15% discount: controlled solder station

ELECTRONICS WORLD

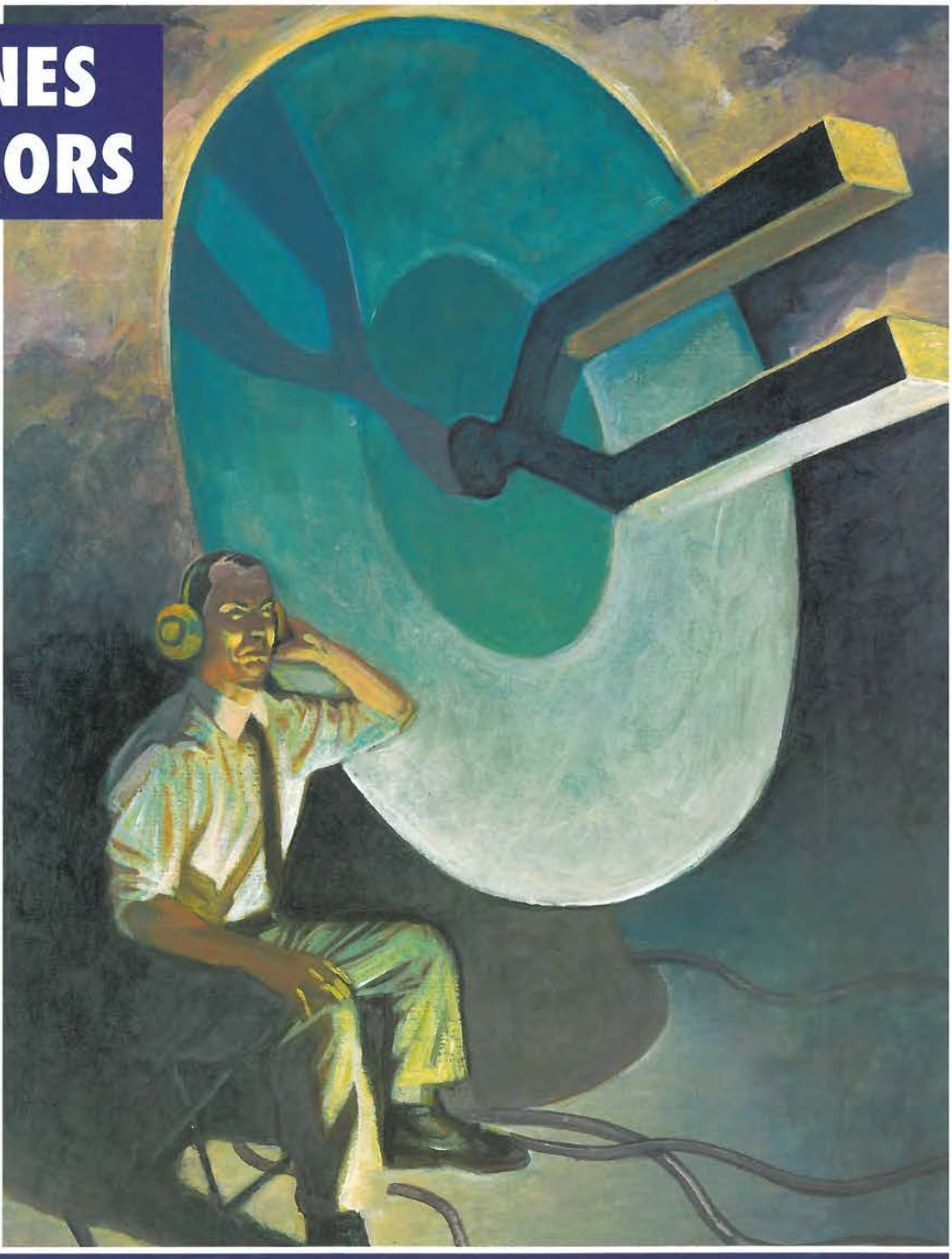
INCORPORATING WIRELESS WORLD

October 1997 £2.35

Austria Asch. 66.00
Denmark DKr. 67.00
Germany DM 18.00
Greece Dra.1100.00
Holland Dfl. 11.75
Italy L. 8800.00
Malta Lm. 1.55
IR £3.30
Singapore S$7.50
Spain Pts. 850
USA $5.95

A REED BUSINESS PUBLICATION
SOR DISTRIBUTION

FIELDS, LINES AND TREMORS



Self exposes speaker cable

Post Office war records

Wideband antenna design

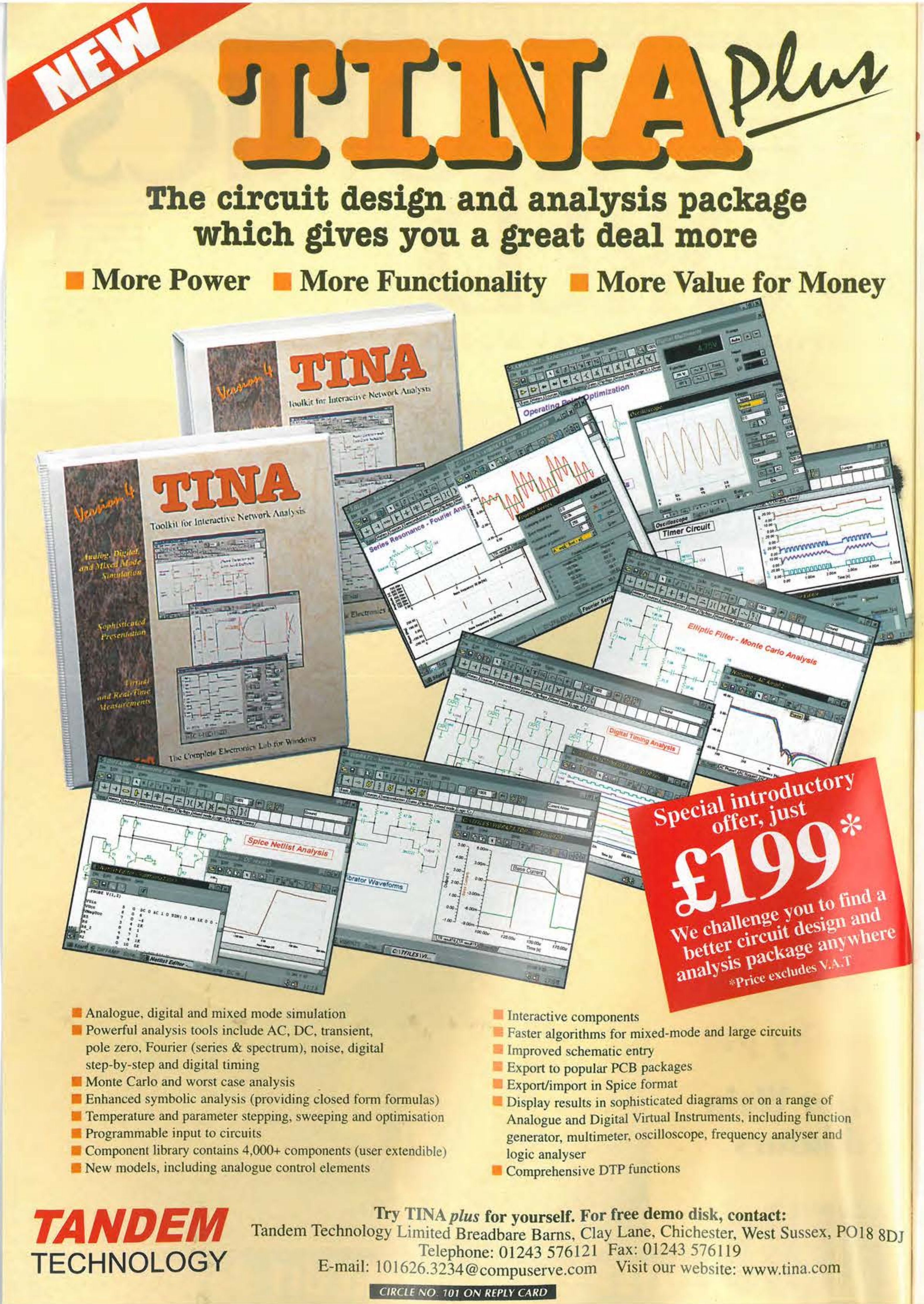
Reduce phase noise

Standby psu

Better Wein oscillators

10>
9 770959 833035

INTERNET FOR CIRCUIT DESIGNERS



Contents

Cover – Hashim Akib

806 FIELDS LINES AND TREMORS
Dominic Di Mario presents three very different VLF detectors for examining your environment. One senses electric fields, one examines magnetic lines and one records unfelt vibrations in the earth.

812 POST-OFFICE WAR RECORDS
Andy Emmerson has been investigating some of the Post Office's strange activities during the war – one of which involved producing vinyl disks.

815 SPEAKERS' CORNER
John Watkinson explains why good loudspeakers are so difficult to produce.

817 KILLING NOISE
Ian Hickman looks at one of the most important nuisances facing designers of oscillators for rf comms – phase noise.

825 DESIGN WIDEBAND ANTENNAS
An impedance-loaded antenna capable of continuous coverage from 6 to 150MHz without tuning or matching illustrates Richard Formato's design procedure.

831 CABLE SONICS?
Have you splashed out on the world's best hi-fi loudspeaker cables? If so, the last thing you need is Doug Self's revealing article.

837 A PROBLEM PHASE
Use text book equations for a Wien oscillator and you could end up with frequency errors of up to 70%. Bryan Hart explains why.

840 HANDS-ON INTERNET
Find better mosfet models, new amplifier designs, improved simulation software and faster search methods on Cyril's pages.

843 WORKING WITH MICROWAVES
Boris Sedacca shows that circuits for working at microwaves need few components and are quite easy to design – provided you can afford the software.

869 STANDBY KEEP-ALIVE POWER SUPPLIES
Efficient psus save energy, but in many applications, power down mode can save even more. Phil Darrington reports.

874 UNDERSTANDING CAN
CAN is a versatile serial bus designed for automotive applications. But it is finding new uses, as Svetlana Josifovska explains.

Regulars

795 COMMENT
HAARP – research or eco folley?

796 NEWS
Assault on batteries, UK traffic policy, Camera on a chip, Driverless cars.

801 RESEARCH NOTES
Nanoguitar, Electronic guide dog, Lower power chips, Mars robot, GPs on PCs.

860 CIRCUIT IDEAS
- NiCd capacity meter
- Sub-woofer filter
- Inductance meter
- 1-180s timer
- Brake for ac motors
- Differentially tuned oscillator
- Mains detector
- Novel headphone amplifier
- Power-on delay
- AC/DC voltage detector

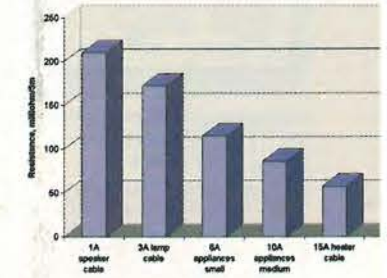
861 LETTERS
Silver and skin effect, Singing preamp, Remote-control enhancement.

863 NEW PRODUCTS

15% reader discount on temperature-controlled soldering stations – see page 817.

Read about a nanoguitar, an electronic guide dog, and this autonomous robot in Research Notes starting on page 802.

HAARP – research or eco folley? Page 795



Doug Self has been analysing cables for connecting loudspeakers. See his conclusions on page 801.

NOVEMBER ISSUE ON SALE 2 OCTOBER

EDITOR
Martin Eccles
0181 652 3128

CONSULTANTS
Jonathan Campbell
Philip Darrington
Frank Ogden

EDITORIAL ADMINISTRATION
Jackie Lowe
0181-652 3614

E-MAIL ORDERS
jackie.lowe@rbi.co.uk

DESIGN
Alan Kerr

ADVERTISEMENT MANAGER
Richard Napier
0181-652 3620

DISPLAY SALES EXECUTIVE
Joannah Cox
0181-652 3620

ADVERTISING PRODUCTION
0181-652 3620

PUBLISHER
Mick Elliott

EDITORIAL FAX
0181-652 8111

CLASSIFIED FAX
0181-652 8938

NEWSTRADE ENQUIRIES
0171 261 7704

ISSN 0959-8332

SUBSCRIPTION HOTLINE
01622 778000

SUBSCRIPTION QUERIES
rbp.subscriptions@rbi.co.uk
Tel 01444 445566
FAX 01444 445447

For a full listing of RBI magazines:
http//www.reedbusiness.com

REED BUSINESS INFORMATION

Research or eco folly?

The High-Frequency Active Auroral Research Project – HAARP – has been a source of increasing controversy as public awareness of the technology increases. In May this year I had the opportunity to present in Brussels on this issue to members of Global International, a group of several hundred Parliamentarians gathered from throughout the world.

"The Earth is delicately balanced, and seeks to restore balance when disturbed. No one really knows how ionospheric experiments will affect that balance, or what the Earth will do in response to try to restore balance." These words are from Rosalle Bertell, Ph.D, of Toronto, Canada, founder of the International Institute of Concern for Public Health. Dr Bertell was commenting on HAARP – a U.S military experiment.

The HAARP project may be the test run for a ground-based 'Star Wars' defence system. Military documents say it is intended to disrupt portions of the ionosphere by beating it with powerful pulsed radio frequency beams. Radiation, in some of the new instruments uses will bounce back to the surface of the planet in the form of extremely low frequency waves of energy.

Intended to be the most powerful ionospheric heater ever built, HAARP's ground-based apparatus – a current array of 48 antennae each powered by its own transmitter – sits in the remote Alaskan wilderness northeast of the city of Anchorage.

HAARP is much more than the auroral and radio communications research project as is claimed by researchers at the University of Alaska's Geophysical Institute and their financial backers – the US Navy and US Air Force. The fact is that HAARP is a military experiment aimed at invasively manipulating the ionosphere by beaming high energy upward from the ground. Such activity could potentially disrupt natural systems on the earth and high above it.

Individual members of the European Parliament are among the growing number of people worldwide who have been startled to hear about HAARP. Voices expressing various levels of concern are being heard in many countries.

For example, in contrast to the cautiously worded comment of Dr Bertell – a researcher in the field of quantum electrodynamics based in Germany – Al Zielinski, paints an apocalyptic picture. He says that HAARP technology could trigger a disaster with a global impact – electromagnetic waves causing destruction "when interacting with protective layers of the earth and its gravitational field."

The concern that this system – and others like it – could effect the geophysical stability of the planet has also been expressed by Brooks Agnew. Brooks is a radio chemist who discovered that some very specific frequencies when broadcast into the Earth could trigger earthquakes even at relatively low power levels.

This same sentiment was expressed by William Cohen, the United States Secretary of Defense on 29 April 1997. Commenting on the possibility of terrorists states he said, "...engaging even in an eco-type of terrorism whereby they can alter the climate, set off earthquakes, volcanoes remotely through the use of electromagnetic waves." These systems we assert, if they are indeed possessed by terrorists as Defense Secretary Cohen has implied, must also be available to the more advanced governmental defence organisations in the world.

The ionosphere is alive with electrical activity, so much so that its processes are "non-linear." This means that the ionosphere is dynamic, and its reactions to experiments are unpredictable. The concept of non-linear is important in

"Is it wise to poke holes in the Earth's electrical umbrella?"

understanding the concerns of independent scientists who are knowledgeable about advanced physics and who warn against brash high-energy experiments on the ionosphere.

Non-linear processes can change suddenly and unexpectedly, or they can increase in power dramatically. Some theorists say that a non-linear process can, under certain conditions, tap into the background energy of space, which is also called "zero-point fluctuations of the vacuum."

Studying radio communications by using a tool as powerful as HAARP is a worthy scientific task but some independent researchers question whether the means justify the end. Is it wise to poke holes in Earth's electrical umbrella? Is it wise to prod a dynamic natural system without knowing how it might react?

HAARP is intended to heat and lift a portion of the ionosphere above a selected location or locations on the planet and to make a huge invisible 'mirror' for bouncing electromagnetic radiation back to the surface of Earth. Why?

The answer is that the US military wants to communicate with its submerged submarines by penetrating the oceans with extremely low frequency radiations. It also wants to penetrate the land with ELF waves in order to search for hidden tunnels or other sites of military interest – a process known as earth-penetrating-tomography. This application is funded under the counter proliferation and non-proliferation of nuclear. weapons in the United States Defense budget.

If the technology is scaled up in size, it could potentially shield a territory from intercontinental ballistic missiles. It could Fry satellites and differentiate between incoming objects such as missiles carrying nuclear warheads and decoys. It could even enhance communications or disrupt communications over a large area of the globe. And of course, it could affect the health of humans and other biological systems. ■

Dr Nick Begich, email: drnick@alaska.net
Internet: http://www.earthpulse,com

Electronics World is published monthly. By post, current issue £2.35, back issues (if available £2.50. Orders, payments and general correspondence to **L333, Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS**. Tlx:892984 REED BP G. Cheques should be made payable to Reed Business Information Ltd
Newstrade: Distributed by Marketforce (UK) Ltd, 247 Tottenham Court Road London W1P OAU 0171 261-5108.
Subscriptions: Quadrant Subscription Services, Oakfield House Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 445566. Please notify change of address. Subscription rates 1 year £32 UK 2 years £43.00 3 years £75.00. Surface mail 1 year £37.00 2 years £60 00 3 years £86.00 Air mail Europe/Eu 1 year £46.00 2 years £73.00 ROW 1 year £56.00 2 years £89.00

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, 475 Park Avenue South, 2nd Fl New York, NY 10016 Tel; (212) 679 8888 Fax; (212) 679 9455
USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. Periodicles Postage Paid at Rahway NJ Postmaster. Send address changes to above.
Printed by BPCC Magazines (Carlisle) Ltd, Newtown Trading Estate Carlisle. Cumbria, CA2 7NR
Typeset by Marlin Imaging 2-4 Powerscrott Road, Sidcup, Kent DAt 4 SDT ,

Assault on batteries

Battery maker Duracell wants battery specifications tightened to avoid the increasing risk of appliances such as smoke alarms failing.

The specifications, defined by the International Electrotechnical Commission (IEC), cover the shape, size and discharge currents of batteries in various applications.

According to Philip Smith, sales manager for Duracell's industrial products division, battery performance is being stretched beyond the IEC's tests and standards. Discharge tests run from 1 to 500mA.

"That doesn't take account of datalogging equipment or cellular phones," said Smith. These drop off either end of the scale, he claimed.

Because the specifications are lax, a battery manufacturer can change the design to a point where it causes problems. For example, at extremely low currents, perhaps tens of microamps in a smoke alarm or datalogger, the battery might even stop working.

But the battery passes the IEC tests because they only go down to 1mA discharge current.

"The IEC is very slow to reflect the changes in the applications," said Smith. "We want the IEC to reflect what is happening in the marketplace with a wider range of discharge currents. A smoke alarm test is needed. Peoples' lives depend on it."

An IEC spokesperson said:"We are trying to respond to the industry's needs. Duracell participates in the process of producing standards and has access to the lobbying procedures."

The IEC is already extending the tests to cover products using very low currents. However, a test is needed for high current products such as mobile phones.

Richard Ball, *Electronics Weekly*

Driverless cars on the road to reality

Researchers in the US say they have successfully completed a series of tests involving cars that drive themselves. The tests were conducted along an eight mile stretch of California motorway.

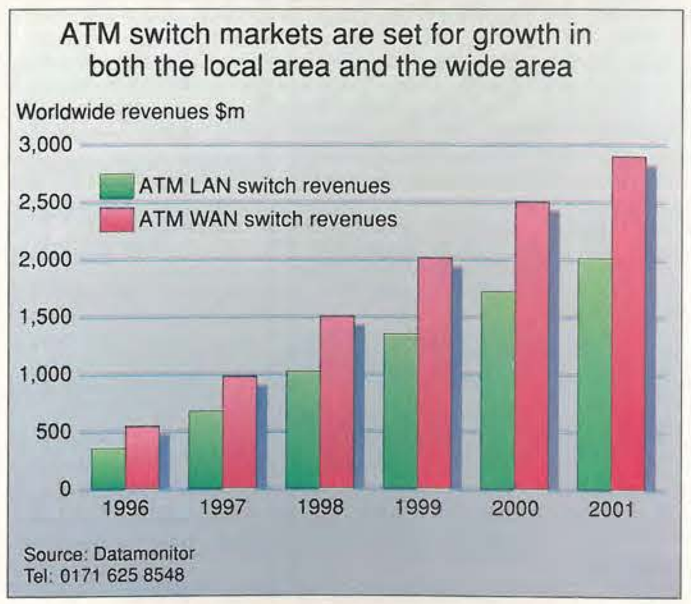
The cars were equipped with on-board computers and video cameras, and were guided to stay within their lanes by small magnets embedded into the roadway at four foot intervals. One driver said that the experience was thrilling for a few seconds but then became "really dull" as he had nothing to do.

The experiments are part of a US national effort to develop fully automated intelligent vehicle highway systems as mandated by a law passed in 1991. A group of high-tech companies have joined with car manufacturers to form the National Automated Highway System Consortium. The group estimates that current motorways could be adapted to handle automated vehicles for as little as $10 000 per mile.

There are still trials in various parts of the US awaiting completion before standards for building cars and automated highways are defined.

1996 saw patents increase

The latest annual report from the UK patent office shows the number of published patent applications, at 11 452, increased by three per cent in 1996. Telecommunications remains the most innovative technological sector: 908 patents were published in the financial year 1996 compared to 819 the previous one. Other patent segments include measuring and testing, with 655; and electric circuit elements and magnets, which at 577, grew by 12 per cent.



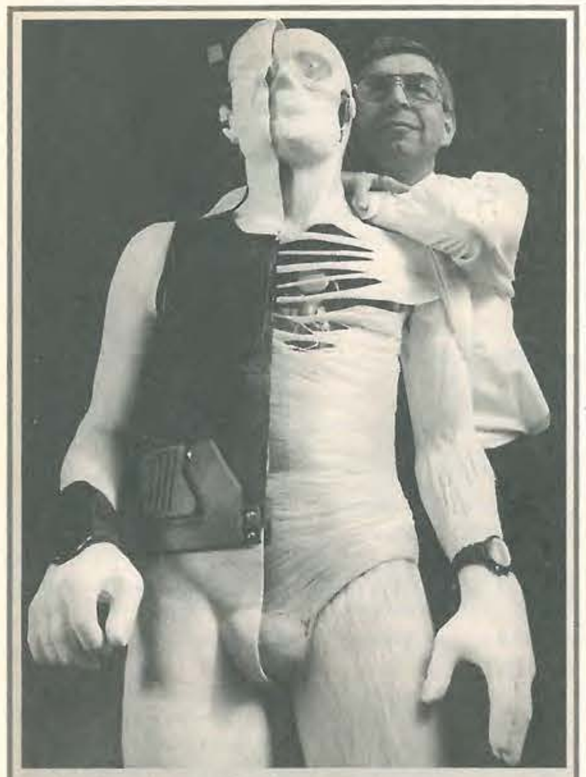
Five times lower power than cmos?

A breakthrough in the way chips consume power, reported by the University of California's Information Sciences Institute (ISI), could dramatically cut power consumption by 80 per cent and make possible new types of highly integrated chip designs.

Researchers within the ACMOS group at ISI have patented a prototype microprocessor called the AC-1, which consumes just one fifth of the power of a similar c-mos processor. The design uses pulsed power and adiabatic charging techniques which recycle some of the power used in the chip's clock cycle.

Researchers are unsure if the same techniques can be applied to commercial microprocessors and other chips, but chip companies will be able to buy licenses for the technology.

Low power consumption is critical to building large, high performance microprocessors where problems of heat dissipation are limiting designs. More exotic types of chips built as a cube could be made possible with low power technologies.



Silicon implants... Researchers at BT Martlesham are working with BT Wired Man – part of a study looking at the feasibility of placing various electronic systems within and on human beings. Embedded items being examined include pacemakers and electronic pain relief modules. The external parts envisaged include an intelligent signet ring, containing data like driving licence and medical details, and a power vest with the potential to generate power from body heat and movement. Shown with Wired Man is Professor Peter Cochrane, head of research at BT Labs.

Firms show interest in intelligent ram

Several DRAM manufacturers have offered to fabricate samples of a novel processor being developed at the University of California, Berkeley.

Dubbed intelligent ram, or i-ram, the IC architecture tackles the increasing discrepancy between processing performance and memory bandwidth by combining a vector processing style of computing with multiple blocks of on-chip store.

According to Professor David Patterson, head of the Berkeley project, "most" of the dynamic memory firms approached – Mitsubishi, Hyundai, LG Semicon, Samsung, IBM and Micron – are interested in manufacturing the IRAM samples.

Patterson expects to work with one or two of the companies, resulting in the first samples by next spring. Envisaged applications for the ICs include PDAs, game consoles, servers and disk drives.

He is confident that intelligent ram will eventually become an established architecture, although he admits that there is plenty that can still go wrong. For him, one promising development indicating i-ram's increasing acceptance is that, "two leading semiconductor companies have started to say negative things about it".

Government warned on traffic policy

Government is urged to establish integrated transport framework to prevent UK missing out on growing telematics markets.

The UK is set to lose out commercially if the government doesn't hurry up with a policy to establish an integrated transport framework based on electronics, telecommunications and IT. So warns ITS Focus, the UK's industrial and institutional voice on transport telematics.

ITS Focus believes that the government must speed up its policy as well as making a direct investment to allow the UK to take a lead and participate in the growing markets of telematics (see 'At a glance' box). Otherwise, countries such as the US will overtake the UK in seizing commercial opportunities.

"The main thing is that this is a rapidly moving area that offers scope for business opportunities and employment but which will continue to stall unless we have organisational, institutional and technical frameworks," said John Miles, public policy adviser to the European Commission's DG XIII and consultant to the ITS Focus.

ITS Focus recommends that an overall framework and analysis of different transport and traffic services is tailored to UK needs for a successful implementation – as has already been undertaken in the US under a legislative $30m funding.

ITS Focus has welcomed the government's research programme on Urban Traffic Management and Control (UTMC) announced last week, and a white paper on integrated transport policy from the newly merged Department of Environment, Transport and Regions (DETR). The research is aimed at developing intelligent control systems to manage traffic in cities.

"We see this this as an important start – it's along the lines we've been setting. However, the resources for the UTMC are quite small for a big ambition. They are £5m over five years," said Miles.

The DETR's white paper will follow an assessment on the motorway tolling trials recently completed at the Transport Research Laboratory in Berkshire.

What is Intelligent Transport System?

- It can be applied to all means of transport: road, rail, air and sea.
- It relies on electronics, telecommunications and IT – collectively known as telematics.
- It combines traffic management and control, travel and traffic information, electronic fee collection, automatic vehicle location, road safety and route guidance and navigation – amongst others.

ITS report, Tel: 01344 770757

New camera-on-a-chip venture

Specialist flash, E2 and PLD house Atmel is to develop a digital-camera-on-a-chip with Polaroid. It hopes to sell the single chip solution on the open market next year.

The ability to integrate this on one chip comes from ES2 (European Silicon Structures), which Atmel bought in 1995.

Atmel's task is to implement imaging sensors in c-mos and integrate them with a dsp and both E2 and flash technologies on a cell-based IC. "Our E2 and flash are easier to mix than other solutions," Atmel's European boss Bob Henderson said.

"Other solutions" means ccds, which are a high power bipolar technology.

Instead of ccds, the Atmel chip will use low power c-mos sensors.

Polaroid will contribute proprietary technology for colour recovery signal processing, colour filter processing and pixel sensing.

As well as its memory technologies, Atmel will be contributing its a-to-d and d-to-a convertor technology, compression circuitry and dsp cores. Last year, Atmel licensed the Oak and Pine dsp cores licensed in turn from DSP Group in California.

Smart cards in Boots

Initially, Boots' smart card will hold shoppers' bonus points, but it has the ability to carry much more information, including people's national insurance number and doctor details.

Boots the Chemist confirms September launch of loyalty smartcard – the first of its kind in the UK. Initially the card, which carries a chip rather than a magnetic stripe, will be used for bonus points but it is expected to be developed further.

"The smartcard may be used as in Germany, for carrying medical information, name, address, doctor's details and national insurance of the person that carries it, and it can be used as a social security card," said one Boots spokesperson.

The smartcard, dubbed Advantage, will carry Siemens' *SLE4442* memory chip with 256byte eeprom and a programmable security code. Up to 80 per cent of the cards will be manufactured by GPT and 20 per cent by Gemplus. The smartcard readers will be supplied by Dione Communications.

Boots says it could have used a magnetic stripe like any other retailer, but it wanted to have more than just a loyalty bonus card by using a chip.

"It allows a greater security, it gives an advantage that the card can be used in any store, the holder can spend the points in any store and the card provides a good platform for other services in the future," said the spokesperson.

Boots is expecting to have up to eight million Britons carrying Advantage by the end of next year.

In 1995, Peter Lilley, then social security minister, told a party conference that the government would adopt a social security smartcard in the near future. But since, the government has abandoned the idea of a chip-based card and instead opted for a magnetic stripe in order to save money in the short term. Its plans are to start rolling out its limited-functionality card in three years time.

Svetlana Josifovska, *Electronics Weekly*

Low distortion a-to-d

Crystal Semiconductor is claiming the lowest total harmonic distortion achieved in an integrated circuit a-to-d converter. Its *CS5396*, aimed at professional audio use, yields greater than 105dB total harmonic distortion plus noise with a 120dB dynamic range.

Mark Taylor, a company spokesman, said: "The only converters that have a similar performance are rack-mounted units."

Within the integrated circuit are two chips, one digital and one analogue. Sample rate of the stereo 24-bit device is 96ksample/s and noise-shaping parameters can be down-loaded during use.

● Asahi Kasei Microsystems, a company that used to have a tie-up with Crystal Semiconductor, has introduced the *AK5352*, a 20-bit stereo a-to-d converter with a 104dB dynamic range.

Euro PC growth is on a high

The European pc market grew by 16 per cent in the second quarter, says a Dataquest report. The quarter growth, measured against the corresponding quarter last year, is the highest in the last 18 months. "It's been a very successful quarter but there are too many counter-indicators to predict another boom," said Steve Brazier, Dataquest's European PC group's associate director. The growth will be welcome news for semiconductor companies such as Intel which, earlier this year, warned that it would not meet second quarter expectations, attributed in part to the downturn of the European pc market. ■

RESEARCH NOTES

Jonathan Campbell

Nanoguitar hits the highest notes

Many people claim to have music in their blood, but how about having a musical instrument in your blood! That claim is at least theoretically possible following an announcement from scientists at Cornell University that they have manufactured the world's smallest guitar.

It has six strings, each about 50nm or the width of about 100 atoms, and if plucked by an atomic-force microscope, for example – the strings would resonate, but at inaudible frequencies. The entire structure is about 10μm long, about the size of a single cell.

The nanoguitar has actually been made for fun rather than function, and is just one of several similar structures.

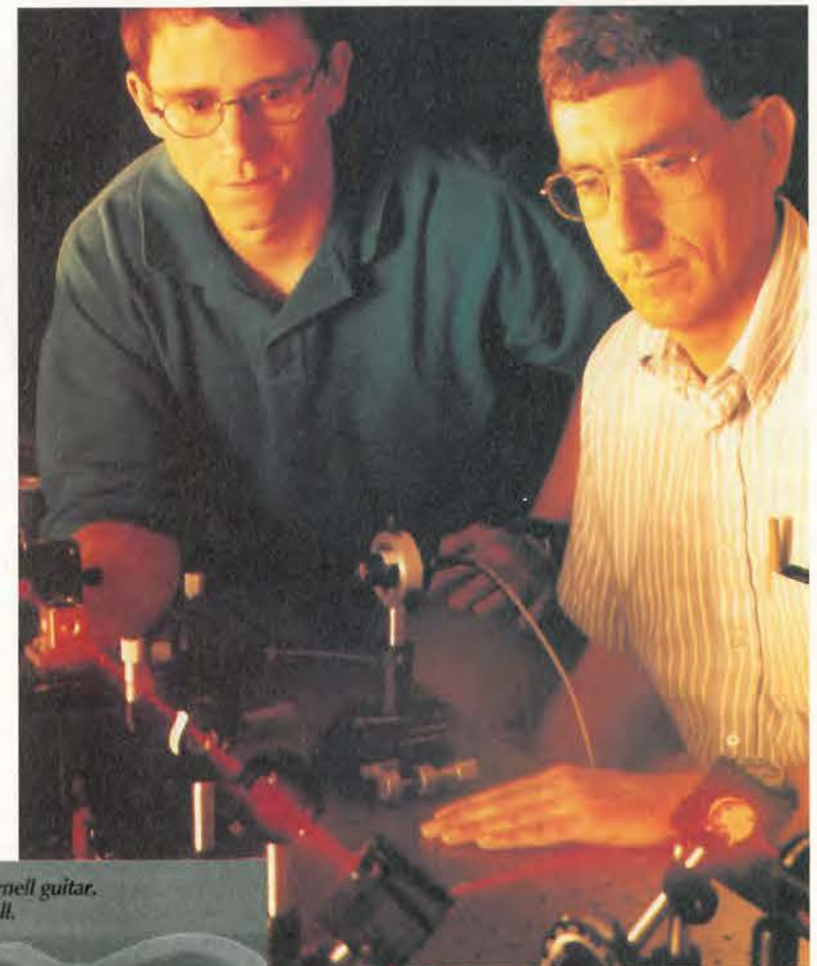
While the guitar makes a good photograph – it won an award for the best micrograph – and illustrates the progress being made in microelectromechanical devices, or mems, it is the other structures and devices being made at the Cornell Nanofabrication Facility that will be of real utility.

