitely thin. In practice, the insulator should not be too thick. Coaxial cable is usually used to feed the antenna, with the centre conductor connected to the monopole and the shield to the ground plane.

Bandwidth of the antenna is determined by the L:D ratio, which is computed by dividing the element length, L, by its diameter, D, in consistent units. Dimensions specified in different units – feet and inches for example – must be converted to the same unit before calculating this ratio.

The term bandwidth means impedance bandwidth, defined here as the range of frequencies where the antenna's input vswr is 2.5:1 or less. Voltage-standing-wave ratio thresholds other than 2.5 can be used, but this value is a good compromise for antennas used for both transmitting and receiving.

Figure 2 plots the monopole vswr relative to 50Ω versus frequency for several L/D ratios. The abscissa is the ratio F/F_0 in percent, where F_0 in megahertz is given by $299.8/(4L)$. Note that L is in metres.

At frequency F_0, L is a quarter-wavelength in free space. For example, if L is 2.5m, the free-space wavelength is 4(2.5), which is 10m, and the corresponding frequency Fo is 299.8/10 which is 29.98MHz. Many of you will recognise this as the formula f(MHz)=300/wavelength(m) with the more accurate value of 299.8 replacing 300. Note that F_0 is not the monopole's resonant frequency, which is slightly lower than F_0.

Figure 2 includes curves for L/D ratios ranging from 2500 to 3.125. The narrowest bandwidth is slightly less than 15% of F_0 and corresponds to the highest L/D ratio of 2500. At L/D=12.5, bandwidth increases to nearly 35%.

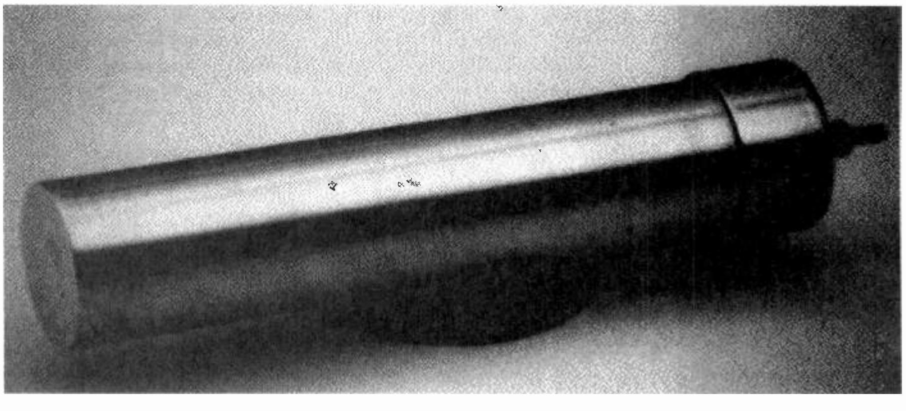
When L/D=5, the bandwidth reaches its maximum value of about 50% of F_0, which is very large for a simple radiating element without any broad-banding components. As L/D falls below five, the bandwidth decreases. At L/D=3.125, for example, it is about 38%.

Figure 2 also reveals an interesting distribution of bandwidth with frequency when L/D=5. Slightly less than two thirds of the available bandwidth is above F_0, and slightly more than a third below it. The lowest frequency where vswr is 2.5:1 is about $0.808F_0$, the highest $1.31F_0$, and the frequency for vswr minimum is about $0.987F_0$.

These observations provide some simple, easy-to-remember and use rules for computing maximum monopole bandwidth (50Ω characteristic impedance, vswr<2.5:1):

- maximum bandwidth occurs when the ratio of monopole length to diameter is five.
- maximum bandwidth is about 50% of the frequency at which the monopole is a quarter-wave long.
- frequency of minimum vswr is about 1.3% less than the quarter-wave frequency.
- approximately two thirds of the bandwidth is above the quarter-wave frequency, and about a third is below.
- vswr minimum is a near-perfect 1.009:1.

As an example, a monopole 43cm long, 8.65cm diameter, covers the frequency range 140.84-228.34MHz, which corresponds to the



Typical maximum-bandwidth monopole with an L/D ratio of 5. It covers approximately 246-393MHz with a vswr less than 2.5.

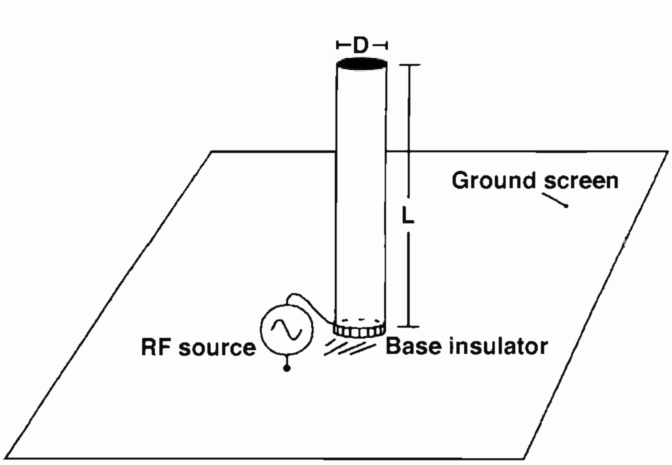


Fig. 1. Monopoles are widely used because of their bandwidth and simplicity. This is a base-fed monopole on a ground plane, whose rf source is connected between the bottom of antenna and ground.

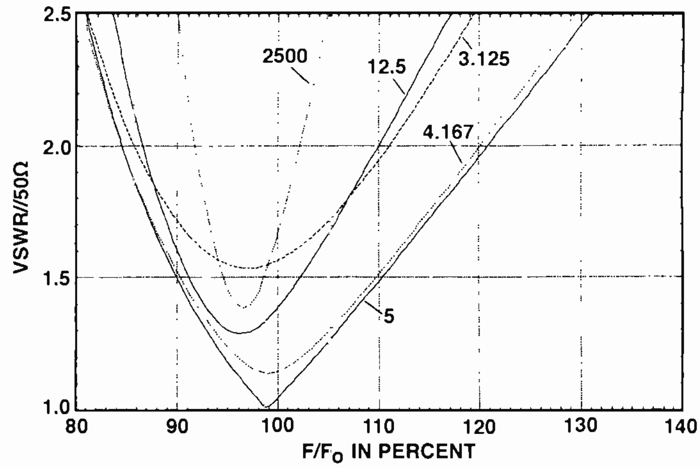


Fig. 2. Voltage-standing-wave ratio for a quarter-wave monopole on a ground plane. Curves for several length:diameter ratios are shown.

2 and 1.25 metre bands. Vswr is less than 2.5 and quarter-wave frequency is 174.3 MHz.

At hf, a monopole 4.35m long and 87cm diameter will cover 13.92-22.57MHz, i.e. 20, 17 and 15 metre bands. Other frequency ranges can be covered by varying the radiating element length and applying the rules above.

Why the monopole's bandwidth behaves this way can be understood by referring to **Fig. 3**. These curves plot the input impedance (resistance R, reactance X) versus frequency for L/D ratios of 2500 ('thin' antenna) and 5 ('fat' antenna).

Resistance variation is similar for both antennas. Value R is comparable at most frequencies and gradually increases with increasing frequency. But when the frequency exceeds 120% of F_0, the behaviour changes. Input resistance of the thin antenna increases quickly above $1.2F_0$, while the fat monopole's flattens out, then starts to decrease.

Input reactance behaves much differently. For the thin antenna, X increases almost linearly with frequency, varying from –60Ω (capacitive) at $0.84F_0$ to +100Ω (inductive) at $1.15F_0$. The thin antenna is resonant (X=0) at only one frequency near $0.96F_0$.

The fat monopole's reactance varies much more slowly over the entire range, 80% to 133% of F_0. While essentially capacitive (–16.5 to –55.5Ω), the antenna is actually resonant at two frequencies: approximately 99.87% and 101.67% of F_0 This is not obvious on the plot.

Because the resistance and reactance, especially X, fluctuate more for the thin antenna, its vswr increases more quickly with frequency than the fat monopole's. Gradual variation of input reactance when L/D=5 is primarily responsible for the fat antenna's large impedance bandwidth.

Building fat monopoles at hf may require special construction techniques because the element diameter is large. Instead of a continuous conducting cylinder as the radiator – a form covered by metal foil, for example – an acceptable, easy-to-build alternative consists of wires parallel to the cylinder axis equally spaced around its circumference. This configuration is sometimes referred to as a 'cage monopole', apparently because of the resemblance to a bird cage. The greater the number of wires, the better the approximation to a continuous conductor. As a rule of thumb, at least eight wires should be used.

Another consideration in building any kind of monopole is the size of the ground plane. Theoretically, it should extend indefinitely in

Monopole applications

Throughout the rf spectrum, the monopole antenna is one of the most widely used. It is probably the only one used at all frequencies ranging from vhf to uhf. Resonant quarter-wave antennas are the rule at mf and above (the AM broadcast band is in the mf range), but other electrical lengths are occasionally employed for special applications. At vlf, the long wavelength forces monopoles to be electrically short.

Communication services supported by these simple antennas are as varied as their operating frequencies. Very-low frequency monopoles support worldwide military communications, especially Navy and Air Force. Base-fed AM broadcast towers are a familiar sight in every part of the country.

At vhf, public services such as the police, fire, and highway departments rely on mobile whip and base station monopoles to provide essential communications. Many private services, for example, taxicabs, trucking, and even remote telephones, also use monopoles in the low vhf range.

At higher vhf frequencies, aircraft communications rely on fuselage-mounted monopoles. And, at uhf, the car phone antenna is probably quickly becoming the most common monopole of all. These systems serve to illustrate how widespread use of the monopole antenna is.

Monopoles are used extensively by the US Navy to support shipboard hf communications, and many of them are broadband systems. Two examples of shipboard monopoles are the US Navy's hf cylindrical cage monopole, consisting of eight wires parallel to the support mast, and the multi-element broadband monopole.

The cage monopole's L/D ratio is greater than 5. This results in less than optimum bandwidth, but nevertheless improves performance over a single, thin element. The multi-element broadband monopole has an L/D ratio much closer to 5:1, which provides better performance than the cage monopole because of the lower L/D ratio.

New communication technologies, especially spread spectrum, frequency-hopping, and frequency-agile systems, demand the widest possible antenna bandwidth. Although these systems are currently used primarily for military communications, commercial applications will follow which generate still more pressure for wider bandwidth. The broadband monopole is an attractive transmit/receive antenna for these new modes because it provides robust performance in a simple, efficient, and inexpensive system.

In the photograph is a typical maximum-bandwidth monopole with an L/D ratio of 5. It is 25.4cm long by 5.08cm diameter and covers approximately 246-393MHz with a vswr less than 2.5. Quarter-wave frequency is approximately 295MHz. This device was fabricated from hard-drawn copper tubing, weather-sealed at the top with brass foil, and mounted on an end-cap containing a centred $^1/_4$-inch threaded brass rod. The rod mounts the monopole on a base feed-through insulator which electrically isolates the element from the ground plane.

Continuous, radial-wire, or wire mesh ground planes can be used. The centre conductor of a 50Ω coaxial cable connects directly to the monopole, and the shield is connected directly to the ground plane. The optimised L/D=5 monopole element thus provides maximum possible bandwidth with no matching or tuning circuitry.

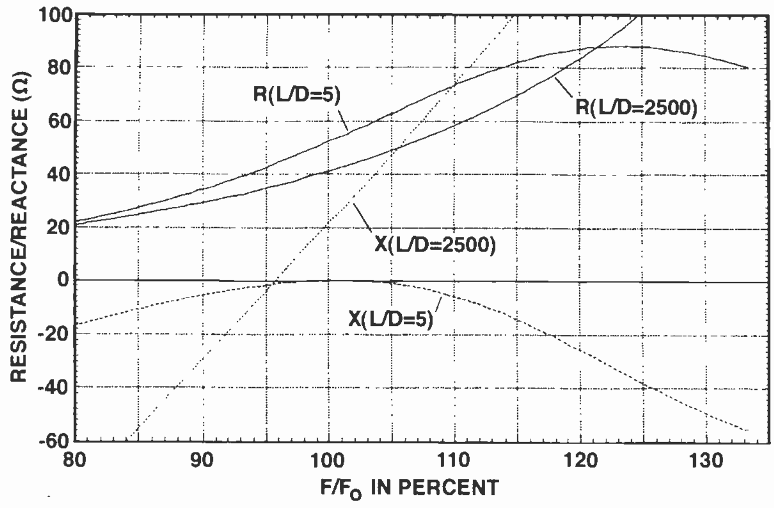


Fig. 3. Input impedances for quarter-wave monopole with ground plane. When frequency reaches 120% of F_o, resistance behaviour starts to change for both thin and fat antennas.

all directions; but, as a practical matter, a circular ground plane of a few wavelengths radius usually works well. The cylindrical radiating element should be a continuous conducting metallic surface; the ground plane should also be continuous if practical.,

Ground planes of wire mesh or radial wires are frequently used. They provide good performance if properly constructed. Mesh openings should be a small fraction of a wavelength, typically eighth-wave or less. If radials are used, a large number is required, at least 16, preferably more.

Predicted monopole bandwidth performance has been verified experimentally, with theoretical and measured data showing excellent agreement, typically within 5%. Measured bandwidth for a 476MHz antenna was actually somewhat greater than predicted. Of course, actual and computed performance will not agree well if modelling assumptions are violated. For example, if the ground plane is too small, or if continuous metallic surfaces are poorly approximated by wire structures, then the agreement between measured and theoretical data will be degraded.

Another potential source of error is making measurements through a long coaxial cable. Resistive losses artificially reduce the vswr and increase the bandwidth at the cable input by dissipating some of the reflected power in an unmatched system. The vswr reference point in this note is the monopole input, so that only data measured at the monopole's base can be compared directly.

These simple design rules should encourage experimentation with broadband monopoles throughout the amateur bands. Multiband or single band antennas are easy to design and build, and can be fed directly with 50Ω coaxial cable without an antenna tuner or matching network.

This article illustrates the importance of the *L/D* ratio in determining the bandwidth of wire radiators generally. Similar considerations apply to other wire antennas, such as dipoles, parasitic arrays like yagis, or active arrays. Even though the monopole design rules are not directly applicable, paying attention to *L/D* should be a design consideration for any wire antenna, because selecting the right value may result in significantly improved bandwidth. ■

Reference:

1. Mathematically, the RF source is a "delta function" or "slice" generator. It drives the antenna by creating voltages +V/2 and -V/2 across the infinitely thin cross section where the monopole base is in electrical contact with the ground plane. In theory this means that the base insulator is infinitely thin; in practice, it should not be too thick.

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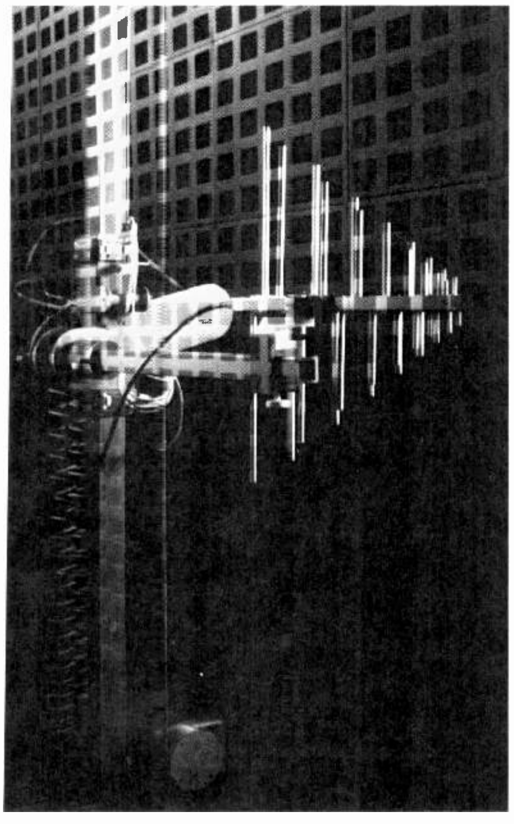


Stable performace in the real world

Norm Dye and Helge Granberg explain how acceptable gain, noise and stability are achieved in the real world of modern high performance transistors. From the book RF Transistors: principles and practical applications.

Making an assumption of unilateral gain has helped in analysing the overall gain of a transistor stage by considering contributions from three parts. But assuming S_{12} (the cause of the interaction of input and output, see November *EW + WW*) has a value of zero ignores the problem of amplifier stability – it also leads to the erroneous conclusion that output matching has no effect on input matching.

Amplifier design calculations which do not include device (and circuit) feedback are only an approximation, and can lead to inaccurate solutions and, possibly, circuit oscillations when the design is realised.

So how are acceptable gain, noise and stability achieved in the real world of modern high performance transistors when S_{12} is other than zero? The answer is straightforward – *S*-parameters, allowing device stability to be calculated by determining a term called the Rollett stability factor *K*.

To make the equation simple, first calculate intermediate quantity $D_s = S_{11}S_{22} - S_{12}S_{21}$. Stability factor *K* is then calculated as:

$$K = (1 + |D_s|^2 - |S_{11}|^2 - |S_{22}|^2)/2|S_{21}||S_{12}|$$

If *K* is greater than unity, then the device will be unconditionally stable for any combination

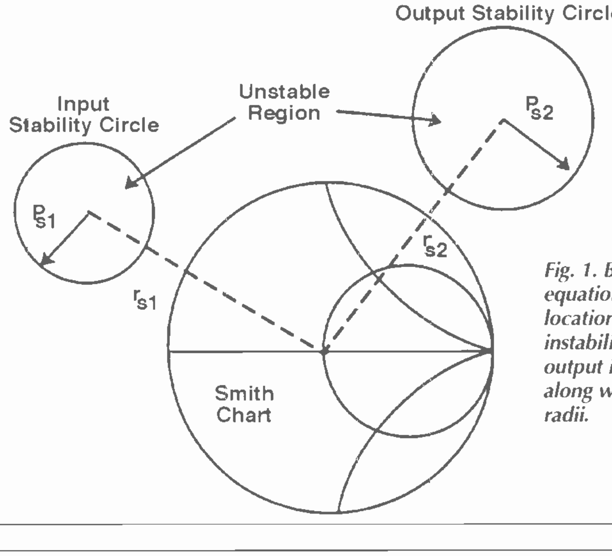
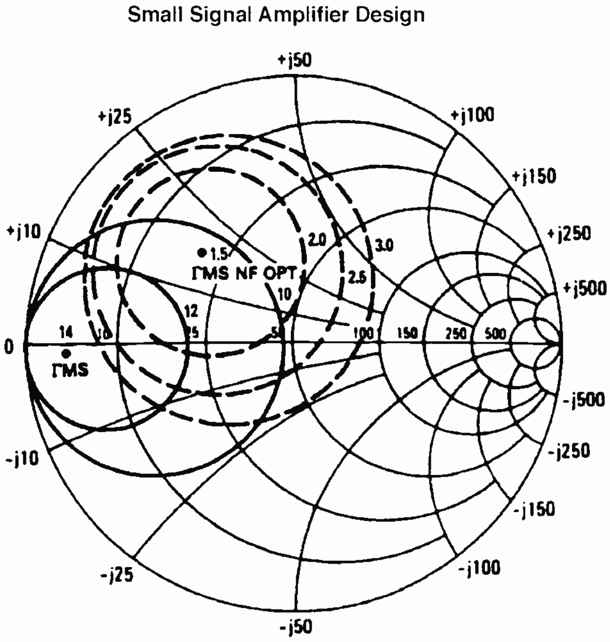


Fig. 1. Bring together the equations for the centre locations of the input instability circle and the output instability circle along with those for their radii.



V_{CE} = 6.0 V, I_C = 5.0 mA
f = 1.0 GHz

f(GHz)	NF OPT(dB)	Rn (Ω)	NF50 Ω (dB)	ΓMS NF OPT
1.0	1.5	7.5	2.2	0.48 ∠134°

ΓMS	ΓML
0.89 ∠ –179°	0.81 ∠66°

Solid circles — Gain
Dashed circle — Noise figure

Fig. 2. Comparing gain and NF circles shows that minimum noise figure cannot generally be obtained at maximum gain.

of source and load impedance. But if it calculates to be less than 1, the device is potentially unstable and will most likely oscillate with certain combinations of source and load impedance.

S-parameters go one step further. They permit the calculation of "stability circles" which can be plotted on the Smith chart and which separate regions of stability and instability. Generally only a portion of the circle will be visible on the Smith chart. When choosing source and load impedances, values that lie within the regions of instability should be avoided.

Manufacturers who supply gain and noise circle data with their transistors also plot regions of instability, typically indicated by dashed lines. Obviously these circles (or portions thereof) will not exist within the Smith chart boundaries for transistors with a value of $K > 1$.

Calculation and plotting of instability circles (**Fig. 1**) are straightforward operations involving S-parameters – though tedious and best performed using a computer. Bringing together the equations for the centre locations of the input instability circle and the output instability circle along with those for their radii r_{s1} gives:

$$r_{s1} = \frac{C_1^*}{|S_{11}|^2 - |D_s|^2}$$

where $D_s = S_{11}S_{22} - S_{12}S_{21}$

$$C_1 = S_{11} - D_s S_{22}^*$$

and r_{s1} is the centre of input stability circle. Also

$$p_{s1} = \left| \frac{S_{12}S_{21}}{|S_{11}|^2 - |D_s|^2} \right|$$

where p_{s1} = radius of input stability circle. Likewise

$$r_{s2} = \frac{C_2^*}{|S_{22}|^2 |D_s|^2}$$

where

$$C_2 = S_{22} - D_s S_{11}^*$$

and r_{s2} is the centre location of the output stability circle. And

$$p_{s2} = \frac{S_{12}S_{21}}{|S_{22}|^2 - |D_s|^2}$$

where p_{s2} is the radius of the output stability circle.

However, determining the proper source and load impedances is simplified to a large extent when the transistor can be treated as a unilateral network. If we have satisfied ourselves about the stability of our circuit, then we will find it beneficial, at least as a first approximation, to treat it in this manner whenever possible during design.

If S_{12} can not be assumed equal to 0, then the equation for power gain G_p can be used to

Plotting a specific power gain circle

A step-by-step procedure for plotting a specific power gain circle would be:

- Select the desired value of G_p;
- calculate $g_p = G_p/|S_{21}|^2$;
- calculate $K = (1 + |D_s|^2 - |S_{11}|^2 - |S_{22}|^2)/2|S_{21}||S_{12}|$;
- calculate $D_s = S_{11}S_{22} - S_{12}S_{21}$;
- determine

$$R_p = \frac{\left[1 - 2K|S_{12}S_{21}|g_p + |S_{12}S_{21}|^2 g_p^2\right]^{\frac{1}{2}}}{\left|1 + g_p\left(|S_{22}|^2 - |D_s|^2\right)\right|}$$

- determine

$$C_p = \frac{g_p C_2^*}{\left|1 + g_p\left(|S_{22}|^2 - |D_s|^2\right)\right|}$$

Once we select a value of Γ_1 from a point on the gain circle, we can then determine $\Gamma_s = \Gamma_{IN}^*$ using

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

Summary of gain/noise figure design procedures

1. Once a transistor and its bias conditions have been selected, the S-parameters should be analysed to determine if the simpler design procedures involving the assumption that $S_{12} = 0$ can be used. Limits placed on the maximum error introduced by this assumption can be seen from:

$$U = \frac{|S_{11}||S_{21}||S_{12}||S_{22}|}{\left(1-|S_{11}|^2\right)\left(1-|S_{22}|^2\right)}$$

and

$$\frac{1}{(1+U)^2} < \frac{G_t}{G_{tu}} < \frac{1}{(1-U)^2}$$

2. Next, use Rollett's stability factor, from $D_s = S_{11}S_{22} - S_{12}S_{21}$ and $K = (1 + |D_s|^2 - |S_{11}|^2 - |S_{22}|^2)/2|S_{21}||S_{12}|$ to identify the possibility of instabilities depending on source and load matching.

3. Subsequent steps depend on the desired results:

- For narrow-band and maximum gain, conjugate match input and output.
- For specific gain at a single frequency, use the gain circles provided by the device manufacturer (or draw the appropriate available gain circle using the equations for g_a, C_1, R_a and C_a). After the gain circle is drawn, select a value for Γ_s and calculate $\Gamma_L = \Gamma_{OUT}$ using:

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L}$$

If $S_{12} = 0$ can be assumed, the "gain" or "loss" can be divided between the input /output matching networks using the equations for d_s, R_s, g_s, G_s, d_L, R_L, g_L and G_L, and appropriate values for source and load terminations determined. In this case the input and output of the amplifier will not be matched to Z_0. So if a low vswr is a requirement, this approach should be avoided.
- If a noise figure and gain at a frequency are needed, use both gain and noise figure circles provided by the device manufacturer and select an appropriate value of Γ_s. Again, calculate Γ_L as previously stated.
- For broadband performance, examine the $|S_{21}|^2$ performance of the transistor over the frequency range of interest and determine the amount of gain or loss that must be provided by the matching networks to keep the overall gain the same at the band edges. Plot these gain circles on a Smith chart using the equations for g_a, C_1, R_a and C_a. By trial and error (or the use of computer optimisation) determine a matching network that will satisfy both "gain/loss" circles simultaneously.

develop a mathematical procedure for determining values of Γ_L and $\Gamma_S = \Gamma_{IN}^*$.

Manipulating that equation allows recognition of constant operating power gain circles having radii R_p of

$$R_p = \frac{\left[1-2K\left|S_{12}S_{21}\right|g_p+\left|S_{12}S_{21}\right|^2 g_p^2\right]^{\frac{1}{2}}}{\left|1+g_p\left(\left|S_{22}\right|^2-\left|D_s\right|^2\right)\right|}$$

where K is the previously-identified Rollett stability factor, $D_s = S_{11}S_{22} - S_{12}S_{21}$ and $g_p = G_p/|S_{21}|^2$.

Locations of the centres C_p of the circles are

$$C_p = \frac{g_p C_2^*}{\left|1+g_p\left(\left|S_{22}\right|^2-\left|D_s\right|^2\right)\right|}$$

where $C_2 = S_{22} - D_s S_{11}$ as before.

Maximum operating power gain occurs when $R_p = 0$, and for this condition and for the case where $K > 1$ (the circuit is unconditionally stable):

$$G_{p.max} = \frac{\left|S_{21}\right|}{\left|S_{12}\right|}\left(K-\sqrt{K^2-1}\right)$$

We have already assumed that $\Gamma_S = \Gamma_{IN}^*$ and, under these conditions $\Gamma_{p.max} = \Gamma_{L.max}$

Because power gain circles involve load reflection coefficients, it is more common to plot constant available power gain circles which involve source reflection coefficients. The process is similar and the equations are only slightly different: $g_a = G_A/|S_{21}|^2$ and $C_1 = S_{11} - D_s S_{22}^*$. Then

$$S_{11} = \text{input reflection coefficient} = \frac{b_1}{a_1}\Big|_{a_2=0}$$

$$S_{22} = \text{output reflection coefficient} = \frac{b_2}{a_2}\Big|_{a_1=0}$$

$$S_{21} = \text{forward transmission coefficient} = \frac{b_2}{a_1}\Big|_{a_2=0}$$

$$S_{12} = \text{reverse transmission coefficient} = \frac{b_1}{a_2}\Big|_{a_1=0}$$

Two-port S-parameter definitions (accidentally omitted from last month's Fig. 1).

$$R_a = \frac{\left[1-2K\left|S_{21}S_{12}\right|g_a+\left|S_{21}S_{12}\right|^2 g_a^2\right]^{\frac{1}{2}}}{\left|1+g_a\left(\left|S_{11}\right|^2-\left|D_s\right|^2\right)\right|}$$

and

$$C_a = \frac{g_a C_1^*}{\left|1+g_a\left(\left|S_{11}\right|^2-\left|D_s\right|^2\right)\right|}$$

where $D_s = S_{11}S_{22} - S_{12}S_{21}$ and $g_p = G_p/|S_{21}|^2$, R_a is the radius of gain circle and C_a is centre of gain circle.

Constant available power gain circles involve Γ_S and, as seen earlier, so do constant noise figures. Thus, both sets of circles can be plotted together on a Smith chart to provide trade-off information between gain and noise figures. These are the curves presented by device manufacturers in their low noise transistor data sheets.

Comparison of gain circles with noise circles (**Fig. 2**) makes clear another fundamental point about low noise amplifier design. Minimum noise figure cannot generally be achieved at the same time as minimum gain. So designing a low noise amplifier becomes a trade-off of gain and noise figure to achieve an acceptable value of each. ■

In a future issue, Norm Dye and Helge Granberg will look at examples intended to clarify these procedures further by working through specific problems.

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Best rf article '95

Following the success of 1994's Writers Award, **Electronics World** and **Hewlett-Packard** are launching a new scheme to run from January to December 1995.

Only articles which have an element of rf design will be eligible for consideration by the judging panel. It is hoped that this year's award will focus writer interest on rf engineering in line with the growing importance of radio frequency systems to an increasingly cordless world.

The aim of the award scheme is to locate freelance authors who can bring applied electronics design alive for other people.

Qualifying topics might include direct digital synthesis, microstrip design, application engineering for commercially available rf ICs and modules, receiver design, PLL, frequency generation and rf measurement, wideband circuit design, spread spectrum systems, microstrip and planer aerials... The list will hopefully be endless.

All articles accepted for publication will be paid for – in the region of several hundred pounds for a typical design feature.

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LETTERS

Letters to *"Electronics World + Wireless World"* Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Braun-ed off

John Powell Riley's article "The man who started ripples in the ether" (*EW+WW*, September, pp. 778-782) was very interesting and contained a lot of information about the early days of wireless. I am also in complete agreement with his words: "No single person invented the radio". But I would like to suggest a reappraisal of the German part of his article.

For me there is too much stress on Slaby and nothing on Braun, who was in fact much more important for the "ripples in the ether".

Karl Ferdinand Braun was a professor in Straßburg, and some of his important work included investigating the semiconductor effect, the voltmeter, the display tube, radar displays, the crystal semiconductor, spark transmitter with primary and secondary system – the type of transmitter used by Marconi for crossing the ocean in 1901 – and ferrite.

Braun was granted a patent for his transmitter in Germany in 1898, and in England, No 1862, in January 1899. Marconi applied in London for the same item in two parts and got the patents in April 26, 1900, and January 7, 1901, one of them number 7777.

In 1899 Braun founded the "Professor Braun's Telegraphie GmbH Hamburg", and in January 1901 he tried to get into business contact with Marconi. Instead he joined with Siemens & Halske in 1901.

Slaby copied from Braun without crediting the origins – Braun, in fact, was really the better inventor – and patent troubles rumbled on until May 1903, when the German Emperor ordered AEG and Siemens & Halske to create a new company, later called Telefunken.

Braun and Marconi together were awarded the Nobel prize for physics in 1909.

I would be very interested to read a copy of the three patents mentioned above, to see their differences. Could any *EW+WW* reader help me?