Researchers have made interferometers for example using parallel mirrors, one of which moves relative to the other. These electrically driven devices can be used to modulate the intensity of the reflected light.

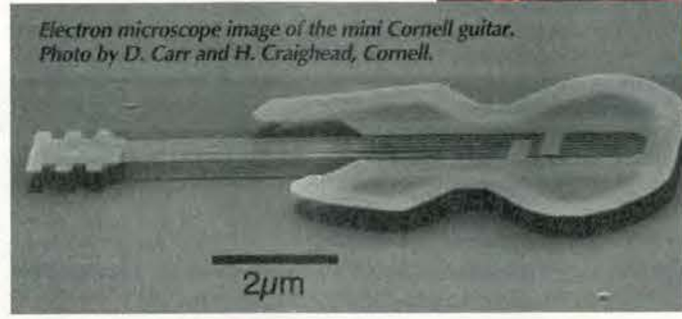
"This could be of interest for light displays," says Harold Craighead, Cornell professor of applied and engineering physics and former director of the Cornell Nanofabrication Facility, a national resource.

"You could have arrays of these things because they're so small, with each one independently drivable. We have tremendous flexibility in what we can build."

In the near term, such nanostructures also can be used to modulate lasers for fibre optic communications. These researchers already have demonstrated the ability to make large amplitude modulation of light signals at high speeds.



Cornell scientists, Harold Craighead on the right, demonstrate one of the world's smallest silicon mechanical devices. A 4μm-wide interferometer too small to see is modulating the red laser light, which could be useful for light displays and other applications. The researchers have made several of the world's smallest silicon mechanical devices at Cornell's Nanofabrication Facility. Photo by Charles Harrington, Cornell University.



Electron microscope image of the mini Cornell guitar. Photo by D. Carr and H. Craighead, Cornell.

"We can make reflected light pulses at a rate of 12million/s," says Craighead.

Such a rate is faster than the bit rate of most ethernet connections.

Most microelectromechanical devices are made by photolithography and chemical etching and have minimum feature sizes of slightly less than 1μm. To build devices with dimensions of nanometres rather than micrometers requires a new fabrication approach.

"I know we can go smaller than this. The question is how small we can go and still have dependable and measurable mechanical properties. That is one of the things we would like to know," admits Craighead.

Using high-voltage electron beam lithography, the Cornell researchers sculpted their structures out of single crystal silicon on oxide substrates. A resist is used to pattern the top silicon layer. The oxide that is underneath this layer can be selectively removed using a wet chemical etch. The result: free-standing structures in silicon crystal.

Robot gets ready to hit the road – on Mars

The Nomad rover will be expected to roam much further than existing planetary rover vehicles.

Nasa scientists were rightly excited at the recent performance of their Mars rover vehicle. But to open up exploration of other planets, researchers really need vehicles able to cope much more happily with travelling relatively large distances.

Now, they hope they may be on the path to developing one in 'Nomad' – a rover that has just set a record by travelling farther than any remotely controlled robot has before over rough territory. The robot's four wheels logged more than 215km across Chile's rugged Atacama Desert, during a field experiment designed to prepare for future missions to Antarctica, the Moon and Mars.

Although the straight-line distance on a map was only about 20km, Nomad had to weave through very difficult terrain, and it made numerous side trips for science and to test the meteorite sensors.

The 750kg robot, developed at Carnegie Mellon and funded by Nasa, also validated the use of colour stereo video cameras with human-eye resolution for geology.

Autonomous driving is critical for planetary exploration because the communications delay between Earth and planets can be many minutes. With autonomous driving, a robot can explore a much greater distance because it doesn't have to wait for a person to decide a safe route. The rover is able to see obstacles and recognise them on its own.

Another first for Nomad is use of an on-board panospheric camera to provide live 360° video-based still images of the robot's surroundings.

The high-resolution video camera focuses up into a hemispheric mirror similar to a store security mirror and takes a 360° picture – one frame per second The video view includes all of the ground up to the horizon in the circle surrounding Nomad.

Electronics take over house calls

The prototype for an electronic house call system that could reduce the number of visits to chronically sick patients made by nurses and GPs has been developed and tested in the US. The system uses established hardware and cable communications to set up a two way link between patient and medic.

Researchers fashioned the prototype from existing computer hardware, with additions such as a multi-function patient monitor – like one used in intensive care units – into which blood pressure cuffs, stethoscopes and other medical devices are plugged.

A commercially available video-conferencing program enables the nurse and patient to see each other and talk throughout the examination. The system also accepts data from a variety of medical devices, such as those registering blood pressure and blood oxygen levels, so that the nurse can listen to heart and lungs and perform an electrocardiogram.

Development and initial testing of the electronic house call system has so far been funded primarily by grants from the Department of the Army and the Georgia Research Alliance.

For the most part, patient feedback has been positive. Patients seem to like the idea of being monitored at their homes. Though the system would mean less face to face contact between medics and patient, the interaction with nurses through the system could actually be greater.

Contact: Michael Burrow, Biomedical Interactive Technology Center, Georgia Institute of Technology, Atlanta, GA 30332-0823, USA. Tel: 00 1 404 894-7034 email: michael.burrow@bitc.gatech.edu

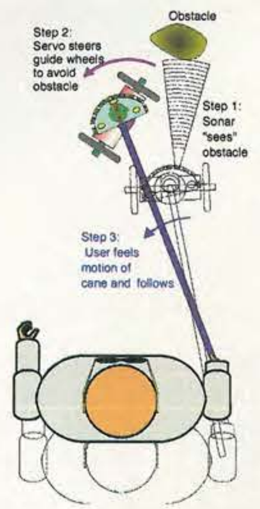
A patient has her temperature taken via computer on a prototype system developed by Georgia Tech and partner researchers. (Photo courtesy Medical College of Georgia)

Electronic guide dog steers round obstacles

You can't pat it and it won't nuzzle up to you. But a new sonar-equipped navigation aid for the blind being developed at that University of Michigan College of Engineering's Mobile Robotics Laboratory claims to be able to detect obstacles in the user's path as well as any dog, and automatically steer around them.

The device, invented by Johann Borenstein U-M research scientist in mechanical engineering and applied mechanics is called a *GuideCane*, and a preliminary version of a working prototype has already been tested by visually impaired individuals. Users reactions are said to have been extremely positive, though more development will be required before the device is ready for widespread commercial use.

The GuideCane will steer a blind user around obstacles.

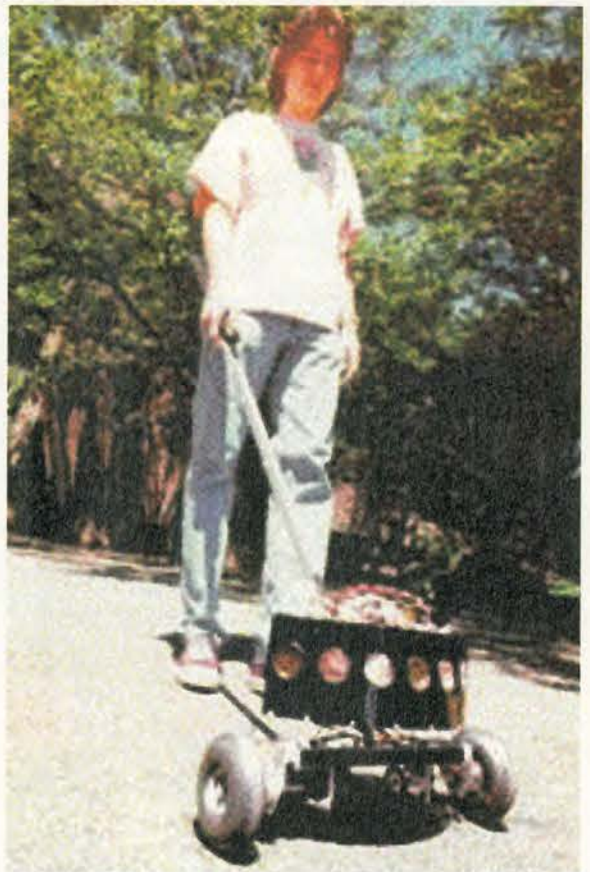


The 3.5kg *GuideCane* consists of a long handle with a thumb-operated joystick for direction control, an array of ultrasonic sensors and a small on-board computer mounted on a two-wheeled steering axle. Users push it ahead with one hand. When the device's ultrasonic sensors detect an obstacle in its path, the computer automatically turns the wheels to steer around the obstacle and resume the original direction of travel.

Steering changes are experienced as a direct physical force through the handle, which should make it easy to follow the GuideCane's path without conscious effort. The body automatically follows the trajectory of the guide wheels just as a trailer follows a car.

Once the obstacle is cleared, the guide wheels resume their original direction."

The Michigan researchers hope that, after a brief adjustment period, most people will become so comfortable that they will be able to navigate around obstacles at their normal walking speed.



Prototype tests have produced good results.

Unlike other navigation aids, there are no complicated acoustic signals to interpret and no extended training period required.

Clock watching can save chip energy

Scientists at the University of Southern California have successfully demonstrated a chip that uses as little as 20% of the energy of conventional models. The new microprocessor – designed by research professor William Athas and colleagues at the USC School of Engineering's Information Sciences Institute (ISI) – recycles energy from the chip's clock, the timing signal normally used to synchronise computing functions.

In conventional chips, the clock consumes a large fraction of the total energy supplied to the chip, all of which eventually winds up as heat.

ISI's experimental chip, called AC-1, has two different clock circuits. It can work with an ordinary clock mechanism, or a flip of a switch will activate circuits that briefly convert the energy of the clock's electric signals into magnetic form.

This captured energy is then reconverted into electrical form and returned to power the data-processing sections of the chip.The energy savings range from 75% to 80%, depending on how fast the clock is set to run (with slower clock settings yielding greater energy savings).

Yet in its energy-recycling mode, the chip is able to perform the very same computing tasks it performed using the conventional clock circuit.

The idea of using the clock to reduce power consumption was first proposed, in 1967, and in the mid-1980s, 'hot clock' chips were proposed and designed for fabrication by the then-standard nmos technology.

Prototypes were made, but nmos was soon eclipsed by cmos, and many researchers thought the hot-clock system would not work. But the ISI scientists hope their success will spark new interest in the idea.

Initial applications of the energy-recycling chips will probably be at the lower-performance, cost-driven end of the chip market. But applications could include portable computers, digital watches, cell phones or GPS position finders. The only difference would be that the batteries would last much longer.

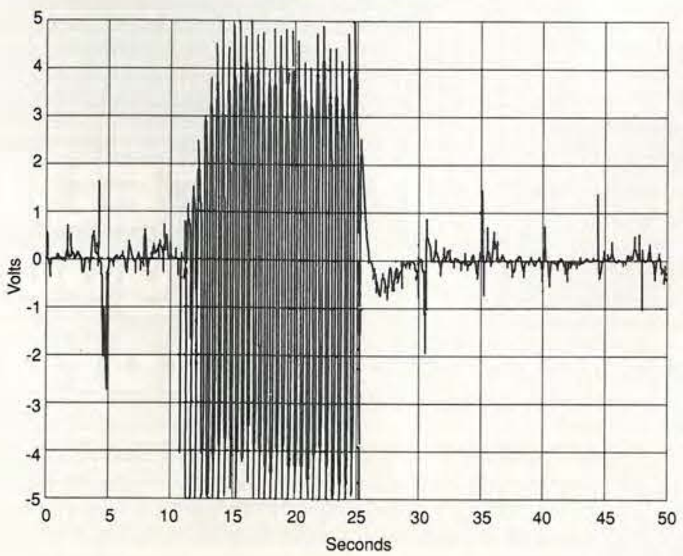
Eventually the researchers expect to double the energy savings with chips that consume only one-tenth the power of conventional chips. ■

For more information contact: William Athas, School of Engineering's Information Sciences Institute, University of Southern California, 3620 South Vermont Avenue, Los Angeles, CA 90089-2538. Tel: 213 740 7600.

Fields, lines and tremors

Dominic Di Mario investigates the world of ultra-low frequencies using three detectors – one indicating electric fields, one for investigating magnetic lines and one for sensing seismic events.

Fig. 1. This anomalous train of waves, detected in Italy, lasts for 14 seconds and repeats at irregular intervals. At times it is rather frequent – one set of waves every 5-10 minutes – but at other times it is absent for days. The frequency is 1.87Hz and it is relatively strong if compared with the other signals in the background. Its origin is anybody's guess, but it is likely that some kind of electric machinery is at work.



Detecting ultra and extra low frequency signals needs something different from a tuned *LC* circuit. Designs incorporating coils weighing several pounds and unyielding capacitors are occasionally implemented by brave experimenters. But there are alternatives if the electric, magnetic and possibly mechanical waves are dealt with separately.

Aperiodic detectors do not tune to a specific frequency and leave the job of frequency determination to other equipment like spectrum analysers. As a consequence, the design can be optimised for best performance in a given application.

Electric fields

Figure 1 shows an oscillation of unknown origin at 1.87Hz. I occasionally detect it with a 3m long antenna just outside my home.

Suspects for the source of the signal are an appliance from a neighbouring house and electric industrial machinery in the area. But, so far I have been unable to discover the source.

When I pushed the detection circuit's sensitivity to its limit, I was able to see all sorts of apparently unrelated signals. The only signal that could be positively identified was a sharp, clean positive peak showing that someone in the area had switched on a television set.

These low-frequency signals are detected by means of a high impedance circuit, **Fig. 2**, based on the *OPA124P*. This difet op-amp is on the expensive side, but its characteristic of stability and low noise make it the ideal choice.

Problems met in the design of this circuit were twofold: interference from the mains and the connection to the antenna. The mains interference problem was solved by inserting a low-pass filter followed by a 50Hz notch filter. Together, these eliminate all mains induced signals – including harmonics.

I took care to make the circuit provide good global transient response with little or no overshoot. Removing or changing one or more of the filter components will change the bandwidth, which is likely to worsen the transient response. Because of the filters, the bandwidth extends to about 10Hz. The system is still useful up to 16Hz, where gain is –10dB, at which point starts to fall off rather rapidly to 50Hz showing, on the simulator, an attenuation of –91dB.

The low-frequency –3dB point is 0.1Hz. This means that dc signals are not amplified. In fact the amplifier is ac coupled. This eliminates all kinds of problems relating to dc stability and coupling, as well as any effect of large electrostatic fields present around the antenna.

It is, however, necessary to balance the dc point of the *OPA124P* in order to compensate for any change induced by the setting of the potentiometer. This is the only adjustment required by the circuit.

When the implementation is complete, switch it on, wait five minutes and then adjust the 100kΩ potentiometer until there is no dc variation when rotating the 220kΩ potentiometer near its maximum. sensitivity.

Catering for high impedance

The second problem was that, due to the very high impedance of the circuit, any length of wire or coaxial cable connecting the input to the antenna would introduce noise and microphonics.

After several tests and trials, I found that the best solution was to connect the antenna directly to the input of the amplifier without any intervening wire or cable. This meant that the first IC had to be mounted directly next to the antenna. Connection to the rest of the circuit is via a five-core shielded cable carrying power, ground, signal and the two connections for the potentiometer.

This solution works very well. The distance between the antenna head-end and the rest of the circuit could be quite long, although I only tested it up to 10m.

Both the setting of the 220kΩ potentiometer and the length of the antenna influence the overall sensitivity of the circuit. It might be convenient to install a telescopic antenna and make the unit portable. This is why I designed the system with two PP3 9V batteries in mind.

Useful signals are detected even with a 5cm antenna, although the interference generated by just walking around is strong enough to bury the signal. A quick test can be carried out using the set up shown in **Fig. 3**. Spikes should be detected whenever one of the poles of a 9V battery is alternatively brought in contact with the screwdriver. The field is enough to generate 1V peaks at the output.

The input of the circuit is protected by low leakage diodes and a series resistor. This is very effective against any overvoltage or mishandling of the input circuitry.

The output goes first through Schottky diodes. This gives a convenient threshold against very low level signals. The result is a zero signal that looks very clean - quite effective if you are using a pen plotter.

Crossover distortion introduced by the diodes is very low and is evident only at very low signal levels. As an alternative you might like to connect the output lead directly to the output of the amplifier. This bypasses the rectifier bridge and gives the full signal – noise included.

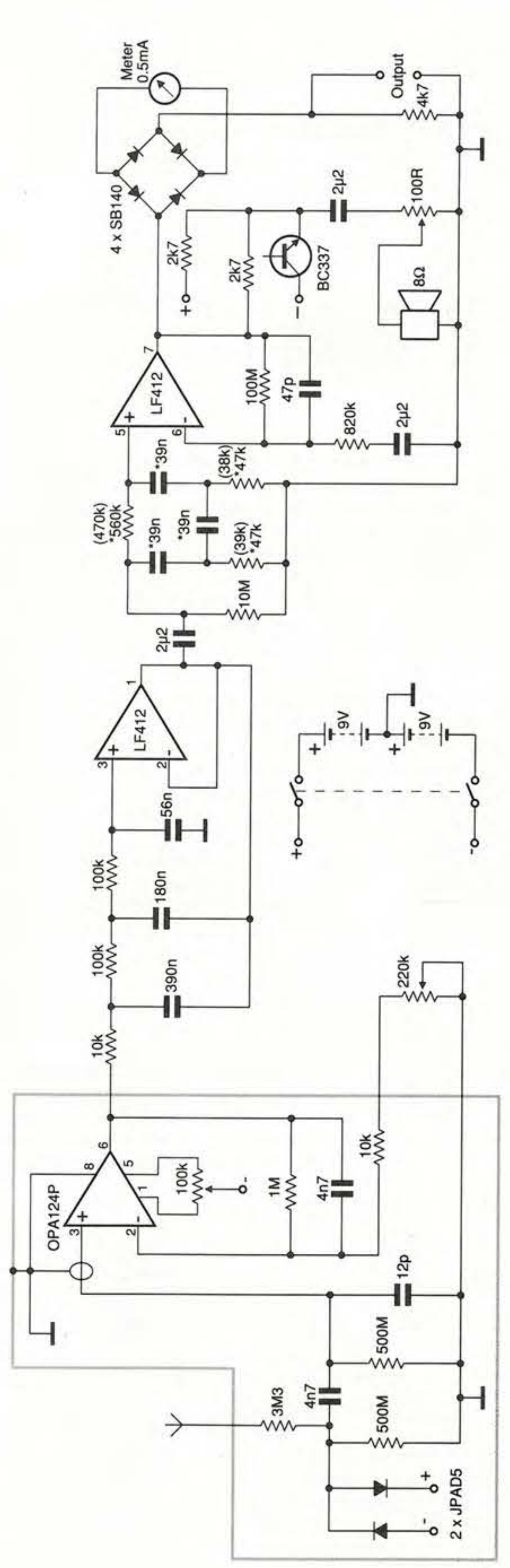


Fig. 2. Complete circuit for the electric field detector. Components marked with an asterisk should be 1%. If 5% components are used, then one of the 47kΩ resistor should be changed to a series of a 39kΩ resistor and a 22kΩ trimmer adjusted for the minimum mains noise. Values between brackets are suitable for a 60Hz mains. An American equivalent to the BC337 for the audio oscillator is the 2N2222.

continued on page 808...

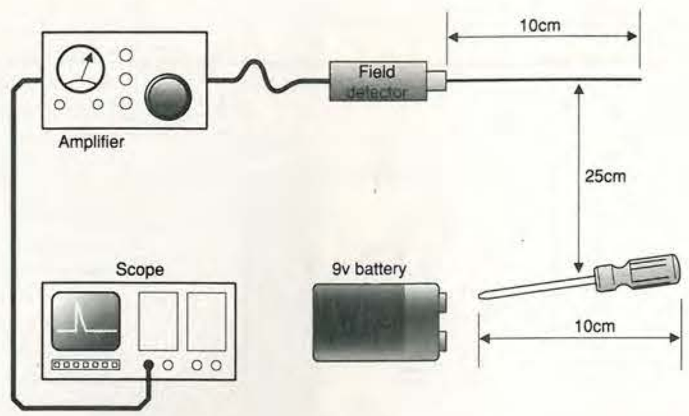


Fig. 3. Test bench for the electric field detector. Charging the screwdriver to the battery potential generates an electric field. This is detected as a 1 V glitch on the scope. The pulse can be positive or negative depending on the battery pole in touch with the screwdriver. The pulse is also detected by the movement of the meter and the change of frequency of the audio oscillator. Both the battery and the screwdriver must be well isolated from the other equipment.

An acoustic signal is available from a frequency modulated audio oscillator. This is useful if you intend to use the meter as a portable instrument.

Note that that mains induced noise can be quite high and could easily saturate the preamplifier. Consequently there is a practical limit to the length of the antenna. In the metropolitan area where I tested the meter, a 3m enamelled copper antenna generated 5V pk-pk at the output of the preamplifier, before the filters. This means that probably a 6m antenna is the longest that could be used under these circumstances. In a rural area the situation will certainly be better, allowing longer antennas.

Implementing the circuit

On a practical note, if you have difficulties obtaining the 500MΩ resistors, connect 100MΩ resistors in series. I recommend that you use high quality input capacitors, mylar or similar, but definitely no ceramic capacitors.

The Burr-Brown data sheet for the *OPA124P* suggests using a printed ground ring around the input lead and connection to ground of the substrate, pin 8. It also suggests direct soldering of the IC to the circuit board.

The preamplifier was assembled in a small metal box with the shielded cable on one side and the antenna wire on the other. If it is to be permanently placed on the outside, a convenient shelter should be foreseen against weather extremes. The temperature range of the IC, from –25° to 85°C, should not cause any problems in normal operating conditions.

Some form of data logger or a pen plotter is essential part of the system. I used the *ADC42* data logger from Pico Technology, connected to the parallel port of a pc. The *ADC42* software also includes a spectrum analyser virtual instrument, so it is a convenient way to analyse the the signal.

I now have several traces of strange signals, oscillations, spikes – even square waves. My next task is to correlate them with known events such as earthquakes or signals that are supposed to make the round trip of the world generating a 7.5Hz oscillation.

Magnetic lines

The sensor that first springs to most people's mind when they think of implementing an electronic compass is a Hall-effect device. But there are alternatives – like the one suggested in **Fig. 4**. This is a classic oscillator running at around 1MHz.

The ferrite rod is almost identical to the one normally found in medium wave receivers – the longer the better. The only difference is the strong magnet attached to one end of the rod.

With the magnet attached, the exact frequency of oscillation now depends on the prevailing magnetic field. If you have a second oscillator, it is possible to detect the beat note which depends on the orientation of the magnetised oscillator with respect to the Earth's magnetic field.

An easy test requires a medium-wave radio tuned to a station, preferably a weak one. Next, tune the oscillator in such a way to generate an audible beat note in the radio. Any minute movement of the rod will now change the audible beat frequency.

I plotted the frequency variation against rotation of the rod, **Fig. 5**. The first curve shows the frequency of the beat note starting from 200Hz and the rod pointing north. The second curve shows the beat frequency, starting from 200Hz but with the rod pointing east.

In both cases, the rotation was clockwise and extended for 180°. For this test, I used a second oscillator because the interference of the transmitting station was too strong.

A complete circuit with two oscillators and ancillary circuitry was the subject of a circuit idea published in *Electronics World+Wireless World* in December 1994.

When prototyping the circuit, take care to use adequate electric shielding around the oscillator. Metal boxes are not suitable. The metal would create loops that would kill the oscillation. A solid plastic box with copper wires or strips, glued on the inside, running parallel to the rod and soldered to ground at one point only is the best solution. A small hole is made in order to adjust the trimmer capacitor with a plastic screwdriver.

As this detector reads the effective direction of the magnetic lines, you will have a change in frequency for any movement of the oscillator both in the horizontal and vertical planes.

Listening to the beat note shows an incredible variety of signals, **Fig. 6**. The situation is much quieter in rural areas, but in a large town you have to wait until late at night – between 1 and 5 o'clock in the morning – if you want to see less frenetic magnetic activity, **Fig. 7**.

Where all this activity comes from is a mystery to me, although I am sure that most of it is man-made. Frequency changes, hence magnetic variations, take place at a very low rate; 4-5Hz is the fastest you may expect. This detector operates from 0Hz, so it will track changes lasting hours, limited only by the long term stability of the oscillator.

The 1MHz oscillation frequency was chosen for convenience. Its radiated field will not be detected by a radio beyond a 2m radius. Nevertheless, a more suitable frequency could be used. Bear in mind that a higher frequency will increase sensitivity but stability might become a problem. Conversely, lowering the frequency makes the circuit less sensitive but more easily stabilised.

Seismic waves

The *OPA124P* and the idea behind the electric field detector was the basis also for the seismic detector, **Fig. 8**. The circuit is straightforward and requires no adjustment. The amplifier is local to the detector element. This solves the problem of connecting a high impedance source to the amplifier without introducing any further degradation.

A three-core shielded cable connects the amplifier to the other elements of the circuit and to the power supply, which is ±15V.

The detector is the piezoelectric element of a kitchen gas lighter. Although its electrical characteristics were unknown to me, it performs well. The problem with this circuit lies in the proper mechanical construction of the detector. It should be on a solid base stuck in the ground, with the detector placed vertically with a 1-1.5kg weight on top. This weight should not have a proper resonant frequency, so a loose package of sand, fine gravel, or even salt, sealed against humidity, works fine.

The weight should be in contact with the detector only and kept in place using vibration absorbing material such as shock absorbant rubber, plastic sponge and the like. The whole set up should be screened against electric fields with a non magnetic screen such as aluminium or copper, and should be enclosed in a sound proof box, shown over.

Of course, you do not have to go to such an extent to see the circuit's basic operation. You can place it on a table if you are happy to see

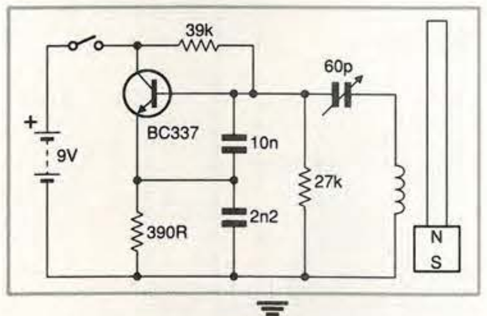


Fig. 4. A basic oscillator detects the Earth's magnetic lines once a magnet is glued to one end of the ferrite. The beat frequency with a medium-wave transmitting station falls in the audio range and can be heard on a radio and measured. Even tiny movements of the rod or the impinging magnetic lines causes an audible change of the beat note. If high stability is required it is advised to build a second identical oscillator instead of using a commercial station.

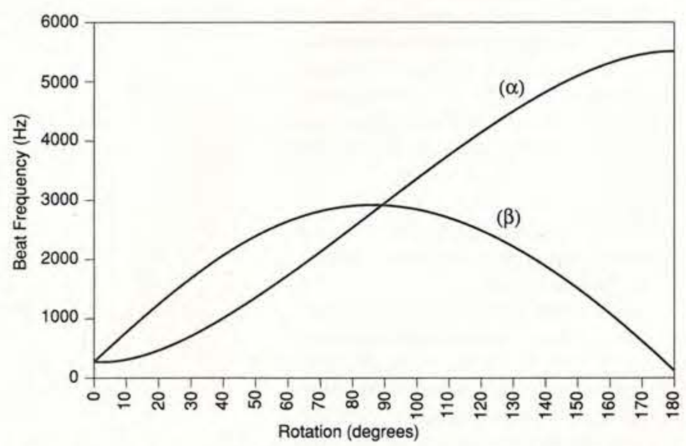


Fig. 5. Curve (a) refers to the north-south orientation of the ferrite rod and curve (b) to the east-west. Rotation takes place on an horizontal plane. If the rotation plane changes, the frequency range of the beat frequency will also change since the magnetic lines are parallel to ground only at the equator. Consequently, an accurate measurement must take in account the latitude, 45° in this case, and the plane.

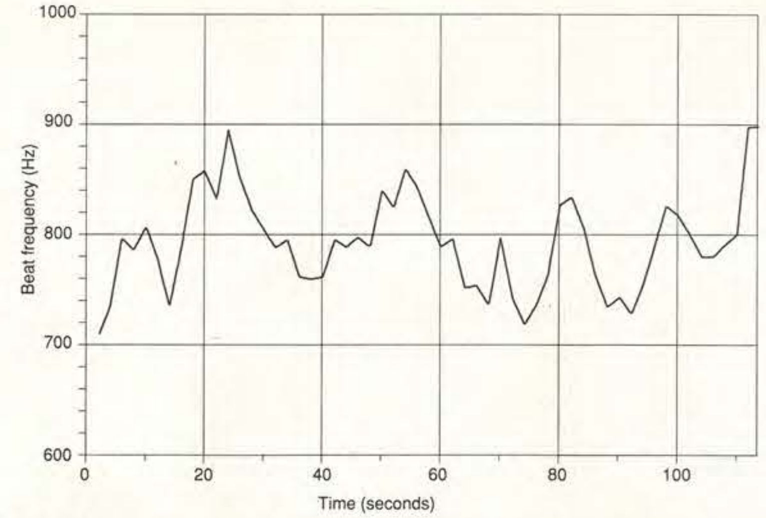


Fig. 6. Printout of from data logger shows the variation of the beat frequency over a period of 113 seconds. Samples where taken at 1s intervals at 6pm in a built-up area. In open country the response is much quieter with several minutes without appreciable changes.

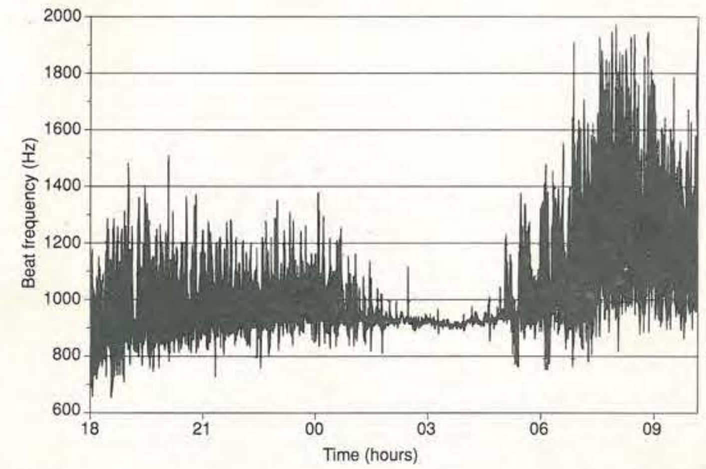


Fig. 7. Data collected from the magnetic detector over a period of 16 hours, from 6pm to 10 am the following day. Large variations take place during the whole day except for a short lull around 4am. This is also the best time to carry out testing of the unit. Data were collected in a large city. The same device shows an almost flat curve only in open country, away from human activity.

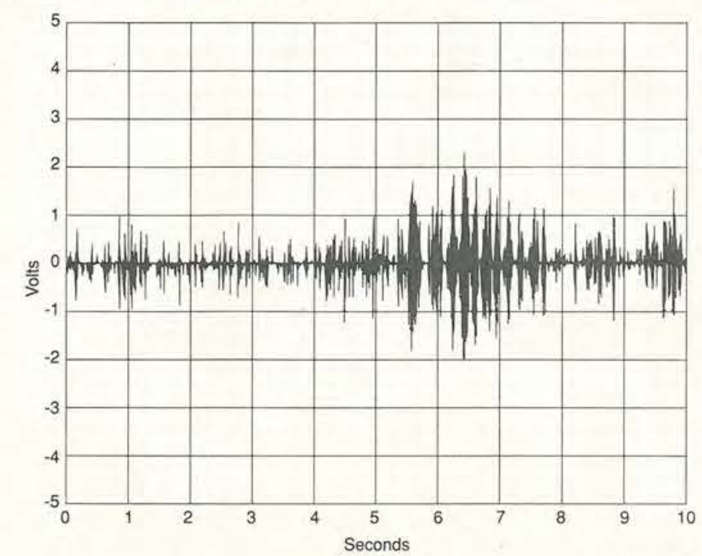


Fig. 9. Seismogram of a lorry, barely heard in the silence of the night, passing in the area and measured over 10s. The frequency range of the signal is mostly between 10 and 50Hz and is bound to interfere with a real seismic wave. This demonstrates that careful choice of the installation site is important if you want to record natural phenomena.