Gregor Ulsamer
Logumer Str 66
D-26723 Emden
Germany

Mosfet input

I found it interesting that Douglas Self should blame the input stage for the disparity in positive and negative slew rates. Unless he was driving the input circuit and VAS stage into an almost-totally saturated condition – provoking a slow recovery which might be exacerbated by a droo in the supply voltages during prolonged driving – then the most likely culprits are almost certainly the ponderously slow output devices and drivers that he has opted to use. This is particularly so since the pnp driver (*MJE350*) has a nominal F_T of 4MHz while its complement MJE340, npn, has an F_T of 10MHz. Moreover the output devices are only rated at 2MHz F_T. These 30A devices probably have a low β and need some considerable driving. At first sight, changing R_{13} in the VAS stage to a lower value might indicate a lack of drive for the pnp driver/output devices at high frequency. The problem is that one erroneously assumes that the output devices are matched at high frequencies. Mostly they are not, and may need a higher current drive than anticipated because of dynamic capacitive effects prominent at high power.

It is this effect which provides a large anomaly between an estimated slew rate, in this case of 40V/µs, and what is actually achieved at full power – which will be considerably less.

One major condition for slew rate, according to the book, is that it refers to the maximum frequency at which the amplifier can deliver a voltage corresponding to its rated power. At 40V/µs, and assuming a peak voltage of 31V, Self's amplifier would be expected to

Hear, Hear

I was very pleased that in *Research Notes* (*EW+WW*, August) Prof Engebretson's point was reported that, technically, we can create sophisticated hearing aids but the question is what do we want them to do?

The fact that so many hearing aids are consigned to the drawer is because they do not restore the hearing to original normality. The situation has now arisen where people can see wars, disasters, sport and so on from the other side of the world, as it happens, and yet a hard-of-hearing person cannot hear properly from the other side of the room.

I never cease to be ashamed of the electronics industry, of which I am a member, when a hearing impaired person makes this sort of comparison to me. I am pleased that the Technical Committee of the charity, "Hearing Concern" whose members are either hard of hearing senior members of the Institution of Electrical Engineers or of the Medical Profession, have taken a more positive line and take the attitude "We want hearing aids which restore our hearing to *normal*". People think that hearing loss can be compensated for by an amplifier, perhaps with some frequency shaping and maybe some amplitude compression. Sadly, this is far from the truth. Some other "distortion" takes place in the impaired cochlea (the part of the ear which converts sound vibration into neural signals for onwards transmission to the central nervous system).

This distortion has yet to be defined. But its effect is that sounds can be loud enough – frequently too loud for the impaired ear – yet the intelligibility of speech remains poor.

It is sometimes suggested that this discrimination deficit is due to loudness recruitment, the curious effect whereby the impaired ear is deaf to quieter sounds, but as they are made louder it seems to recover the loss until frequently, at high levels, sounds can be heard at the same loudness as a normal ear.

Prof Engebretson also made reference to this effect.

In experiments in the laboratory we corrected this problem and yet patient's speech discrimination was improved by very little, if at all. Unfortunately the technique corrected the loudness distortion with the addition of some harmonic or intermodulation distortion and noise

In the course of many years in otological work I have questioned hard of hearing people – including acoustically and technically experienced people – as to what it is that makes speech difficult for them to understand. But nothing they say gives us a real clue as to how to reverse the effect electronically in a hearing aid.

If you are hard of hearing and know the answer then for goodness sake don't keep it to yourself.

The advent of digital signal processing has given the search for the answer a new impetus. Kent University, Electronics Department, has designed chips which mean that all sorts of dsp facilities can be provided in a hearing aid and use less current than the equivalent analogue circuit. That is great for aid users, but what they, and we, and Prof Engebretson, are waiting for is to know how to program the thing to produce real benefits.

Edward Trinder
Institute of Laryngology and Otology
University College London

'Hearing Concern", 711 Armstrong Road, London W3 7JL.

Digital Old Master

With reference to "Painting Old Masters in pixels" (*Research Notes*, October, p.802), I understand there is no possibility of restoring the Mona Lisa to its original state, as opposed to conserving it without trying to turn the clock back.

Some say, too, that the subtlety of the Sistine Chapel ceiling has been obliterated in restoration. So in the long term, digital records will be the only valid ones, though masterpieces that have deteriorated grotesquely will obviously still have high value attached to them.

Bernard Jones
London

deliver full power at over 200kHz.

Of course it is deliberately rolled off within the audio bandwidth to cater for stability margin and to reduce SID. Since it is rolled off, the amplifier cannot slew at the rate suggested.

The point is that to maintain stability etc the amplifier is probably beginning to roll off at about 15kHz at full power into a 4Ω load. The net result is an amplifier that sounds comparatively dull in the treble, compared to mosfet designs.

The dynamic capacitive effects of bi-polars at high power always force the designer into compromises between a reasonable bandwidth, stability and distortion – not to mention the poor ruggedness of higher F_T types.

Naturally, lower power amplifiers will benefit in these respects. But as you go up in power and use multiple output devices paralleled-up, this capacitive effect is increased so that using faster and faster devices becomes mandatory. The ready-made solution to these problems is to use mosfets which, when used properly, have few vices, provided certain strict rules are observed.

In the February issue (*Letters*, 1994) Douglas Self, in reply to Ivor Brown's letter, suggests that mosfets are "so depressingly non-linear". This statement is contrary to text books which make a case for lower distortion, particularly in terms of the third-harmonic content, giving more favourable intermodulation and cross modulation figures than a bi-polar device. So for mosfet designs, less negative feedback is required, giving better stability margin for wider bandwidth. In fact Self said as much in his article *Sound mosfet design* (*EW+WW*, September, 1990) where he shows in his hybrid circuit a distortion figure of 0.05% open loop.

The linearity of mosfets is proportional to drain current. But if you set the quiescent current as Self has done in his hybrid circuit to 45mA, the mosfets will initially wander since the change-over from positive to negative temperature co-efficients occurs at around 80mA. Devices will gravitate towards this by temperature changes during driven conditions and this effect may be the cause of his non-linearity – it is why Hitachi suggests an 80mA quiescent minimum. Parallelling up devices improves linearity, as does increasing the quiescent current.

As to the charge that they will not drive complex loads, this is the sort of comment of someone who has not understood the relationship between the gate-source voltage to drain current, and who has placed a low voltage protection zener across the input then expects it (the mosfet) to deliver full current.

I find it disappointing that after 20 years, mosfets still seem to be regarded with deep suspicion by some parties due, it seems, to a lack of understanding. For my part, given all the difficulties, I cannot understand why designers are still playing with bi-polar audio output stages.

Some 15 years ago David Hafler brought out his *DH200* mosfet amplifier, a fully symmetrical design of 100W+/channel. That is still absolutely superb and is what I call blameless: updated with a better power supply it is probably unbeatable!

My last comment concerns the incessant need for the cognoscenti to express what we hear, in terms of sine waves, as an expression of hearing limitation. I know of few people who have actually taken the trouble to find out how the ear, and more importantly the brain/ear interface responds.

The ear has its own non-linearity which produces overtones.This effect and the fact that the brain is capable of interpreting extremely fast changing wavefronts is one of the major reasons why elderly people can detect tones well up into the audio bandwidth – top hats (cymbals), triangles, etc – and yet will have a severely limited response to sine waves. The way in which the brain perceives this information tends to make a mockery of those experts who say "you can't hear this or that because...": differences are already perceived between amplifiers despite their speed being supposedly more than adequate to encompass the highest audio signal. It would be nice if this anomaly in relation to the hearing mechanics could be cleared up once and for all by someone that really knows what he/she is talking about.

I prefer mosfet amplification because I have yet to hear a bi-polar that produces anything like the detailed clarity and openness that it affords – particularly using a 0.75in dome super-tweeter, power for power!

V J Hawtin
Middlesex

References
The Audio Handbook, Gordon J King; *Sound and Hearing*, Charles Gramer, and *The Instruments of Music*, Robert Donington.

D. Self replies

I regret that Mr Hawtin needs to reread the article a little more carefully.

I did not blame the input stage for unequal positive and negative slew-rates; the major cause of this effect – which I expounded on at some length – is capacitive feedthrough in the VAS current source, which reduces the current available during positive slew.

I was indeed driving the input stage into saturation, because this is the basis of slew-limiting, but was careful not to drive the VAS into clipping; the two constraints are quite different. There are no slow-recovery effects because there is no clipping, and the supply rail has no significant influence.

The power devices are indeed relatively slow compared with the small-signal stages, though this has no direct effect on slew-limiting, this is why many of the tests were done on a 'model' amplifier with a TO-92 *Class-A output stage, to minimise the possible complications. Possibly Mr Hawtin has overlooked these paragraphs.*

I do not assume that the output devices are matched at hf, or anywhere else. If the output stage had a truly massive beta-mismatch between the top and bottom halves then it might in theory be possible to have a slew limit that was slightly different above and below the zero axis, but this is not a realistic condition.

The amplifier does not have its closed-loop gain rolled-off at 15kHz, and it most certainly does not sound dull into a 4Ω load.

I stand by my previous statement that fets are non-linear by comparison with bipolar junction transistors. I would have thought that the wingspread gain diagrams (wingspread is a handy term borrowed from a slightly different field of technology) in Part 4 of Distortion in Power Amplifiers puts the matter beyond doubt – especially since they are backed up by practical measurements. I don't know what textbook he is talking about, but possibly it refers to rf mixers, where fets do have their advantages; however, this has nothing to do with power amps.

Continued over page

Feed-forward feedback

I read with interest Giovanni Stochino's article (*Audio design leaps forward*, *EW+WW*, October, pp. 818-824) on the application of feed-forward in Audio.

But I believe that the concept can be developed much more clearly and simply. First, we have his block diagram showing how power amplifier error is extracted, determined and isolated via a criterion network comprising α, γ and *SC*; and the formula $V_o = V_p' - V_a$.

Next, Mr Stochino introduces E_p and E_a (the error components in the main and auxiliary amplifiers), which is confusing as these are not shown in the figure. It is much simpler to assume that all errors, noise etc. are contained in the terms G_p and G_a, describing the two amplifiers. The expression is then developed straightforwardly by substitution:

$V_p = V_i.G_p$;
$V_p' = \lambda.V_p$;
$V_p' = \lambda.V_i.G_p$.

Similarly,

$V_a = V_e.G_a$;
$V_e = V_i.\alpha + V_p.\gamma$,
$V_e = V_i.\alpha + V_i.G_p.\gamma$.

thus:

$V_a = G_a.V_i(\alpha + \gamma.G_p)$.

The output V_o then becomes:

$V_o = V_i.G_p.\lambda - G_a.V_i.(\alpha + G_p.\gamma)$.

The system gain is:

$V_o/V_i = G_p.\lambda - G_a.\alpha - G_p.G_a.\gamma$.

Clearly, the system is independent of the main amplifier if:

$G_p.\lambda - G_p.G_a.\gamma = 0$.

This leads to the condition (assuming $\lambda = 1$)

$G_a.\gamma = 1$,

which is equivalent to Mr. Stochino's result, but is much easier to understand.

From this result it can also be concluded that in this case the system gain becomes:

$V_o/V_i = -G_a.\alpha$.

Furthermore, the condition $G_o = G_o' = G_o''$ is not necessary.

However, $\alpha = 1/\gamma$, which follows from Figs. 2 and 5.

By the way, I wonder whether a copy of the patent application would be available. What is it exactly that Mr.Stochino wishes to patent?

Jan Didden
Zevenbergen
The Netherlands.

As for Mr Hawtin's last comment, I am left breathless at his apparently total ignorance of the entire field of psychoacoustics, which has addressed itself to ear and brain for at least the past century, involving thousands of scientific workers.

There are more mis-statements in letter, but commenting on them all would be too tedious. I need hardly say that I am deeply grieved to learn that Mr Hawtin considers I am only playing with power amplifiers, and look forward in humility to the first article from him showing how it really should be done.

Sound question

I found Douglas Self's article *High speed audio power* (*EW+WW*, September, pp. 760-764) very interesting. But one question remains unanswered. Does Mr Self's modified design *sound* better or worse than his original design?

I think we should be told.

Phil Randal
Worcester

D.Self replies:
The answer to this question is very simple: it sounds exactly the same. I think the article made it clear that I was examining the slew-rate issues both from the commercial point of view, where faster is better, and also as an illustration of just how far astray simplistic theory can lead you. I certainly don't claim that the slew-rate increases described open new vistas of audio ecstasy.

Subjective bias

Concerning his view that: "Audio electronics circuits are built and sold to be listened to." Jerry Mead informs us (*Letters, EW+WW*, July 1994) that:" The research [he] cited... is [his] own" and that "It is culled from a career spanning 25 years of talking and listening to customers in the broadcasting, sound recording, live sound and music markets around the world". This is not scientific research, this is market research.

Unfortunately for Jerry the great bulk of data collected through scientific research does not support his hypotheses.

Jerry says his 'research' is "culled from a career...". But what is of great importance is what he has culled out. The fact that proof of his hypotheses is so elusive should indicate to him that he might be on the wrong track, especially when there already exists such a large body of research that proves the unreliability of his methods.

An essential part of scientific research is that data and procedures (experimental, computational, etc) be open to scrutiny. It keeps researchers honest and accurate.

During my career in electronics and audio (which spans more than 20 years) I have never seen nor heard of 'subjectivist' views being vindicated by rigorous scientific research. I for one would like to see it if exists. Until then I will be guided by the information that already exists and will disregard the garbage put up by the mostly ignorant 'subjectivists'.

Jerry suggests that because audio electronics is for listening to we should use our ears to test it. As any competent audio designer knows, human hearing is far too unreliable, non-uniform and too easily fooled to be used often for testing purposes. This is why 'objectivists' place more faith in simple tests for thd, imd, frequency response, etc.

I am sure we would all prefer to be able to test our designs just by listening to them; unfortunately there are too many extraneous influences (including personal prejudice) to make this practical and reliable.

Jerry has courageously offered to take part in some independent listening tests, and he has laid out his requirements for his part in these tests. He wishes to choose his own source material. However this would show us more about his musical preferences than the performance of the equipment. It has been said that choosing program material represents one of the most obvious opportunities for prejudicing listening test results.

Jerry wishes also to choose his own listening levels. But if he is asking to be given a volume control to change the levels while testing, that would invalidate the results.

Unless the frequency response of

the system is maintained to a small fraction of a decibel, and the listening level does not change more than 0.5dB (possibly even less) then audible differences will emerge, due to the nature of our hearing. This fact has been known for over 20 years and is usually forgotten or ignored by subjectivists, and is the reason many amplifiers have loudness and tone controls. You can hear the effect simply by turning the volume up slowly from zero and listening as the music becomes brighter and fuller.

I should also like to respond to Charles Friell and R L Tufft (*Letters*, *EW+WW*, August, 1994).

Friell is correct in thinking "...that there is rather more involved than logic and mere engineering principles when discussing sound reproduction".

Like him I am not convinced of the need for amplifier distortion to be vanishingly small. Scrutinising data given for loudspeakers and microphones, especially frequency response graphs and distortion figures, shows that they are indeed quite inferior to most amps and pre-amps. Their performance weaknesses have long been known and, along with the listening room and human hearing, contributes the greatest limitation of most audio systems.

In response to R L Tufft's call for suppliers and manufacturers to furnish us with information about their cables, proving that some non-existent thing does not exist can be extremely difficult. I would warn everybody to be wary of any such information. This is like asking a Christian about the existence of God.

Finally, let me congratulate Doug Self for winning the *EW+WW*/HP writer's award. His "Distortion in Audio Power Amplifiers" is a solid contribution to the practice of power amp design and has a legitimate place beside Peter Baxandall's great "Audio Power Amplifier Design" from the 1970s.

Phil Denniss
Dept of Plasma Physics
University of Sydney

I agree – DS.

Musical components

I welcome the common sense attitude of Stephen Merrick (*Letters*, July) who says: "The only purpose for audio amplifiers is the reproduction of music".

There is no absolute rating of amplifier performance, only a consensus. But that consensus does show a statistical bias towards products which are designed with the enjoyment of music foremost.

The engineer must start by listening to amplifiers, identifying those that make the most positive contribution to the music and seeking out the design features that they share.

On this basis I would like to make some comments on Douglas Self's design published in February *EW+WW*. (To meet Douglas' challenge that critics should offer their own rival amplifiers, I would point to my Apex and Virtuoso designs, published in the UK around five years ago. These were designed to reproduce music, although one version of the Virtuoso was measured by a leading technical reviewer at 0.001% distortion at 2/3 power at 1kHz.)

If the standard low-cost metal film resistors are replaced with semi precision types (eg Holco *H8*) the sound quality would be noticeably better. Further upgrading with ultra precision types such as Vishay *VSRJ* bulk foil would give quite a surprising sonic gain, but at high cost. (Try R_8 and R_9 first.)

Try replacing R_1 with a polyester or polypropylene capacitor, or place a similar capacitor in parallel with R_2. I would prefer to design the feedback resistor to be high enough to use a polypropylene capacitor for C_2, if this can be done without creating a high dc offset at the output. I do not recommend a dc servo – they sound worse than a polyester capacitor. Remove C_4 and C_5 unless this affects amplifier stability. The amplifier is bound to sound better without.

There are audio transistors available which have smoother gain *versus* current graphs and can improve sound quality. The Darlington output arrangement is the best, but some readers might like to compare its sound with a single Darlington pair such as *MJ11015/6*. Insulating pads with lower thermal resistance between transistors and heat sink can improve sound quality slightly as well as reliability.

Constant current sources is good design. Removing C_{11} should degrade the sound slightly, but try replacing R_{21} with a constant current diode (*J507* or *J508*).

Cascode transistors above the collectors of Tr_2, Tr_3 and Tr_4, and separate power supplies for the low and high current sections offer further scope for improving sound quality.

On the subject of cables, previous correspondents are right when they imply that both copper and silver degrade sound quality, but silver degrades less than copper. To test the validity of this statement, obtain a suitable collection of high quality cables and listen to them.

In fact every single component in the audio chain degrades the signal. The skill of good design in both amplifiers and cables is finding what degrades it the least.

Graham Nalty
Audiokits Precision Components
Derby

Before buying expensive cables and components, I suggest you get someone to demonstrate the claimed improvements to you – Ed.

Magic numbers

I have received several letters concerning the golden mean, golden section or golden ratio mentioned in my article "Magic numbers in electronics" (September, pp.730-733).

G C A of Dorset enclosed a photocopy of a brief entry in *Colliers Encyclopaedia* which gives the standard geometrical construction and states that the latter is dealt with in Euclid Book VI, *Proposition 30*. This led the Pythagoreans to a realisation of the existence and significance of "incommensurables" or surds.

A C of Newcastle-upon-Tyne wrote to say that the Greek mathematician Eudoxus was the first man to investigate the golden mean, at first empirically and then more formally. His letter states that the Greek letter *phi* came to be associated with the golden mean after the artist Phidias, who used it extensively to proportion his sculptures. A C added that the ratio of successive numbers in the Fibonacci series approaches the golden ratio as the series progresses.

A H D of Hampshire wrote to say he was intrigued by the connection of the golden mean to the amplitude response of filters, but says it is not called *K* (or for that matter *phi*, but *tau*, being 'the Fibonacci number'. He kindly encloses a photocopy of a four page article which appeared in an issue of *Scientific American* of about a quarter of a century ago, concerning Fibonacci and his series, the first two terms of which are both 1, with each subsequent term being the sum of the previous two. It gives an explicit formula for the *n*th term, based upon *K*, *phi*, *tau* or call it what you will, and a long list of intriguing properties of the series.

Ian Hickman
Hampshire

Noisy response

It was very kind of Ian Hickman to go to such high-powered lengths in his reply to my letter (*Letters*, August, November).

To consider just one simple point though: why do I wonder what $^1/_f$ noise is? From books, I read that it is the same as '"Schottky noise" or "flicker noise" which, as M G Scroggie states in his book *Foundations of wireless and electronics* occurs at very low frequencies (a few kHz) and is called $^1/_f$ noise because it is inversely proportional to frequency.

So we seem to have an equation: $I_n = K/f$ where I_n is the noise current in a given low frequency circuit, f its operating frequency, and K a constant.

How then, I ask, is this equation developed, derived and demonstrated? I am unaware of any studies of this theory. Perhaps Ian Hickman could enlighten me.

Peter Dawe
Oxford

No conspiracy

As a practising design engineer with a physics background, I have often been entertained by the more novel views on scientific orthodoxy propounded by *EW+WW*. But there seems to be a tendency towards charges of conspiracy (eg A Goldberg, *Letters*, July 1994), with the implication that refusal to accept any outlandish idea is a sign of bigotry or self interest.

I would like to reply to those charges using the words of Richard Dawkins, whose books are a model to anybody with a valid case to argue:

"There have been times in the history of science when the whole of orthodox science has been rightly thrown over because of a single awkward fact. It would be arrogant to assert that such overthrows will never happen again. But we naturally, and rightly, demand a higher standard of authentification before accepting a fact that would turn a major and successful scientific edifice upside down, than before accepting a fact which, even if surprising, is readily accommodated by existing science". The Blind Watchmaker

Some readers are confusing this widely held establishment view with something altogether more sinister and, to my mind, non-existent. I suggest that those with an interest in heterodox ideas attempt a more coherent statement of their position, supported by objective evidence and repeatable experiment, and bearing the above passage in mind. If they do one tenth as good a job as Dr. Dawkins they will gain my ear, and my respect, although I suspect not my conversion.

Pete Davis
Birmingham ■

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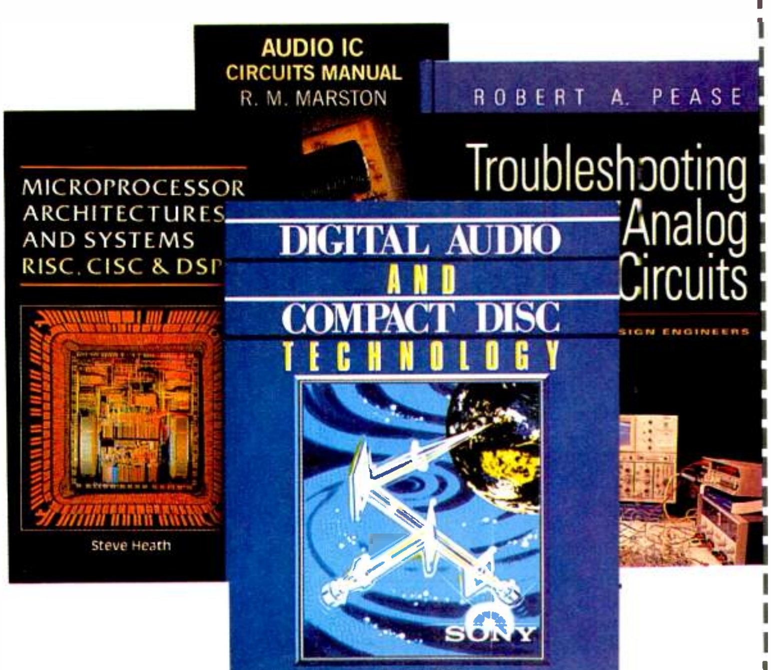

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Digital synthesizer generates quadrature-matched sines and cosines

Functions performed by Raytheon's *TMC2340* are waveform synthesis, modulation, and demodulation. Designed to drive a d-to-a converter, the synthesizer generates quadrature-matched quadrature-matched pairs of 16-bit sine and cosine waves.

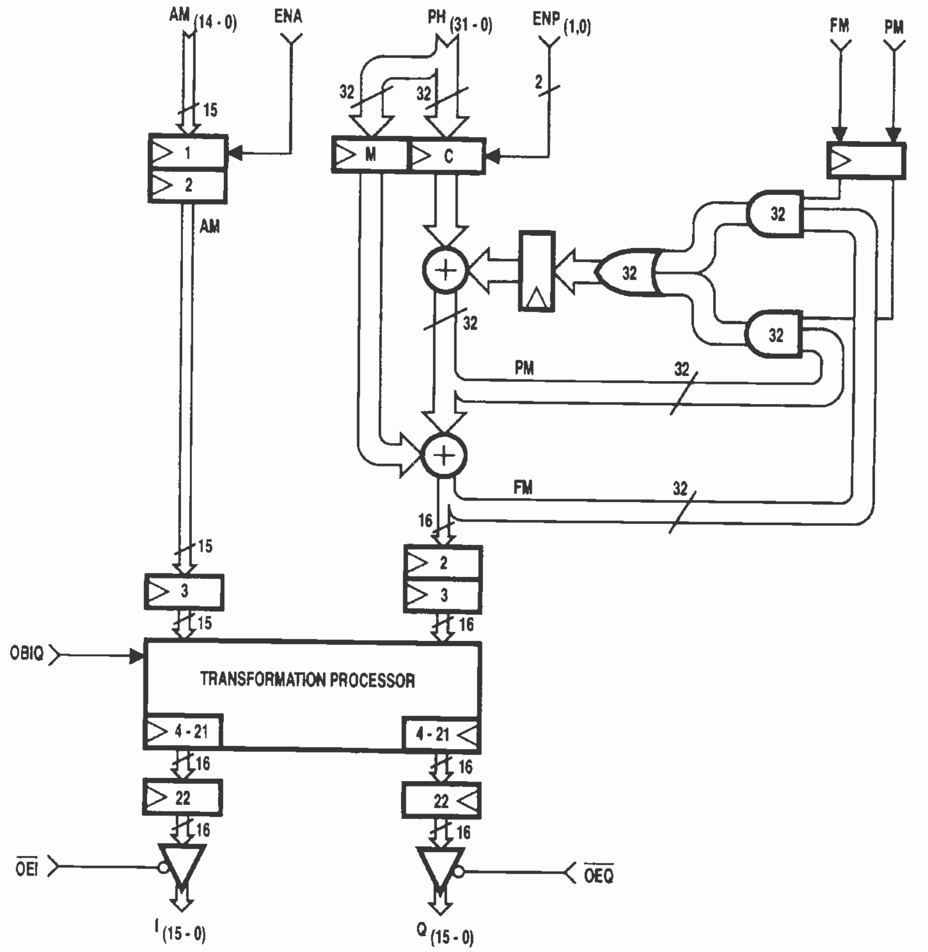
As described in the *TMC2340* data sheet, inputs needed are a a ttl clock signal and user-selected 15-bit amplitude and 32-bit phase increment values. Output is in 16-bit offset binary format, and the device can operate at up to 25 mega-operations per second.

These waveforms are easily phase or frequency-modulated on-chip, and the amplitude input facilitates gain adjustment or amplitude modulation. Digital output frequencies are restricted only by the Nyquist limit of clock rate/2, with frequency resolution of 0.006Hz at the guaranteed maximum 25MHz clock rate.

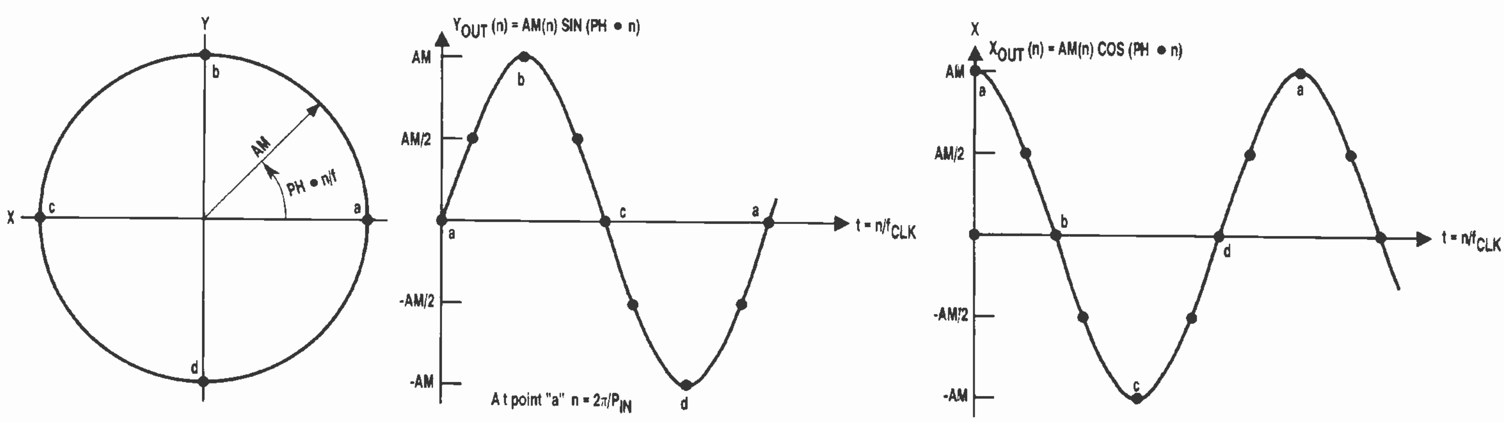
A new data-word pair is available at the output every clock cycle. All input and output data ports are registered, with a user-configurable phase accumulator structure and input-clock enables to simplify interfacing. The phase data range over a full 2π radians. All signals are ttl compatible.

Polar data – i.e. phase and magnitude – is converted into rectangular, cartesian, format. The first transformed result is available at the outputs 22 clock cycles after startup, with new output data available every 40ns.

Input clock enables simplify system bus connections. Input ports accept 15-bit amplitude and 32-bit phase data. The output ports produce 16-bit rectangular data words in either 16-bit offset binary or 15-bit unsigned magnitude format.



Elements of Raytheon's 25Msamples/s digital synthesizer for producing waveform sine or cosine pairs with accurately matched quadrature.



Input to output relationship for sinusoid generation.

Phase/amplitude to sine/cosine conversion

The *TMC2340* performs a coordinate-space transformation according to the familiar trigonometic relationships shown. With constant amplitude and phase increment values and either fm or 'pm high', the *TMC2340* outputs a series of complex number pairs. These represent the horizontal and vertical projections of a vector rotating about the origin, i.e., a cosine wave and a sine wave.

Via the device's internal phase accumulators, it is easy to generate high-accuracy digital quadrature sinusoidal waveforms with minimal support. The 32-bit data path ensures negligible cumulative error in most applications. Accuracy of the transform is limited only by the truncation of the result to 16 bits prior to the transform processor and the +1lsb maximum error of the transform algorithm.