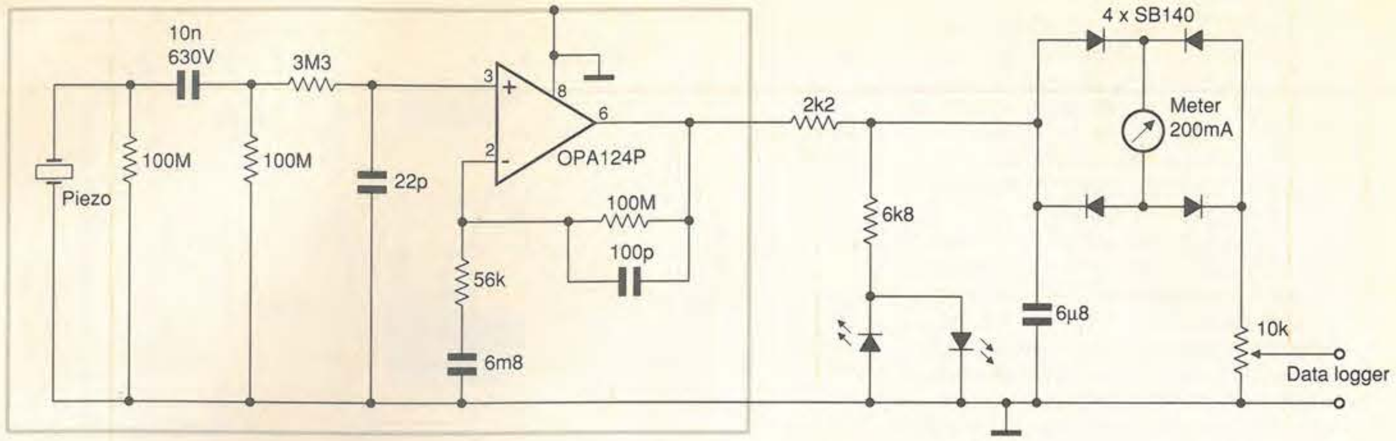


Fig. 8. There are no settings or adjustments in this seismic detector. The screened portion is placed together with the piezoelectric detector and connected to the remaining part of the circuit with a three-core shielded cable. Piezoelectric gas lighters are economical and readily available. The element from any such lighter should be suitable for the detector.

the resonant frequency of the table – and probably your house. Two leds flash alternatively when a seismic wave is detected and the overdriven meter gives a clear reading, even with small signals.

Road and railway traffic are the main interfering signals. They could mask a real seismic wave. You can learn to distinguish the signature of interfering signals but it will still be a problem to recognise a seismic wave simply because they do not happen frequently enough – fortunately.

The frequency range of the circuit is from 0.44 to 11Hz with a –10dB point at 24Hz. In seismological terms, this is a narrow band detector, as opposed to a 0.1 to 100Hz alternative, which is defined as a wideband detector. This detector has a smooth filter so signals outside the nominal bandwidth will be detected. But as most of these signals are human artefacts, **Fig. 9**, a certain amount of attenuation is preferable.

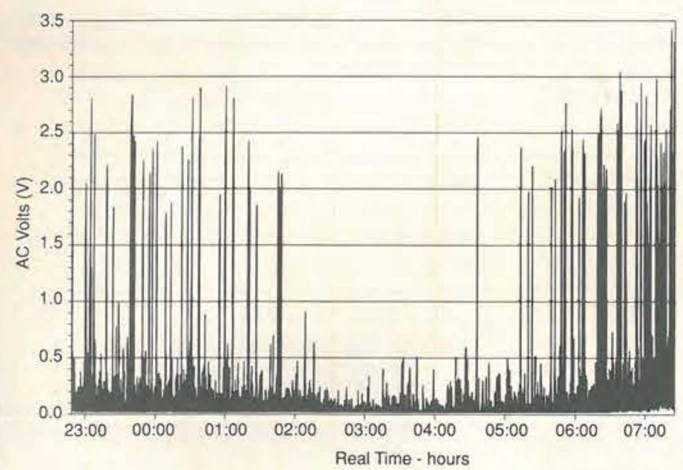


Fig. 10. Printout of data logged from 11pm to 8am the following day. Readings where taken at 1sample/s and shown as an alternating voltage. The 10kΩ potentiometer was adjusted to accommodate the peak readings of the local tramway. Also in this case there is quiet period around 3-4 am underlining the need to install the detector away from cities, roads and railways.

Morning quiet

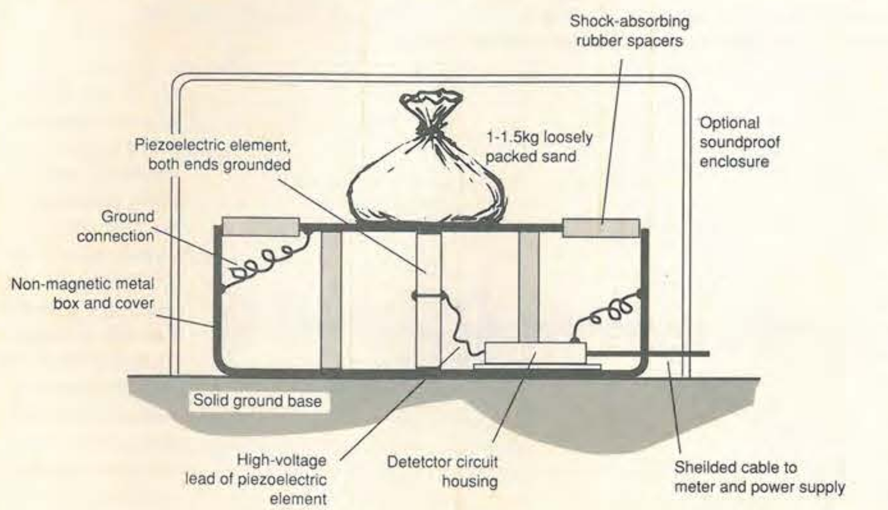
As with the magnetic detector, I noticed a quiet period during the early hours of the morning with the seismic detector, **Fig. 10**. At other times, it is as frantic as the magnetic or the electric detector. There is probably a relationship that only a multichannel recorder could confirm.

Sensitivity seems to be at the right level. With an overall gain of 60dB, the circuit amplifies a hiccup in close proximity, but I did not attempt to properly calibrate the device.

The seismic waves detected are the vertical component of the surface wave. Adding two further detectors, rotated by 90°, would also allow you to record the horizontal components.

Take care when handling the detector. Hitting it even gently will generate hundreds of volts – thousands of volts in the worst case. So take safety precautions to protect yourself, and the input circuit, where the only protection is via the 3.3MΩ series resistor. This protection is sufficient for normal handling of the detector but it is advisable to keep the input shorted during installation.

If the purpose of the above circuits is to detect natural phenomena, then they should be operated away from cities or industrial sites, unless of course, you are looking for man-made activity. But wherever you are, you are bound to discover many of the mysteries, quirks and anomalies of the world of ultra-low frequencies ■



Practical implementation of the seismic detector. Most of the circuit is housed together with the detector. There are no critical components. The rubber that keeps the weight in place must not transmit vibrations coming from the ground. Spongy rubber works fine. The sound proof enclosure prevents low-frequency sound – typically from helicopters – from interfering.

His Majesty's Voice – or how the GPO became involved in the record business during World War II – researched by Andrew Emmerson

Post Office war records

More by accident than by design, the British Post Office became involved in many strange activities during World War II. Its work devising the Colossus, Cobra and Tunney devices for breaking enemy codes at Bletchley Park is now fairly well known, whereas similar work producing high-speed computers for co-ordinating anti-aircraft gunnery and for timing bomb release has not received the same recognition.

Other less well-known episodes include the freight trains run by the Post Office tube railway in London. These carried spoil from new bunkers under construction and brought it to the surface away from the sensitive location. And there was the assistance the Post Office submarine cable team provided for the design of 100 miles of hollow three-inch cable which carried vital fuel across the Channel to support the Allied armies in the invasion of Europe.

But perhaps the most unlikely of all unusual assignments was the Post Office's involvement in the development of hi-fi recording techniques. During World War II, materials for recording radio programmes were so scarce in Britain that the Post Office was forced to commandeer a factory to produce blank recording discs. In so doing, its boffins made significant improvements to recording techniques.

"The BBC alone used 7000 blank discs a week"

The BBC relied on vast quantities of blank discs for sound recording. The corporation alone used 7000 a week for making transcriptions of its own programmes and the enemy propaganda monitored by its listening stations. The armed forces – and the secret services, it is rumoured – also made heavy demands on the supply of direct-cut discs; they were used for covertly recording the conversations of prisoners-of-war and also for capturing weak radio signals. By replaying the weak signals many times, it was hoped that they would eventually be read correctly.

The reliance on these discs was so great that it was felt necessary to put their manufacture under closer direction. To safeguard supplies, the Post Office took control of one of the two factories which produced them. They and the BBC also improved record cutting head technology and a microgroove recording cutter using these blanks was first used on VE-Day.

This is to anticipate events, however. The story really starts back in the mid-1930s when Cecil Watts invented a sound recording technique using a metal disc coated with cellulose nitrate lacquer. People often call these discs 'acetates' but this title is a misnomer.

"With its high signal-to-noise and power versus bandwidth ratios it has not been improved upon"

This direct recording disc was and is a remarkably high quality sound recording medium. With its high signal-to-noise ratio and power versus bandwidth ratio it has not been improved upon. Surface noise is low and in fact the technique is still used today for mastering some albums. The discs themselves were fairly soft, the disc cutter being a sapphire cutting tool.

But this was not the only method. In 1939 the commercial record companies were still using wax for mastering – an inferior process. The BBC, did however, use these cellulose nitrate discs for short-term sound archiving. In addition, they had two other recording systems at their disposal. These were the Marconi-Stille – which recorded magnetically onto steel tape – and the Philips-Miller – which made an optical sound track on a material similar to cinema film.

With the war under way, these other two systems hit problems. The special steel for the Marconi-Stille recorders came from Sweden but supplies ceased in August 1940, while the optical film for the Philips-Miller machines came from Gevaert in Antwerp. After this city had fallen to the Germans in May 1940, supplies of this tape also ceased.

Alternative sources were tried for both technologies but, none was ideal. This left the BBC dependent on the direct recording cellulose nitrate discs alone at a time when the demands on recording for broadcast purposes was mushrooming.

"In effect, the manufacture of these discs was nationalised"

Unprepossessing they may look but blank discs – similar to these – for one-off recordings played a vital role during WWII. So important were they to the war effort that the Post Office commandeered a factory to safeguard supplies (photo from National Sound Archive).

There were only two sources of the blank discs, EMI and Cecil Watts' firm, MSS Recording Company. In the summer of 1941 it was decided that the MSS factory should come under the control and management of the Post Office, which also provided additional capital to boost output. In effect, the manufacture of these discs was nationalised. Two scientists from the Dollis Hill research station, J.F. Doust and F.G. Hopwood, were put in charge of this activity.

In a separate initiative, the BBC devised an improved recording cutter and this cutter used these MSS blanks. The frequency response extended beyond 10kHz and so it could be called high fidelity; later on it could and did cut microgroove records. The new system was first used on VE-Day.

As well as blank discs, Cecil Watts's MSS company had made disc-cutting machines since the mid 1930s. These also came under the control of the Post Office when they took charge in 1941.

Improvements to the MSS disc cutters were also made during the period of Post Office management. It can truly be said that the Post Office played a part in improving recording technology – even if it was more by historical accident than by previously planned intention. The Post Office did not, however, invent the microgroove record, as some people have alleged.

"The story of the GPO's wartime involvement in acoustics doesn't end here..."

The story of the GPO's wartime involvement in acoustics doesn't end here, though. There were many other interesting projects. Special noise-cancelling microphones and headphones were devised by the Post Office for tank use.

Deep below the Post Office's Dollis Hill research station, a special chamber with deafening sound effects was constructed to simulate a tank's interior. Producing the sound effects with which to test these devices was another matter; recordings on 78rev/min gramophone records would not last long enough and the chosen solution made good use of other Post Office technology.

First of all, recordings of a tank rumbling past a microphone were made on direct-cut disc. Ten-second recordings made in this way were transferred to 35mm film by the Crown Film Unit – a new name for the old Post Office Film Unit – and a two-second long section finally transferred to glass disc, for playing continuously on a speaking clock machine. Apparently the joint in the recording and the two-second repetition were not noticeable in use.

Similar machines were made for the Royal Air Force; these generated continuous background aircraft noise effects for training radio operators. Four different aircraft types were covered, at normal speed and at absolute maximum.

Yet another machine was developed for training fighter pilots; this was a twin-channel simulator, providing continuous aircraft noise and spasmodic machine-gun effects, as and when required.

At the end of hostilities all this development activity came to an end. The MSS factory was handed back to Cecil Watts and all these matters were effectively forgotten. At this distance in time nonetheless they form a fascinating sideline in the history of sound recording and reproduction.

My thanks go to Peter Copeland of the National Sound Archive for his assistance in preparing this article. Some details were taken from the book 'BBC Engineering 1922-1972' by Edward Pawley, and others from an IEE paper by Arnold Lynch entitled 'Some Derivatives of the Speaking Clock.' ■

SPEAKERS' CORNER

Reaction to John Watkinson's iconoclastic view of loudspeaker design in 'Loudspeakers Exposed', June 1997 issue, revealed how much interest there is in this topic. In this new monthly column, John continues his fresh and often controversial look at loudspeaker technology and looks forward to exploding a few myths along the way.

Loudspeakers are everywhere. So it's not hard to study and compare them. Mostly I compare them with the original sound. Even casual inspection reveals that, with rare exceptions, most of today's loudspeakers fall far short of what is desirable – and look awful into the bargain.

There is no technological reason for this state of affairs; in fact we have never been better served with advanced materials and the means to form them.

Looking on the bright side, if the vision exists, it only requires some good industrial design to have much better sounding and looking loudspeakers than are generally available. In this series I intend to explore some possible avenues.

The only fundamental criteria we have for loudspeaker performance are purely subjective. It follows that a loudspeaker designer must understand psychoacoustics, quantitative as well as qualitative. As human beings vary, we have also to understand statistics in order to appreciate that psychoacoustic data are not absolute like classical physics but are subject to variation.

If psychoacoustics helps us to define what performance our loudspeaker must have, it does not tell us how to achieve it. We can stir the air in a variety of ways, but seldom directly. Generally we need to cause some object to vibrate, and subsequently transmit those vibrations to the air. The study of what happens next is known as acoustics.

Causing precisely controlled vibration requires a knowledge of how masses move. Control theory tell us enough about electromagnetic and electrostatic actuation of diaphragms. To follow what goes on inside the speaker we need to know thermodynamics because we are compressing a gas, and it matters whether we do that isothermally or adiabatically.

As listeners we are primarily interested in what happens outside the speaker; in the far field to be precise. It is the job of the cabinet to separate these two results: it must be utterly inert and resist the reactions from transducers and the internal pressures without any movement or flexing. Most of today's loudspeaker cabinets are very poor in that respect, yet a basic knowledge of structural engineering and materials science will help to design a suitably rigid structure.

In order to anticipate and control the high frequency sound output of a loudspeaker, it is important to understand the wave or diffraction theory of acoustics. The concepts of coherence, phase and interference, even chaos theory, become crucial to stereo imaging, yet are widely neglected. In fact light and radar behave in exactly the same way as sound – although on different scales – because they are wave motions. A knowledge of wave optics is extremely useful in loudspeaker design because it allows the diffraction behaviour of cabinets to be predicted.

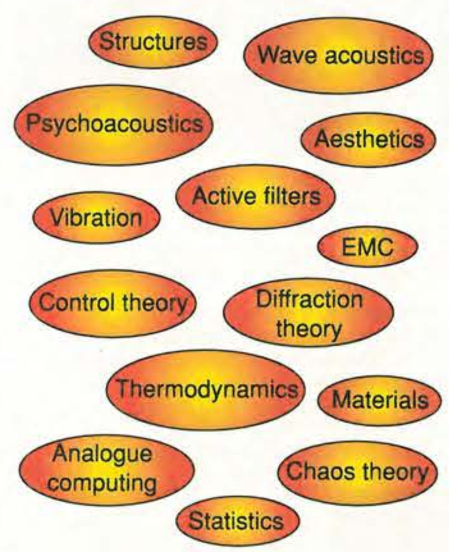
As passive speakers simply cannot meet modern criteria, the ability to design electronic systems has become essential. This encompasses not only amplifier design, but also active filter and analogue computing implementations. Emc awareness is important too, but knowledge of speaker cables can be safely set aside as we won't be using any.

Although I consider the above list of skills to be a minimum for the development of advanced loudspeakers, such combinations are rare. Little wonder that the average loudspeaker today is so primitive. Many are developed using pure empiricism. The mechanism is not understood but the parameters are changed until it sounds best.

Unfortunately a speaker designed in this way may only sound good on a certain type of music. The design may then appear in a cookbook to be copied parrot fashion with even less understanding of the principles.

Often the designer is forced by economics to use sub-optimal techniques. Here the home constructor may be at an advantage because he can fabricate complex shapes which could never be produced economically. ■

John Watkinson, FAES



The disciplines required for advanced loudspeaker design are shown here. Their combination into a working product allows so many possibilities that the process has to be described as an art.

ng
SE

ystal, the circuit operated much as , once the tank circuit was retuned from vious value of 10.4MHz to the series nt frequency of the crystal – a shade 10MHz. The series resonant frequency e equivalent series resistance R_s and parameters of the crystal were deter- beforehand, as detailed in the panel.

tz to the rescue

utput of the oscillator, now crystal con- l, was connected to the phase noise dis- ator as shown in Fig. 2 of the earlier . The only change made to the phase discriminator was the addition of an 2.7m of coaxial cable to the delay cable. vas to bring its length back to $^{15}/_4$ at the 10MHz operating frequency.

dentally, the length of the cable used for rlier 10.4MHz oscillator was quoted in n the previous article as 108.4m. That n fact the equivalent length in air: the al length calculated out as 72.3m. Thus ble represented the best part of what was ally a 100m reel.

nnected the crystal oscillator of Fig. 1 to ase noise discriminator. Everthing was set up and left working for a while to reach equilibrium. The phase noise out to 5kHz offset from the oscillator frequency was then measured using the *HP3580A* audio frequency spectrum analyser, and the trace stored in its digital store.

As in the previous article, a 56kΩ resistor was then connected in parallel with R_1, resulting in some 10dB increase in output level, with the oscillator transistor bottoming each cycle. The phase noise out to 5kHz was again measured and displayed, this trace and the stored trace both appearing simultaneously on the HP3580A.

The resulting display was photographed, **Fig. 2a**). As you can see, the two traces are virtually identical. Indeed, it is only the two

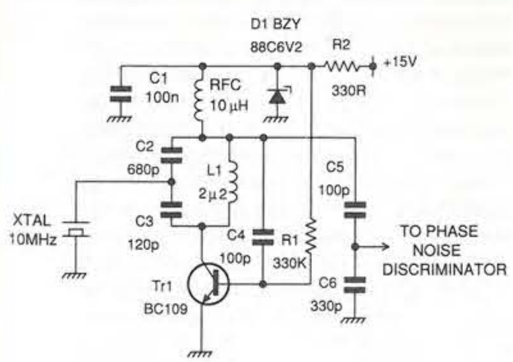


Fig. 1. The LC oscillator of Ref. 1 modified for crystal control.

My earlier article showed that when the resonator is an *LC* 'tank circuit', with the limited value of Q thus obtainable, the phase noise is lower if the design is such that the transistor maintaining amplifier is not allowed to bottom. I was curious to know if this was still the case with a much higher Q resonator, such as a quartz crystal.

To enable a direct comparison to be made, I used the same oscillator circuit as in Fig. 2 of the previous article, but modified to incorporate a 10MHz crystal, **Fig. 1**. In the earlier circuit, the junction of the two capacitors tuning the inductor in a Colpitts oscillator was grounded directly. In the new circuit of Fig. 1, it is grounded via a 10MHz crystal.

Given the low equivalent series resistance of

Killing NOISE

Oscillators with very low phase noise are crucial in modern, fast data radio links with high spectral efficiency. Relatively easily achieved in a crystal oscillator, low phase noise is a much more daunting design task in a wide range vco, as Ian Hickman explains.

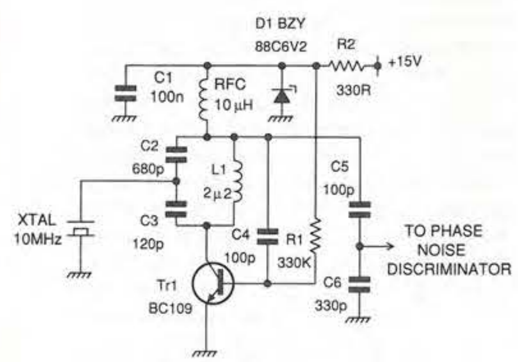


Fig. 1. The LC oscillator of Ref. 1 modified for crystal control.

How do you make a 'quiet' – i.e. spectrally pure – oscillator? In an earlier article[1], I highlighted the need for oscillators with very low close-in phase-noise sidebands. The characteristics of an oscillator to fill this requirement are well known, and include the following:

- a high-Q resonator – tuned circuit, crystal, SAW resonator or microwave cavity;
- a low - noise transistor or other active component as the maintaining amplifier – especially with low $^1/_f$ noise, as this cross modulates onto the output;
- a high-gain maintaining amplifer stage so that it can be lightly coupled to the resonator which thus works at close to its unloaded Q;
- a buffer stage for the output, again enabling the resonator to work at close to its unloaded Q;
- a large amplitude of oscillation to ensure the stored energy at the desired frequency is greatly in excess of circuit noise;

My earlier article showed that when the resonator is an *LC* 'tank circuit', with the limited value of Q thus obtainable, the phase noise is lower if the design is such that the transistor maintaining amplifier is not allowed to bottom. I was curious to know if this was still the case with a much higher Q resonator, such as a quartz crystal.

To enable a direct comparison to be made, I used the same oscillator circuit as in Fig. 2 of the previous article, but modified to incorporate a 10MHz crystal, **Fig. 1**. In the earlier circuit, the junction of the two capacitors tuning the inductor in a Colpitts oscillator was grounded directly. In the new circuit of Fig. 1, it is grounded via a 10MHz crystal.

Given the low equivalent series resistance of the crystal, the circuit operated much as before, once the tank circuit was retuned from its previous value of 10.4MHz to the series resonant frequency of the crystal – a shade below 10MHz. The series resonant frequency F_s, the equivalent series resistance R_s and other parameters of the crystal were determined beforehand, as detailed in the panel.

Quartz to the rescue

The output of the oscillator, now crystal controlled, was connected to the phase noise discriminator as shown in Fig. 2 of the earlier article. The only change made to the phase noise discriminator was the addition of an extra 2.7m of coaxial cable to the delay cable. This was to bring its length back to $^{15}/_4$ at the lower 10MHz operating frequency.

Incidentally, the length of the cable used for the earlier 10.4MHz oscillator was quoted in error in the previous article as 108.4m. That was in fact the equivalent length in air: the physical length calculated out as 72.3m. Thus the cable represented the best part of what was originally a 100m reel.

I connected the crystal oscillator of Fig. 1 to the phase noise discriminator. Everthing was set up and left working for a while to reach equilibrium. The phase noise out to 5kHz offset from the oscillator frequency was then measured using the *HP3580A* audio frequency spectrum analyser, and the trace stored in its digital store.

As in the previous article, a 56kΩ resistor was then connected in parallel with R_1, resulting in some 10dB increase in output level, with the oscillator transistor bottoming each cycle. The phase noise out to 5kHz was again measured and displayed, this trace and the stored trace both appearing simultaneously on the HP3580A.

The resulting display was photographed, **Fig. 2a)**. As you can see, the two traces are virtually identical. Indeed, it is only the two

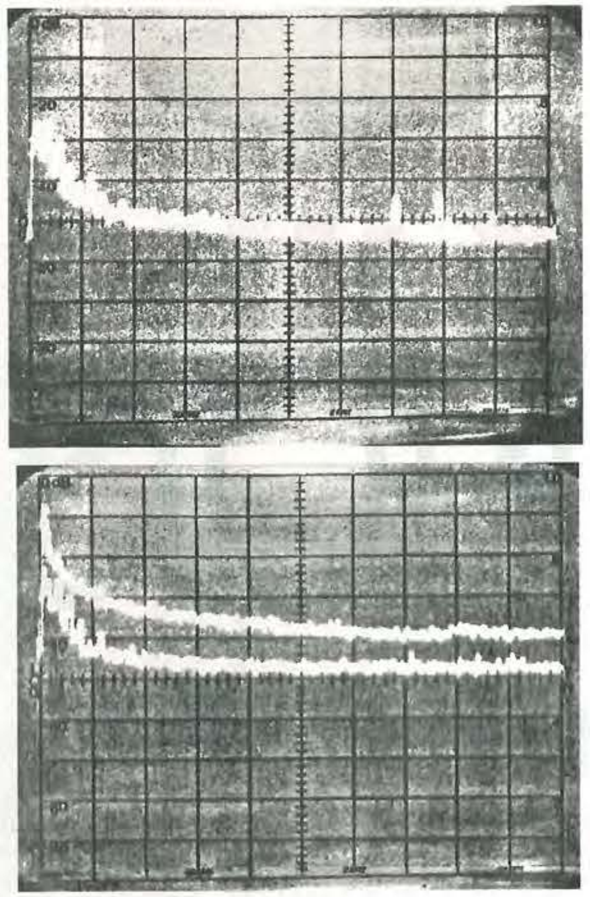


Fig. 2a). Close-in phase noise of the crystal controlled 10MHz oscillator, out to 5kHz offset. Span 0-5kHz, reference level, top of screen, is -60dBV, 10dB/division vertical, analyser bandwidth 30Hz, post detector smoothing maximum, sweep speed 100 seconds per horizontal division. The two traces – transistor bottoming or not bottoming – are superimposed, being nearly identical. b) The same oscillator without crystal control. Upper trace, bottoming, lower trace not bottoming. Spectrum analyser settings as in a).

odd spikes of mainsborne interference on one of them, three quarters of the way across the screen, that show there really are two separate, superimposed, traces.

Furthermore, both traces show lower phase noise than the non-bottoming version of the oscillator without crystal control – at least above 1kHz offset. This was shown in the previous article, and is reproduced here as Fig **2b)**, lower trace.

The conclusion, then, is clear. A resonator with a Q of 40 000 can make up for the deficiences of a mediocre maintaining oscillator. But on the other hand, the tuning range of a crystal controlled oscillator is limited to a few parts per million, or a few tens of parts per million at most.

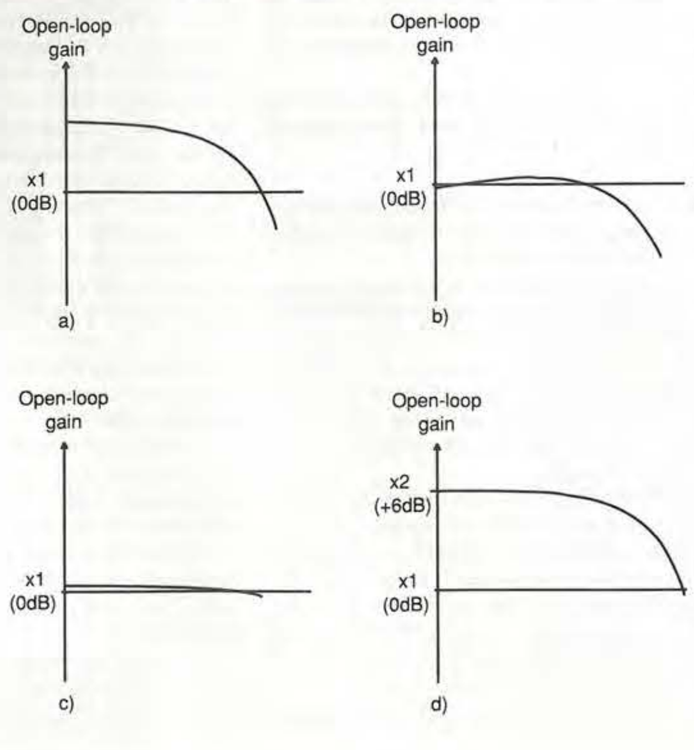


Fig. 3. Various types of open loop gain versus amplitude characteristics which may be encountered – see text.

Where a greater tuning range is needed, as is usually the case in the voltage-controlled oscillator of a synthesizer, then a non-crystal controlled *LC* oscillator is required, and a non-bottoming maintaining amplifier is much to be preferred.

Avoiding the noise?

Even when the maintaining transistor of an oscillator does not bottom, it still operates in a non-linear mode, cutting off for part of each cycle. This results in noise – particularly the device's $^1/_f$ noise – cross-modulating onto the desired sinusoidal output. The lower order non-linearity of the non-bottoming circuit results, as has been shown, in lower close-in phase noise.

Normally, oscillators rely on nonlinearity to stabilise the amplitude of the oscillation. Initially, at switch-on, as the amplitude builds up from nothing, the transistor operates linearly. The loop gain at the operating frequency is in excess of unity. But as the amplitude increases further, the device is driven into non-linear operation, and the loop gain at the fundamental falls to unity. This is illustrated in **Fig. 3a)**, which shows the open loop gain of the stage at the centre frequency of the tuned circuit.

In principle, this can be plotted by opening the loop and applying a gradually increasing external signal at the closed loop frequency of oscillation. But of course, to be representative, the signal must be applied from the same source impedance which the transistor sees in closed loop operation.

Likewise, the loading of the transistor's output impedance on the tank circuit must also be representative of closed loop operation. Both of these conditions could be met by breaking the loop and embedding the oscilator in a long chain of identical stages, as shown in **Fig. 4.**

The highlighted stage sees the appropriate source and load impedances, and if the chain is long enough, it will operate at unity gain. For the earliest stage will selectively amplify the noise at the centre frequency of the *LC* tank circuit, and subsequent stages will further amplify and band limit the signal, so that the boxed stage runs in limiting at the same amplitude as if it were operating closed loop.

If you think such a long chain of identical circuits is too expensive a proposition, even conceptually, a much shorter chain – with the output of the last stage connected back to the input of the first – will do as well or better. For by definition, each stage operates at unity gain and zero phase shift. A ring of just three, or even two such stages illustrates the idea.

Figure 3a) shows the open loop gain versus amplitude of a surefire oscillator, one where the gain at low amplitude is well in excess of unity. A badly designed oscillator with a characteristic like Fig. 3b) is occasionally encountered, and infuriating it is, too. It usually starts up all right, due to the switch-on transient kicking it into life. But just occasionally it fails to start – usually just when it's needed most.

It is tempting to think that an oscillator

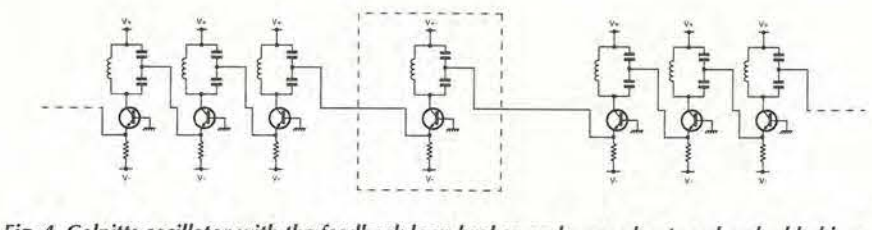


Fig. 4. Colpitts oscillator with the feedback loop broken and opened out, and embedded in a long chain of similar stages.

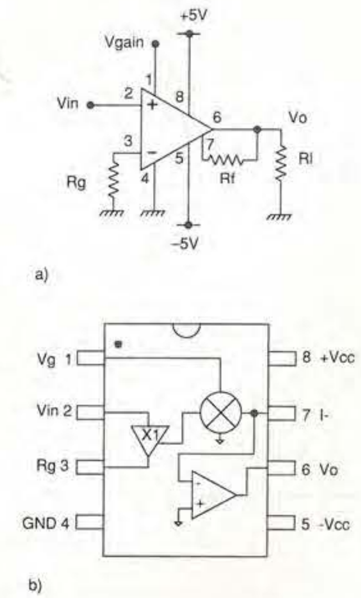


Fig. 5a). The CLC5523 conected as a non-inverting amplifier.
b) Device pin-out. This is the same for both the DIP and SOIC versions, but note that when using the DIP version, pin 4 must be grounded via 25Ω rather than directly.

where the transistor runs in a very nearly linear regime, would show very low close-in sideband noise. Such an oscillator would show a very low level of harmonic content, not merely on account of the filtering effect of a high-Q tank circuit, but as a result of the (near) linear operation.

The absence of non-linearity should avoid cross-modulation of the device's low frequency noise onto the carrier, as close - in noise sidebands. The open loop gain versus amplitude characteristic of such an oscillator is shown in Fig. 3c).

Unfortunately, the very shallow angle at which the characteristic crosses the unity gain line means that the amplitude is not well defined. Such an oscillator is thus extremely sensitive to any external influences, such as signals in the vicinity of the centre frequency. Adjust the loop gain down a little until the stage just barely oscillates, loosely couple in a 'sling-out' aerial consisting of a few feet of wire, plus a diode detector, and you have quite a sensitive radio receiver.

A bright idea tested

In my earlier article, I mentioned the Comlinear *CLC5523* variable-gain amplifier from National Semiconductor, together with its possible application as the maintaining amplifier in an rf oscillator. The scheme proposed was to use the device enclosed in an automatic level control loop, so that the amplitude of oscillation was controlled without any non-linear operation on the part of the maintaining amplifier.

Within the *CLC5523* are a closed-loop input buffer, a voltage controlled gain cell and an output amplifier. The input buffer is a transconductance stage, whose gain is set by the gain setting resistor R_g, **Fig. 5a)**. The output amplifier is a current feedback op-amp and is configured as a transimpedance stage whose gain is set by, and equal to, R_f.

With its 1800V/μs slew rate, the *CLC5523* offers a 250MHz bandwidth and adjustable gain. Figure 5a) shows the device conected as a non-inverting amplifier, the maximum voltage gain – with V_g=+2V – being set by the ratio R_f/R_g.