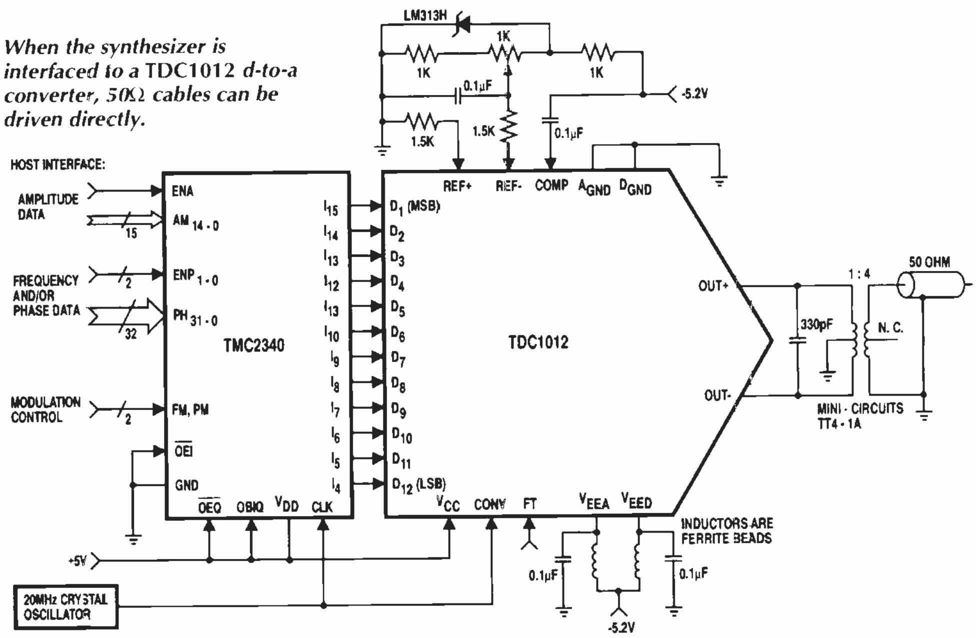
Amplitude modulation is performed simply by varying the amplitude input. Either frequency or phase modulation can be realized by configuring the synthesizer as shown in the two configuration diagrams.

Connection of the *TMC2340* to a *TDC1012* d-to-a converter is straightforward. As shown in the circuit diagram, the converter data lines are connected to either the I or Q outputs. Both outputs may be used, with two *TDC1012*s for quadrature synthesis.

Full design details and equations for waveform generation/modulation are also included in the data sheet, as is a description of the control circuitry.

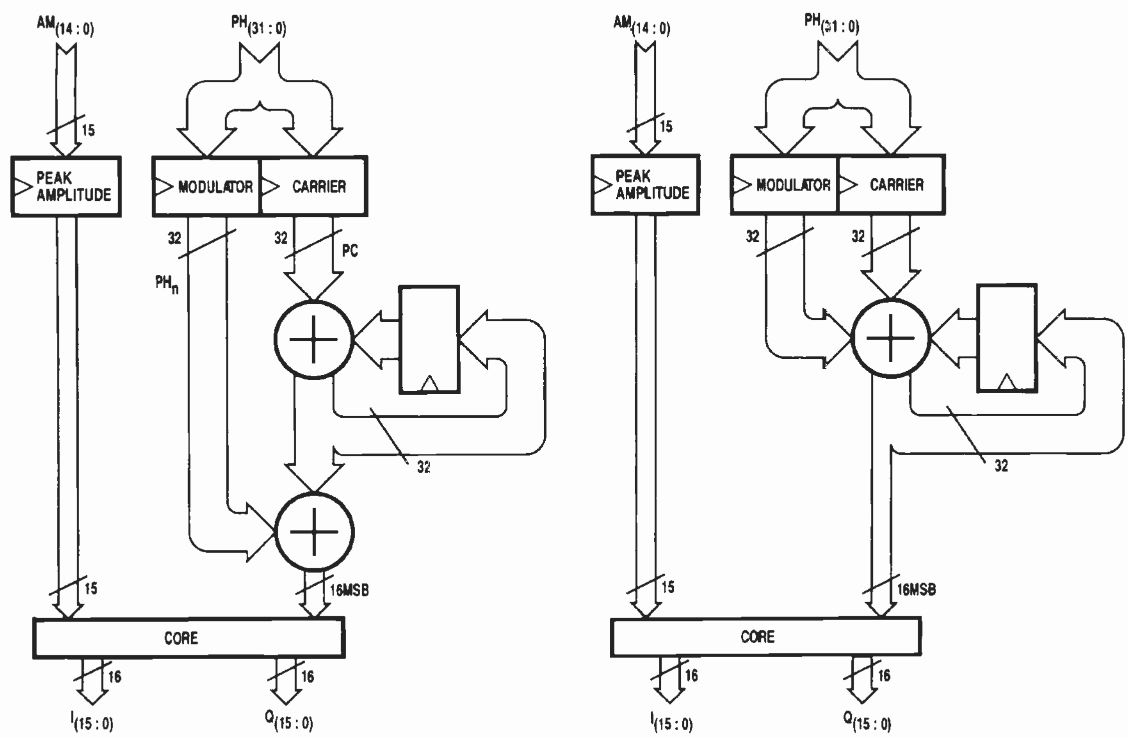
***Raytheon**, Ambar Components, 17 Thame Park Road, Thame, Oxfordshire OX9 3XD. Tel. 0844 261144, fax 0844 261789.*

When the synthesizer is interfaced to a** TDC1012 **d-to-a converter, 50Ω cables can be driven directly.



Note: To use two TDC1012's in quadrature, connect second TDC1012 to Q_{15} (MSB) to Q_4 and ground $\overline{OEQ}$.

Configuring the** TMC2340 **for phase modulation, left, and frequency modulation, right.



Current-conveyor ICs for high-end audio

Two ICs, each consisting of an accurate current conveyor, a current mirror, and two unity gain buffer amplifiers, are described in a preliminary data sheet from Phototronics.

Designed specifically for high-quality audio applications, these ICs are the *PA630* and *PA630A*. The devices are identical apart from two additional pins on the *PA630A*, **Fig. 1**. These are used to interface with two external junction fets to enhance conveyor performance.

The devices are fabricated on an advanced complementary bipolar process that produces npn and pnp transistors with equally high bandwidth and current gain. Although functionally equivalent to the schematic shown, the circuitry of the mirror and current conveyor comprises a novel connection of Wilson current mirrors. It also incorporates an emitter de-generation compensation scheme to optimise the transient response and stability of these mirrors, and a novel output mirror arrangement to enhance output impedance.

In professional audio, virtual ground inputs can be implemented without the global negative feedback required by most other circuits such as op-amps. The ICs are therefore inherently free of dynamically induced distortion mechanisms such as transient-intermodulation and slewing induced distortion.

Bandwidth of the current conveyor is 18MHz while bandwidth of the buffer amplifier is 50MHz. Current conveyor distortion is rated at 0.02%.

Each unity gain buffer amplifier consists of four emitter followers and two current sources as shown in **Fig. 2**. The quiescent operating point of each can be set independently via an external resistor, R_{set}. Pins 1 and 3 may be left open to save current if this buffer is not required, but since the other buffer provides internal bias it must be powered up with an output stage current of no less than one tenth of the conveyor quiescent current.

As the current sources are actually Wilson mirrors, $2V_{be}$, and the output devices are five times larger than the input transistors, the quiescent current in the output stage can be calculated as $5[V_{cc}+V_{ee}-2.8V]/R_{set}$.

Additional pins on the *PA630A* can be used to interface with two external junction fets which buffer the output, as shown in **Fig. 3**. This arrangement provides extremely high output impedance, improved accuracy, and lower distortion.

The precision rectifier circuit of **Fig. 4** provides glitch free performance up to 1MHz due to the fact that the diodes are current driven and there is no feedback. In contrast, even fast op-amps (30V/μs, 20MHz) will produce distortion spikes well below 100kHz when used as precision

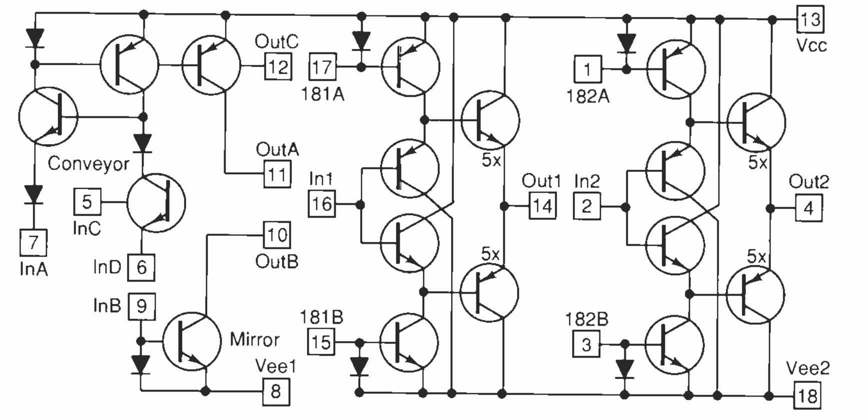


Fig. 1. Internal equivalent of the **PA630A** *current-conveyor IC for professional audio. Since the device can be implemented without global feedback, two dynamically-induced distortion mechanisms are eliminated.*

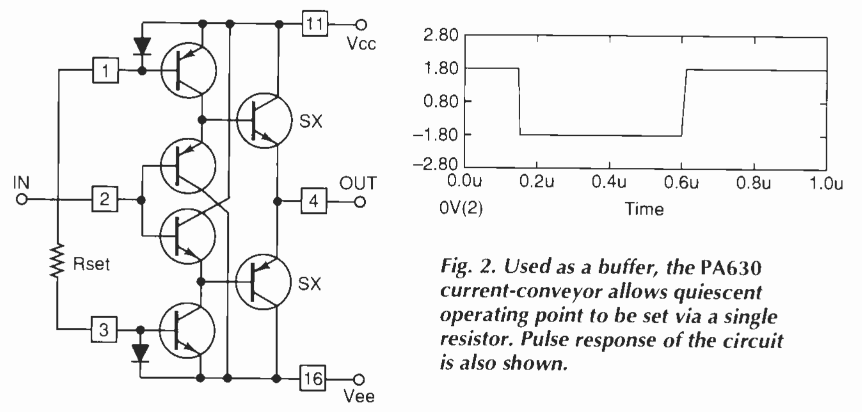


Fig. 2. Used as a buffer, the **PA630** *current-conveyor allows quiescent operating point to be set via a single resistor. Pulse response of the circuit is also shown.*

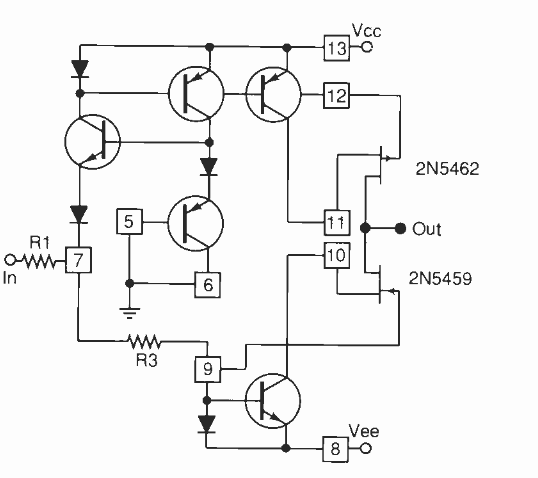


Fig. 3. Two additional pins are provided on the **PA630A** *to allow junction fets to be connected. This arrangement provides high output impedance, improved accuracy and lower distortion.*

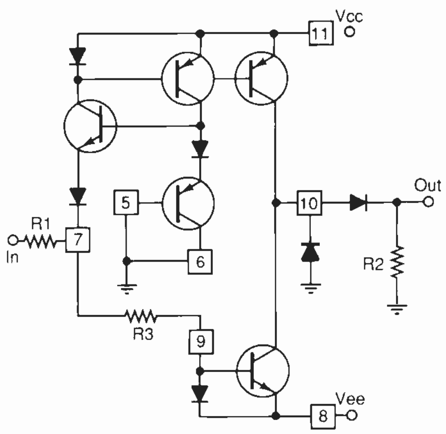


Fig. 4. Acting as a precision rectifier, the current conveyor provides glitch-free performance up to 1MHz.

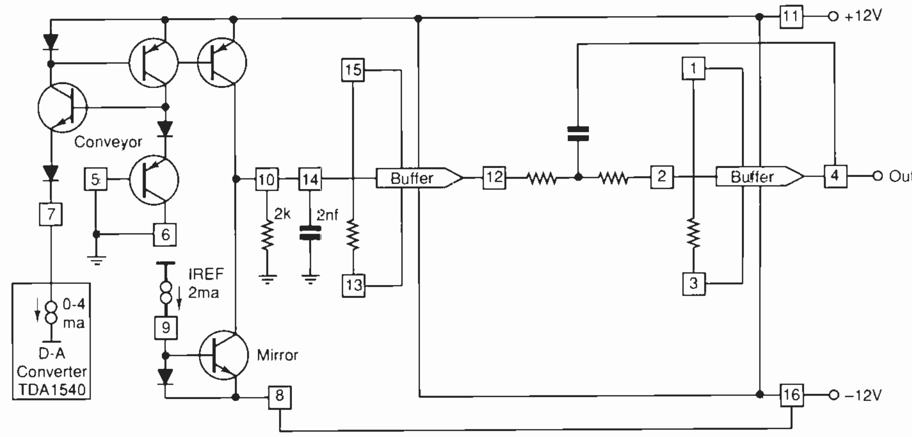


Fig. 5. CD player analogue output stage using a current conveyor IC designed especially for audio. The IC implements the virtual ground needed by the data converter, in addition to the second-order linear-phase filter.

rectifiers, due to the slew rate limitations that result from having the diodes inside the feedback loop.

Figure 5 shows the *PA630* in the analogue section of compact-disc player. The IC implements the virtual ground required by the d-to-a converter, as well as the second order linear phase filter. This configuration completely eliminates the potential for the generation of dynamically induced distortion.

***Phototronics Co.**, Box 977, Manotick, Ontario, Canada K0A 2N0. Tel. 613 692 2247, fax 613 692 2605.*

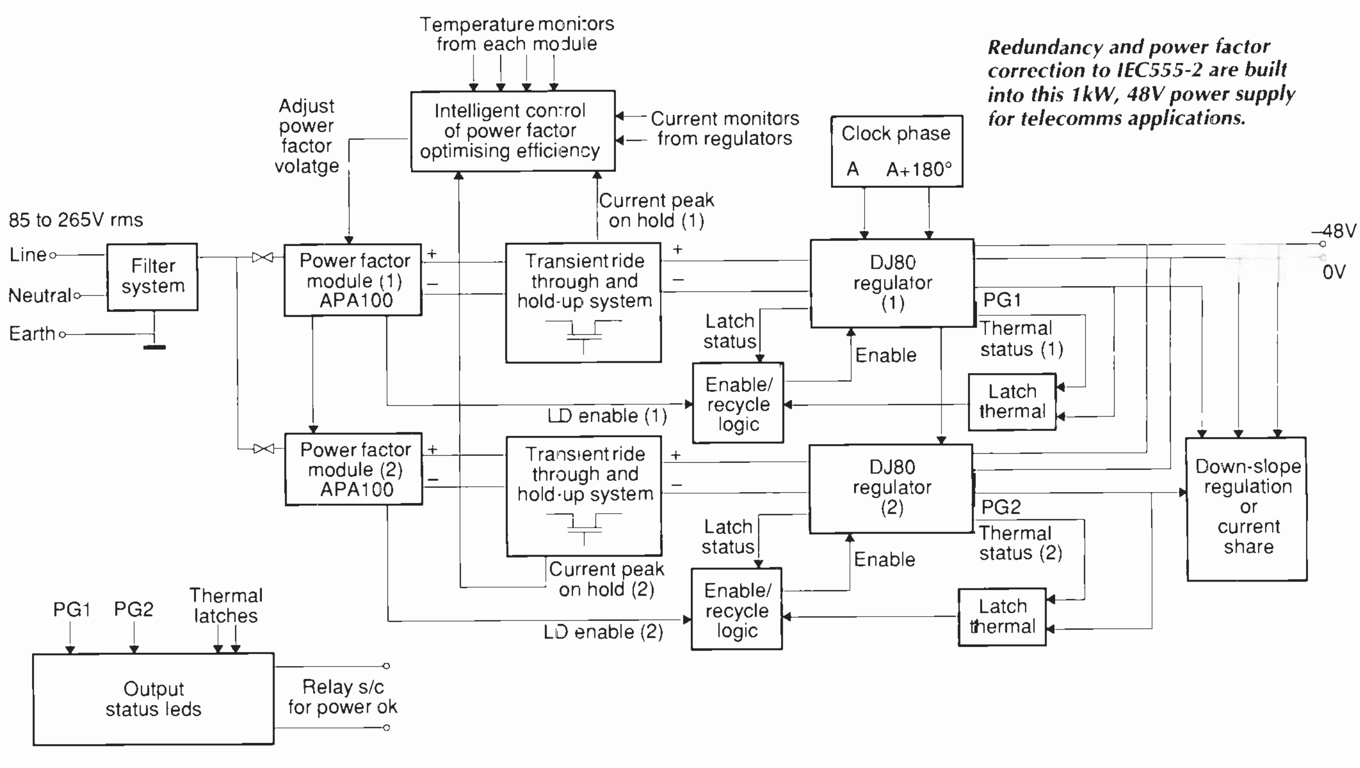
1kW, 48V supply with power factor correction

Three diagrams and few bullet points outline a 1kW modular power supply in Astec's *PFC.PM4 AMPSS Application Note*. Intended primarily for telecomms applications, the 48V supply features power-factor correction and n+1 redundancy ñ a method of increasing power supply reliability by parallelling a 'redundant' regulator module.