As V_g is reduced to zero, the gain falls to a minimum, with up to 80dB of gain reduction possible, depending on the operating frequency. In linear terms, i.e. gain measured in V/V, most of the gain change occurs over the range 0.8V<Vg<1.2V, while in logarithmic terms, i.e. gain measured in decibels, the gain variation with V_g is linear from 10dB below maximum gain, i.e. +1V, downwards.

Figure **5b)** shows the device pin-out. This is the same for both the DIP and SOIC versions, but note that when using the DIP version, pin 4 must be grounded via 25Ω rather than directly, Fig. 5a).

After some considerable time and development effort, I arrived at the circuit of **Fig. 6.** The *CLC5523* is set for a maximum gain of ×10, determined by the ratio R_{22}/R_{20}. Oscillation is maintained via a two turn feedback winding L_2, L_1 having some 14 or 15 turns.

Considerable care is needed where the automatic level control, or alc, loop is concerned.

Quartz crystals for frequency control

A quartz crystal makes a high-Q resonator, which also exhibits a very high degree of frequency stability with time and changes in temperature.

When excited at its resonant frequency, the crystal vibrates – rather like a tuning fork but usually at a much higher frequency. Like a tuning fork, when the excitation ceases, the vibration slowly dies away, over many cycles. The higher the 'Q' or quality factor, the slower the decay.

In fact, Q is the ratio of the energy stored to the energy lost per radian. Thus if the Q of a crystal were say 10 000 – a very modest figure – then 0.01% of the stored energy would be lost per radian. After 10 000 radians, or 1592 cycles, the energy left would be down to $^1/_e$ or 37% of the original amount.

A high quality crystal resonator would have a Q of 100 000 or more; the very best approaching 500 000. The best raw natural quartz exhibits a bulk Q of over a million, but adding the electrode arrangements always reduces this.

The plate of quartz behaves like a series tuned resonant circuit, the resonant frequency being determined by the values of the motional inductance L_m and the motional capacitance C_m. These are the electrical equivalents of the mass and compliance – springiness – of the quartz.

Equivalent inductance and capacitance are shown in **Fig. A**, together with R_s. This represents the lossiness of the vibratory material, determining how quickly the vibration dies away in the absence of excitation. It is known as the equivalent series resistance esr.

The crystal can be maintained in a steady state of vibration by applying an electrical signal across it. The piezo-electric effect causes minute changes in the crystal dimensions at the frequency of the applied signal. The amplitude of the vibrations is negligible except at the natural resonant frequency of the plate of quartz, the resonant frequency being determined mainly by the size and thickness of the plate.

In addition to L_m, C_m and R_s, Fig. A also shows the capacitance C_0. This represents the capacitance between the two areas of metalisation, one on either side of the plate. These serve two purposes. They couple the electrical stimulation to the crystal, and there are also wires bonded to them which serve to support the crystal.

The mounted cystal is usually enclosed in a container of glass or metal, filled with air, dry nitrogen or – preferably – a vacuum.

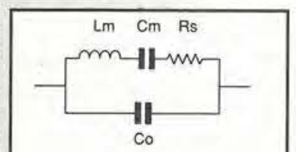
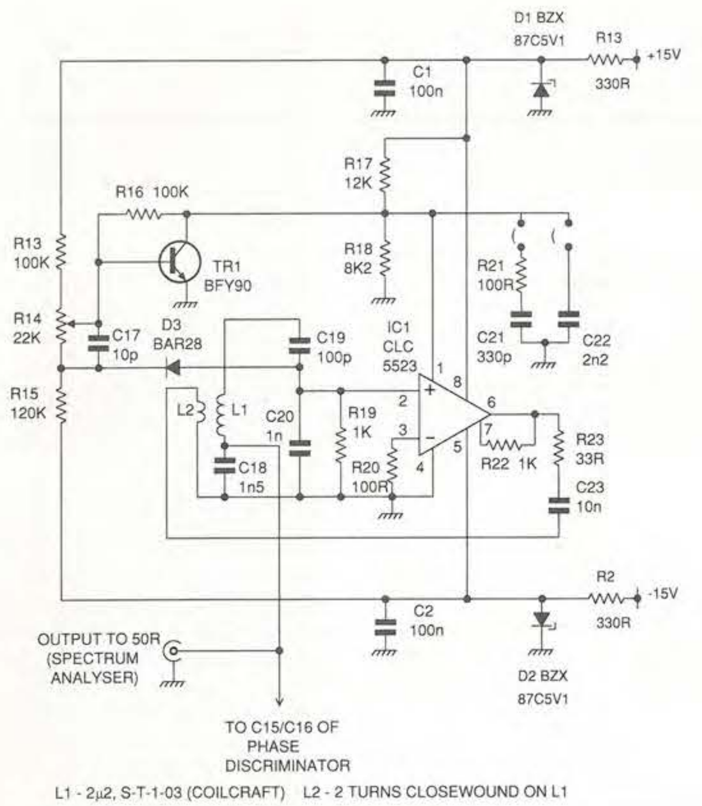


Fig. A. Equivalent circuit of a quartz crystal.

Fig. 6. Circuit of a 10.4MHz oscillator with a linear maintaining amplifier, the amplitude of oscillation being controlled by an ALC loop.



Any attempt at adding an rf bypass capacitor following the diode detector D_3 resulted in 'squegging', i.e. oscillation of the level-control loop resulting in bursts of rf, each followed by a recovery dead period. Any low-pass filter, whatever its time constant, must follow the loop error amplifier Tr_1. Thus the dc level out of the schottky diode detector D_3 still shows a considerable level of rf ripple.

Both the ripple and the dc level – the latter adjusted by an offset control R_{14} – are applied to the loop error amplifier Tr_1. This compares the adjusted dc level with 0V ground and reduces the gain-control voltage V_g applied to pin 1 of IC_1 to maintain the desired level of oscillation. Note that even if Tr_1 cuts off completely, the potential divider formed by R_{17} and R_{18} prevents V_g exceeding +2V.

A requirement for proper operation of the device is that V_g should not exceed +2.5V. If it does, the gain control circuitry may saturate, and the gain may actually be reduced. With the circuit shown, the loop was very effective. There was no measurable change in output level as the frequency was tuned, by means of the core of L_1, over the range 9.5-11.5MHz.

Note the apparent similarity in Figure 3 between a), and d), even though they are in fact very different. In a) the fall in gain with increasing amplitude is due to the device cutting off and probably also bottoming – it is physically incapable of supplying as much gain at the larger amplitude. In d) on the other hand, the gain falls only because the level control loop causes it to: the device could provide more gain at that amplitude if asked to.

Thus the characteristic crosses the unity gain line at a steep angle as in a), resulting in a closely defined amplitude. And yet the amplifier is working linearly, more so even than the device illustrated in c), which exhibits very poor amplitude control.

The circuit of Fig. 6 was designed to supply a 50Ω output load directly, and this is connected across C_{18}. The reactance of this capac-

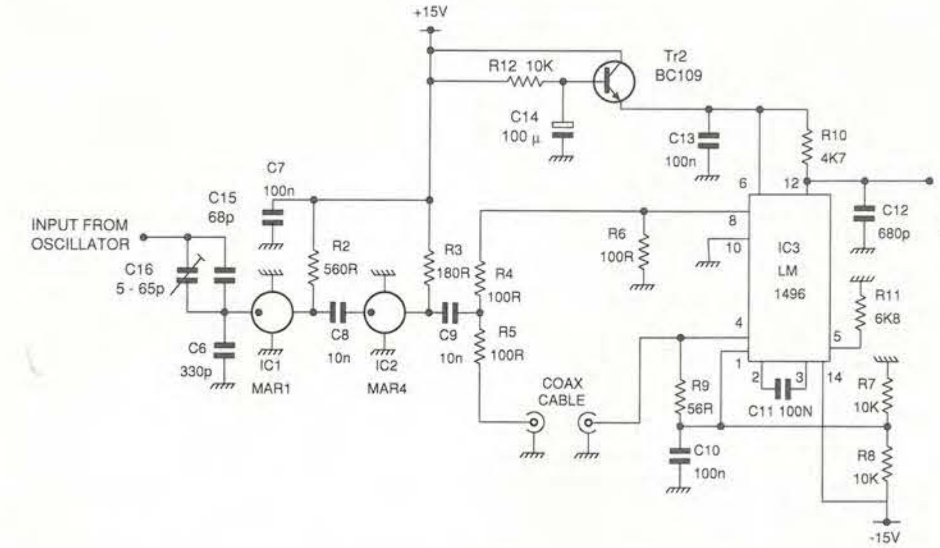


Fig. 7. The phase noise discriminator used is slightly modified from the earlier circuit, by the addition of an active smoothing circuit associated with Tr_2.

Measuring crystal parameters

Crystal manufacture is a specialised business, although traditionally amateur radio enthusiasts used to grind their own, to get the exact frequency they required.

A crystal manufacturer has the specialised equipment to measure all of the parameters shown in Fig. A, and others – such as temperature coefficient, ageing, etc. Such equipment is not generally available in the usual electronics development laboratory. But if you do want to know the parameters of a given crystal, as I did in connection with a 10MHz crystal oscillator described in this article, it is not a difficult matter to measure them. The usual laboratory rf instruments, plus a simple jig involving a little purpose built circuitry, will do the job.

Figure B shows the arrangement I used to characterise a 10MHz crystal dredged from my stock. It was made by an East Anglian crystal manufacturer which is no longer in business. It was salvaged in the early sixties from a naval missile tester. There, it formed the reference in a purpose-designed digital voltage/frequency/period/etc. meter that was the heart of the system. Naturally, in the circumstances, detailed information on it is no longer available; only the maker's name, the frequency '10000O' and a batch number stamped on the solder-seal metal can.

I measured the series resonant frequency F_s using the crystal as the shunt leg in a tee-attenuator pad, the series legs being 100Ω resistors. These resistors should be as small as possible and certainly not wirewound. Chip resistors would be fine.

For accurate results it is important that the impedance seen looking 'each way' from the crystal, should be purely resistive. To the left, this is ensured by the 10dB pad, the 50Ω splitter and the 50Ω output impedance of the signal generator. My signal generator is a home-brew affair, using the now discontinued Plessey *SP2002* direct digital synthesis chip, providing a range up to 300MHz with sub-1Hz resolution.

Any signal generator may be used, though a good slow motion drive is needed. On a cheap non-synthesised generator, the easiest way to achieve exceedingly fine tuning resolution, is to slightly tweak the fine output level control. Note that the resulting changing load on the oscillator pulls the frequency slightly.

To the right, the input return loss s_{11} of the Mini Circuits *MAR1* rf amplifier is 23.1dB, angle 164°, at 100 MHz – the lowest frequency at which it is listed. Thus at the frequencies up to which crystals are readily available, the impedance is very close to purely resistive, given the added effect of the 100Ω resistor.

With the 10MHz crystal connected, the resonant frequency F_s at which the output dipped to a minimum was found to be 9.996709MHz. The crystal was then replaced by a resistor, selected to give the same amplitude as the crystal minimum at the monitor. This required a 22 and a 75Ω resistor in parallel, giving an equivalent series resistance of 17Ω.

Next, the two 100Ω resistors were removed, and the crystal connected in their place, as indicated in Fig. B. Again, the frequency of the minimum was measured, giving the parallel resonant – or 'antiresonant' – frequency F_p as 10.010938MHz.

As with the resonant frequency, there were several much smaller resonances at slightly higher frequencies, as is commonly the case. But it was clearly evident which was the main resonance.

The series resonant circuit L_m plus C_m looks inductive above resonance, and at some frequency, this effective inductance becomes parallel resonant with C_o, giving the antiresonant frequency. The value of C_o was measured on an audio frequency bridge[2], where the motional impedance of the series circuit is so high as to be out of sight.

It turned out to be 7.4pF. From this, the effective inductance of the L_m, C_m branch to resonate at F_p=10.010938MHz with 7.4pF calculates out as 34.15537µH – a reactance X_p of 2148.3925Ω. Reactance X_p is the difference between the reactance of L_m and of C_m at F_p.

Now the frequency ratio of the antiresonant to the resonant frequency $F_p/F_s=K$ is 10.010938/9.996709=1.00142337. So,

$$2\pi F_s K L_m - \frac{1}{2\pi F_s K C_m} = X_p \qquad (1)$$

Using,

$$C_m = \frac{1}{4\pi^2 F_s^2 L_m}$$

to eliminate C_m from equation (1), and knowing F_s, K and X_p, then L_m calculates out to be 12023.687µH, or to take a realistic view of experimental accuracy, say 12mH. This is a much larger inductance than could be realised at 10MHz with any practical coil, and resonates with C_m=0.02pF. It is the large ratio of C_o to C_m which accounts for the antiresonant frequency being so close to the series resonant frequency.

The Q of the series resonance is the ratio of the inductive, or capacitive, reactance divided by the equivalent series resistance, or $2\pi F_s L_m / R_s$.

Putting in the figures,

$$Q=2\pi 9.996709\times 0.012/17=44337$$

and around 40000 is a reasonable figure for the Q of a metal-can crystal.

Like most crystals for use at frequencies below 30MHz, the crystal was designed for use at parallel resonance, with additional parallel capacitance totalling – including C_o – around 30pF. This would result in the antiresonant frequency, i.e. the designed operating frequency, being lower than 10.010938MHz.

In practice, the total shunt loading capacitance would be adjusted to give the desired exact 10.000MHz, with the crystal operating at the designed 70°C in an oven.

Crystals for use above 30MHz are usually designed for a total parallel load capacitance of 20pF, or alternatively, for use at series resonance. In the latter case, final adjustment is by means of a trimmer capacitor (or inductor) in series with the crystal.

The adjustment or 'pulling' range available with series operation is only about a tenth of that possible with parallel resonant operation.

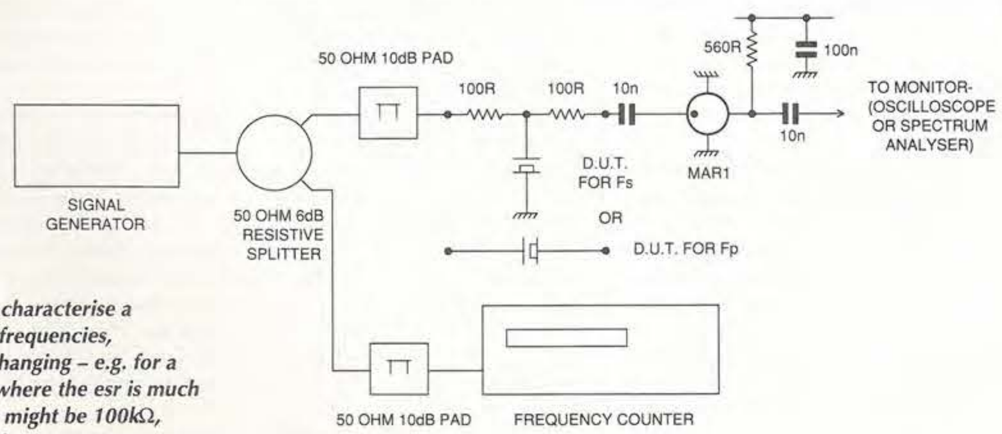


Fig. B. Test set-up used to characterise a 10MHz crystal. For other frequencies, components might need changing – e.g. for a 32768kHz clock crystal, where the esr is much higher, the 100Ω resistors might be 100kΩ, feeding a virtual earth fet-input opamp.

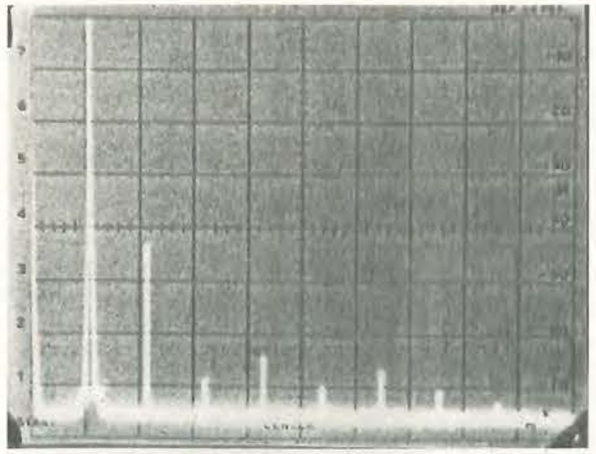


Fig. 8. The spectrum of the output of the linear rf oscillator using the CLC5523. Reference level (top of screen) 0dBm, 10dB/division vertical, Span 0 - 100MHz, IF bandwidth 100kHz, video (post detector) filter max, analyser front-end attenuator set to 20dB.

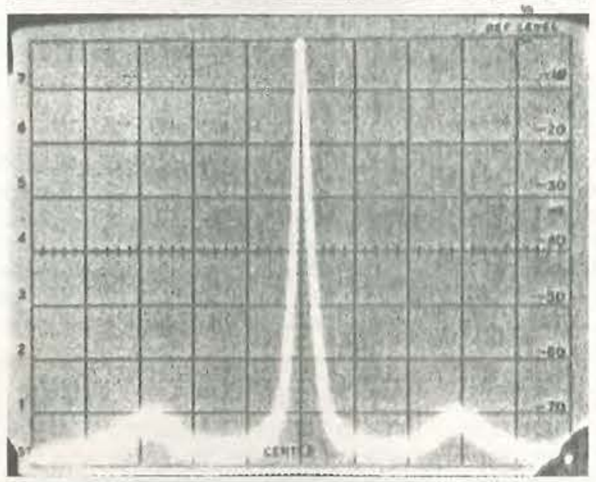


Fig. 9. The spectrum of the output of the oscillator using the CLC5523. Reference level (top of screen) 0dBm, 10dB/division vertical, centre frequency 10MHz, 500kHz/division horizontal, IF bandwidth 10kHz, video (post detector) filter max, analyser front-end attenuator set to 20dB.

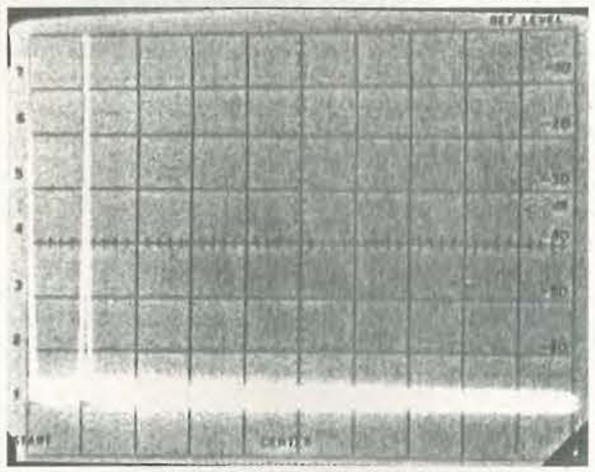


Fig. 10. The spectrum of the output of the linear rf oscillator using the CLC5523, with C_{22} in circuit in place of R_{21},C_{21}. Reference level (top of screen) 0dBm, other settings as for Fig. 8 except analyser front-end attenuator set to 30dB.

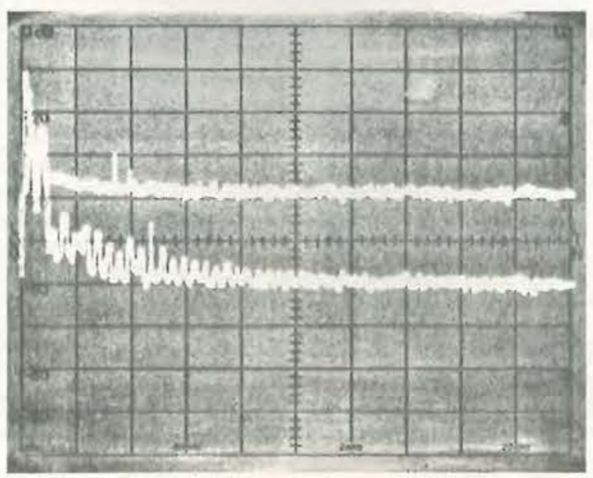


Fig. 11. Lower trace: spectrum of the output of the phase noise discriminator with its input grounded. Upper trace: spectrum with the output of the CLC5523 oscillator connected as shown in Figs 6 and 7. Both cover 0-5kHz, reference level –60dB, 10dB/division vertical, resolution bandwidth 30Hz, smoothing maximum, 100s/division sweep speed.

itor at the operating frequency is around 10Ω, preventing excessive loading by the output of the tuned circuit.

The oscillator runs from ±5V rails, derived from the ±15 supplies powering the phase noise discriminator. The spectrum of its rf output, with R_{21}, C_{21} of Fig. 6 in circuit, is as in **Fig. 8**. The second harmonic was over 40dB down on the fundamental output, with all other harmonics over 70dB down.

The auto level control loop is a negative-feedback loop like any other, and must be properly proportioned to ensure stability. But in this application, merely ensuring absence of loop oscillations is not enough.

Way beyond the critical loop unity gain or "corner" frequency, the loop phase shift will, at some point become positive. At this frequency there will be an enhancement of the sideband noise, and this can just be seen in Fig. 8 as shoulders either side of the fundamental output, at something over 70dB down.

In combination with C_{21}, R_{21} provides the designer with some control over the position and amplitude of these noise sidebands. The smaller C_{21}, the further out they occur. With the value of C_{21} shown, they are about 70dB down and situated at ±1.4MHz from the carrier as shown at the wider dispersion of 500kHz/division in **Fig. 9**.

The level of harmonics observed in Fig. 8 would be regarded as excellent in any conventional rf oscillator, but seem perhaps excessive if the maintaining amplifier really were running in a purely linear mode. And indeed this proved to be the case.

They are in fact due to the small residual rf ripple at the collector of Tr_1 modulating the gain of IC_1 – for the bandwidth of the gain control input of the *CLC5523* is itself equal to 95MHz typical.

Figure 10 shows the 0-100MHz spectrum of the output when R_{21}, C_{21} in Fig. 6 were disconnected, and C_{22} connected in their place. Such harmonics as were then visible on the spectrum analyser display turned out to be due to distortion in the analyser itself. Its front end attenuation was therefore increased by 10dB to 30dB, the IF gain being adjusted to retain a top of screen reference level of 0dBm. This of course results in the noise floor in Fig. 10 being 10dB higher than in Figure 8.

All harmonics are clearly well over 60dB down, but the noise peaks mentioned earlier are not visible in this photo. They were in fact at –65dBc, at 600kHz either side of the carrier. As the value of C_{22} is increased, the noise sideband humps move closer to the carrier and become more pronounced. Finally, at some ridiculous value around 1μF, squegging sets in.

How about the phase noise?

Clearly, as an oscillator with low harmonic content, the circuit is a great success. But of course, the whole *raison d'être* of the work was to seek an oscillator with reduced close-in phase noise. This was measured using the phase noise discriminator shown in Fig. 7.

The circuit used was similar to the earlier version, but with the addition of an active smoothing circuit, R_{12}, C_{14} and Tr_2. The intention was to reduce the mains related harmonics which can be seen on phase noise measurements taken with the circuit. However, the improvement proved small.

On investigation, I found that although the mains related spectral lines are 100Hz apart, they are not harmonics of power supply full-wave rectifier ripple, but odd harmonics of 50Hz. They are therefore doubtless due to the leakage flux of mains tansformers. The phase-noise discriminator was surrounded on the lab bench by various instruments – spectrum analysers, oscilloscope, freqency counter and power supplies.

I recorded a run covering 0-5kHz with the oscillator conected to the phase noise discriminator. Both were powered up, but with the $C_{16,17}$ input of the phase noise discriminator connected to ground.

This set-up is shown as the lower trace in **Fig. 11**, which represents the lower limit of phase noise measurement capability. As such, it is a more meaningful measurement than the lower trace in Fig. 4a) of the previous article. That showed the level with the supplies switched off. It thus represented the noise floor of the *HP3580A* low-frequency spectrum analyser working from a 4.7kΩ resistive source, representing a noise figure of about 5dB.

The oscillator was then conected to the phase noise discriminator as indicated in Figs 6 and 7 – with R_{21}, C_{21} in circuit and C_{22} disconnnected – and another run recorded. Both traces are shown in Fig. 11.

Did it work?

In a word, no – at least, not as well as hoped. Compare the linear oscillator performance, upper trace in Fig. 11, with that of the *LC* oscillator covered in the earlier article, reproduced here as Fig. 2b). You can see that at 2.5kHz and upwards the former is as noisy as the bottoming *LC* oscillator, and worse than the non-bottoming version.

At low frequencies however, it does not show such a marked rise in noise. Indeed, below 500Hz, it is better than the bottoming *LC* oscillator and comparable with the non-bottoming version. This gives me hope that with some further development, an improved performance may be achieved.

A difficulty arising from using the *CLC5523* in this application, is that it has a low impedance output. This is quoted as typically below 30mΩ at dc and still not much more than an ohm at 10MHz. In the circuit of Fig. 6, this has been padded up to 33Ω. But via L_2 this still reflects damping across the tuned circuit, resulting in an estimated working Q of only 10.

The next task clearly is to achieve a tank circuit working Q of close to the unloaded Q of the inductor, resulting – hopefully – in an oscillator with very low close-in sideband noise. Watch this space. ■

References

1. Hickman, Ian, 'Understanding phase noise, *Electronics World*, August 1997, pp. 642 - 646.
2. Hickman, Ian, 'LCR measuring transformed,' *Electronics World*, November 1994, pp. 931-934.

Design wideband antennas

Richard Formato's guidelines for helping you optimise your impedance loaded wideband antennas include a design example that gives continuous coverage from 6 to 150MHz without resorting to tuning and matching.

Adding impedance loading – resistance and reactance – to an antenna is one of the most effective ways to increase its bandwidth. Since the early 1980s, this approach has received progressively more attention, and today it is state of the art for wideband systems.

Choosing a suitable loading profile can maintain high radiation efficiency while providing a remarkable increase in bandwidth. One of the dipole designs in this article, for example, provides continuous coverage from 6 to 150MHz with no tuner or matching network.

The question is, of course, how to select an optimum loading profile. Selecting the wrong profile can produce dismal results. Indiscriminately adding resistance to an antenna obviously causes performance to deteriorate, and the reduction can be severe.

This article analyses several key design parameters for selecting wire antenna loading profiles. Design guidelines are then developed to achieve the greatest antenna bandwidth.

Some of the results are rather surprising, but they can lead to substantial performance improvements for loaded antennas.

Resistive loading

The simplest loading scheme is to insert a resistor in the antenna. In the early 1960s, Altshuler[1] built a centre-fed dipole, or cfd, whose input impedance was, for practical purposes, flat over a 2:1 frequency range as a result of added resistance. The antenna was loaded with a single 240Ω resistor in each arm a quarter-wavelength from the end.

Because resistance reduces the radiated power, the dipole's bandwidth was increased at the expense of radiation efficiency, which was reduced by about 50%. This is the inevitable trade-off in designing impedance-loaded wire antennas.

Bandwidth is best in heavily loaded antennas, but the resulting penalty in radiation efficiency may be too high to provide acceptable power gain – i.e. the product of directive gain and efficiency.

Fig. 1. Elements of the centre-fed dipole, comprising two wire radiating elements with half-length h and radius a.

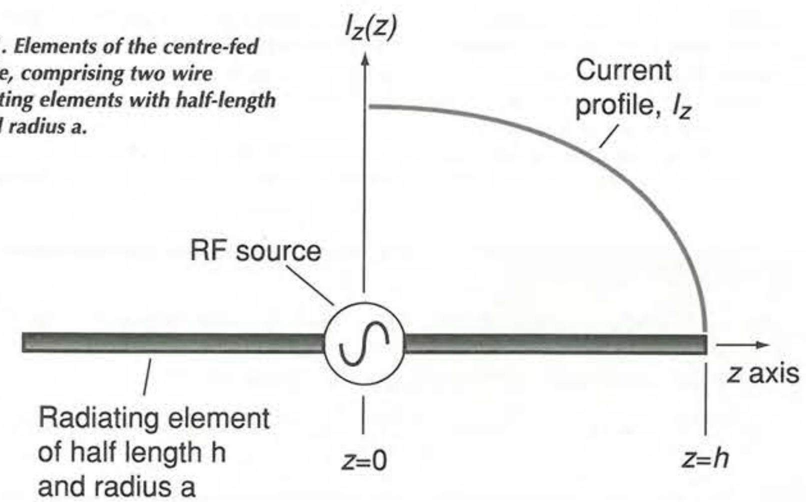


Table 1. Resistance-only profile for the centre-fed dipole example.

Distance (m)	Resistance (Ω)
0.00	0.00
0.76	3.86
1.52	4.36
2.28	4.98
3.03	5.76
3.79	6.75
4.55	8.05
5.31	9.82
6.07	12.30
6.83	15.99
7.59	21.89
8.34	32.31
9.10	53.90
9.86	113.47
10.62	438.06

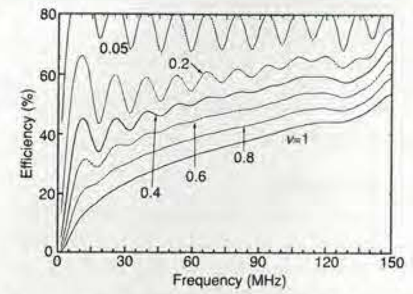


Fig. 2. Demonstration of how radiation efficiency varies with different values of v for the centre-fed dipole design example.

A few years after Altshuler's work, Wu and King[2] published a theoretical model of the loaded centre-fed dipole. Unlike the discrete resistor approach, the Wu and King loading profile varied continuously along the antenna and could be implemented with a conductive surface layer of different materials – aluminium and carbon, for example – of varying thickness. Profiles based on the Wu-King theory, which requires a travelling wave current mode with a linear amplitude decay, provide good bandwidth but relatively poor radiation efficiency because the antenna is heavily loaded.

An example of what can be achieved is the wideband field probe for sampling impulsive electromagnetic fields, built by Kanda[3]. A centre-fed dipole probe was designed using the Wu-King profile. It was resistively and capacitively loaded by depositing on a glass rod substrate a segmented, conductive thin-film of varying thickness.

Narrow rings were burned away using an argon laser to separate the conductive segments, thereby providing the reactive component of the loading profile. Although the loaded centre-fed dipole was useful as a field probe, it was not useful as a transmitting antenna because it was so inefficient. Its transfer function was typically below –22dB.

Attempts have been made to increase the efficiency of impedance-loaded antennas. Rama Rao and Debroux[4,5], for example, developed a more efficient hf monopole antenna by using a 'fractional' Wu-King profile. Efficiencies of 15-36% with a standing-wave ratio of two or less were achieved from 5-30MHz in a 35ft-high antenna. The monopole was continuously loaded with a profile equal to 30% of the Wu-King profile, and a fixed, lumped-element matching network was inserted at the antenna feed point.

Still better bandwidth and efficiency can be achieved by using an improved loading profile derived from a travelling-wave current mode with a power law amplitude decay, instead of the linear decay required by the Wu-King profile. The derivation of the improved profile and its relationship to the Wu-King profile have been developed in the March 1997 issue[6].

The nonlinear amplitude decrease of the current along the antenna results in a higher average antenna current. In turn, this increases the radiated fields and total radiated power. The antenna's radiation efficiency is higher because it radiates more of the input power.

Several parameters influence how well a particular loading profile performs. There is no one 'best' profile. Important design parameters include: value of the power law exponent; design frequency; wire length-to-diameter ratio; and number of antenna segments, i.e. the number of discrete resistors. Certain parameters are more important than others – in the sense of having a relatively greater impact on performance – and for some parameters the results are unexpected.

These four design parameters are discussed below for typical hf/vhf centre-fed dipole antenna designs. Radiation efficiency and standing-wave ratio, or swr, are examined for rf sources between 2 and 150MHz – the upper limit of the computer model. There are, of course, other important measures of antenna performance, including power gain and pattern. But these are not examined in detail because they are usually acceptable in an impedance-loaded antenna with a good loading profile.

Continuous loading profiles

Figure 1 shows the centre-fed dipole antenna. It consists of two wire radiating elements with half-length h and radius a. The total dipole length is $L=2h$, and its diameter is $D=2a$.

Amplitude of the current profile is plotted schematically along one element's length. Maximum current occurs at the rf source at the feed point, and it decreases along each arm until reaching zero at the end.

The centre-fed dipole's bandwidth is increased by symmetrically loading it with an internal impedance profile, i.e. resistance and reactance. The profile is given by:

$$Z^i(|z|)=R^i(|z|)+jX^i(|z|), |z|\le h$$

where Z^i is the complex internal impedance per unit length, in Ω/m, consisting of lineal resistance R^i and reactance X^i, and where $j=\sqrt{-1}$.

The resistance and reactance per unit length for the improved power-law impedance profile are given by:

$$R^i(z)=60v(h-|z|)^{v-2}\left\{\psi_R-\frac{(1-v)\psi_I}{2k_o(h-|z|)}\right\} \quad (1a)$$

$$X^i(z)=60v(h-|z|)^{v-2}\left\{\psi_I+\frac{(1-v)\psi_R}{2k_o(h-|z|)}\right\} \quad (1b)$$

where $k_o=2\pi/\lambda_o$ is the wave number and λ_o is the free-space wavelength corresponding to the design frequency, which is designated f_o. v is the power-law exponent ('profile exponent') for a travelling-wave current distribution that minimises resonance effects.

The first equation recovers the 100% Wu-King profile when $v=1$. The general case corresponds to $v\le1$. The derivation of equation 1 and its relationship to previous work are discussed in Formato[6].