Using the modules shown, power-factor correction is to IEC555-2. The scheme is said to offer protection against severe line transients to 450V peak for 20ms, and low output ripple due to module interleaving. Fet switches are used to isolate the *DJ80* regulator modules from transients.

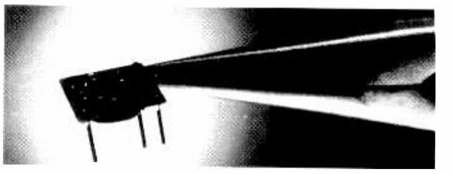
In addition to the overall system diagram shown, the note contains more detailed schemes for the power factor correction and transient protection sections.

***Astec Standard Power**, High Street Wollaston, Stourbridge, West Midlands DY8 4PG. Tel. 01384 440 044, fax 01384 440 777.*



Redundancy and power factor correction to IEC555-2 are built into this 1kW, 48V power supply for telecomms applications.

Transmitter-receiver pair for *remote control*



Measuring 13 by 13mm, this low-power transmitter for remote control send to distances over 50m under the right conditions.

Designing a uhf transmitter to meet the EMC requirements of the DTI's MPT1340 standard is notoriously difficult. In addition to a having low-distortion circuitry, the transmitter needs a very carefully designed pcb, enclosure and antenna.

Much of the design effort is removed by using a ready-made transmitter module like the *LPRSTX-418* transmitter from Low Power Radio Solutions. Provided the module's applications guidelines are followed, it should be possible to make a complete transmitter capable of gaining approval at the first attempt – saving considerable design effort, time and costs.

When designing a uhf remote-control link, cleaning up the transmitter signal enough to obtain type approval can be an expensive and time-consuming business. One solution is to use ready-made modules.

Measuring 13 by 13mm, the *LPRSTX-418* is designed for applications such as radio keyfobs and wireless security systems. The module incorporates a saw filter and is intended for on-off signalling, via amplitude modulation. It is optimised for pcb loop antennas, and is compatible with most encoding integrated circuits operating from 5V to 12V. Typical applications include key fob designs, car alarm 'blippers', garage door openers and lighting controllers.

Note that the transmitter module is not type approved and any equipment that incorporates it will need type approval. A 433.92MHz version for Europe will be available soon.

The *LPRSTX-418* integrates the main rf components required for a low power radio transmitter. The designer only needs to interface a tuned loop antenna and set the drive level to the module from the chosen encoding integrated circuit.

Antenna matching has been optimised so a transmitter can produce the maximum permitted output of –6dBm on a compact loop antenna without drawing excessive currents or exceeding the spurious emission limits.

Figure 1 shows a typical circuit for a design using a loop antenna. A 1 to 5pF ceramic trimmer is used to tune tine loop for maximum output. **Figure 2** shows a typical layout for a keyfob transmitter. The dimensions are not critical but excessively small or large loops should be avoided as these will affect antenna matching and efficiency.

When using a printed antenna on the pcb you should specify a good quality fibreglass material. Lower cost materials such as srbp will cause excessive losses at uhf.

Modulation drive level

Output power of the *LPRSTX* is controlled by varying the resistor in series with the modulation input to the module. Power and spurious emission increase with rising drive current.

The table below gives typical values for a loop antenna The actual value for a particular antenna and supply voltage will need to be determined by experiment. It is advisable to aim for effective-radiated-power levels 3dB or more below the limits set in MPT1340 to allow for measurement errors.

Drive voltage	5V	8V	11V
Drive resistor	150kΩ	330kΩ	470kΩ

The actual level required will depend on the antenna efficiency. Increasing the drive in order to overcome excessive antenna losses will also increase the spurious emissions.

Circuit-board layout and decoupling

To achieve the results described above, good layout is needed. The loop antenna should be free of any components or tracks except for the module and tuning capacitor. Avoid routeing tracks under the module. Failure to observe these precautions may cause excessive spurious emissions.

Encoding and other circuits should be mounted over a ground plane to screen them from the rf. The ground plane should not extend under the module. Decoupling has been provided on the module which is adequate for most applications. If extra decoupling is needed use ceramic capacitors of 220 to 1000pF from the supply and modulation pins to ground.

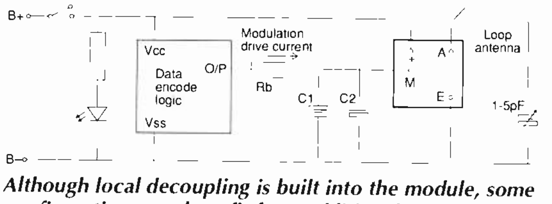
Hybrid receiver module

Operating on 418MHz, the RF290A-5S am hybrid receiver module is fully compatible with the LPRSTX-418 transmitter. Receiver front-end and logic circuits have separate supply pins for easy interfacing to decoding logic operating from a different supply voltage.

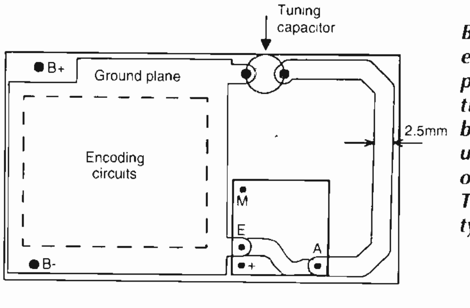
Measuring 16.5 by 36.1mm, the superregenerative receiver is a single-in-line module designed for direct pcb mounting. Pin pitching is a standard 2.54mm.

More details on page 1051...

LPRS transmitter/receiver module pair for under £12. For details of the special offer, see page 1051.



Although local decoupling is built into the module, some configurations can benefit from additional ceramic capacitors from ground to the transmitter's positive supply and modulation inputs.



By adding simple encoding logic and a power supply, the transmitter module can be used to emit a unique code for, say, opening a garage door. Tuning capacitance is typically 1-5pF.

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

500ksample/s, 12-bit a-to-d. Samples of the Linear Technology *LTC1278* analogue-to-digital converter are now available. The 12-bit device samples at 500ksample/s and consumes only 75mW from a single 5V supply. It features a sinad ratio of 70dB and exhibits a thd of 74dB at its Nyquist frequency. On chip are a 300ns sample-and-hold, a precision reference and a clock internally trimmed for a 1.6µs conversion time, which synchronises to each sample command. Micro Call Ltd. Tel., 0844 261939; fax, 0844 261678.

Palette D-to-A converters. *RGB528/514/513* are three palette d-to-a converters by IBM, offering the kind of video performance normally associated with workstations and required for applications in desk-top publishing and multimedia use.

The 32-bit pixel port uses a 2Mbyte video memory to deliver up to 16.8 million colours on screens with up to 1600 by 1280 resolution, the *513*'s clock generator operating at 170MHz.

Working at 250MHz, the *528* has a 128-bit pixel port and is meant for workstation graphics. Its two programmable clocks allow the graphics controller and pixel frequencies to be optimised. This is claimed to be the first palette d-to-a to enable flat panels to display graphics without sacrificing performance.

Main claim of the *514* is its space saving, being contained in a 144-pin quad flat pack. It has a 64-bit pixel path, packed 24-bit pixels, triple monotonic 8-bit d-to-a converters and a 220MHz clock. Blue Micro Electronics. Tel., 0604 603310; fax, 0604 603320.

5V, 10-bit d-to-a converters. Maxim's *MAX503/504/515* are claimed to be the first low-power, voltage-output, 10-bit d-to-a converters to operate from a single 5V supply; *503* and *504* will also accept ±5V. *503/504* have internal references and either unipolar or bipolar outputs, while *515* uses an external reference and draws only 150µA from the supply. All three have compatible upgrades to 12 bits (*MAX530/531/539*) and will work as four-quadrant multipliers. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

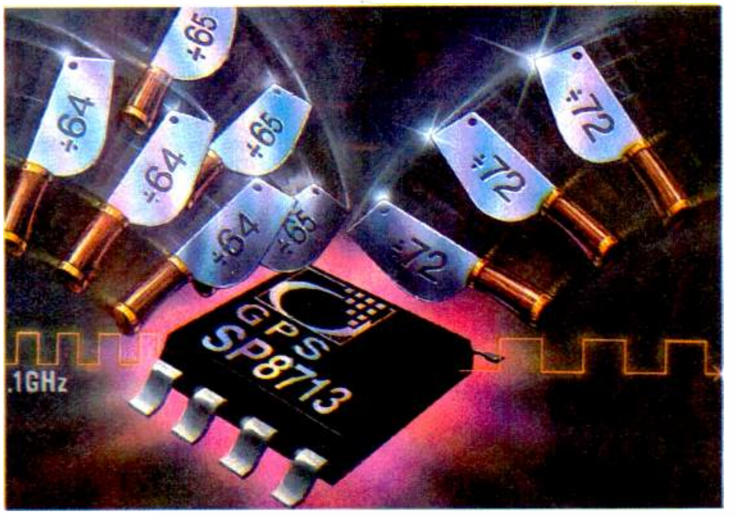
'Fastest' flash converter. Analogue-to-digital converters in the Signal Processing Technologies' *SPT7750/55* family have guaranteed speeds up to 750Msample/s Unusually, each comparator is provided with a preamplifier to buffer the capacitive comparator inputs and to reduce feedback of switching current into the inputs and reference ladder. With a 50MHz input, s:n ratio is 47dB and thd –46dB. At 250MHz, total dynamic error is –38dB Ambar Cascom Ltd. Tel., 0296 434141; fax, 0296 29670.

Discrete active devices

Zero-bias Schottkys. Low-cost, surface-mounted, zero-bias Schottkys by H-P are intended for use in passive and active identification and tagging in the current standard frequency bands of 915MHz, 2.45GHz and 5.8GHz. They are used to detect radiated energy and either radiate harmonics back to the interrogator or rectify the energy to drive other circuitry. Characteristics of these p-type diodes include 100mV forward voltage, 0.3pF at 1GHz, tangential sensitivity –57dB at 2.45GHz and voltage sensitivity 30mV/µW at 2.45GHz. Hewlett-Packard Ltd. Tel., 0344 362277; fax, 0344 362269.

Digital signal processors

Video-CD chipset. TI's *TMS320AV220* video full-motion cd chipset consists of three ICs: an MPEG-1 decoder; MPEG audio decoder; and NTSC encoder. Together with a 4Mb dram and a d-to-a converter for audio, the chipset turns a cd player into a video-CD player. It decodes, synchronises and decompresses audio and video data encoded to the MPEG standard, providing an NTSC video output and audio. Texas Instruments. Tel., 0234 270111; fax, 0234 223533.



MPEG-2 decompression First in a family of MPEG-2 chips from Blue Micro is a single-chip decoder for the full-motion digital video market, processing the compressed bit streams to obtain high-quality video. An error-concealment feature fully exploits the good data available. Multiple dram configurations store coded data and partially decoded and reference pictures are supported. Filtering and display control allow aspect-ratio changes. Blue Micro Electronics. Tel., 0604 603310; fax, 0604 603320.

Linear integrated circuits

Dual/quad op-amps. *LT1361/62* dual/quad op-amps by Linear Technology are 50MHz gain/bandwidth devices with slew rates of 800V/µs and draw only 4mA per amplifier. *LT1364/65*, also dual/quad devices, are 70MHz types slewing at 1000V/µs and taking 6mA. These devices are in LT's *C-Load* family and will drive 10,000pF loads without oscillating. Micro Call Ltd. Tel., 0844 261939; fax, 0844 261678.

Single-supply uhf amplifier. Anglia's *RF2103* is a medium-power linear amplifier working from a single positive 3-6.5V supply as the final stage in uhf transmitters in the 450-1000MHz range or as an exciter in higher-power equipment. Peak output is 800mW in cw or 400mW average for a two-tone input on a 6.5V supply.

Triple-modulus frequency divider. Three division ratios in GPS' *SP8713* prescaler instead of the usual two solve the difficulties of gapless band coverage. These are caused by having to lower the synthesiser loop division ratio to cope with increasing system comparison frequencies as fractional *n* techniques come into use to improve locking performance. With dual-modulus prescalers, the only answer is to use low division ratios, which require high input frequencies; the trouble is that fast, low-ratio prescalers are not available. Using the *SP8713* gives a low overall ratio but relatively high individual ratios in the three dividers. Used with the GPS *NJ88C50* synthesiser, the *SP8713* forms a fractional-*n* PLL on two chips.

Switchable ratios are 64, 65 and 72 and the device works at up to 1.1GHz. Current drawn is 4.7mA maximum from 2.7V-5.5V and there is a pin-selected economy mode in which only 20µA is needed. Times for power-up and power-down are about 2ms. GEC Plessey Semiconductors Ltd. Tel., 0793 518510; fax, 0793 518582.

PCMCIA data acquisition. Adept Scientific has a data acquisition interface to connect to a mobile computer's pcmcia slot. The *PCM-DAS08* device connects to transducers to measure current, temperature, pressure, flow or other process variables when used with software such as Labtech Notebook or VisSim *DACQ* for Windows, which effectively converts a PC into a measuring instrument.

PCM-DAS08 card and socket services installation software handles both addressing and the link between computer and card, so that cards can be changed quickly when required. The device is i/o-mapped and is controlled by standard software. ComputerBoards Universal Library, which is provided, supports QuickBasic, VisualBasic for dos, C and Turbo Pascal. Adept Scientific Micro Systems Ltd. Tel., 0462 480055; fax, 0462 480213.

Maximum cw output at 3V is 135mW. Total gain is 25dB, depending on output matching. Anglia Microwaves Ltd. Tel., 0277 630000; fax, 0277 631111.

Logic building blocks

Interface logic. TI's *Widebus+* bus interface logic provides a 36-bit interface and is sufficiently integrated to eliminate the need for discrete components. With 5ns propagation delay and low consumption and noise, the devices possess a range of data communication functions – transparent, registered, latched, clocked and universal bus functions. Bus hold allows the devices to hold the last state of the bus, so pull-ups are not needed. Pin layouts allow the devices to be mounted on either side of a board to share the same control lines; non-powered boards may be plugged into a powered backplane. Texas Instruments. Tel., 0234 270111; fax, 0234 223459.

Memory chips

16M drams. Volume shipments of second-generation 16M drams organised as 1M by 16 constitute a first for NEC. *μPD4216160* and *μPD4218160* are made in 0.45μm technology and are available with access times of 70ns or 80ns and in 3.3V or 5V versions. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

512K-by-8 Superflash. The *Silicon Storage Technology 28SF040* is the second member of the *SuperFlash* family, which combines the reliability of an eeprom with a small-sized cell. It is compatible with standard eprom, flash eprom and eeprom and includes sector erasability. Time to erase and rewrite a page is under 10ms. Ambar Components Ltd. Tel., 0844 261144; fax, 0844 261789.

Microprocessors and controllers

Datacomms processor. IDT's *R3071* and *R3071E* are mips-based data communications risc microprocessors operating at speeds up to 50MHz with up to 20KB total cache; the *3071E* also has a memory-management unit. Both devices are compatible with other members of the risc controller family, so that a single design can be up-graded without a major redesign. Versions available operate at 33MHz, 40MHz and 50MHz. Micro Call Ltd. Tel., 0844 261939; fax, 0844 261678.

Low-power controller. Epson has two 4-bit, low-power microcontrollers, meant for use in metering and remote control. Core cpu used in the devices is the *SMC6200A* and on-chip peripherals include rom, ram, lcd driver, supply voltage detector and a-to-d converter. *SMC621C* has an 108-instruction set and a 32kHz clock, with 455kHz or 1MHz as mask options. *SMC6292* has 100 instructions and 32kHz/1MHz clock. Rom is 2K by 12bits and ram 128 by 4bits. It also has an 8-bit i/o port, 4-bit Schmitt input port and 4-bit output. Epson. Tel., 0442 227331; fax, 0442 227244.

Embedded PC. The Ampro *CoreModule/PC* is a new member of the family of embedded microcontrollers, which contain the equivalent of a PC motherboard and expansion cards, this one intended to sell for under £100 in quantity. The board has a 9.8MHz *8088*-compatible cpu, 256Kbyte of dram, a series/parallel controller, keyboard port and speaker interfaces, together with an on-board bootable solid-state disk. Power consumption is 0.6W from 5V. Diamond Point International Ltd. Tel., 0634 722390; fax, 0634 722398.

Mixed-signal ICs

Voltage detectors. Using 90% less current than their rivals, Panasonic's *MN1380* series of cmos voltage detectors monitor power supply voltages to computers and other systems, provide the reset for initialisation at power-up and prevent runaway when the power supply varies. The devices provide a choice of cmos n-channel open-drain and inverted cmos output to match devices from any maker. Current consumption is 1μA for a drain voltage of 5V. Panasonic Industrial (Europe) Ltd. Tel., 0344 863444; fax, 0344 861656.

Oscillators

Rubidium oscillator. A miniature atomic frequency standard, the *Ball-Efratom FRS-C* rubidium oscillator, is now available in the UK. Intended for use as a local oscillator or time-base reference, the oscillator exhibits a long-term accuracy of 1 in 10^{-9}/year after ageing and temperature effects are considered. Annual adjustment to compensate for ageing in rubidium oscillators is extended to five years in the FRS-C. Output options include 5MHz sine, 10MHz sine, 10MHz ttl and 2.048MHz ttl. Warm up to full accuracy is six minutes. Sematron UK Ltd. Tel., 0734 819970; fax, 0734 819786.

SMD oscillator. Contained in a surface-mounted package measuring 10.5 by 5 by 2.4mm, the *Seiko Epson SG-636SCE* is a cmos-compatible oscillator for 3.3V devices, working in the range 2.2187-70MHz. Current consumption is 10mA and 1μA in standby. Temperature range is –10°C to 70°C, with a stability of ±100ppm/°C and ageing at ±5ppm/year. Advanced Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

Data logging

Portable logger. *Bitlogger* by Logic Beach Inc. is a portable, weather-proof instrument intended for site work with later analysis on a PC. No programming knowledge is needed and user-configuration is possible by means of a menu-driven approach. Plug-in units allow sensors for temperature (thermocouples, rtds and thermistors), voltage, current, 4-20mA current loop, frequency, count, events, pH and more. A modem may be used to collect data from a number of loggers with polling software for control. *Quickgraph* software allows downloading to a PC. Vidas International Marketing. Tel., 0428 606222; fax, 0428 606676.

Radio comms test set. Marconi Instruments's *2945 TestMate* is said to be the lightest fully-featured radio communications service monitor available at 11.4kg, possessing a test speed up to four times as fast as its rivals. Features include a digital spectrum analyser with an update speed fast enough for live monitoring of transmissions; a very easy-to-use user interface; a bright, high-resolution lcd visible under all lighting conditions; and fully protected inputs. A pcmcia memory card interface allows results to be saved for later examination. Marconi Instruments Ltd. Tel., 0727 859292; fax, 0727 857481.

Power semiconductors

60A Schottky. IR has the *MBR6045WT*, which is a centre-tap Schottky rectifier exhibiting a maximum forward voltage drop of 0.5V per leg at 30A and 125°C. Maximum DC reverse voltage is 45V and d*v*/d*t* is 10,000V/μs. InternationalRectifier. Tel., 0883 713215; fax, 0883 714234.

PASSIVE

Passive components

Vernier dial drives. Jackson's new range of vernier dial drives come in three sizes, having overall diameters of 43mm, 50mm and 70mm, with front-of-panel depths of 18mm, 20mm and 22mm. Standard coupling is for 0.25in or 6mm spindles and the drives come with a 100-division scale over 180°. Variations of shaft size and scales are available to order. The drives use a ball-bearing system instead of the more common friction drive to confer freedom from slippage and backlash. Jackson Brothers Ltd. Tel., 081-681 2754; fax, 081-681 3728.

SM fuses. The Schurter *OMF* range of surface-mounted fuses is extended to encompass the 63mA-5A current range. To speed production, the fuses can be supplied reeled or boxed with the fuse inserted, for fitting to the board in one operation. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Dielectric filter. AVX announces the *PDFC* series of dielectric filters meant for use in telecomms, particularly in the DECT sector. Frequency range is 1.8-2GHz, insertion loss is 3dB and size is 6.5 by 5.5 by 3mm, making the products compatible with the latest equipment. Filters to provide lower insertion loss and improved stop-band attenuation are available to order. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Connectors and cabling

Space-saving connectors. On a 0.1in pitch, *FCN-790* connectors by Fujitsu are designed to fit into tight spaces, the series including both straight and right-angled types standing only 8.6mm and 4.6mm off the board respectively. They are available with contacts numbering between 10 and 40. Making a reliable ribbon contact, a locking feature ensures that the contact is not broken by wear, vibration or shock. Devlin Electronics Ltd. Tel., 0256 467367; fax, 0256 840048.

Phono connectors. Two ranges of audio connectors from Deltron incorporate a T-slot outer contact spring mechanism to improve mating with RCA-type sockets. A professional range of plugs comes in colour-coded anodised aluminium bodies with a large internal cable clamp and a strain-relief spring. The centre pin is gold-plated, as is the T slot contact spring; colour coded sockets are also offered. A screened range of connectors use silver plating and bright nickel and also have the contact spring, being produced with or without a six-colour coding. Electrospeed. Tel., 0703 644555; fax, 0703 610282.

Displays

Flat touch screen. The *intersys tp200* flat display colour and monochrome touch screens replace hard-wired panels such as push-button pilot lights, selector switches, digital readouts and message displays. A colour leaflet describing them is available from Contraves. Contraves Industrial Products and Systems Ltd. Tel., 0604 493201; fax, 0604 670779.

Filters

High-current filters. Dale Electronics' Model *TJ* high-current filter inductors are toroidal and vertically mounted to reduce interference. Seven models (*TJ3-TJ9*) cover the inductance range 1.2μH-5600μH at current ratings up to 20A. Vishay Components (UK) Ltd. Tel., 0915 144155; fax, 0915 678262.

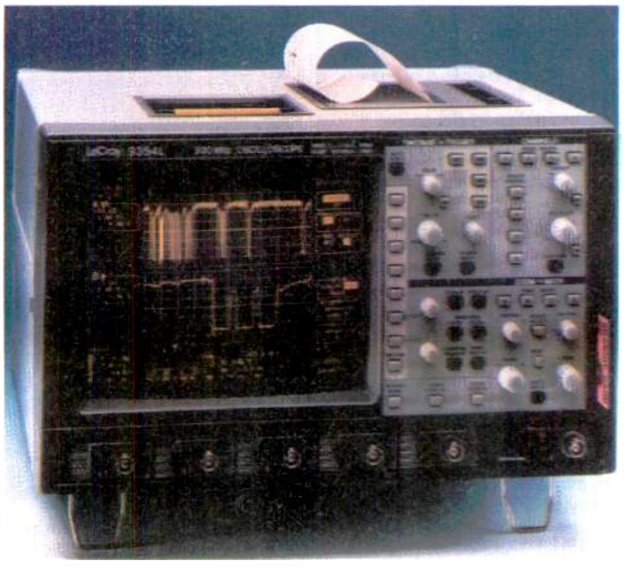
Instrumentation

EMC signal analyser. From Thurlby Thandar, the Model *8010* EMC signal analyser allows both large and small companies to carry out conducted emission tests to final compliance level.

Combined with the TTI *LISN1600* line impedance stabilisation network and linked to a PC, the *8010* forms a complete system, conforming to CISPR-16 for equipment drawing up to 16A, and suitable for use in normal lab. conditions.

In effect, the instrument combines a receiver and spectrum analyse with the PC's control, storage and display functions. It operates over 10kHz-30MHz with 200Hz and 9kHz bandwidths and true peak, quasi-peak and average detection. A preselector improves dynamic range to over 105dB and am and fm demodulation with audible monitoring is provided. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

PC-based instruments. Three more PC-based instruments from Pico are announced. The *SLA-16* is a logic analyser about the size of a cigarette packet, offering 16-channel operation and an 8K trace buffer. Internal and external clocking modes up to 50MHz are supported; its software provides state listings and waveform displays. *ADC-100* turns the PC into a dual-channel dso, spectrum analyser, frequency meter and voltmeter and, with extra software, a long-term data logger and chart-recorder emulator. For data logging, the *ADC-22* provides 22 input channels, its specification being otherwise the same as the existing *ADC-11*. Pico Technology. Tel., 0954 211716; fax, 0954 211880



More memory for DSOs. LeCroy's *9350/54* digital storage oscilloscopes, a six-member family of 500MHz, 500Msample/s single-shot/channel instruments, are now accompanied by memory upgrades from 25 and 100Kbytes/channel to 100Kbytes and 2Mbytes/channel, which means that a lower cost is possible for the initial purchase. LeCroy Ltd. Tel., 0235 533114; fax, 0235 528796.

Remote analogue measurement. Remote indication by analogue sensors to 16-bit accuracy is the function of the IMS *ADAM-4014D* module, which also has a local display. Measurement is transmitted up to 1200m by RS485 digitised Ascii output. The instrument has its own processor to convert transducer signal to engineering units; high or low limits are settable and initiate an alarm if exceeded, two digital outputs triggering other equipment in that case. Up to 256 units can be daisy-chained on one RS485 link. Integrated Measurement Systems Ltd. Tel., 0703 771143; fax, 0703 704301.

SDH/PDH test set. *Themis* is Schlumberger's telecomms data analyser to work with both synchronous digital hierarchy and plesiosynchronous digital hierarchy networks or any combination of the two. It is based on an open VMEbus architecture enabling it to accommodate future forms of network testing. A simple user interface and auto-configuration are combined with an extensive menu for the experienced user. Schlumberger Technologies. Tel., 01252 375111; fax, 01252 370792.

Electrometer. Having 5×10^{-15}A of input bias current, less than 1mV of burden voltage and over $2\times10^{14}\Omega$ input impedance, Keithley's *Model 6512* electrometer has full autoranging on all ranges while measuring current (2fA-20mA), resistance (100mΩ-200GΩ), voltage (10μV-200V) and charge (10fC-20nC). The instrument is digitally calibrated and the display uses exponential notation or engineering units. Control is front-panel or by IEEE-488 interface. Keithley Instruments Ltd. Tel., 0734 575666; fax, 0734 596469.

Interfaces

FLXibus i/o. The German company esd GmbH has an industrial i/o module for the FLXibus in Force computers. *Eagle-811* provides eight analogue input channels for ±10V to 12 bits, each being scanned with a sample rate of 29kHz and the results stored in fifo memory; an 8-bit wide output port at ttl level is provided to shape external input filtering for the analogue channels, which drive 12-bit d-to-a converters to the four analogue outputs. Eight digital inputs take 5V-30V levels and are opto-isolated from other module sections. esd GmbH. Tel., +49 511 37 29 80; fax, +49 511 63 36 50.

Multibus II i/o. *IO CBX/PAC* by Concurrent Technologies is a carrier board for *IndustryPack* modules and provides Multibus II users with a range of i/o interfaces. Modules available include counters, quadrature decoders, data converters, servo loop controllers, stepper controllers etc. Concurrent Technologies Ltd. Tel., 0206 752626; fax, 0206 751116.

Phone line/codec interface. *MT9196 C-Phone* by Mitel is a single IC for use in digital telephone equipment, incorporating a CCITT-compliant filter/codec, digital gain pads and a dtmf generator and ringer. Complete interfaces are provided for connection to the handset and speakerphone transducers. Internal registers are accessed through a serial port compatible with standard microcontrollers, the digital interface being STI and ST-bus compatible. It is programmable in μ-law or A-law; transmit and receive gains are programmable from –24dB to 21dB in 3dB steps; side tone levels are –9.96dB to 9.96dB in 3.2dB steps. Mitel Semiconductor. Tel., 0291 430000; fax, 0291 430400.

Literature

Robot listing. The Association for Robotics and Automation has relaunched its *Datafile*, which is a complete listing of all robots available in the UK. Each of 100 robots from 17 suppliers has a page of detailed description of its vital statistics to do with reach, payload and the work it can do. Company members of the BRA get one free, individuals pay £30 and non-members £50; updates cost £15 and £20. The Association for Robotics and Automation. Tel., 021–628 1745; fax, 021–628 1746.

Component placement. Component placement and insertion machines for leaded and sm devices are described in a Panasonic catalogue, which also covers *Panasert* adhesive dispensers, solder paste printers and automatic soldering equipment. Panasonic Industrial (Europe) Ltd. Tel., 0344 853277; fax, 0344 853803

Rechargeable batteries. Yuasa has

Flush-mount switches. EAO-Highland has a range of adaptors and actuators for its push-button switches to allow flush mounting on the panel, with stainless steel or aluminium surrounds. The adaptors also fit a range of keylocks, lever switches and indicators and will take a lens so that metal buttons can be internally illuminated. EAO-Highland Electronics Ltd. Tel., 0444 236000; fax, 0444 236641.

Please quote *"Electronics World + Wireless World"* when seeking further information

a catalogue and guide to the company's *NP* range of valve-regulated, sealed lead-acid batteries for industrial use. Capacities range from 1Ah to 150Ah in 4V, 6V and 12V types. Yuasa Battery Sales (UK) Ltd. Tel., 0793 612723; fax, 0793 618862.

Materials

Screening membrane. ITO-coated polyester film with earthing tags, made by Wasp, provides rfi screening of membrane switch panels, meeting all international rfi-suppression requirements. It will also screen apertures with little light reduction. Wessex Advanced Switching Products Ltd. Tel., 0705 453711; fax, 0705 473918.