The reactance computed from equation (1b) can be positive, i.e. inductive, or negative, i.e. capacitive. The lineal inductance in H/m and lineal capacitance in F/m are given by $L^i=X^i/2\pi f_o$ and $C^i=(2\pi f_o X^i)^{-1}$ for $X^i>0$ and $X^i<0$, respectively, where the design frequency f_o is in hertz.

In equation 1, $\psi=\psi_R+j\psi_I$ is a complex quantity known as the expansion parameter,[1,2] its real and imaginary parts being subscripted R and I, respectively. Symbol ψ represents the ratio of the centre-fed dipole's vector potential to current, which is approximately constant along its length. The expansion parameter is defined[2] in the panel headed 'Expansion parameter equations'.

Because ψ is frequency dependent, it is usually evaluated at the antenna's fundamental half-wave resonance, that is, $\lambda_o=4h$ (see Wu and King[2] for details). As discussed below, however, this choice does not necessarily provide the best antenna performance. The design frequency f_o, in hertz, and the wavelength λ_o (in metres) are related by $f_o\lambda_o=c$, where $c\cong2.998\times10^8$m/s is the free-space velocity of light.

The improved loading profile specified by equation 1 in general contains both resistance and reactance. But adding reactance to a wire antenna – especially capacitive reactance – can complicate construction.

As a consequence, many practical designs

Expansion parameter equations

$$\psi=2\left[\sinh^{-1}\left(\frac{h}{a}\right)-C(2k_oa,2k_oh)-jS(2k_oa,2k_oh)\right]+\frac{j}{k_oh}\left[1-\exp(-j2k_oh)\right] \quad (2)$$

C and S are the generalised sine and cosine integrals[2,7] given by,

$$C(b,x)=\int_0^x\frac{1-\cos W}{W}du \quad (3a) \qquad S(b,x)=\int_0^x\frac{\sin W}{W}du \quad (3b) \quad \text{where } W=(u^2+b^2)^{1/2}$$

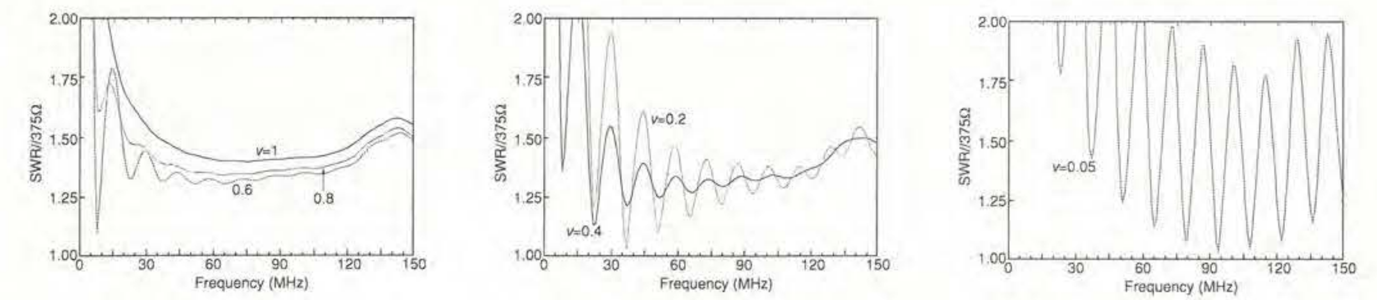


Fig. 3. Standing-wave ratio curves for various values of v. Significantly, in a), swr is generally lower with decreasing v.

employ only resistive loading, because excellent results are often achieved even without the loading profile's reactive component (see Rama Rao and Debroux[4] and Rama Rao[5], for example). In the centre-fed dipole designs discussed below, only resistive loading is considered.

Discrete loading

Building a continuously loaded antenna can be a formidable task. This is especially so if some exotic technique is required – like vapour deposition of a conductive thin-film layer – or if reactive loading is included. An effective alternative is to construct an approximation to the continuous loading profile using discrete resistors at intervals along the antenna.

The antenna is constructed of highly conducting – for practical purposes, perfectly conducting – cylindrical wire segments that are connected together by resistors. Single resistors can be used or, for large-diameter radiators, multiple resistors may be used.

This type of discrete loading profile can provide excellent bandwidth and efficiency, better even than a continuous profile, and it does so without the problems associated with continuous profiles or reactive loading. Because of these advantages, only discrete loading profiles are considered in this article.

A discrete profile may be determined by first dividing the centre-fed dipole into an odd number of equal length segments, N. The centre segment, which contains the rf source, is not loaded. All other segments are loaded with a lumped resistance placed at the segment centre.

The value of the resistor is computed as the product of the segment length – in metres – and the value of the continuous loading profile evaluated at the segment centre, R^i in Ω/m from equation (1a). This approach provides a piecewise linear, or step, approximation to the continuous loading profile. There is, of course, any number of other discrete approximations, but only this uniform step approximation is considered here.

As an example of a typical discrete profile, consider a centre-fed dipole with the following design parameters:

L=22m
D=10cm
f_o=7MHz
v=0.4
N=29
ψ=8.822–j2.464.

There is no reactive loading or feed-point loading, and the full 100% profile is used. The discrete resistance-only profile, computed as described above for $z\ge0$, appears in **Table 1**. Distance is measured from the origin, and the loading is symmetrical in each arm of the centre-fed dipole.

The loading resistance increases slowly. It rises from 3.86Ω on either side of the rf source, ±0.76m from the centre, to just over 438Ω in the last segment, located ±10.62m from the source. This very gradual increase in resistance is typical of more efficient loading profiles.

In the sections that follow, several antenna design parameters are investigated by examining the computer-modelled performance of a typical 22m-long centre-fed dipole using discrete resistance-only loading.

Power law exponent

Probably the most important design parameter in determining radiation efficiency is the value of the profile exponent, v, which determines how quickly the current amplitude decays along the dipole.

Slower decay, i.e. lower v results in a higher average antenna current. This increases the radiated fields and consequently the efficiency. The improvement in efficiency can be quite dramatic. The trade-off is that decreasing v increases the peak standing-wave ratio, or swr, and causes it to fluctuate more with frequency.

The influence that v has on radiation efficiency and swr is illustrated in **Figs 2** and **3**. These plots are based on computer-modelled data for a 10cm diameter, 22m-long centre-fed dipole with N=29 and a 100% resistance-only loading profile computed at a design frequency of 7MHz, which is approximately the fundamental resonance. Efficiency and swr are plotted as a function of the rf source frequency from 2 to 150MHz. Calculations were made every 1MHz.

The profile exponent v has a very significant effect on radiation efficiency, with lower values resulting in higher efficiencies. The curve in Fig. 2 for v=1, which corresponds to the 100% Wu-King loading profile, shows that the efficiency increases from about 1% at 2MHz to about 54% at 150MHz.

The variation with frequency is smooth and monotonic. But as v decreases, the efficiency increases progressively more rapidly, especially at lower frequencies. When v=0.2, the efficiency increases from about 14% at 2MHz to more than 65% at 10MHz, a 51% increase in a span of only 8MHz.

Beyond 10MHz, the efficiency fluctuates more or less periodically, with a gradually increasing trend until it reaches a maximum above 75% at 150MHz. For v=0.05, the efficiency exhibits a pronounced quasi-periodic fluctuation, but its minimum value is more than 68%, and the maximum is well above 80%.

Figure 3 plots swr parametric in v for an rf source characteristic impedance of 375Ω. If a different feed system impedance is used, an appropriate broadband transformer would be required. For v=1.0, the swr varies smoothly from a maximum of greater than 2:1 at 12MHz to a minimum of about 1.45 near 67MHz. It then increases gradually above 67MHz, with a slight dip near 150MHz. The curves for v=0.8 and v=0.6 show the same general trend. But, significantly, the swr is generally lower with decreasing v, even though it fluctuates more at lower frequencies.

Figure **3(b)** plots swr for v=0.4 and v=0.2. The swr is generally lower for v=0.4 than it is for v=0.6, but the variability with frequency is much greater, and the peak values are higher at some frequencies.

For v=0.4, the swr exceeds 2:1 between about 12 and 16MHz, but it is below 2 for v=0.6. As v decreases to 0.2 and then to 0.05, Fig. **3(c)**, the swr fluctuation becomes more pronounced, and the peak values are higher. The minimum swr values, however, are generally lower, and, on the average, the swr is still well below 2:1.

The best choice for v is evidently the lowest value that provides acceptable swr at frequencies of interest. Choosing v in this way ensures the highest possible radiation efficiency, and the improvement is usually very substantial.

Design frequency

The design frequency f_o is another important parameter in determining a good loading profile. Although it appears to be accepted practice to choose f_o close to the centre-fed dipole half-wave resonance frequency (see Wu and King[2] for example), this choice is not necessarily the best.

Because the expansion parameter, which plays a major role in determining the loading profile, is frequency-dependent, the actual choice of design frequency must be based on how much a given loading profile improves bandwidth while still providing good radiation efficiency.

There is no other sensible scheme for determining f_o because there is no theoretical basis for choosing one value over another. The best approach is therefore empirical, which is the

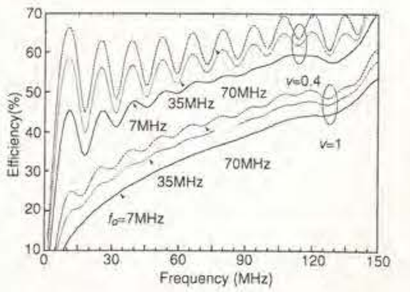


Fig. 4. Radiation efficiency against two values of v at three important frequencies for a 22m long 10cm diameter centre-fed dipole with N=29.

approach adopted here.

Radiation efficiency and swr plots for three design frequencies and two values of the profile exponent v are shown in **Figs 4** and 5. These plots are for a 22m-long, 10cm diameter centre-fed dipole with N=29 and a 100% resistance-only loading profile.

Values of 7MHz, which is approximately the half-wave frequency, and 35MHz, and 70MHz were used for f_o, with v=1.0 and v=0.4 at each frequency. Except for the fundamental resonance, f_o was chosen arbitrarily. Other choices would yield different results, but the observations made here are still generally applicable.

As expected, the lowest efficiencies in Fig. 4 result from the most heavily loaded profile, at v=1.0. The efficiency increases with frequency and shows slightly more variability at the higher design frequencies. The less heavily loaded profile (v=0.4) is much better – especially between 2 and 10MHz. The higher values of f_o increase the fluctuation in the efficiency, but the variability is not great.

The most important feature of the efficiency data is that higher design frequencies result in substantial improvements. Frequency f_o has a major impact on radiation efficiency, and its influence is greater for profiles with lower values of v. Because the efficiency increases with decreasing v, the influence that f_o has becomes even more important.

Thus, even though the 22m centre-fed dipole has a fundamental resonance near 7MHz, choosing a design frequency that is ten times greater provides better performance.

For example, as the curve for v=0.4 shows, the radiation efficiency at 10MHz is about 45% when f_o=7MHz, but it increases to 67% when f_o is increased to 70MHz. Choosing a higher design frequency thus results in a much better antenna.

The advantage of a higher design frequency is also evident in the swr plots of Fig. 5. It is quite significant that selecting f_o=70MHz when v=1.0, Fig. **5(a)**, results in generally the lowest swr across the entire 2 to 150MHz band.

When v=0.4, Fig. **5(b)**, choosing f_o=70MHz results in swr ≤ 2 across most of the band. The variability is greater, and the swr is not consistently lower with increasing f_o, as it is when v=1.0. These effects are minor, however, and better overall performance usually results from higher values of f_o.

Radiator length-to-diameter ratio

Increasing the element diameter is a standard broadbanding technique for wire radiators. It is therefore not surprising that a larger diameter, impedance-loaded centre-fed dipole exhibits better overall performance than its thin counterpart.

In order to investigate the effect of L/D, I computed the radiation efficiency and swr from 2-150MHz for radiating element diameters of 0.1cm, 1cm, and 10cm. A 22m long centre-fed dipole was modelled, yielding L/D ratios of 22000, 2200 and 220, respectively, which represent antennas ranging from 'extremely thin' to 'thin'. Profile exponents of v=1.0 and 0.4 were used with 29 segments and a 100% resistance-only loading profile computed at a design frequency of f_o=7MHz.

The larger diameter centre-fed dipoles have better radiation efficiency at all frequencies for both values of v. The improvement in efficiency becomes progressively greater at higher frequencies, and it approaches a factor of two at the high end of the band. For the profiles with v=0.4, increasing the element diameter also reduces fluctuations in the radiation efficiency, but this effect is not evident in the heavily loaded profiles when v=1.0.

The largest diameter element provides the best standing-wave ratio performance, especially at lower frequencies. Its swr is below 2.5 at all frequencies above 6.5MHz, and below two above approximately 12MHz. The swr decreases quickly up to about 30MHz and flattens out at less than 1.5 for most of the rest of the band.

By contrast, the standing-wave ratio for the very thin element with D=0.1cm, is high, being above two throughout the band, and above 2.5 below 30 MHz. The very thin radiator thus fails to provide acceptable standing-wave ratio even though it is very heavily loaded.

Similar standing-wave ratiobehavior is evident when v=0.4. The fattest element provides the best performance. Its standing-wave ratiois below 1.5 over most of the band, and below 2.15 at all frequencies above 6MHz.

Decreasing the diameter to 0.1cm increases the standing-wave ratio, but not as much as it did for the more heavily loaded profile with v=1.0. When v=0.4, however, the standing-wave ratio variability becomes much more pronounced for smaller element diameters.

Building an antenna with a low L/D ratio, that is, making it 'fatter', may be difficult if too large a diameter conductor is required. A continuous cylindrical surface can sometimes be approximated by a sufficient number of parallel wires uniformly spaced around the cylinder's circumference. A large number of wires may be required, however, depending on how good the approximation must be.

Such a dipole structure, sometimes called a 'cage dipole' because of its resemblance to a bird cage, can offer a convenient and effective alternative to large diameter cylinders for low L/D designs.

Segmentation

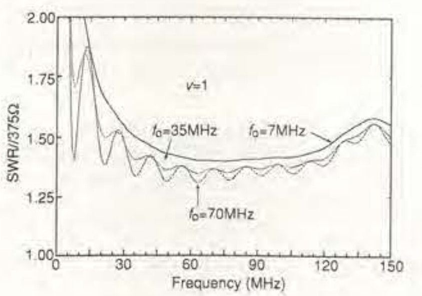
The segmentation used, that is, the value of N, also influences how well the loaded centre-fed dipole performs. I studied its effect by computing radiation efficiency and standing-wave ratio for three segmentations. The results were somewhat unexpected. A 22m long, 1cm diameter centre-fed dipole was modelled with N=29, 59, and 119 segments. A 100% resistance-only loading profile with v=1.0 and 0.4 was computed at f_o=7MHz.

The efficiency data show that the least segmented antenna, N=29, provides the best overall performance. As segmentation increases, radiation efficiency generally decreases, although the change is not great from N=59 to 119. This result is somewhat surprising, since increasing segmentation presumably provides a better approximation to the continuous loading profile.

However, the data show quite convincingly that the net effect of adding more discrete resistance is to increase the i^2R – Joule heating – losses more than the radiated power, resulting in lowered efficiency.

This effect occurs for both values of the profile exponent. As is typically the case, the efficiency fluctuates more with frequency as v decreases, and the variability is greatest at the low end of the band. One effect of increasing N is to reduce the fluctuation somewhat, but the change is not pronounced, and it occurs only when v=0.4.

The standing-wave ratio data are not as clear cut, but the general conclusion is still that the lowest segmentation probably provides the best overall performance.For v=1.0, the ratio is lowest for N=29 at frequencies below approximately 90MHz – about a factor of 12 greater than the fundamental resonance. In the same frequency range, it is not significantly different from the N=59 or 119 values even when v=0.4.



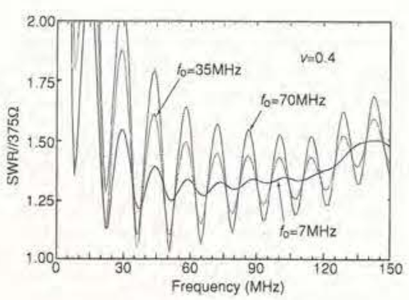


Fig. 5. Standing-wave ratios at three frequencies for v=1 and v=0.4. Lowest swr across the entire 2-150MHz band is achieved for f_o=70MHz when v=1, curve a). Antenna dimensions are as in Fig. 4.

Above 90MHz, the antenna with N=59 performs best for both values of v, but the 119-segment design is very close.Nevertheless, in that same frequency range, 'eyeball average' standing-wave ratios for N=29 are about 1.6 for v=1.0 and 1.65 for v=0.4, which are very good indeed.

Thus, the least segmented antenna provides very robust swr performance at all frequencies from 2-150 MHz.

In summary

This article has investigated four design parameters for impedance-loaded wideband wire antennas. Using the following guidelines to select these parameters when designing an impedance loading profile should provide a near-optimum wideband antenna.

First, the profile exponent v is very important in determining radiation efficiency. Lower values of v result in higher efficiencies, and the improvement is very significant. The optimum value for v is the smallest value that provides acceptable swr at frequencies of interest.

Secondly, because the design frequency f_o is usually chosen close to the fundamental centre-fed dipole resonance, it is somewhat surprising that higher frequencies generally result in much better radiation efficiency, especially for loading profiles with smaller values of v. Higher values of f_o also usually give better overall swr performance. The optimum value for f_o is the highest value that provides acceptable swr at frequencies of interest.

Thirdly, the lower the L/D ratio, the better. Large diameter radiating elements provide much more bandwidth than thin ones, even without impedance loading. A large diameter radiator makes it easier for a loading profile to provide the greatest possible bandwidth. If necessary, large diameter conductors can be approximated by multiple parallel wires.

Finally, reducing the segmentation, that is, the number of discrete resistors used to approximate the theoretical continuous loading profile, results in slightly better radiation efficiency. The swr is not particularly sensitive to segmentation at 'low' frequencies, and it is slightly better with increased segmentation at 'high' frequencies.

For a thin antenna, 'low' frequencies are less than about 12 times the fundamental resonance, and 'high' frequencies are greater. The design guideline for segmentation is to use the smallest number of discrete resistors that meets the swr objectives. ■

Software on disk

Richard's PC software relating to this article, for computing loading profiles, is available on a 3.5in disk. Send a postal order or cheque for £10, payable to Reed Business Information Group, to Wideband Disk, Electronics World Editorial Offices, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Please don't forget to mention your name and address – some do!

References

1. Altshuler, E. E., The Travelling-Wave Linear Antenna, *IRE Transactions on Antennas and Propagation*, July 1961, p324
2. Wu, T. T., and King, R. W. P., The Cylindrical Antenna with Nonreflecting Resistive Loading, *IEEE Transactions on Antennas and Propagation*, May 1965, pp. 369-373 (see also Corrections, p998, *IEEE Trans. Ant. Prop.*, Nov. 1965)
3. Kanda, M., Time Domain Sensors for Radiated Impulsive Measurements, *IEEE Transactions on Antennas and Propagation*, May, 1983, p438
4. Rama Rao, B., and Debroux, P. S., Wideband HF Monopole Antennas with Tapered Resistivity Loading, IEEE Military Communications Conference, Monterey, California, 30 September to 3 October 1990 (MILCOM '90)
5. Rama Rao, B., Optimised Tapered Resistivity Profiles for Wideband HF Monopole Antenna, 1991 IEEE Ant. & Prop. Soc. Symposium, London, Ontario, Canada
6. Formato, R. A., Improved Impedance Loading for Wideband Antennas, *Electronics World*, (UK) March 1997, p. 102.
7. Abramowitz, M., and Stegun, I., Handbook of Mathematical Functions With Formulas, Graphs, and Mathematical Tables, U.S. Department of Commerce, National Bureau of Standards Applied Mathematics Series 55 (AMS 55), Ninth Printing, November 1970, p231.

Cable sonics?

Audio expert Douglas Self reveals why he uses Woolworths' mains cable to drive his speakers in this, the second of three articles on the interface between the power amplifier and loudspeaker.

Loudspeaker cables are relatively simple things – notwithstanding the haze of controversy that often surrounds them. In the drive to adopt ever more complex pulse-testing techniques, most of them of very questionable relevance, the basic cable properties appear to have been overlooked. This is emphasised by the striking fact that the proponents of unlikely cable-hypotheses rarely even mention fundamental parameters like the resistance and inductance of their test specimens.

I looked at the first-order effects first, and only then examined the details. This approach at least has the merit of novelty. Firstly, it is beyond doubt that an audio cable is not a transmission line.[1] Transmission line effects can have nothing to do with audio, unless you're in the telephone business.

In real rf work, such effects are usually considered negligible until the frequency is high enough for the length of the line to reach one-sixth of a wavelength. The wavelength of 20kHz is 15km, so unless your loudspeakers are over a mile away from the amplifier, there is no need to fret.

A length of cable has series resistance, series inductance, and shunt capacitance as its major parameters. A simple cable model is shown in **Fig. 1** and in **Fig. 2a)**, where the series resistance and inductance, and the shunt capacitance, are represented as lumped components. The capacitance is shown concentrated after the resistance and inductance, where it will cause the most response variation; this is actually a very slight rise rather than a roll-off. This is due to interaction with the inductance.

In practice the effect of this capacitance is tiny compared with the roll-off due to the series inductance. A typical response change at 20kHz is a fraction of a thousandth of a decibel, so the effect is negligible anyway. The amplifier is assumed to have zero output impedance unless otherwise stated.

It seemed worthwhile to confirm that this position for lumped capacitance really is the worst case, and this was done by simulating the cable models in Fig. 2. A first approximation to a distributed model is Fig. **2b)**, where the cable is split into two halves. A slightly more accurate version is Fig. **2c)**, with the cable now divided into four. Simulation confirms that Fig. 2a is the worst case for capacitative high-frequency lift; but the effect is tiny in all three cases. Further thoughts on cable capacitance are to be found in a separate section below.

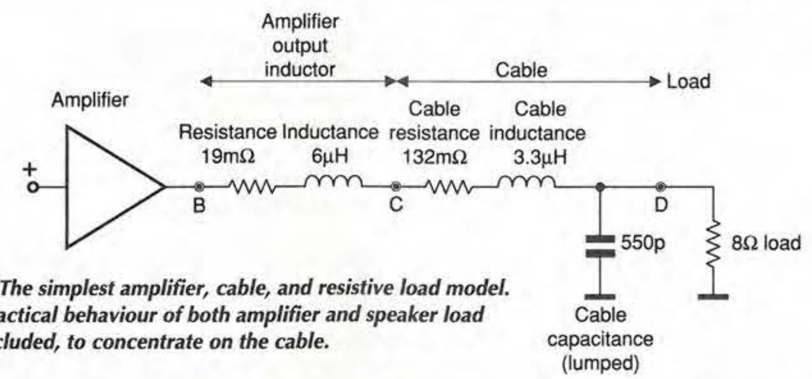


Fig. 1. The simplest amplifier, cable, and resistive load model. The practical behaviour of both amplifier and speaker load are excluded, to concentrate on the cable.

Cable Table 1. Common domestic electrical cables from a DIY store compared.

	Recommended use	Cross section mm^2	Calculated resistance, mΩ/5m	Construction
1A	Speaker cable (sic)	0.4	210	2-core
3A	Lamps	0.5	172	3-core
6A	Appliances small	0.75	115	3-core
10A	Appliances medium	1.0	86	3-core
15A	Heater cable	1.5	57	3-core
35A	Cooker cable	6	14.4	Twin & earth
45A	Shower heaters	10	8.6	Twin & earth

Cable resistance effects

The first basic parameter is cable resistance, determined by the conductor material and its total cross-sectional area. Copper is the obvious material, though silver can be used if you are rich and gullible enough. Silver is the most conductive metal but despite its cost has only 7% less resistivity than copper. It also presents some awkward problems with non-conductive tarnish films, namely silver sulphide.

If you want to reduce your cable resistance slightly, this is not an economical way to do it, compared with a minor increase in copper cross-section. I have seen mercury cables suggested, in the form of quicksilver-filled hosepipes. In view of the insidiously poisonous nature of mercury vapour, one can only hope this wasn't taken seriously by anyone.

Having rejected silver as ridiculously expensive, and settled for copper, it is then just a matter of using enough area to get a suitably low resistance per metre. The main consequence of unwanted series resistance is frequency response perturbations due to the

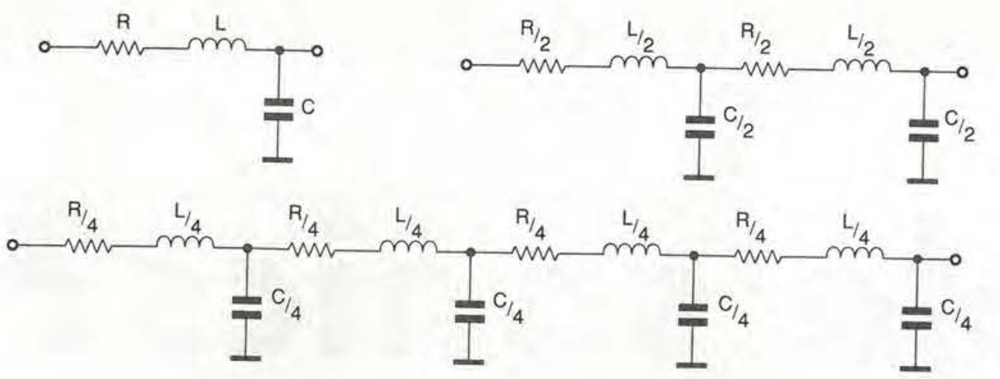


Fig. 2. Speaker cable models: 2a) Circuit elements combined into lumped components. 2b) Cable resistance, inductance and capacitance are split into two halves, the first approach to a distributed model. 2c) Cable elements are split into four quarters, more closely approaching a distributed model.

speaker impedance varying with frequency.

The second basic parameter is cable inductance, determined both by the conductor spacing and their diameters. The equation for the inductance of parallel spaced cylindrical conductors is:

$$L = 4\times10^{-7}(1+4\log(\frac{D}{\sqrt{R_1 \times R_2}})) \qquad (1)$$

in henries per metre. Variable D is the spacing between centres of two cylindrical conductors while R_1 and R_2 are the radii of the conductors.[2]

The main consequence of this series inductance is a high-frequency roll-off with an 8Ω load. Cables that deliberately space the two conductors increase the inductance considerably, and this is not a good idea.

Fortunately, inductance increases slowly with spacing. There is the theoretical possibility of some sort of special interaction between cable inductance and loudspeaker behaviour, but the practical inductances are so small – at least two orders of magnitude below crossover inductor values – that this does not seem likely to be a problem in practice.

Cable investigations

As usual in audio, if you want some hard facts you have to get them yourself. As a first step, I took myself to the local DIY superstore to find out what was on offer to those who looked no further for their speaker cables – ie 99% of the population.

The data I gathered is shown in **Cable Table 1**, where the fact that the skimpiest cable of the lot was labelled 'Speaker cable' shows just how remote Subjectivist cable neurosis is from the rest of the world.

The cables measured in **Cable Table 2** were obtained from various sources. The resistance values calculated here are slightly higher than those measured.

I then measured some cables, standardising on 5m lengths as this should be adequate for even large rooms. I used a Wayne-Kerr *4210* LCR meter at 1kHz. The two smallest cable sizes looked too weedy to bother with, and I stuck with my initial thought that the easiest and cheapest solution to hooking up your loudspeakers is the now famous lawnmower cable. This is usually 6A two or three-core, each core being 30/0.18, ie 30 strands of 0.18 mm diameter copper, with PVC insulation.

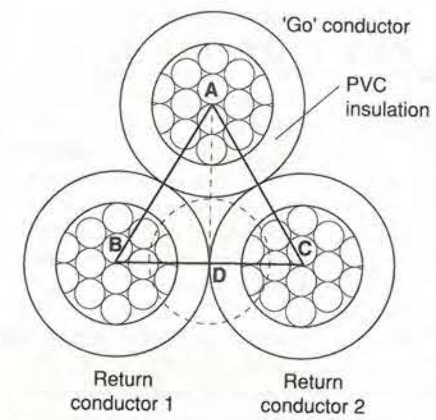


Fig. 3. The geometry of three-core cable usage, showing how inductance reduction works by reducing the effective conductor spacing.

The first sample was 6A two-core, and it really was cut from a defunct lawnmower. This has a basic go-and-return (ie total loop) resistance of 178mΩ and an inductance of 3.2μH. If we use the amplifier/output-inductor combination examined in the first part of this article,[3] with a total dc resistance of 24mΩ, the extra cable resistance degrades the so-called damping factor from 330 to 40. Note that all cable parameters quoted are for the total go-and-return path.

This gives a flat attenuation of 0.20dB, due to the cable resistance forming a potential divider with the 8Ω load, and a main hf roll-off due to the cable inductance forming an *LR* low-pass filter with the 8Ω load, giving a further loss of –3dB at 398kHz. This corresponds to –0.01dB at 20kHz, which is negligible.

Add an amplifier output inductor of 6μH to a cable inductance of 3.3μH and the roll-off becomes –3dB at 145kHz, and –0.086dB at 20kHz. This demonstrates that even a maximal-value output inductor combined with twin-core cable does not pose a problem with an 8Ω resistive load. However, increasing either inductance further might be unwise.

What to do with the spare conductor

The second lawnmower cable was three-core. I suggest that the thing to do with the spare conductor is to parallel one of the other two; this obviously reduces the total resistance to three-quarters of the original value, and a bit less obviously – to me anyway – also reduces the total inductance. I always parallel the ground return rather than the hot conductor, on the philosophical basis that earths should be more solid than signal paths. Here, of course it makes little difference.

The reduction in inductance is explained by the geometry of the situation in **Fig. 3**, suggested to me by George Chadwick.[4] If the normal go and return conductors are A and B, then when third conductor C is connected, the return current divides in two. This is equivalent to the original current in a notional conductor at a point D halfway between the B and C.

Distance A-D is √3/2, or 0.87, of A-B, and so the effective conductor spacing is reduced. Putting this into equation 1 shows that inductance is reduced by 1.4 times. Measured reductions in inductance as in Cable Table 2 are 1.32 times, which is reasonable considering that the physical dimensions are subject to tolerances.

Using two conductors only the resistance is 189mΩ and the inductance 3.3μH, and the loss and roll-off much as the two-core sample. Connecting the third conductor reduces the round-trip resistance to 142mΩ and the inductance to 2.5μH. The damping factor improves from 40 to 50.

The total shunt capacitance between go and return conductors is 488pF for the two-wire connection, and 655pF for the three-wire case. With three conductors, the frequency-insensitive loss is reduced to 0.153dB, and the inductive roll-off is now –3dB at 520kHz, which is a mere 0.007dB at 20kHz.

The third sample was 13-Amp three-core cable, as used for connecting electric fires. It is an interesting economic point that 13A cable only has 1.5 times the copper cross-sectional area of 6A, though a factor of 2.2 would be needed for the same current density; presumably a slightly higher voltage-drop per metre is accepted.

Using only two cores of 13A cable, five metres has a resistance of 132mΩ and inductance of 3.3μH. Moving up to the larger cable has reduced resistance exactly as predicted from the increased cross-sectional area, but inductance is unchanged. The flat resistive loss is reduced to 0.14dB, and the roll-off stays at 0.011 dB at 20kHz.

The improvement over 6A cable with three cores used is minimal. Using three cores of

13A, resistance falls to 98mΩ, the lowest so far, and inductance is 2.5μH; once more unchanged from the 6A cables.

Cook cable

The fourth sample, and the next step in cable sophistication in my view, is the stuff you wire cookers up with. Cooker cable has a flat format with two heavy-gauge conductors and a much lighter earth connection between them. This is usually called 'twin-and-earth'. For 5m, the go-and-return resistance is 21mΩ, which is an order of magnitude lower than the lawnmower cable. The inductance is higher at 4.1μH, due to the greater spacing of the main conductors.

Connecting the central earth conductor reduces total resistance slightly to 19mΩ and inductance usefully to 3.1μH. With this rather serious cable, the frequency-insensitive resistance loss is reduced to 0.02dB, and the inductive roll-off not greatly altered.

Finally, I examined 50Ω RG58 coaxial cable, which has been recommended for speaker cabling[5]. Coaxial cable is not made with anything like the weight of copper in cooker cable, so its resistance is relatively high at 158mμ. The 5m inductance at 1.3μH is less than half that of the ordinary cables used two-core style, though three-core usage closes some of the gap. The point is that the effects of cable inductance appear to be an order of magnitude less important than cable resistance, so the lower inductance of coaxial cable is not the deciding factor.

Cable Table 2. Comparison of common domestic electrical cables, a coaxial alternative, and purpose-designed speaker cables.

	mΩ	μH	pF	Res loss dB	Inductive roll-off dB@20kHz	Basic construction Nxdia.
Lawnmower1 6A 2core	178	3.2	569	0.2	0.010	31×0.18mm
Lawnmower2 6A 2core	189	3.3	488	0.2	0.011	30×0.18mm
Lawnmower2 6A 3core	142	2.5	655	0.15	0.007	30×0.18mm
13-A cable 2 core	132	3.3	494	0.14	0.011	25×0.24mm
13-A cable 3 core	98	2.5	674	0.10	0.010	25×0.24mm
Cooker cable 2core	21	4.1	547	0.02	0.018	7×1.4mm
Cooker cable 3core	19	3.1	357	0.02	0.010	7×1.4mm
Coax RG58C 50Ω	158	1.3	388	0.17	0.002	19×0.18mm
Van den Hul	120			0.13		7×21×0.15mm
Sharkwire	400			0.42		129×0.10mm
Heywire	1200			1.21		1×0.6mm

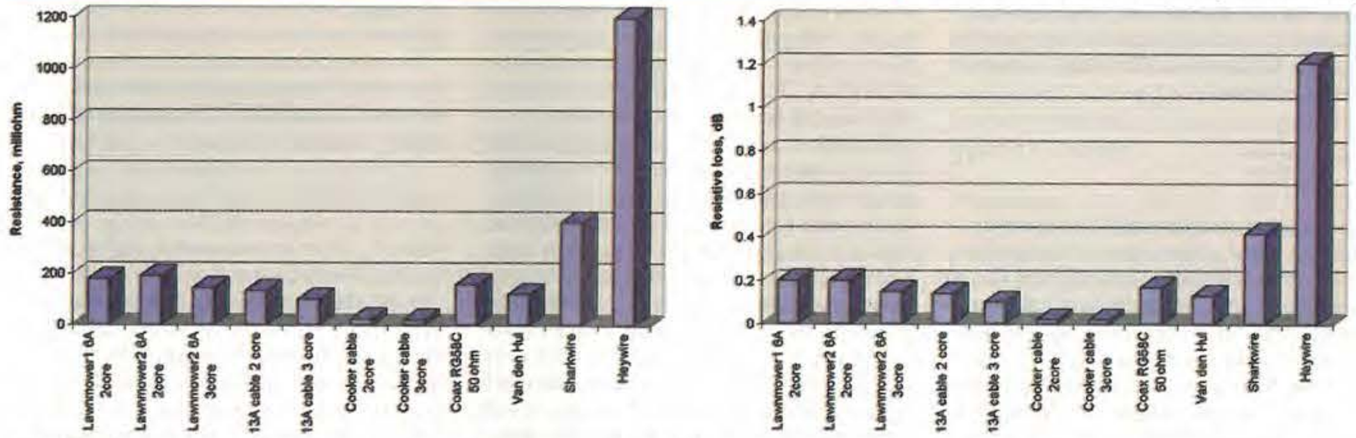
All data are for 5m round-trip, measured at 1kHz. Inductive roll-off is ref 1kHz, ie this is the *extra* attenuation due to inductance. Data for the last three commercial speaker cables are from Eric Ford[6]; no inductance values were given.

Referring to **Cable Table 2**, the loss and roll-off data given here applies only to 8Ω resistive loads. This is unrealistic, but gives a feel for the size of the problems. It would be desirable to produce similar results for a real speaker load, but they come in almost infinite variety – which one would you use?

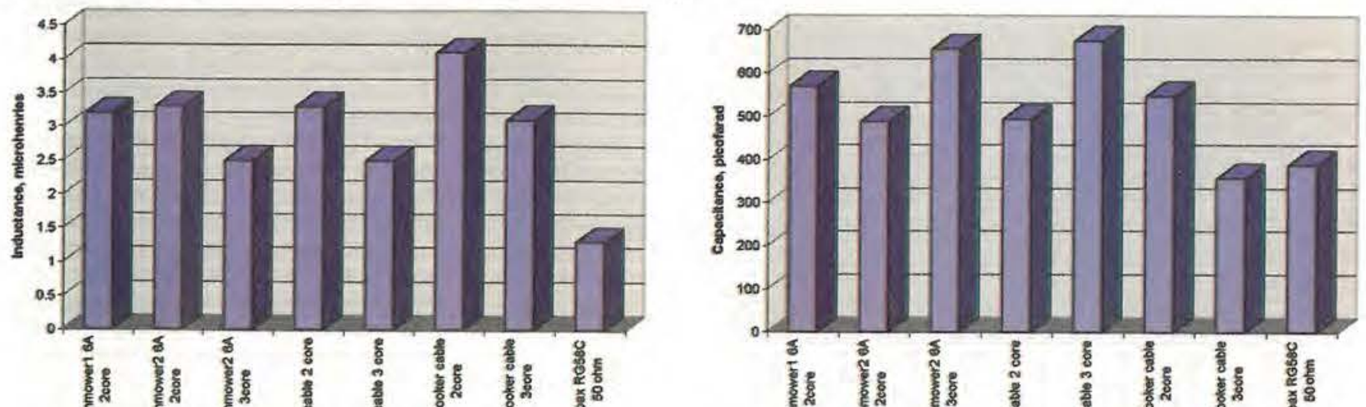
Amplifier stability

There is anecdotal evidence that when it became fashionable to use allegedly better-sounding cables, several makes of 'high-end' amplifier showed hf instability. This has been ascribed to the greater shunt capacitance of some of the peculiarly constructed cables.

We will probably never know the truth of the matter – much as we will probably never know if there really was a 'transistor sound' in the sixties, allegedly due to crossover distortion. But the shunt capacitance of even a very strange cable is so much smaller than the values that cause trouble, usually 100nF and up,



All cables in Table 2 compared for resistance and loss, above, left and right.



Inductance, capacitance and inductive roll-off charts for the cables detailed in Table 2.

that it is very unlikely to affect the stability of an amplifier.

A much more likely explanation is that all the amplifiers involved lacked output inductors, but the old cables had enough series resistance to isolate the amplifier from the capacitative components of the loudspeaker impedance. The new cables – being in general much thicker – did not.

Omitting the output inductor is not a good idea; so long as it is included, instability from any capacitance should be impossible.

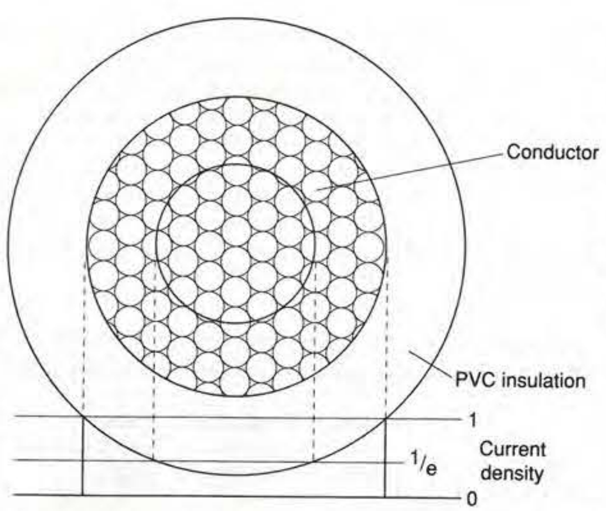


Fig. 4. Skin effect in action. The effective conducting area is defined by the point where current density falls to 1/e. At audio frequencies it is much less severe than shown here. The conductor is shown as stranded, and it is assumed the strands are intimate contact.

Skin effect

An unfortunate complication to the study of audio cables is that their ac resistance is greater than that at dc. This is not due to cable inductance. What we are talking about here is real resistance, not just an increase in impedance magnitude due to additional inductance.

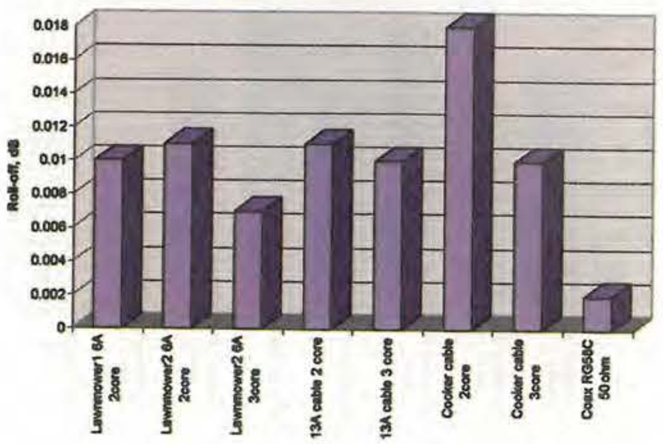
This increase in ac resistance is due to skin effect. The ac current in a conductor is not uniform across its cross-section, but concentrates in the outer part of it. This is because for an elementary filament carrying current, the impedance at the conductor centre is greater than that at the surface, so the current is diverted towards the surface. You will have to take this on trust unless you want to explore some serious vector calculus.

The result is that the ac current distribution across a conductor dies away exponentially with distance below the surface. At some skin depth d the current has fallen to $1/e$ of its maximum, and the situation is equivalent to a uniform current distribution in a thin outer shell of the same thickness d. Skin depth d may be determined from,

$$d = \frac{1}{\sqrt{\pi f \mu \sigma}} \qquad (2)$$

where μ is permeability and σ is resistivity.

Having found d from this well-established equation[7,8] it is simple to calculate the area of the annulus representing the equivalent conducting part of a circular cross-section cable, and thus find the increase in resistance, **Fig. 4**. Note that this is a true frequency-dependent resistance – not an inductance – and does not have the phase shift that would accompany extra inductance.



Taking 13A copper mains cable as an example, equation 2 shows that below 13kHz, the skin depth is greater than the cable radius, and there is no significant effect. At 20kHz, the skin depth is 0.47mm, and the conductor radius 0.6mm. The cross-sectional area is 1.08mm^2, of which only the central 0.06mm^2 is unused. Therefore resistance per metre only increases from 152mΩ to 160mΩ, and this can have no appreciable consequences.

Some writers appear to believe that the presence of skin effect must mean that the conductor is acting as a transmission line; this is quite untrue.

Cable cross-sections other than circular give different results, but since the circular version has the minimum perimeter for a given area, all other shapes give more surface, and so less skin effect. This is why high-power rf stages often use flat copper strap connections.

Skin effect is relatively more significant for low-resistance cables, as current can move laterally more easily. For loudspeakers however, the real criterion is the total absolute resistance, which is always reduced by using larger cable. The conclusion must be that even in the worst case there will be no problem, and there is no justification at all for Litz-type cables.

Skin effect is more significant in other technologies; it is vital at rf because the skin-depth is so small, and hence silver-plating is well worthwhile. It is also important in 50Hz power transmission, because of the low impedances involved, and in some circumstances multiple parallel busbars are required to control the extra resistive heating.

Proximity effect

One further aspect of cable theory is the proximity effect. This results from the interaction between the adjacent go and return currents. Proximity effect is similar to skin effect, but in this case the interacting currents are moving in opposite directions, causing the current distributions in each conductor to crowd towards each other. As with skin effect, this reduces the effective cross-sectional area of the conductor and increases its resistance. Again, the effect is an increase in apparent resistance as frequency increases, and not an increase in inductance.

The only reference that I have found to this deals rather disconcertingly with resistance in ohms per thousand feet[9]. The ratio between ac and dc resistance varies as,

$$\textit{proximity factor} = \sqrt{\frac{f}{R_{dc}}} \qquad (3)$$

where R_{dc} is the dc resistance of 1000ft, which for 5m/200mΩ lawnmower cables is 10.85Ω.

At 20kHz, the proximity factor is therefore 42.9, which yields from[9] an ac resistance about 7% greater than the dc value. This is of the same order of magnitude as skin effect, and should be equally negligible.

Once again, the effect is relatively more significant for low-resistance cables, as lateral current mobility is greater, and once more the greatest problems arise in AC power transmission.

Biwiring

In recent years, a loudspeaker cabling practice known as 'biwiring' has become fashionable. There is no evidence that it does any good, but naturally this has not impeded its popularity.

Assuming we have a two-way speaker with the hf and lf sections of the crossover separately accessible, biwiring uses two separate cables from the same amplifier output to drive two separate crossover inputs, as shown in **Fig. 5**.

The conceptual origin of this practice is disclosed by the similarity of the word to 'bi-amping' where separate amplifiers would be used

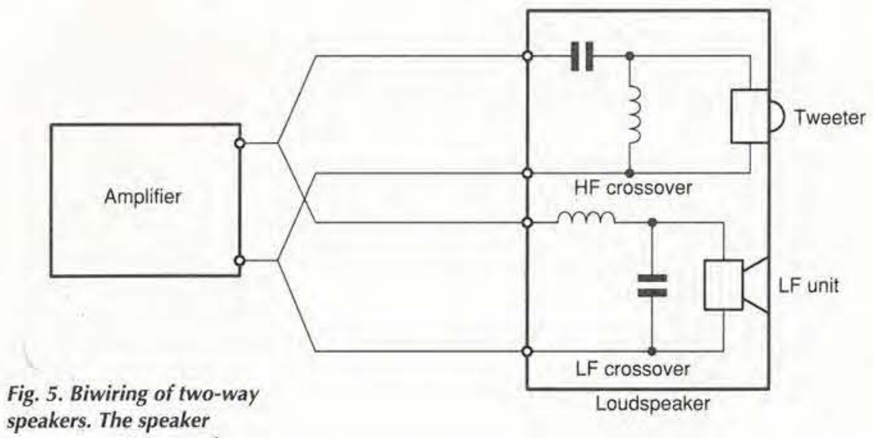


Fig. 5. Biwiring of two-way speakers. The speaker crossover unit is greatly simplified.

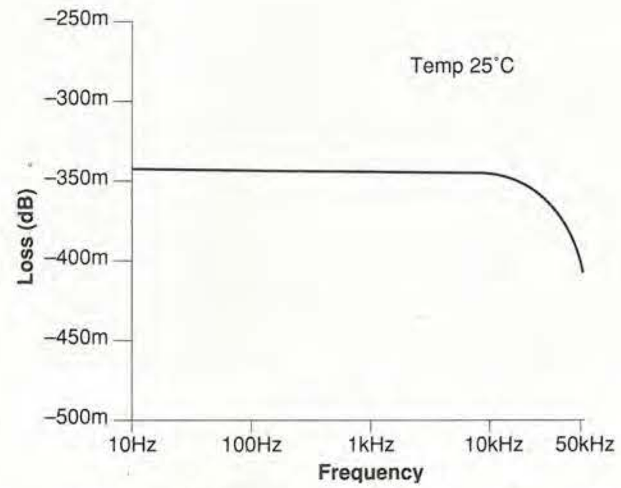


Fig. 6. Showing a resistive loss of 0.34dB, the start of the inductive roll-off, and the minimal importance of cable capacitance. The effect of 655pF may be just visible as a slight thickening of the trace between 20 and 50kHz.

for the high and low-frequency sections. Bi-amping has a rationale in terms of reduced amplifier intermodulation, the possibility of sophisticated crossovers, and so on, but none of this applies to solid copper, so biwiring seems pointless.

It is conceivable that a loudspeaker could exist which interacted in some undesirable way with cable inductance, so that separate inductances in separate cables eased the problem. However, the low inductance values shown by normal cables seem to make this very unlikely.

Cable capacitors

Cable Table 2 shows that the shunt capacitance in normal cables is low – at less than 150pF/m. Simulations were done on the circuit models in Fig. 2, using some worst-case cable parameters of 0.32Ω resistance, 3.3μH inductance, and 655pF shunt capacitance.

If these R,L,C parameters are treated as 'lumped' components, the worst case is Fig. 2a, with all the capacitance lumped after all the inductance. This gives a rise of 0.00028dB at 20kHz. Figure 2b gives 0.00021dB and Fig. 2c 0.00018dB; as the model more closely resembles a distributed system the effect becomes less.

These tiny response deviations appear to demonstrate beyond doubt that cable shunt capacitance is so small its effects are completely irrelevant. Therefore all the speculation about second-order capacitance effects – such as dielectric absorption – must be completely pointless.

Figure 6 shows the resistive loss of 0.34dB and inductive roll-off resulting from the above cable parameters and an 8Ω load. Both zero and 655pF capacitance versions are plotted. You can see that the two curves are virtually identical.

Crystal structure

There has been much debate about the crystal structure of copper, and whether this or its oxygen content can affect anything. Most of it has shown what can only be called an abysmal ignorance of the most basic physics and metallurgy.

Crystal boundaries in copper cannot and do not act as mysterious diodes, and the simplest of experiments proves this beyond doubt.[10] Much of the suspicion directed towards the oxides of copper probably stems from faint memories of copper-oxide rectifiers; their operation relied on a carefully-grown cuprous oxide film in pressure-contact with a lead washer, and this has no relevance to the bulk copper in cables.

In summary

All the effects described here are real, though unlikely to ever be audible under even the most critical conditions. The use of a resistive load only is obviously open to objections, which will be addressed in a forthcoming piece on speaker impedance curves and their implications.

It emerges that even a bit of cable can be complicated if examined in detail. For more details on some of the issues raised here, see reference 11. There is, however, no sign of anything mysterious, incomprehensible, or indeed audible. If you examine a real first-order effect, such as response variation due to cable capacitance, and find it far below any possibility of perception, then speculative second-order effects such as dielectric absorption in this capacitance can surely be dismissed.

There is no justification for treating cables as transmission lines, and no need for special cables that cost insulting sums per metre. Silver appears to have no possible justification for its use. The only factors that count are series resistance and inductance. The first can only be minimised by heavy-gauge cables. Inductance offers more possibilities for its reduction; keep the go and return conductors at minimal spacing, use multiple conductors, or resort to a coaxial structure.

From the considerations given here, the best speaker cable would seem to be 13A cable with three cores used. Significant reduction in resistance requires cooker cable, which is bigger and much less flexible, and so harder to hide. If cable inductance is your main concern, then RG58C coaxial cable certainly has less of it, but its resistance is nearly twice that of 13A cable.

Just for the record, my usual speaker connections are 13A mains cable, with neutral and earth paralleled to reduce resistance and inductance, purchased from Woolworths in 1972. Let it never be said that I am not at the leading edge of cable technology. ■

References

1. Greiner, 'Amplifier-loudspeaker interfacing,' *JAES*, May 1980, pp. 310-315.
2. Fink & Carroll (Ed), *Standard Handbook for Electrical Engineers* 10th edn, McGraw-Hill 1968. Section 2-81, inductance of parallel wires.
3. Self, D, 'Amplifier Outputs & Speakers' *Electronics World*, Sept 1997, p. 717.
4. Mr George Chadwick MIEE, Private communication.
5. Bateman, C, 'Measuring speaker cables,' *Electronics World*, Jan 1997, p. 56.
6. Ford, E, 'Speaker cables pulse tested,' *Electronics World*, Dec 1996, p. 930.
7. Fink, D, ed., *Standard Handbook for Electrical Engineers*, 10th edn, McGraw-Hill 1968. Section 10-425, skin effect.
8. Morrison, R, *Grounding & Shielding Techniques in Instrumentation*, Wiley Interscience 1986, p166. ISBN 0-471-83805-5
9. Fink, D, ed., *Standard Handbook for Electrical Engineers*, 10th edn, McGraw-Hill 1968. Section 10-426, proximity effect.
10. Self,D, Letter: 'Twisted cables,' *Electronics World*, Mar 1997, p. 260. Impedance measurements, no micro-diodes.
11. Davis, F, 'Effects of cable, loudspeaker, & amplifier interactions,' *JAES*, June 1991, p. 461.

A PROBLEM PHASE

When designers tackle Wien bridge oscillators, they often overlook phase shift in the amplifier. But simply relying on textbook equations can result in frequency errors up to 70%. Bryan Hart explains the importance of amplifier phase shift and presents a practical way of dealing with it.

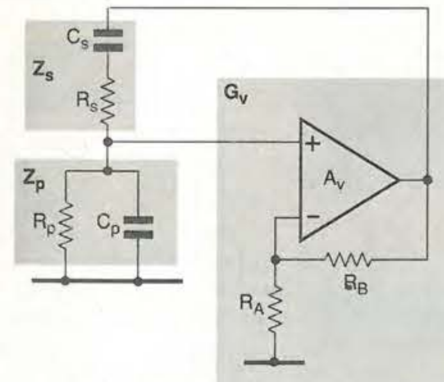


Fig. 1. Wien-Bridge oscillator using an op-amp with gain A_v as the gain block.

The effect of amplifier phase shift on the oscillation frequency of a Wien-bridge oscillator, incorporating an operational amplifier with a single-pole frequency response, is related to the unity-gain frequency of the op-amp by a simple formula. For practical purposes, this formula can be regarded as exact.

In a standard text book analysis of the Wien-bridge oscillator, which is widely used in the design of commercial signal generators, it is assumed that the amplifier used is ideal in that it exhibits zero phase shift.[1] Consequently the oscillation frequency is shown to be solely dependent upon the passive Wien RC network parameters.

But, all practical amplifiers exhibit a finite, even if sometimes small, frequency dependent phase shift.

In this article, I present a simple expression for the oscillation frequency for the common practical case of an amplifier with a single-pole frequency response. I also discuss the significance of the expression in Wien-bridge oscillator design.

Wien bridge analysis

Figure 1 shows the circuit of a Wien-bridge oscillator with a voltage amplifier, G_V, whose voltage gain is G. Amplifier block G_V comprises an op-amp, represented by A_V, with open-loop voltage gain A, and closed-loop gain-setting resistors R_A and R_B.

Amplitude stabilisation is achieved by making R_B a thermistor with a negative temperature coefficient.

The Barkhausen criterion for sustained oscillations to occur at the output of G_V is,

$$\frac{GZ_P}{Z_S+Z_P}=1\angle 0 \qquad (1)$$

Negative feedback around A_V ensures that,

$$G=\frac{A}{1+Ab_o} \qquad (2)$$

where

$$b_o=\frac{R_A}{R_A+R_B} \qquad (3)$$

Initially, A_V is assumed to have an infinite input impedance. As shown by the linearised Bode gain-plot in **Fig. 2**, the dc and low-frequency gain is A_o, the cut-off frequency is ω_c, and the unity-gain frequency is ω_T, which is approximately $A_o\omega_c$.

Hence,

$$A=\frac{A_o}{1+j\left(\frac{\omega}{\omega_c}\right)} \approx \frac{1}{\left(\frac{1}{A_o}\right)+j\left(\frac{\omega}{\omega_T}\right)} \qquad (4)$$

Substituting this value of A in (2) gives,

$$G=\frac{1}{b+j\left(\frac{\omega}{\omega_T}\right)} \qquad (5)$$

where,

$$b=b_o+\frac{1}{A_o} \qquad (6)$$

For the Wien network,

$$Z_P=\frac{R_P}{1+j\omega C_P R_P}$$

and

$$Z_S=R_S+\frac{1}{j\omega C_S}$$

so, after routine algebraic manipulation, it follows that,

Putting it into practice

In practice, equation (10) may be regarded as exact since the non-zero input capacitance and non-infinite resistance of G_v can be taken as included in C_P and R_P respectively. Also, $A_o\omega_C$ differs from ω_T by less than 0.005% for A_o greater than 100.

Typically, A_o is greater than 10 000. In the case of op-amps with the input-circuit architecture of the classic 741 op-amp, Ao is very large – at 100 000 – but subject to a large uncertainty. Nevertheless, ω_T is still closely-defined, being determined by the input stage bias current and a capacitor included to give dominant-lag frequency-compensation[2].

Table 1 shows some spot values of ω_O/ω_R calculated from (13, 14). Clearly, for a value of u comparable with unity, the effect of amplifier phase shift on ω_o can far exceed the effect of component tolerances. From (8c), a ±1% resistor tolerance and a ±2% capacitor tolerance leads to a 3% tolerance on ω_R, and hence ω_O.

For operation at around 1MHz, it is not possible to meet the condition $\omega_O/\omega_R \approx 1$ using available voltage-feedback op-amps because it requires an f_T of around 100MHz.

Discrete/hybrid component amplifiers, or possibly current-feedback op-amps, are required.

When u is much less than unity b is approximately 0.333. In this case, a binomial expansion of (13) that discards all but the first two terms yields the approximation,

$$\frac{\omega_O}{\omega_R} \approx 1 - 4.5\frac{\omega_R}{\omega_T} \qquad (15a)$$

or

$$\frac{\omega_O - \omega_R}{\omega_R} \approx -4.5\frac{\omega_R}{\omega_T} \qquad (15b)$$

Table 1. Spot values, calculated from equations (13) and (14), rounded up to three decimal places.

$u(=\omega_R/\omega_T)$	b	ω_O/ω_R
0	0.333	1.000
0.001	0.333	0.996
0.01	0.333	0.958
0.1	0.317	0.717
0.2	0.288	0.570
0.5	0.185	0.331

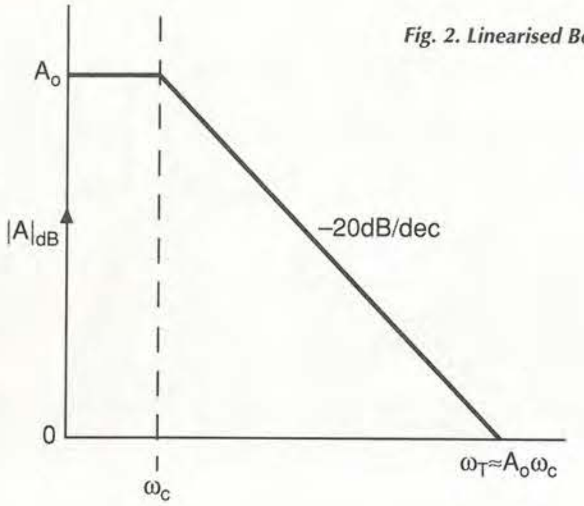


Fig. 2. Linearised Bode gain-plot of A_v.

$$\frac{Z_P}{Z_S+Z_P} = \frac{1}{m + jn\left[\left(\frac{\omega}{\omega_R}\right) - \left(\frac{\omega_R}{\omega}\right)\right]} \qquad (7)$$

where,

$$m = 1 + \left(\frac{R_S}{R_P}\right) + \left(\frac{C_P}{C_S}\right) \qquad (8a)$$

$$n = \sqrt{\left(\frac{C_P R_S}{C_S R_P}\right)} \qquad (8b)$$

$$\omega_R = \frac{1}{\sqrt{C_P C_S R_P R_S}} \qquad (8c)$$

Frequency ω_R represents the resonant point of the Wien RC network and the frequency of oscillation for an amplifier with zero phase shift.

Using (5) and (7) in (1), the condition for oscillation in the present case is,

$$\left[b + j\left(\frac{\omega}{\omega_T}\right)\right] \times \left\{m + jn\left[\left(\frac{\omega}{\omega_R}\right) - \left(\frac{\omega_R}{\omega}\right)\right]\right\} = 1 + j0 \qquad (9)$$

From this equation, it follows that oscillations can be sustained at a frequency $\omega = \omega_O (\neq \omega_R)$ given by,

$$\frac{\omega_O}{\omega_R} = \sqrt{\frac{nb}{nb + mu}} = \frac{1}{\sqrt{1 + \frac{mu}{nb}}} \qquad (10)$$

$$\text{where } u = \frac{\omega_R}{\omega_T} \qquad (11)$$

Also from (9), the value of b to be used in (10) is found from the positive root of the quadratic equation,

$$mnb^2 + b(m^2u - n) + mu(nu - 1) = 0 \qquad (12)$$

The negative root of this equation has no physical significance since b is, by definition, a positive number (see, (3) and (6)).

The popular practical choice,

$R_P = R_S = R$, $C_P = C_S = C$

gives,

$m = 3$,
$n = 1$

and

$$\omega_R = \frac{1}{CR}$$

Then, (10) takes the simpler form,

$$\frac{\omega_O}{\omega_R} = \sqrt{\frac{b}{b + 3u}} = \frac{1}{\sqrt{1 + \frac{3u}{b}}} \qquad (13)$$

Similarly, (12) becomes,

$$3b^2 + b(9u - 1) + 3u(u - 1) = 0 \qquad (14)$$

■

References

1. Millman. J, and Grabel. A, 'Microelectronics' (2nd edition) McGraw-Hill, 1987, pp 663-4.
2. Op-cit., pp. 637-639.

Hands-on Internet

Better mosfet models, new amplifier designs and improved circuit simulation software are the topics of Cyril Bateman's Internet column this month. Cyril also discusses a new search engine enhancement that should save you a lot of wasted surfing time.

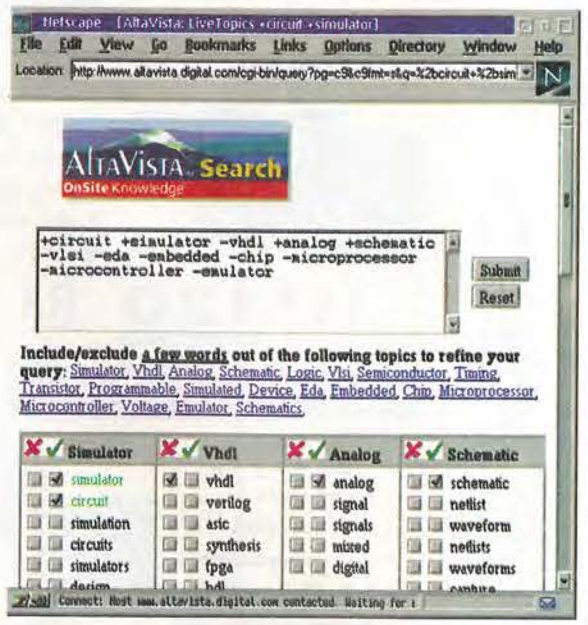


Fig. 1. AltaVista's 'Live Topics' interactive Search Wizard. My initial search using +circuit +simulator resulted in 9000 hits. The refined search using 'Live Topics' produced 100 high quality results.

The rapid growth of industrial and commercial Internet web pages, together with the fashion for large, more complex sites, continues apace. This growth has negated the speed recently gained from using faster modem access. Perhaps more importantly it has much increased the difficulty in locating data.

To counter this, the keyword search engines continue to add both storage and functionality to their systems. The inevitable result is that a search now returns with ever more matches.

Most keyword search engines encourage use of Boolean controlled keywords. Thus a search can become well focused if your chosen control words match those required by the search engine. If unmatched, the search result can be utterly useless.

AltaVista,[1] which always provided Boolean controls, has now added their 'Live Topics' facility. Live topics, also called 'Search Wizard', can be used with text only or Java enabled browsers. It provides a list of matching keywords to select or de-select for a more refined search.

These live topics words are not pre-defined, but are 'dynamically categorised' by the search engine, from analysis of all keywords found in the documents identified in your original search result.

Where to look

1	Alta Vista Search Engine	http://www.altavista.digital.com
2	The Human Search Engine	http://www.humansearch.com
3	Deep-submicron MosFet Research	http://rely.eecs.berkeley.edu:8080/researchers/sinitsky
4	BSIM3 Group	http://rely.eecs.berkeley.edu:8080/bsim3www
5	Silvaco SmartSpice	http://www.silvaco.com
6	Penzar Development	http://www.penzar.com/topspice.htm
7	Beige Bag Software	http://www.beigebag.com/b2spice2.htm
8	LSI Logic	http://www.lsilogic.com
9	Pass Laboratories	http://www.passlabs.com
10	High Fidelity Engineering	http://www.decware.com/zenamp.htm

Using Live Topics

To illustrate the new search tool, I ran a normal AltaVista Boolean search using two keywords, +circuit +simulator. This search found documents containing 541 426 occurrences of circuit and 115 958 for simulator, spread over 9000 document pages – far too many for me to digest.

While live topics is withheld for finds less than 200 documents, for this search it was of course offered me. Immediately under the 9000 document statement, I was given three choices, 'tables' for JavaScript enabled browsers, 'text-only' for any browser or 'help'.

Choosing tables resulted in a JavaScript interface of a box showing my original search words and tables of selectable sub-keywords, grouped under major headings. AltaVista advise selecting only one or two words, then repeat searching.

I chose to include schematic and analog, then exclude vhdl, vlsi, eda, embedded, chip, microprocessor, microcontroller and emulator. This revised search returned only 100 document pages – much more suited to my needs, **Fig. 1**.

Had I performed a conventional search, I would not have thought of many of these keywords. I would certainly not have deselected. This dramatically demonstrates the value of the live topics words, and the power of this new technique, which is unique to AltaVista.

Using the latest browser versions, a slightly different Java Interface should be available which downloads an applet to your system. It works in a similar way to that shown, but further improves the versatility of the software by providing a choice between table or graph based lists of the live topics words.

Should all the on-line search techniques described to date fail, help is to hand in Internet's equivalent of your friendly local librarian. A new website, HumanSearch[2], was started by seven students of the University of Rhode Island.

Intended to overcome the search problems which result from the rapidly increasing Web document base, HumanSearch offers use of its mostly unpaid volunteer team – of researchers free of charge – to search for any topic on your behalf. One particular speciality search they can perform is lost person tracing.

To use this unique search offering, simply fill in the one page questionnaire. Their team of skilled searchers will work on your request and within a couple of days, e-mail you with the results, **Fig. 2**.

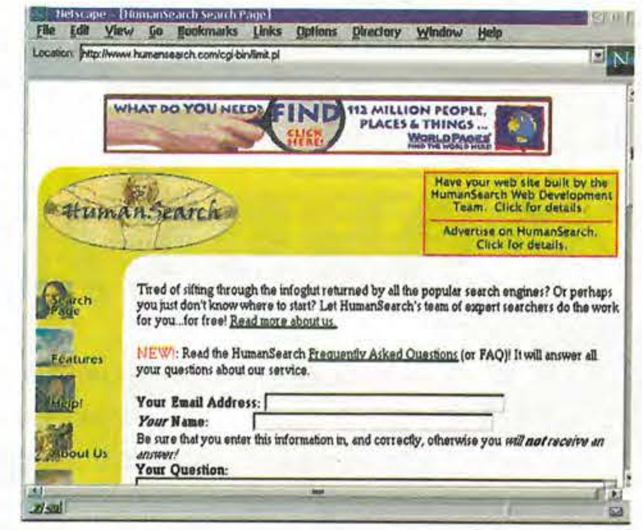


Fig. 2. Your friendly access, the HumanSearch web librarians. When automated search engines fail, why not try a search via humans?