Flexible casting compound. *TRA-CAST 3010* by the US company TRA-CON, is a clear, flexible, low-viscosity compound for potting and impregnating heat-sensitive components, having low exotherm and low curing shrinkage. It is repairable and suitable for potting PCBs. TRA-CON Inc. Tel., +1 (617) 391-5550; fax., +1 (617) 391-7380; e-mail tcepoxy@ aol.com.

Power supplies

6kV modules. In a package measuring 38.1 by 63.5 by 15mm, nine members of the Brandenburg *569* series of metal-cased voltage converters provide outputs from 100V to 6kV at 3W from inputs of 4-12V or 5-15V. Output is proportional to input and regulation is better than ±5% up to full load. Temperature coefficient is 200ppm/°C and short and long term drift 300ppm. Ripple is 0.005% to 0.2% of output. Brandenburg. Tel., 01384 393737; fax, 01384 440777.

Radio communications products

Vlf-to-hf receivers. Covering the 10kHz-30MHz frequency range, the Rohde & Schwarz *EK 895/6* are developments of the earlier *890/1*. The use of digital signal processing is responsible for new features such as 11 bandwidths from 150Hz to 8kHz, five selectable delay times, double notch filter, suppression of two interfering signals, noise blanking automatically matched to noise characteristics, syllabic squelch, rf input protection to 100VEMF, ISB and fsk/afsk demod. and i/o channel outputs for further dsp. Rohde & Schwarz UK Ltd. Tel., 0252 811377; fax, 0252 811447.

Circuit protection

Lightning barrier. A pcb-mounted lightning barrier from Hunter, the *ESP TN/Z* provides protection for one end of a twisted pair by restricting transients up to 5kV caused by lightning to 200V and voltage 'let-through' to interface circuitry to below the level at which damage is likely to occur Response time is less than 10ns; working voltage is 145V at 300mA. Hunter Electronic Components. Tel., 0628 75911; fax, 0628 75611.

Switches and relays

Illuminated switches. Fujisoku illuminated switches come in a choice of colours, styles, shapes and types of illumination. Type *LP* has single or dual-colour lighting by red, yellow and green leds with 12mm by 12mm or 12mm by 15mm lenses, and with snap-action mechanisms switching up to 3A at 250V ac (gold for signal switching). *LTM/LTR* switches are 7.5mm or 10mm round or square; *TM/TR* types can be supplied with no lighting, and *DP4/5* models are dual-pole types for pcb or panel mounting. Devlin Electronics Ltd. Tel., 0256 467367; fax, 0256 840048.

Elastomer keypads. The *EECO* Switch division of Transico Inc. has a range of low-cost conductive elastomer keypads using silicone rubber, available in a wide range of sizes, keytops and colours in opaque or translucent form and with a coating on the legend. Transico Inc., EECO Division. Tel., 01954 781818; fax 0194 789305.

Telecomms switch. AT&T's *LH1529* solid-state relay is contained in an 8-pin s-bend package, combining an ssr with an autopolarity optocoupler. It is for use in pcmciaType 2 form factor in applications that combine the switch-hook function and ring detection, although is equally suitable for designs combining other switching and isolation functions. It is rated at 350V/120mA with a typical 20Ω of on-resistance. The s-bend pins allow insertion into board cut-outs to conform to a pcmcia card thickness. AT&T Microelectronics. Tel., 0734 324299; fax, 0734 328148.

Little microswitches. Measuring 14.7 by 5.4 by 6.8mm, Cherry's *DK Series* of single-pole, double-throw microswitches handle 10mA-2A at 12V dc (ac on request), depending on the model, life expectancy being up to 500,000 operations. Contacts are the gold crosspoint type and there is a range of actuator types and terminals. Cherry Electrical Products Ltd. Tel., 0582 763100; fax, 0582 768883.

Reed relays. Made by Coto Wabash, *9090* mini-sip reed relays measure 3.81 by 15.24 by 6.6mm and come in single-pole normally cpen form in hermetically sealed glass and epoxy moulded lead-frame packs. Coils take 5V or 12V and the devices switch 10W. Diodes and magnetic shielding are offered. Avnet Time. Tel., 0462 484444; fax, 0462 488646.

Windows data acquisition. IMS has a data acquisition and control package for Windows – *Genie*. It presents a simple front-end for rapid setting up and configuration of scientific and engineering measurement and control applications without the need to be a programming expert. Strategy Editor uses a GUI using icons from a screen library, connected together to form a sequence of commands, the icons representing data acquisition, control, maths computation, file i/o, etc. Operating conditions in the facilities represented by the icons are changeable by dialogue boxes on screen. Display Editor allows the user to customise the presentation, creating graphical objects such as instrument panels. *Genie* supports Windows Client and Server DDE. Integrated Measurement Systems Ltd. Tel., 0703 771143; fax, 0703 704301.

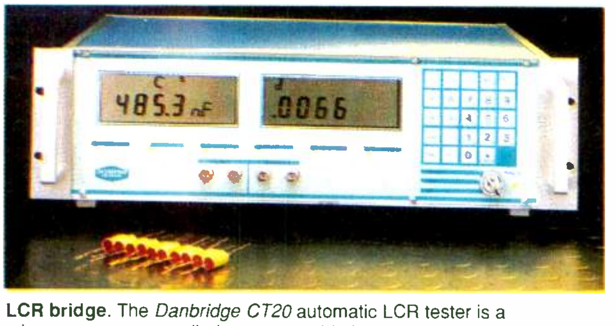
Transducers and sensors

Rotary position sensor. The *M-22* resistive absolute rotary position sensor by Control Transducers is a small device for slow or hand-operated equipment where accurate positioning is needed. Electrical travel is 320° or 340° and independent linearity 0.5% or 0.25%. It is available in servo and bush mounting in an aluminium and plastic case of 22mm diameter. Control Transducers. Tel., 0234 217704; fax, 0234 217083.

Remote charge converter. Allegro has a ruggedised remote charge converter to plug directly into signal conditioners and analysers with built-in 4-20mA current sources. The Endevco *2771 AM3* is a low-noise, two-wire, single-ended, solid-state device for use with piezoelectric transducers designed for systems complying with MIL-STD-740-2. Attached to the BNC output of the conditioner/analyser, the RCC transforms the charge output of a high-impedance accelerometer to a low-impedance voltage. Frequency response is 1-50kHz and gain 1mV/pC. Endevco UK Ltd. Tel., 0763 261311; fax, 0763 261120.

Hall sensors. Multiplexed Hall-sensor ICs from Allegro sense magnetic fields or switch status and send the reading over a two-wire power/signal bus. The *A3054KU/SU* are digital sensing devices intended for use in multiple-sensor systems where minimum wiring is required. Addressing is sequential by factory-programmed address, up to 30 sensors being accommodated on the same bus. Allegro MicroSystems Inc. Tel., 0932 253355; fax, 0932 246622.

Temperature sensor. Analog's *AD22100* temperature sensor measures in the –50°C to 150°C range, using a single 4-6V supply, no negative rail being needed for sub-zero temperatures. The IC is ratiometric, producing a voltage output proportional to temperature and the supply voltage, so that supply voltage variation does not affect the measurement. Analog Devices Ltd. Tel, 0932 253320; fax, 0932 247401.



LCR bridge. The *Danbridge CT20* automatic LCR tester is a microprocessor-controlled programmable instrument meant for use in production, incoming test and quality control. In addition to *L*, *C* and *R*, the bridge also measures *Q*, *D*, R_s and R_p and G_p for *C* and *L*, and L_s and C_p for resistance to accuracies of 0.05% on the main ranges. GPIB and RS232-C interfaces are fitted and a dual-frequency mode allows the measurement of loss factor and capacitance simultaneously at two of the three frequencies provided. The keyboard is dust-proof. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Sensor signal processing. MFP GmbH claims to have overcome all problems associated with inductive-transducer signals. Its sensor interface for PCs, the *IT 40 PC*, works entirely in the digital domain and requires no signal conversion or range shifting, and provides a 16-bit resolution. There is a PC add-on card, a connection box for four probes, a 15m cable and software for automatic measurement and an open programming interface. Ulrich Hartmann. Fax, 0309 691301.

COMPUTER

Computer board-level products

Graphic cards. Three new boards from Volante, one for the PCIbus and two for the Vesa Local bus, are accelerators to speed up operations in Windows, OS/2 and cad. *Warp VL* and *Warp PCI* use the Tseng *ET4000W32P* graphics engine and there is a feature connector for multimedia capture devices and monitor power management. Hero include Corel Draw with the boards. Hero Electronics Ltd. Tel., 0525 4055015; fax, 0525 402383.

Computer systems

Industrial Pcs. For use in hostile conditions, the *Sight Systems* range of industrial Pcs consist of a 19in 4U chassis, a 14in or 17in svga monitor in an enclosure, and a narrow-width keyboard with or without a trackerball built into a 1U drawer unit. Its chassis accommodates 8 to 20-slot passive backplanes or the new active backplane range or a range of motherboards. It is complete with an approved 250W power supply and the passive backplane type will take any of the Sight Systems all-in-one cpu cards from *386* to the 100MHz *486DX4*. Sight Systems Ltd. Tel., 0273 430993; fax, 0273 410380.

Data communications

EIA/TIA-562 transceivers
LTC1385/6 transceivers for portable computers draw 200µA from 3.3V, both having two drivers and two receivers working at speeds up to 120kb/s into 1000pF and 3kΩ. They are protected against repeated ±10kV esd strikes. The *1385* works in normal, shutdown and driver-disabled modes, current requirement in shutdown being 0.2µA; the *1386* works in normal mode only. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

ITU-T V.34 chipset. AT&T's earlier version of the *V.34 Complete Modem Chipset*, available since February, is now in its final form to the ITU 28.8kbit/s standard; the earlier version can be software upgraded –a feature that helps users to cope with non-compatible modems. The device supports V.34 28.8kbit/s data, V.17 14.4kbit/s fax, AT command set, V.24 and MNP4 error correction, V.24bis and MNP5 data compression, Class 1 send and receive fax and flash download. AT&T Microelectronics. Tel., 0734 324299; fax, 0734 328148.

PCMCIA fax/modem. *Portable Add-ons* announces its combined fax and modem on a pcmcia card for portable PCs. Since telephone connectors are not an international standard, the company also offers single connectors or packs of connectors to cover the continents. In all, 70 are available. Portable Add-ons (UK) Ltd. Tel., 0483 440777; fax, 0483 452304.

Mass storage systems

1.5Gbyte optical drive. A multi-function unit operating with both 1.5Gbyte rewritable and 1.4Gbyte worm (write-once-read-many) disks, Panasonic's *LF-7300* half-height optical disk is also backwards compatible with 1Gbyte and 500Mbyte rewritable disks and 470Mbyte and 940Mbyte worm disks. It uses double-sided disks and is in the form either of a bare drive or desk-top unit, working vertically or horizontally. There is a 512Kbyte cache and access time is under 45ms. A SCSI-2 interface allows connection to computers running dos, Windows, OS/2 or Novell 386. Panasonic Industrial (Europe) Ltd. Tel., 0344 863444; fax, 0344 861656.

Software

SpiceAge for Windows 4. Version 4 of Those Engineers' *SpiceAge for Windows* has many new features, prominent among which are reflection coefficient analysis and noise analysis.

There is now a DDE link with the utility program Modelmaker, which synthesises op-amp, transformer, attenuator and transistor library models. Reflection coefficient analysis in the complex plane option gives scattering matrix diagonals s_{11}, s_{22}, etc. In noise analysis, each component is measured and total s:n ratio or total noise obtained.

There is a time-optimising mode for digital circuit analysis, in which a relaxed circuit makes no further contribution to the calculations until it does something, speeding the process by about ten times while missing no event.

Automatic time step interval is now provided, which TE claims to be an improvement on that in some other programs. It can be switched off, if required.

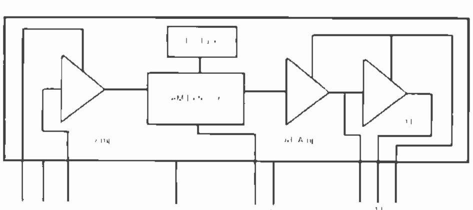
Further upgrades include faster quiescent analysis and improvements to Spice operation. Windows .WAV files import and export and there is better screen and print presentation. Those Engineers Ltd. Tel., 0181 906 0155; fax, 0181 906 0969. ■

Special-offer – low-power tx/rx for less than £12

For more details of the LPRS transmitter/receiver pair, see page 1046.

Low Power Radio Solutions is offering an am transmitter and receiver module pair at £11.74 inclusive to the first 500 EW+WW readers sending in the coupon below. This exclusive offer is a reduction of well over a third of the list price of £19.50.

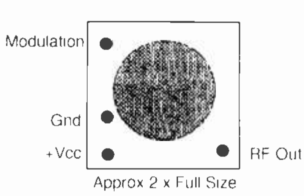
Communication distance is heavily dependent on circumstances including environment and antenna design, but it can exceed 50m in free space. Even inside a building, communication can be up to 25m. ■



Receiver pin-out

1	5V
2	Gnd
3	Rf in
7	Gnd
10	+5V
11	Gnd
13	Test point
14	Data out
15	+5-24V

Hybrid receiver module for 418MHz, above. In open air, this module can receive at distances of more than 100m in free air when used with the LPRSTX-418 transmitter.



Left, pin-out of the 418MHz low-power transmitter, showing its outline at roughly twice full size.

Features and specifications

Transmitter module
- Pcb mounting oem module
- Optimised for pcb antenna
- Small size – 13mm×13mm×5mm
- 418MHz saw controlled am

Supply voltage	5 to 12V
Supply current	2.6mA typ.
Frequency	4·8MHz
Radiated power (erp)	–6dBm
2nd harmonic erp	–56dBm
ERP above 1GHz	–60dBm max.

Receiver module
- Hybrid module
- Single-in-line package
- Regenerative superhet design

Sensitivity	Better than 2.5mV (–100dBm)
Rf pass band	±1MHz (3dB)
Lf pass band	2kHz square wave max.
Supply, rf circ.	5V, 5mA
Supply, o/p circ.	5-24V, 2mA

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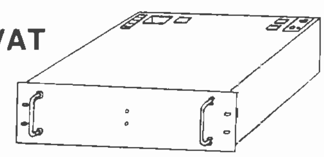

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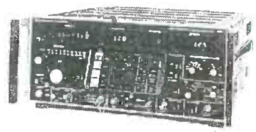

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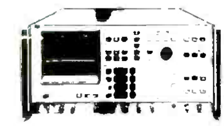

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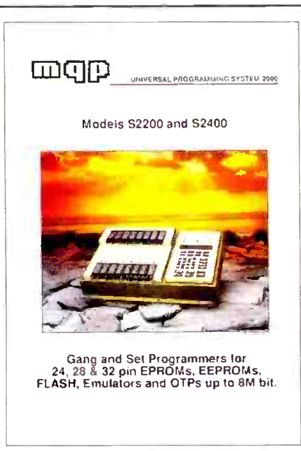

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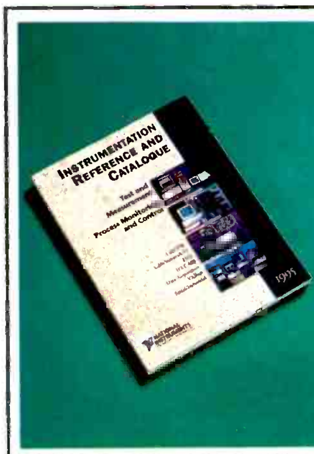

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