Simulation and design software

My AltaVista live-topics search revealed a possible answer to Ian Hegglun's mosfet question, also three Spice based circuit simulators not previously mentioned.

In the May issue, Ian asks whether level-three mosfet models and/or *Electronics Workbench 5*, can accurately model in the crossover region. All Spice simulator-engine core software originates from The University of California at Berkeley. The latest Berkeley Spice is 3F5, used in this *Electronics Workbench* version.

For historical reasons, mosfet models, together with relevant core simulation software, are less well developed than those for bipolar devices. Berkeley currently has two mosfet simulation projects. A research project to investigate energy balance modelling for accurate current drive prediction of deep-submicron mosfets[3].

A second program under the generic heading of BSIM3[4] is responsible for actual BSIM implementations in Spice. While the 3F5 core has been BSIM compatible for some time, the BSIM3 level 3V3.1 mosfet feature was added only in February this year. As a result, it is necessary to carefully check the exact status of 3F5 based simulators.

Silvaco[5] produces the *SmartSpice* simulator and advanced Spice models. *SmartSpice* fully supports BSIM3V3.1, has enhanced both the physical model used and its numerical implementation. Thus Silvaco claims leadership in advanced Spice model technology.

The *SmartSpice* simulation results show their BSIM3V3 model eliminates certain negative capacitances found using

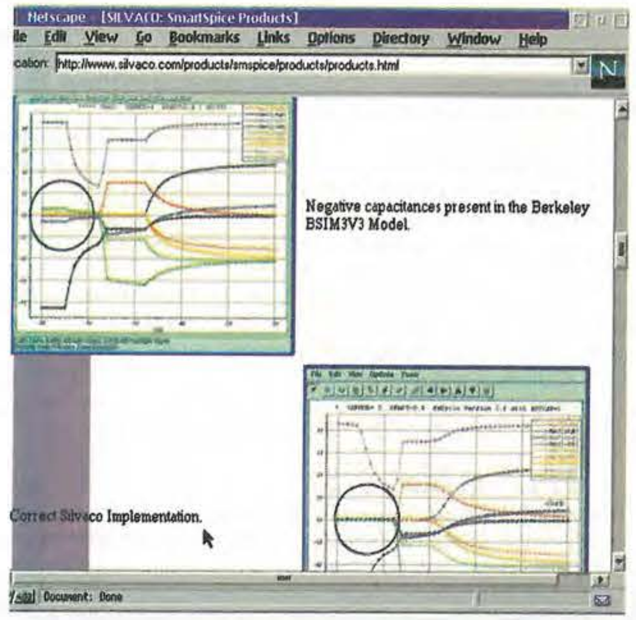


Fig. 3. Silvaco claims leadership in Spice Model Technology. This company's version is clearly improved, compared to standard BSIM3V3.

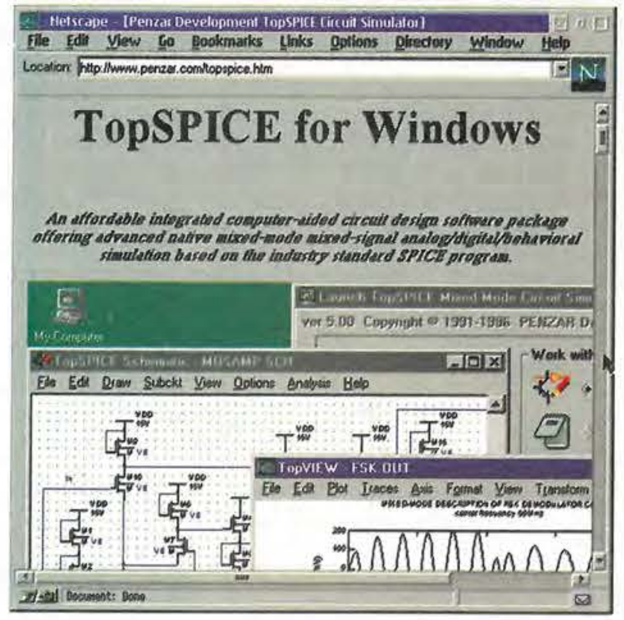


Fig. 4. TopSpice, the affordable Spice based analogue simulator. Unusually, TopSpice provides Smithchart and can use 'S' Parameters.

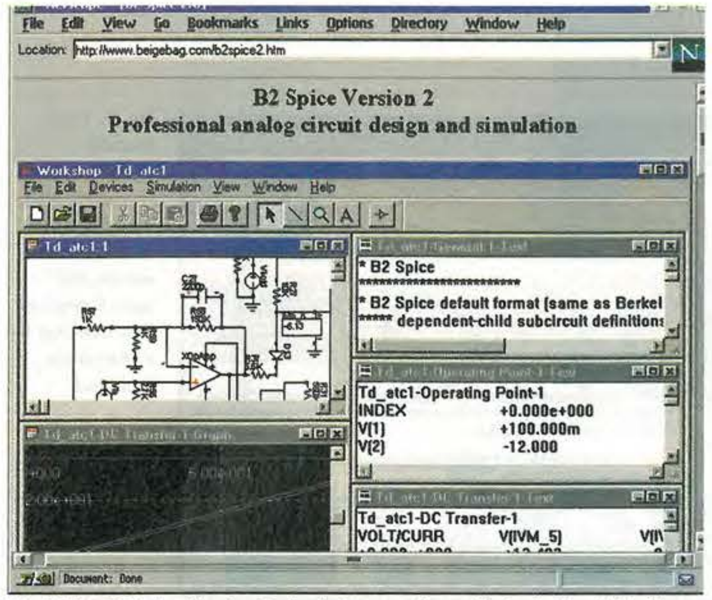


Fig. 5. The B2 Spice Version 2 simulator complete with Symbol and Model editors. Download and evaluate the trial version of this professional simulator.

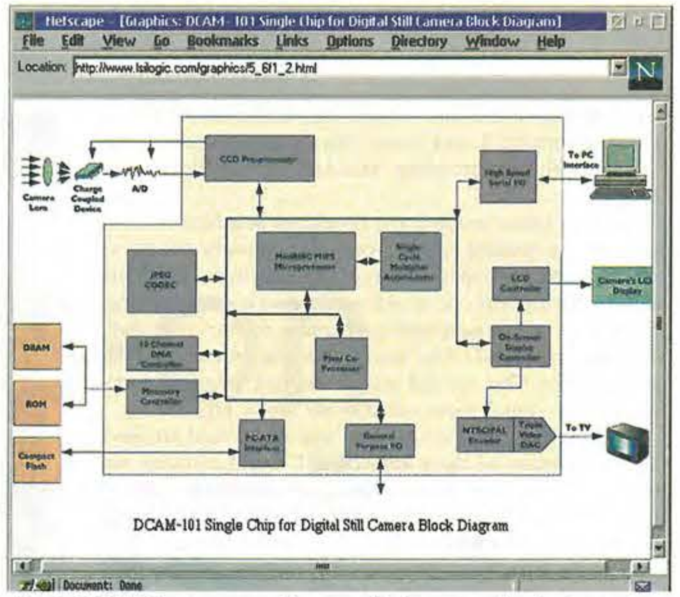


Fig. 6. Digital Camera on a chip – result of co-operation between two technologies. The Minolta LSI Logic co-operation commenced in 1996.

the Berkeley version. The company has also enhanced the impact ionisation current model for both short and long channel devices, **Fig. 3**.

TopSpice for Windows[6] is a 32-bit simulator offering analog/digital/behavioural mixed-mode simulations, together with schematic capture. Its built in logic simulator dramatically reduces mixed-mode simulation times. Unusually, this affordable simulator includes a Smith chart post processor able to handle 'S' parameters, **Fig. 4**.

Version 2 of B2 Spice[7] is available both for Windows and Macintosh. This simulator helps overcomes model availability problems by including both model and symbol editors. A 4Mbyte download trial version is available. The Windows version requires that either Access 7 be preinstalled, or their DAO program, which is also available to download, **Fig. 5**.

Circuit applications

As explained by Leslie Warwick in the July issue, although video still cameras offer the ideal method to input illustrations into engineers reports, those offering better than VGA resolution are expensive compared to comparable quality conventional film cameras. In part this cost results from digital functions requiring a number of integrated circuits. Following collaboration between the Minolta Company Ltd. and LSI Logic,[8] this looks set to change.

The *DCAM-101* chip incorporates all the major digital functions needed to make a high resolution camera, reducing costs and providing real-time operation rather than the delays inherent in other designs.

This chip handles high resolution charge-coupled device arrays up to 2000 by 2000 pixels, can capture, compress to JPEG, and store 3.3 million pixels per second at 24 bits/pixel. But it consumes less than 1.2W. It has the potential to allow lower cost cameras but with improved resolution **Fig. 6**.

Following the recent strong interest in audio topics and Professor Leach's low transient intermodulation amplifier design in the September issue, I now include two quite different designs found on Internet.

Nelson Pass, of PassLabs,[9] has considerable experience in audio amplifier design. His Web page includes a do-it-yourself section offering a free choice from four differing variations on his 'Zen' amplifier. These choices are the Class A, A-40, A75 and Citation12 designs.

Since recent published designs have tended to involve several stages with resultant complexity, I have chosen to show his minimalistic 'Zen' approach, **Fig. 7**.

Finally for those who prefer their active devices to run visibly hot, another novel approach is the 'single-ended tube amp' from High Fidelity Engineering.[10] Design considerations and circuit for this ongoing design can be found on Decware's page. This amplifier trades off the number of active stages, against dc current design complications in the output transformer. ■

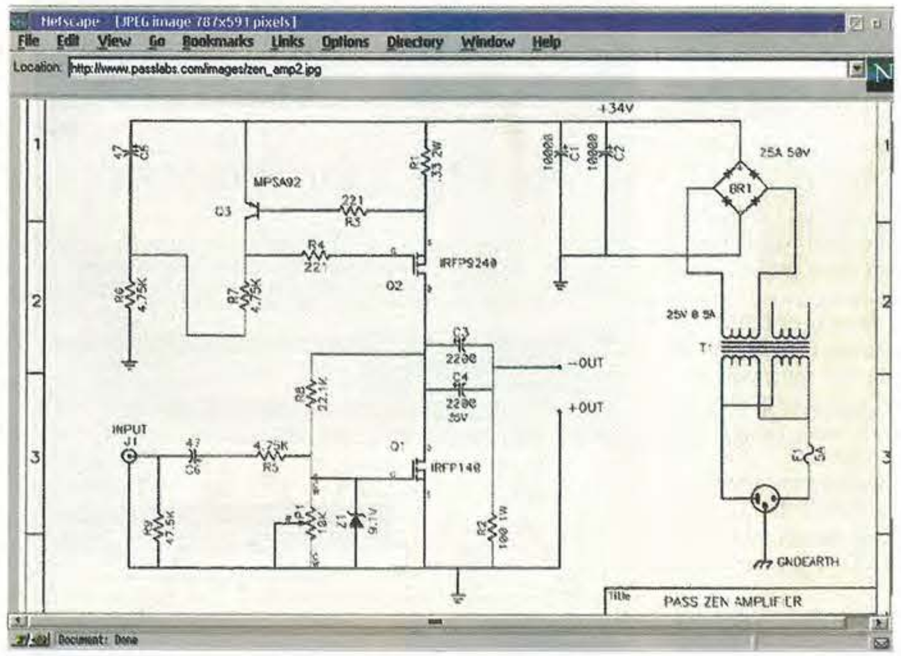


Fig. 7. The Nelson Pass 'Zen' minimalist mosfet amplifier. – one of eight quite different DIY designs offered by PassLabs.

Working with microwaves

Boris Sedacca looks at the problems involved with developing circuits for microwave frequencies – where capacitors and inductors become tracks on the pcb.

The ultimate test of any electronic gadget is whether or not it works when it is switched on. Before it gets to that stage, there is an exhaustive list of equipment to empty the deepest pockets. This is particularly true with microwave gadgets.

But if you can beg, borrow or steal time on a scalar network analyser and sweep generator, you can build something fairly cheaply. Most of the work is in the design and software simulation, and computer time is relatively plentiful.

Obtaining the right software again depends on the depth of your pocket. Microwave design software tends to be expensive, but you can often have a package for a trial period from the supplier.

Introduction

Why would you want to get into microwaves? Well, for a start you can eliminate many components.

If you design a receiver that works at 5GHz and beyond, then you can use microstrip. The advantage of microstrip technology is that the geometric shape of track on a substrate determines whether it behaves as a capacitor, inductor, filter, attenuator, mixer – or even an antenna.

Fewer physical components means lighter weight too. And microstrip is compatible with surface mount technology.

The downside? Producing a design is a lot more complicated. With the right design calculations, you can go straight into a CAD package like *AutoCad* and produce a mask, but before you do that, a microwave simulation package will prove immensely helpful. Conventional electronics CAD packages are of little use at microwave frequencies.

The two leading contenders in the microwave simulation software market, are Compact Software's *Serenade 7* and Hewlett Packard's *Series IV*. First a little background.

Transmission lines

If you strike an 'A' tuning fork – i.e. one with an order of magnitude of 440Hz – and place it next to a piano or guitar string that is also tuned to A, the string will vibrate in tune because of the sound pressure waves propagated by the tuning fork. The surrounding air acts as a transmission line.

In a similar way, a pair of open wires carrying an alternating signal can propagate an electromagnetic wave through the air that can excite ac through an aerial at a particular distance away from the current source.

When a transmission line is terminated in an impedance other than its characteristic impedance, you get reflection.

Distributed amplifiers

To widen bandwidth, rf design engineers today have a method available called distributed amplification. Although distributed amplifiers involve microstrip technology, they are easy to set up on monolithic microwave integrated circuits, or MMICs, and much more difficult on hybrid circuits. MMIC technology did not exist of course, when Percival[1] originally patented his distributed amplifier in 1937, but apart from that, very few modifications of the original concept have surfaced.

The distributed amplifier produces relatively lower gain than the multiplicative amplifier but yields significantly larger bandwidths through the ingenious use of the active devices' parasitics. Unlike conventional amplifiers, the higher the frequency it operates at, the better it works.

The problem is that it needs matched transmission lines and this is difficult if not impossible to implement in hybrid form. It only works efficiently if the parameters of every one of those transistors are identical and that can only be guaranteed by having them all from the same GaAs substrate. Even if they were identical the surface mounts are unlikely to be uniform.

GaAs fets capable of operation up to 100GHz have useful amplifier gains limited to around 10GHz. Ayasli[2] has provided a simplified equivalent-circuit model for a four-stage fet travelling wave amplifier. In this circuit, microstrip lines are periodically loaded with the complex gate and drain impedances of the fets, forming lossy transmission line structures of different characteristic impedance and propagation constant, **Fig. 1**.

An rf signal applied at the input end of the gate line travels down the line to the other end, where it is absorbed by the terminating impedance. However, the gate circuits of the individual fets dissipate a significant portion of the signal along the way. The input signal sampled by the gate circuits at different phas-

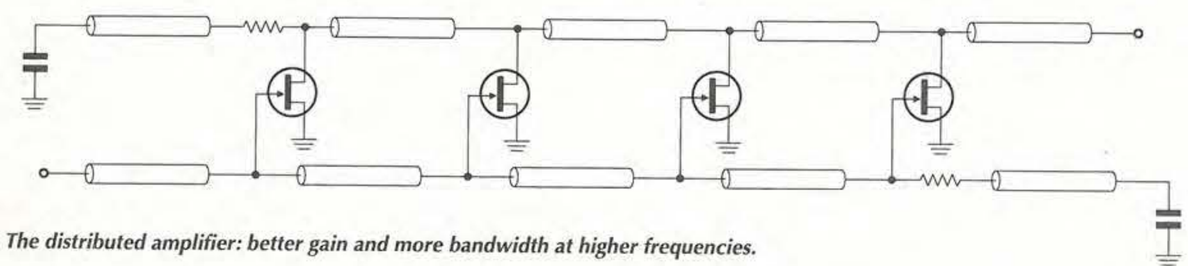


Fig. 1. The distributed amplifier: better gain and more bandwidth at higher frequencies.

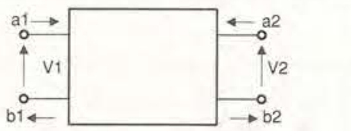


Fig. 2. The FET as a two-port, four-terminal device.

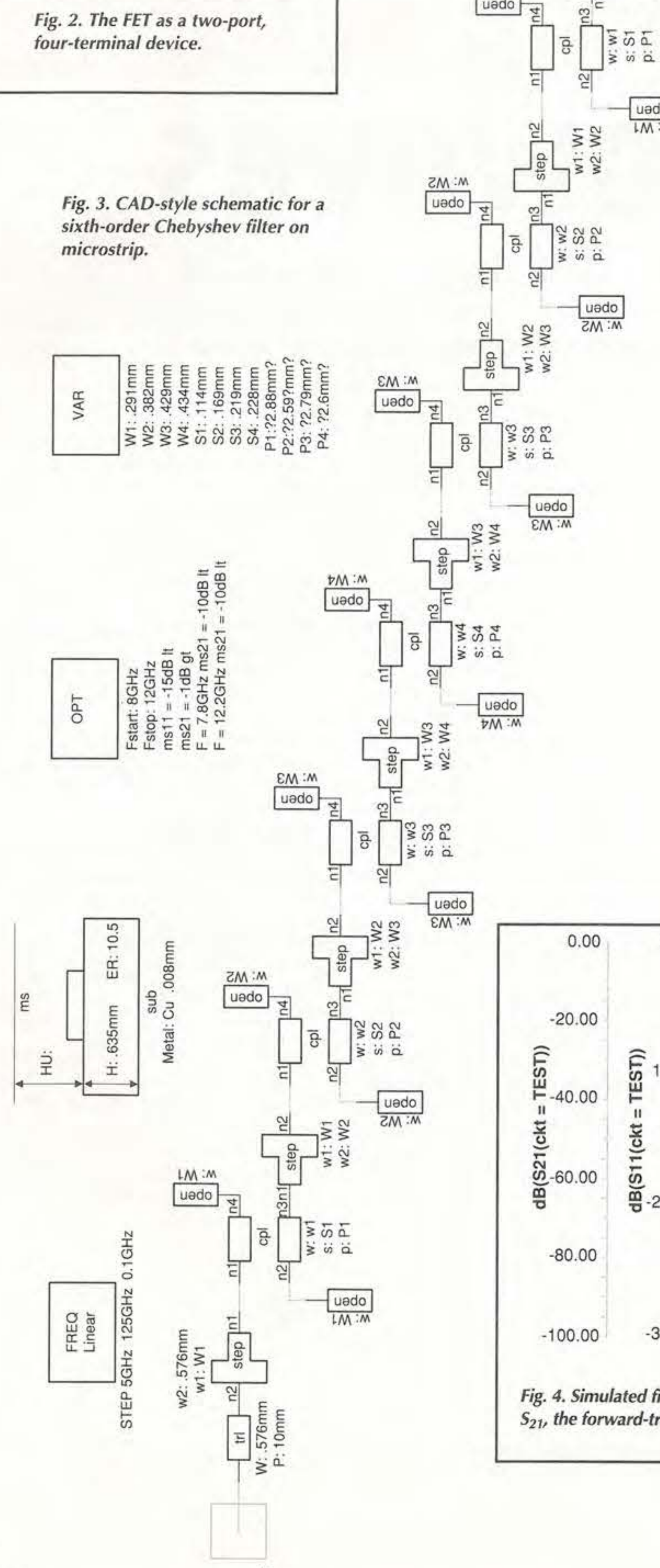


Fig. 3. CAD-style schematic for a sixth-order Chebyshev filter on microstrip.

es – and generally at different amplitudes – is transferred to the drain line through the transconductance of the fets.

If the phase velocity of the signal at the drain line is identical to the phase velocity of the gate line, then the signals on the drain line add. The addition will be in phase only for the forward-travelling signal.

S-parameters

A microwave transistor's S-parameter specifications is essentially a table of its behaviour at different frequencies, which allows us to interpret it as a two-port, four-terminal device where a_1 and a_2 are waves entering the device at ports 1 and 2, and b_1 and b_2 are waves leaving the device, **Fig. 2**.

The main columns of the individual rows of the table contain the following information where,

Input reflection	$S_{11}=b_1/a_1$
Forward gain	$S_{21}=b_2/a_1$
Output reflection	$S_{22}=b_2/a_2$
Reverse gain	$S_{12}=b_1/a_2$

The first three are the most important, with the last, S_{12} usually being so close to zero that it can be ignored.

These relationships can be represented in the form of an equation [b]=[S][a] and this can be extended into a matrix,

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

For maximum power transfer, we need conjugate matching of complex impedances such that,

$$\begin{bmatrix} S_{11} & S_{21} \\ S_{12} & S_{22} \end{bmatrix} \begin{bmatrix} S_{11}^* & S_{12}^* \\ S_{21}^* & S_{22}^* \end{bmatrix} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$

As an example, the Mitsubishi *MGF1303B* GaAs fet which retails around £5.50, has the following S-Parameters in the 10GHz row of the table,

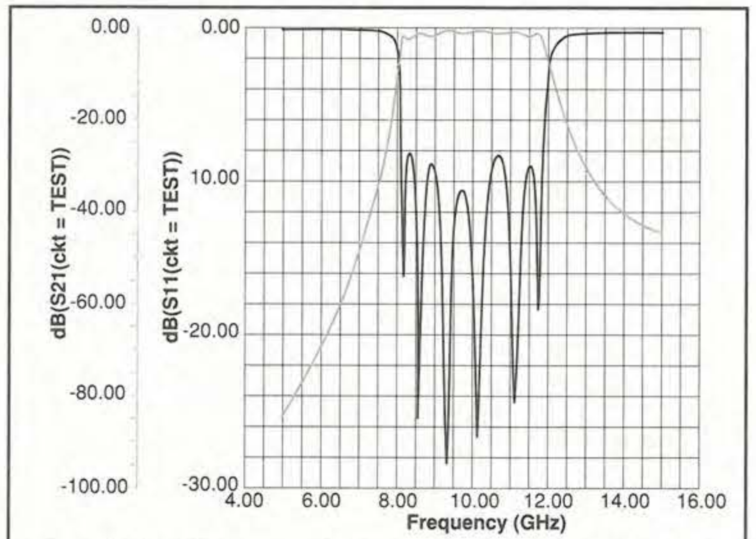


Fig. 4. Simulated filter response for the above schematic, Fig. 3. Grey curve is S_{21}, the forward-transfer gain while the black curve is S_{11} – input reflection.

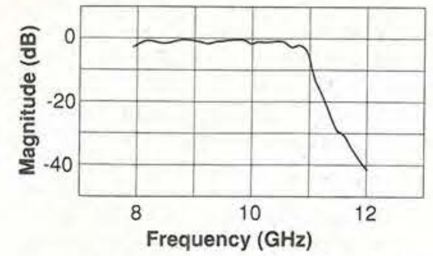


Fig. 5. Actual measured filter performance for gain, S21.

$$S_{11} = 0.704\angle -163.1^\circ \quad S_{21} = 1.712\angle -17.1$$
$$S_{12} = 0.082\angle 20.1^\circ \quad S_{22} = 0.584\angle -138.2^\circ$$

These are vectors normalised to a characteristic impedance of 50Ω. For maximum power transfer, matching input and output networks are needed,

$$S_{11}^* = 0.704\angle 163^\circ \text{ and } S_{22}^* = 0.584\angle 138.2^\circ$$

respectively. But before you do that, you need bandpass filters at both ends – at the input to stop unwanted frequencies being amplified, and at the output because the amplifier itself will introduce spurious signals.

Filter design and simulation

As wideband amplification is being looked at here, consider a centre frequency f_0 of 10GHz and a passband of 8 to 12GHz for this example. In addition, assume that the signal is to be attenuated to less than –10 dB at 7.8 GHz and 12.2GHz respectively, and that the passband ripple is to be less than 1dB.

This gives a fractional bandwidth,

$$\delta = \frac{f_2 - f_1}{f_0} = \frac{12-8}{10} = 0.4$$

and a frequency transformation from the low-pass prototype filter to the bandpass filter,

$$\frac{\omega_i'}{\omega_c} = \left[\frac{2}{\delta}\cdot\frac{f_i - f_0}{f_0}\right] = \left[\frac{2}{0.4}\cdot\frac{7.8_i - 10}{10}\right] = 1.1$$

so,

$$\frac{\omega_i'}{\omega_c} - 1 = 0.1$$

For a full explanation of these equations, refer to Edwards[3].

The last value gives us an X-axis co-ordinate for reading off a chart of Chebyshev filter characteristics. Matthaei[4] et al have produced charts and tables for passband ripple from 0.01dB to 3dB. In this case we are looking at a ripple of 1 dB, and reading off –10 dB attenuation on the Y-axis gives the order of the filter n=6. From that we get seven g-values from a table of prototype elements, plus a value for g_0=1, as follows:

g=(1 2.1456 1.1041 3.0634 1.1518 2.9367 0.8101 2.6599)

At this stage, the calculations become incredibly tedious, so I recommend the use of a maths package like *MathCad* or *MatLab*. You can even use a spreadsheet but a problem here can be that the equations are normally hidden from the user. I used *MathCad*, but for heavyweight engineering work, you may want to use *MatLab* instead.

As I am discussing using parallel-coupled, or edge-coupled, microstrip lines instead of inductors and capacitors, the above values are converted to seven admittance inverter, or J, parameters using a characteristic impedance of Z_0=50Ω as follows. For the first coupling structure,

$$J_0 = \frac{1}{Z_0}\cdot\sqrt{\frac{\pi\delta}{2g_0g_1}}$$

for the intermediate coupling structures with k=1...5,

$$J_k = \frac{1}{Z_0}\cdot\frac{\pi\delta}{2\omega_c'\sqrt{g_kg_{k+1}}}$$

and for the final coupling structure,

$$J_6 = \frac{1}{Z_0}\cdot\sqrt{\frac{\pi\delta}{2g_6g_7}}$$

so,

J=(0.011 8.147×10^{-3} 6.833×10^{-3} 6.69×10^{-3} 6.833×10^{-3} 8.147×10^{-3} 0.011).

Note the symmetry of the coupled sections. Now the seven odd and even-mode impedances can be worked out with k=0...6 as follows,

$$Z_{0o_k} = Z_0\left(1 - J_k \times Z_0 + (J_k)^2 \times (Z_0)^2\right)$$

and,

$$Z_{0e_k} = Z_0\left(1 + J_k \times Z_0 + (J_k)^2 \times (Z_0)^2\right)$$

giving,

Z_{0o} = (37.58 37.929 38.74 38.87 38.74 37.929 37.58)

and,

Z_{0e} = (91.582 78.666 72.918 72.319 72.918 78.666 91.582)

From this you can obtain the widths and spacings in millimetres, as well as the odd and even-mode permittivities. In turn you can then work out the lengths of the individual microstrip coupled-section by looking them up in a table, or by keying the figures into a specific microwave design package such as *Serenade 7* or *Series IV*. The following figures are from *Serenade 7*'s transmission lines utility,

w=(.291 .382 .429 .434 .429 .382 .291)
s=(.114 .169 .219 .226 .219 .169 .114)
ε_o=(5.8465 5.9098 5.9585 5.9648 5.9585 5.9098 5.8465)
ε_e=(7.0714 7.2556 7.3437 7.3529 7.3437 7.2556 7.0714)

These figures are based on *RT Duroid* – a

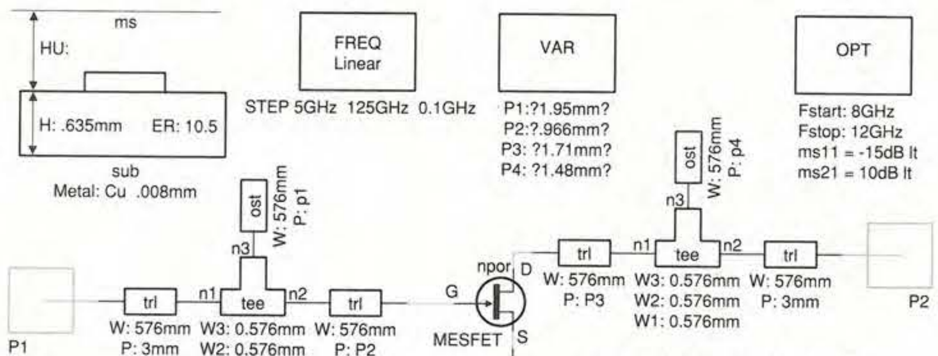


Fig. 6. Schematic for a single-stage GaAs fet amplifier

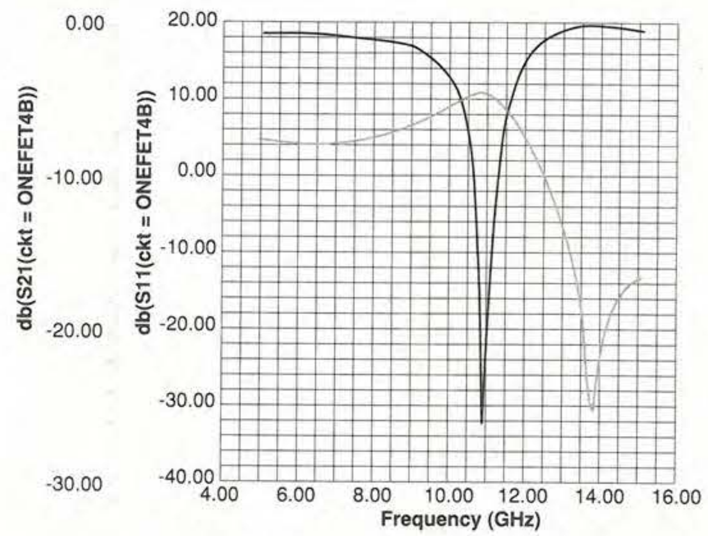


Fig. 7. Simulated MGF1303B GaAs fet amplifier response. Grey curve S_{21} is forward-transfer gain while the black one is S_{11}, the input reflection.

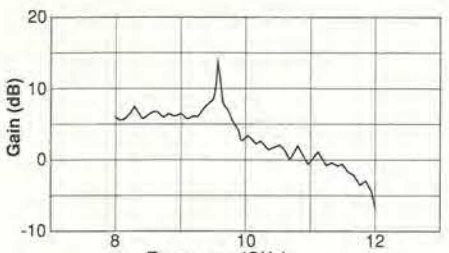


Fig. 8. Measured performance for amplifier gain (S21)

copper-clad substrate that can be etched in a similar way to conventional pcbs. Although marketed as a prototyping material for microstrip, *RT Duroid* is good enough for many mainstream production applications.

In this case, the substrate's thickness – height h – is 0.635mm, copper track thickness, t, is 0.008mm, and relative permittivity ε_r is 10.5.

It is assumed that centre frequency f_0 is 10GHz, the free space wavelength,

$$\lambda_0 = \frac{c}{f_0} = \frac{3\times10^8}{10\times10^9} = 0.03\text{m} = 30\text{mm}$$

where c is the velocity of electromagnetic waves in free space.

Now you can work out the odd and even-mode substrate wavelengths as follows,

$$\lambda_{o_k} = \frac{\lambda_0}{\sqrt{\varepsilon_{o_k}}} \text{ and } \lambda_{ek} = \frac{\lambda_0}{\sqrt{\varepsilon_{e_k}}}$$

giving,

λ_o=(0.012 0.012 0.012 0.012 0.012 0.012 0.012)

and

λ_e=(0.011 0.011 0.011 0.011 0.011 0.011 0.011)

Hence the average substrate wavelength λ_{AV}=0.012m=12mm. The microstrip sections are quarter wavelengths and need to be shortened to take account of discontinuities like the open-end effect and step widths, involving more tedious calculations which the reader will probably be relieved to know are not included here for lack of space.

The *Serenade 7* schematic in **Fig. 3** shows the final dimensions of the filter in the variables block, or VAR. Starting with the control blocks at the top, FREQ sets the sweep generator from five to 15GHz in increments of 50MHz, SUB specifies the substrate, and OPT sets the optimisation goals.

In the VAR control block, W, S and P are the widths, spacings and lengths respectively of each filter section. The question marks surrounding the P values signify that these are the variables to be optimised.

The microstrip circuit elements consist of the input and output ports P_1 and P_2. There are also two 10mm lengths of transmission line TRL with characteristic impedance Z_0 of 50Ω – in the case of this substrate, *Serenade 7*'s Transmission Lines utility output a width of 0.576 mm. The STEP elements takes account of step changes in widths, CPL microstrip elements of the coupled sections, and OPEN elements of the open-circuit end effects.

After optimisation, the output shown in **Fig. 4** was obtained, showing a near-perfect filter characteristic – in theory at least. The S_{11} parameter is shown in black, S_{21} in grey. Now similar results were obtained with *Series IV*. However, *Series IV* does not need open-end elements because it has specific coupled filter elements, MCFIL, which include open-end effects. Furthermore it has a layout utility included in the product, whereas although *Serenade 7* has its own layout utility based on AutoCAD, it uses a third party add-on called *S2A*.

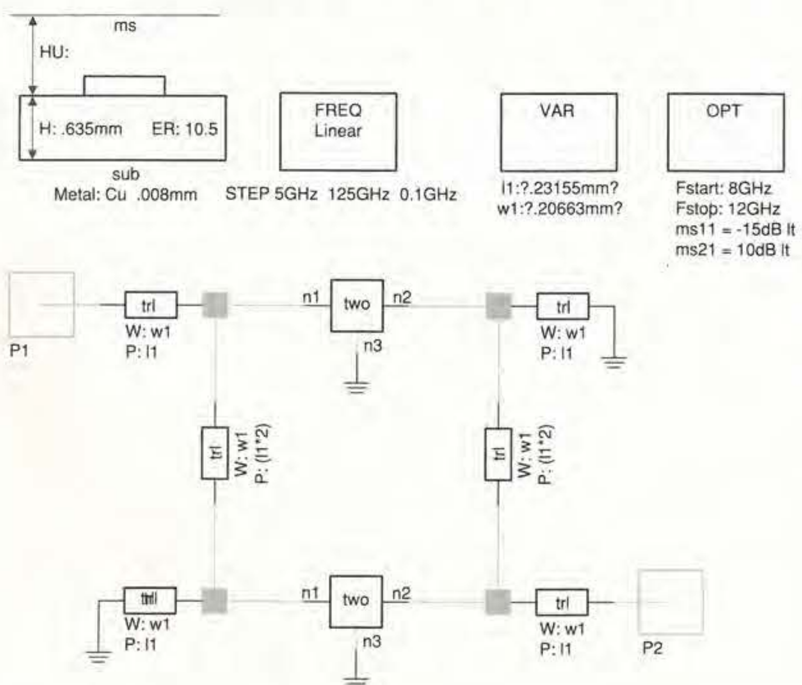


Fig. 9. Schematic for a two-stage distributed amplifier

Series IV does not like working in metric units, so the dimensions have to be converted to thousandths of an inch. Personally, I prefer to work in metric units.

So much for the theory, but what happens in practice? Well, it matches theory quite accurately, in fact. I built a microstrip filter and at the time, I was not aware of the effect of step width discontinuities, so the bandwidth has been pulled lower down the frequency range.

The dimensions obtained from earlier simulations were used to produce an *AutoCAD* drawing from which a mask was made. There is no space here for more detail about frabrication, so the following chart is a preview. The data was output from a Hewlett Packard vector network analyser and imported into Lotus 1-2-3, **Fig. 5**.

Amplifier design and simulation

Designing any amplifier involves a trade-off between conflicting objectives of maximising gain, maximising bandwidth and minimising noise – among others.

Earlier, I mentioned the Mitsubishi *MGF1303B* GaAs fet S-parameters at 10GHz. For maximum gain at that frequency, resonant matching is needed for S_{11} and S_{22}. Using *Serenade 7* or *Series IV*'s transmission-lines utility again, you get an effective permittivity ε_{eff}=7.36, from which you can calculate the substrate wavelength,

$$\lambda_s = \frac{\lambda_0}{\sqrt{\varepsilon_{eff}}} = 11.1\text{mm}$$

At this point, the Smith chart becomes invaluable for transmission line design, both for lumped components and microstrip. This facility is available on both products, but I found it easier to use the conventional hardcopy version. I have not included it here for lack of space.

Polar co-ordinates of $S_{11}=0.704\angle{-163.1°}$ correspond to the complex rectangular form 0.18–j0.145 on the chart, representing the source impedance Z_S, the inverse of which is then converted to a source admittance Y_S=3.37+j2.71. The distance between these two values represents the diameter of a circle with its origin at the centre of the chart.

By extending the diameter from Y_S to the edge of the chart, you can read off the wavelength as 0.226. To get a resonant match, first move clockwise along the circle until it intersects the unity admittance circle, to get a value for Y_1=1–j2. Drawing a straight line from the centre of the chart through Y_1, you can read off the wavelength as 0.313.

Now the length of microstrip can be found from the transistor to the stub, or distance d_s=(0.313–0.226)λ_s=0.966mm. Next, cancel the imaginary part by drawing another straight line from the centre of the chart through +j2 at the edge of the chart, giving a wavelength of 0.176. Therefore the length of the stub l_s=0.176λ_s=1.95mm.

The same procedure is applied for $S_{22}=0.584\angle{-138.2°}$, representing the load impedance Z_L, giving,

dL=(0.325–0.192)λs=1.48mm

and,

lL=0.154λs=1.71mm.

You can now simulate this design as shown **Fig. 6**. From this the output shown in **Fig. 7** was obtained, again with S_{11} shown in blue and S_{21} in red. Note the resonant peak of about 11dB in S_{21} just below 11GHz. This matches published maximum available gain at that frequency for the *MGF1303B* fet.

Now the filter simulation matched reality quite accurately and the fact that it is a passive design makes all the difference. When it comes to an amplifier, this is an entirely different ball game because as an active device it needs power, which means that the gate and drain lines need to be biased using bias pads.

The measured results are shown in **Fig. 8**. This is nowhere near as close to the simulation as the results from the filter, but it does show a clear resonant peak at about 9.5GHz. Apart from the biasing, the variance may also be accounted for by the fact that the tracks on the circuit where narrower than those on the mask because of over-etching. This fact was established when the circuit was inspected using an Alpha step machine.

There are various output formats available from *Series IV*'s layout facility including Gerber, AutoCAD and HP Graphics Language.

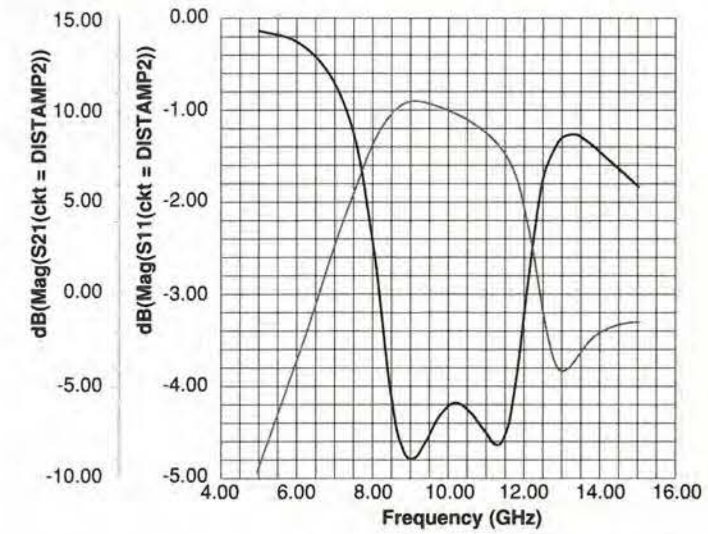
So much for resonant or narrow-band amplifier matching. As I mentioned earlier, to increase the bandwidth you can use a distributed amplifier setup such as that in **Fig. 9**.

Note the absence of stubs – just uniform lengths of transmission line from the ports and to the grounds, and twice the length between the amplifier stages – two in this case. For convenience I have used three-terminal, two-port models here to represents the GaAs fets, but the effect is the same.

Now the length and width of the line is crucial, and there are no straightforward analytical design formulas one can apply to derive these measurement. The only practical way is

continued on page 871...

Fig. 10. Simulated performance for distributed amplifier. Grey curve S_{21} is forward-transfer gain while the black one is S_{11}, input reflection. Note how the bandwidth is spread.



ELECTRONICS WORLD — READER INFORMATION SERVICE

INCORPORATING WIRELESS WORLD

For more information about any of the products or services in this issue of **ELECTRONICS WORLD**, simply ring the relevant enquiry number.

Enquiry numbers may be found at the bottom of each individual advertisement.

101	102	103	104	105	106	107	108	109
110	111	112	113	114	115	116	117	118
119	120	121	122	123	124	125	126	127
128	129	130	131	132	133	134	135	136
137	138	139	140	141	142	143	144	145
146	147	148	149	150	151	152	153	154
155	156	157	158	159	160	161	162	163
164	165	166	167	168	169	170	171	172
173	174	175	176	177	178	179	180	181

Name

Job title

Company Address

Telephone OCTOBER 1997

Only tick here if you do not wish to receive direct marketing promotions from other companies. ☐

Newsagent order form

Pass this order form to your newsagent to ensure you don't miss the next issue of *EW.*

To ....
(name of Newsagent)

Please reserve me the November issue of *Electronics World*- on sale 2nd October - and continue to order every month`s issue until further notice

Name..

Address ....

....

....

....

Thank you

Subscribe today!

Guarantee your own personal copy each month

Save on a 2 year subscription

ELECTRONICS WORLD
INCORPORATING WIRELESS WORLD

Postage will be paid by licensee

Do not affix postage stamps if posted in Gt Britain, Channel Islands, N Ireland or the Isle of Man.

Business Reply Service
Licence No. CY711

ELECTRONICS WORLD
Reader Information Service
Reed Business Information
Oakfield House
Perrymount Road
Haywards Heath
Sussex RH16 3BR

21

SEE OVER!

ELECTRONICS WORLD
INCORPORATING WIRELESS WORLD

SUBSCRIPTION CARD

Please enter my subscription to ELECTRONICS WORLD. I enclose Cheque/Eurocheque to the value of £ ______ made payable to **Reed Business Information**
Please charge my Mastercard/Visa/Amex account

With £ ______ Expiry Date ______

Signature ______

Name ______

Job Title ______

Address ______

______ Postcode ______

Tel: ______ Country ______

SUBSCRIPTION RATES

UK 1 year	£32
UK 2 years	£58
Student rate(proof required)	£21
Airmail	
Europe 1 year	£46
Europe 2 years	£83
Rest of the world 1 year	£56
Rest of the world 2 years	£101
Surface mail 1 year	£37

Post to:
ELECTRONICS WORLD
P.O. Box 302
Haywards Heath,
West Sussex RH16 3DH UK.

CREDIT CARD HOTLINE
Tel: +44 01444 445566
Fax: +44 01444 445447

Please tick here if you do not wish to receive direct marketing-promotion from other companies ☐

049

ELECTRONICS WORLD
INCORPORATING WIRELESS WORLD

SUBSCRIPTION CARD

Please enter my subscription to ELECTRONICS WORLD. I enclose Cheque/Eurocheque to the value of £ ______ made payable to **Reed Business Information**
Please charge my Mastercard/Visa/Amex account

With £ ______ Expiry Date ______

Signature ______

Name ______

Job Title ______

Address ______

______ Postcode ______

Tel: ______ Country ______

SUBSCRIPTION RATES

UK 1 year	£32
UK 2 years	£58
Student rate(proof required)	£21
Airmail	
Europe 1 year	£46
Europe 2 years	£83
Rest of the world 1 year	£56
Rest of the world 2 years	£101
Surface mail 1 year	£37

Post to:
ELECTRONICS WORLD
P.O. Box 302
Haywards Heath,
West Sussex RH16 3DH UK.

CREDIT CARD HOTLINE
Tel: +44 01444 445566
Fax: +44 01444 445447

Please tick here if you do not wish to receive direct marketing-promotion from other companies ☐

049

CIRCUIT IDEAS

Over £600 for a circuit idea?

New awards scheme for circuit ideas

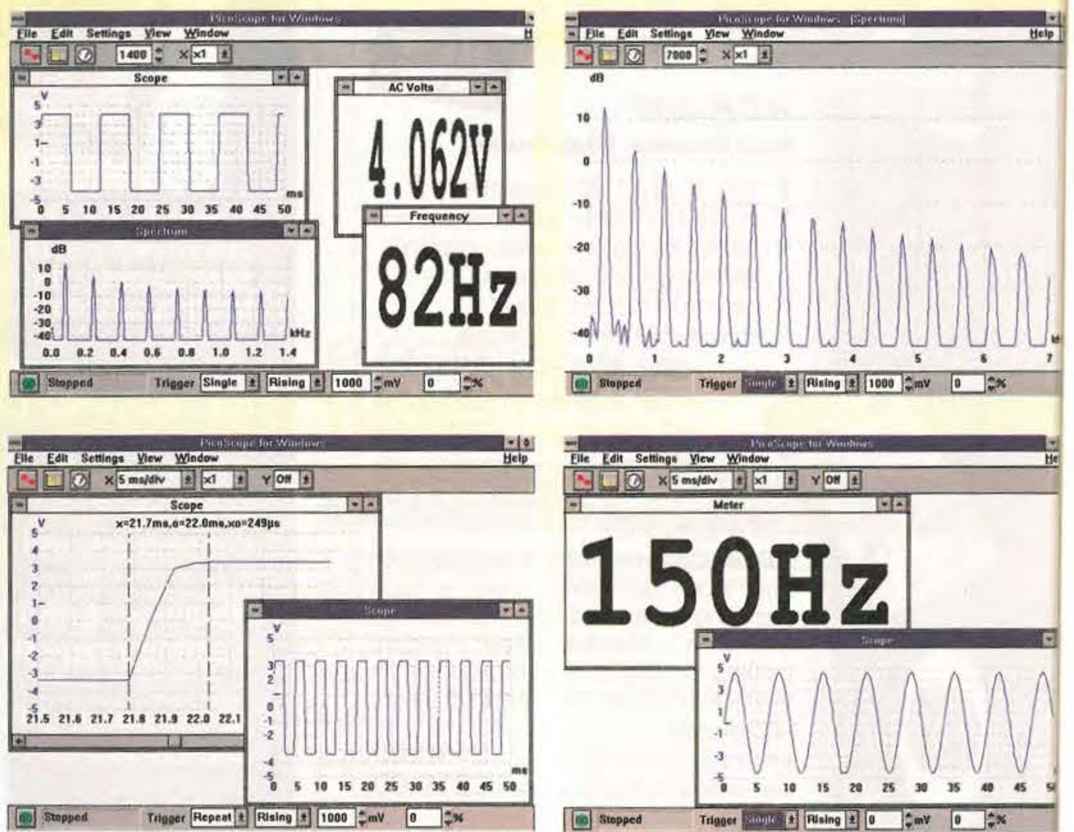
- Every circuit idea published in *Electronics World* receives £35.
- The pick of the month circuit idea receives a Pico Technology ADC42 – worth over £90 – in addition to £35.
- Once every six months, Pico Technology and *Electronics World* will select the best circuit idea published during the period and award the winner a Pico Technology ADC200-50 – worth £586.

How to submit your ideas

The best ideas are the ones that save readers time or money, or that solve a problem in a better or more elegant way than existing circuits. We will also consider the odd solution looking for a problem – if it has a degree of ingenuity.

Your submission will be judged on its originality. This means that the idea should certainly not have been published before. Useful modifications to existing circuits will be considered though – provided that they are original.

Don't forget to say why you think your idea is worthy. We can accept anything from clear hand writing and hand-draw circuits on the back of an envelope. Type written text is better. But it helps us if the idea is on disk in a popular pc or Mac format. Include an ascii file and hard-copy drawing as a safety net and please label the disk with as much information as you can.



Turn your PC into a high-performance virtual instrument in return for a circuit idea.

The ADC200-50 is a dual-channel 50MHz digital storage oscilloscope, a 25MHz spectrum analyser and a multimeter. Interfacing to a pc via its parallel port, ADC200-50 also offers non-volatile storage and hard-copy facilities. Windows and DOS virtual instrument software is included.

ADC42 is a low-cost, high-resolution a-to-d converter sampling to 12 bits at 20ksample/s. This single-channel converter benefits from all the instrumentation features of the ADC200-50.

ADC-42 WINNER

NiCd capacity meter

Unusually, this capacity meter draws power from the NiCd battery under test. It discharges battery packs of from four to ten cells, a charger being connected after the test. As an example, a four-cell pack applies 3.5V to the circuit after allowing for current sensing and reverse polarity protection.

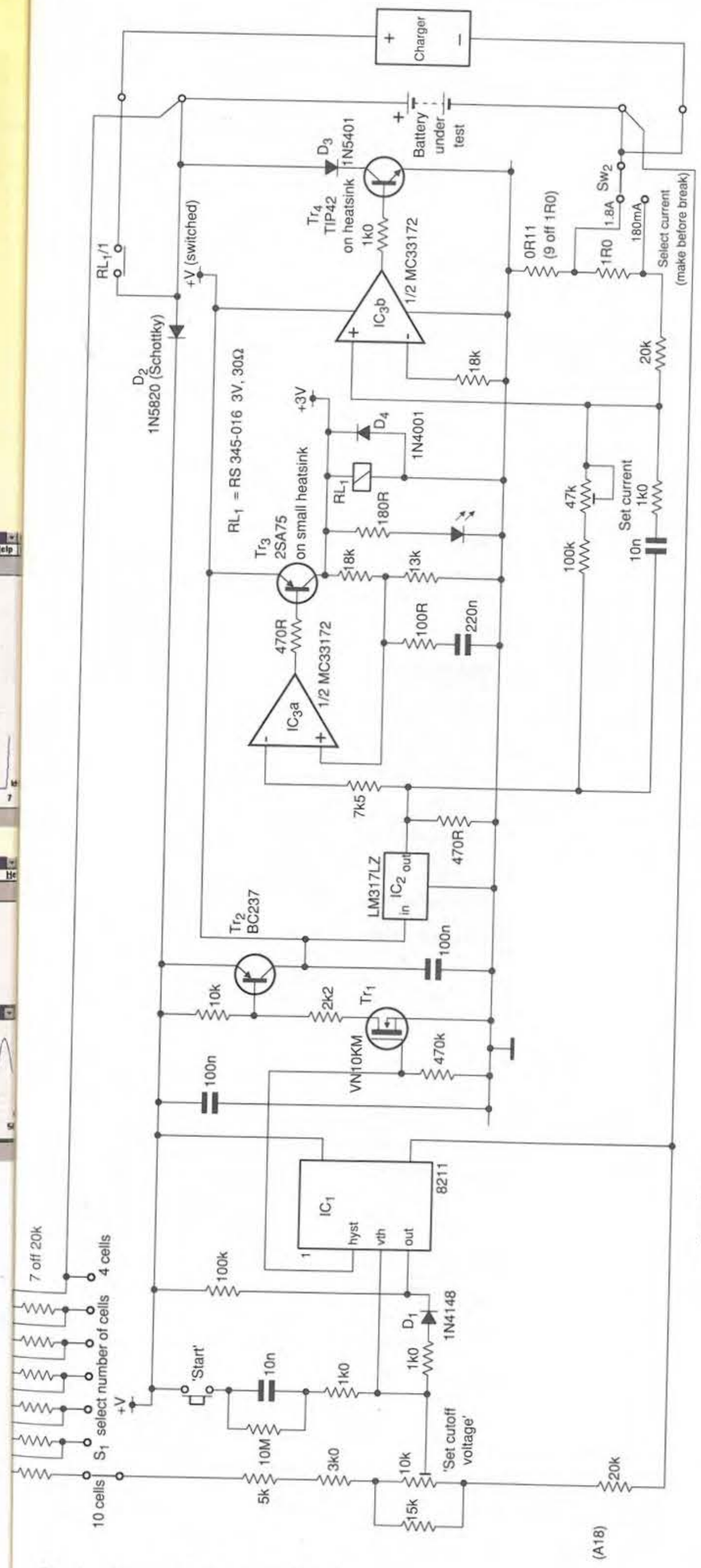
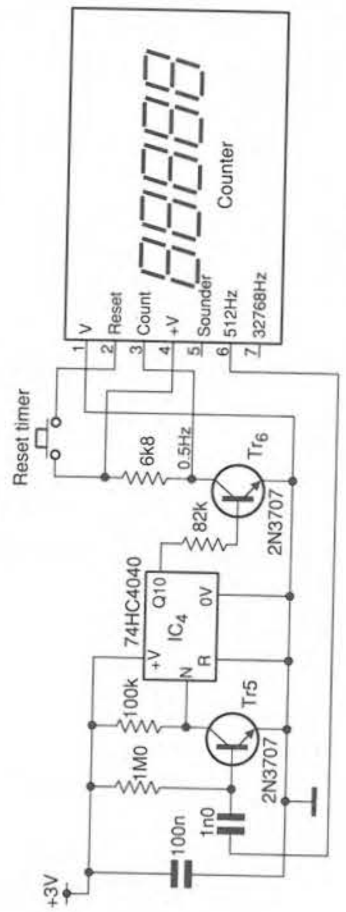
Latching comparator IC_1 starts the discharge when the start button is pressed, discharge ending when battery voltage falls to 1V/cell. In this case, the functions of output and hysteresis pins are reversed, since at low currents their threshold voltages differ by a few millivolts, the output being higher and preventing clean switching. Output current from the hysteresis pin is low and is amplified in Tr_1 and Tr_2, the resulting switch being capable of supplying over 100mA.

The discharge circuit is formed by IC_{3b} and Tr_4 and is switchable to 180mA and 1.8A by S_2. Very good power supply rejection in the *MC33172* (IC_{3b}) renders the discharge current largely immune to battery-voltage changes; the prototype's discharge current varies by less than 0.2% for a battery-voltage change of 4-25V.

To measure discharge time, the *Maplin FS13P* counter accepts its own 512Hz output, divided by the *4040* counter, to increment at two-second intervals while the battery is discharging. This is equivalent to 0.1 or 1mAh per count on the low and high current ranges. Power for the counter timer comes from IC_{3a} and Tr_3, which form a low dropout regulator to supply the relay.

Lower discharge currents may be used; PP9 batteries at 18mA can be handled, but the relay must be omitted. Higher currents or more cells will necessitate a heat sink for Tr_4.

John R Hunt
Middlesbrough

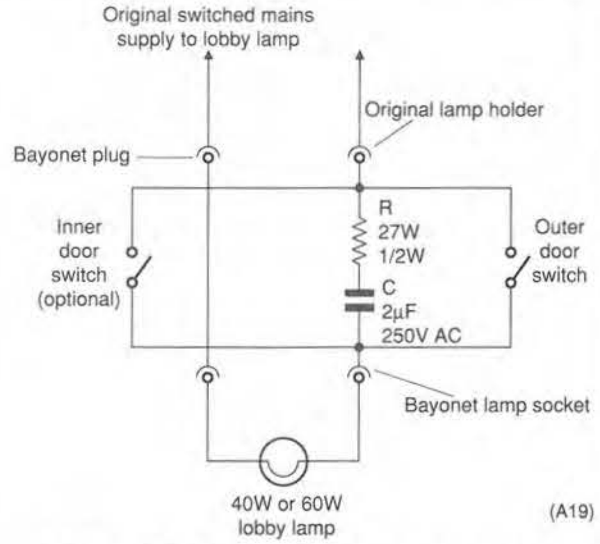
NiCad tester discharges batteries, timing the discharge to measure capacity and indicating the result digitally. Batteries are then recharged automatically.

Porch light saver

Anyone with a porch having inner and outer doors may have a 40W or 60W light in the porch both to show who is calling and to light up the keyhole of the inner door. It may also be a good idea to leave the light on much of the time to make people think you are at home when you are not. The trouble is, electricity is expensive and lamps last a very short time when on permanently.

You can avoid these problems by connecting a 2µF, 250V capacitor in series with the lamp, which will then be dim and possibly too dim. However, a door-switch mounted so that its contacts are open when the outer door is closed, can bypass the capacitor; when the door is opened, the light goes to full brightness. A 27Ω resistor in series with the capacitor will reduce inrush current. There could be a second such switch on the inner door.

H T Wynne
Glasgow



A lobby light left on permanently will cost less and last longer with this switching circuit.

Electronic control for dc motors

Smooth and gentle start and stop, speed control and a 'kick' pulse to start sticky motors are the hallmarks of this inexpensive dc motor controller.

Designed for applications in which starting and stopping of a dc motor must be smooth over 0.5-5s this controller costs little and uses pulse-width modulation for speed control. It will take motors of between 5W and 50W although, at low acceleration rates, it is happiest when the load is the same for each start.

For speed control, the voltage on C_1 is compared with the sawtooth on C_2, produced by the *555* 300Hz oscillator IC_1, in IC_{2b}. The result is a 300Hz pulse at the output of IC_{2b} whose width is roughly proportional to the voltage on C_1.

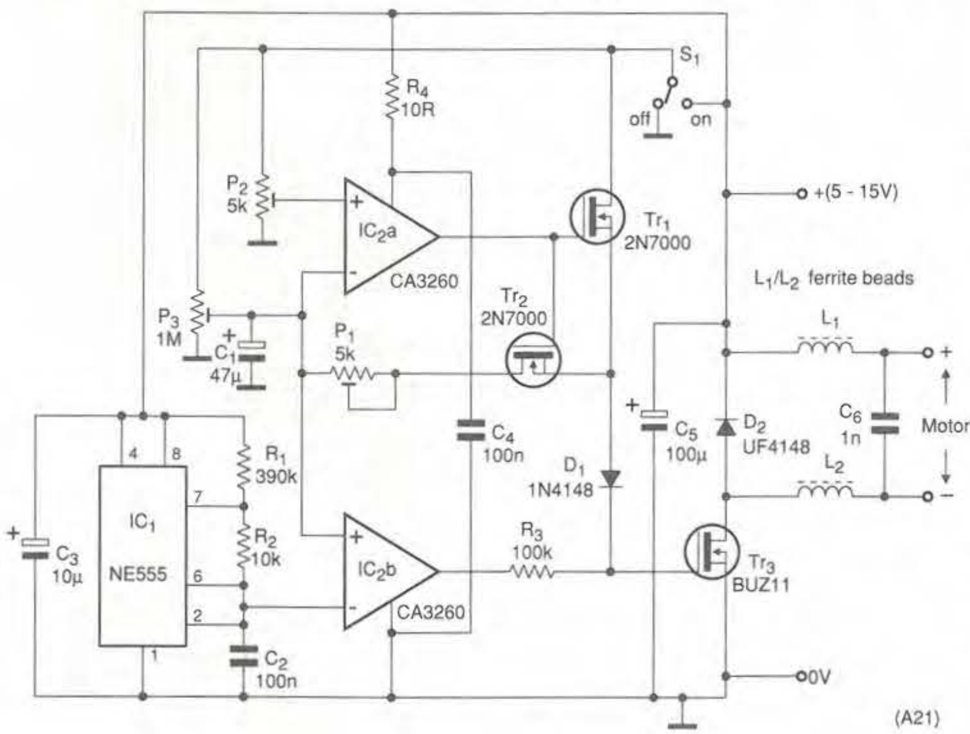
To start, S_1 is set to on; since C_1 has not begun to charge, the output of IC_{2a} is high, Tr_3 is conducting and power is applied to the motor. The voltage on C_1 rises rapidly, charging current coming via S_1, Tr_1, Tr_2 and P_1, until it reaches the voltage on pin 3, whereupon power to the motor is cut off. This pulse of power to the motor lasts between 50ms and 200ms, set by P_1 to suit the motor in use.

In acceleration, P_3 is set to produce a pulse width, after the start pulse, that runs the motor slowly. Since the output of IC_{2a} is low, C_1 charges further, but more slowly through P_3, to the voltage set by P_3. The motor accelerates to a speed set by the adjustment of P_3, output pulses now being wide enough at maximum speed to produce, effectively, the full dc level. To make adjustment of P_2 and P_3 easier, a 47µF capacitor across C_1 slows everything down to make acceleration and deceleration take about 1s.

Decelerating, when S_1 is off, C_1 discharges towards 0V by way of P_3, the output pulse width becoming shorter and the motor slowing. When the voltage on C_1 reaches the voltage on pin 3 of IC_{2a}, the output again goes high. C_1 now discharging rapidly through the same path as for rapid charging. The output pulse is now zero and the motor stops cleanly with no jitter.

As regards emc, the circuit has been tested; emi from the circuit was unmeasurable, since it was masked by noise from the motor brushes, itself within limits due to the use of the $L_{1,2}$ ferrite beads and C_6.

Rolf Schmidt
North Kessock
Inverness



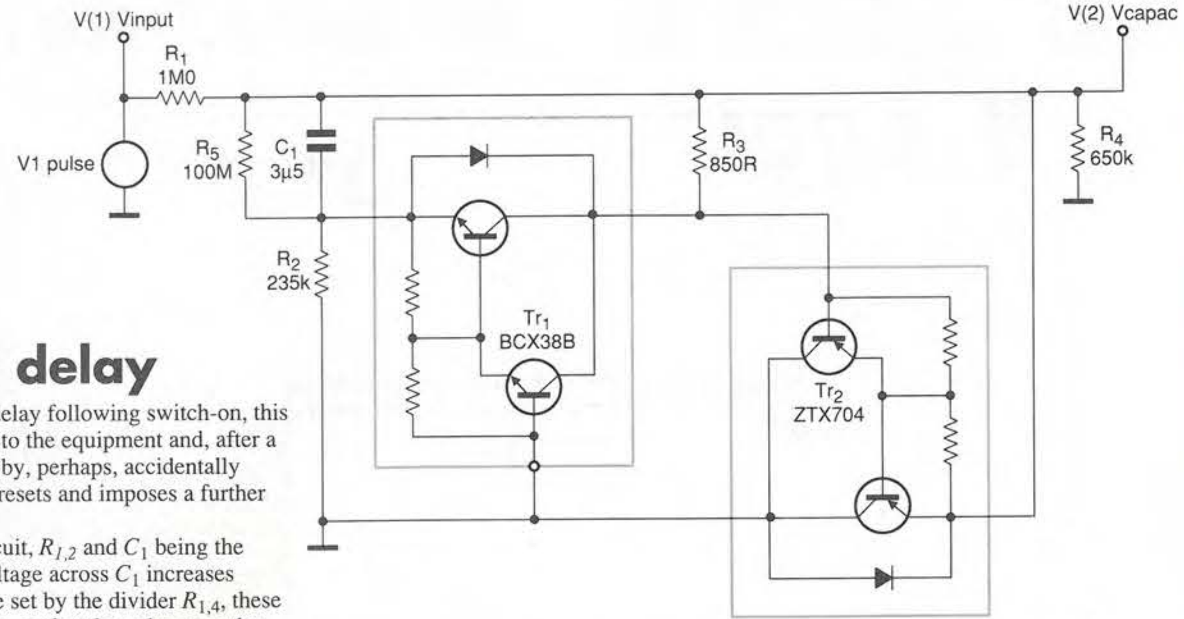


Fig. 1. Delay timer to avoid, for example, loudspeaker thump on switch-on and to reset quickly after a power glitch or accidental switch-off.

Power-on delay

After several seconds delay following switch-on, this circuit applies power to the equipment and, after a break in the input caused by, perhaps, accidentally switching off, the circuit resets and imposes a further delay.

Figure 1 shows the circuit, $R_{1,2}$ and C_1 being the relevant components. Voltage across C_1 increases exponentially to a voltage set by the divider $R_{1,4}$, these values having been chosen so that the voltage on the capacitor at the output reaches a value large enough to activate a comparator or buffer after the required time. The rest of the circuit is to avoid the necessity to make $R_{2,4}$ small for a rapid reset, which would result in a very small voltage on C_1.

The Darlington transistors and R_3 perform this function by providing an alternative discharge path for C_1. Low voltage at the input causes C_1 to start discharging through $R_{2,4}$, causing a voltage drop across R_2 and turning on Tr_1.

Current flows through the low-value R_3 and C_1 discharges rapidly, Tr_2 assisting the process by switching on in response to the voltage developed across R_3 and maintaining Tr_1 in conduction for the rest of the discharge time.

To obtain a usable delay, it is only necessary to use a comparator at the output designed to switch at 3.5V. The use of Darlington transistors and the inclusion of Tr_2 are only needed when the discharge must take place in a few tens of microseconds.

My thanks to Richard Wessen of University Technology Sydney for simulating the circuit.

Martin Gosnell
Warriewood
New South Wales
Australia

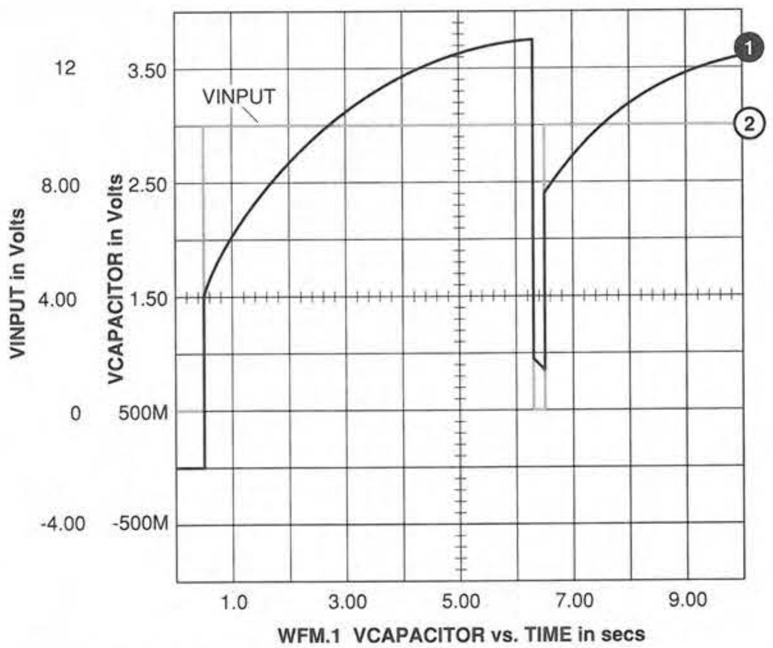


Fig. 2. Response of the circuit to initial switch-on, an interruption in supply and delay after the interruption.

Ac/dc voltage detector/isolator

When the ac or dc input to the circuit shown is over 50V and below about 250V, output is 5V; otherwise, it is zero.

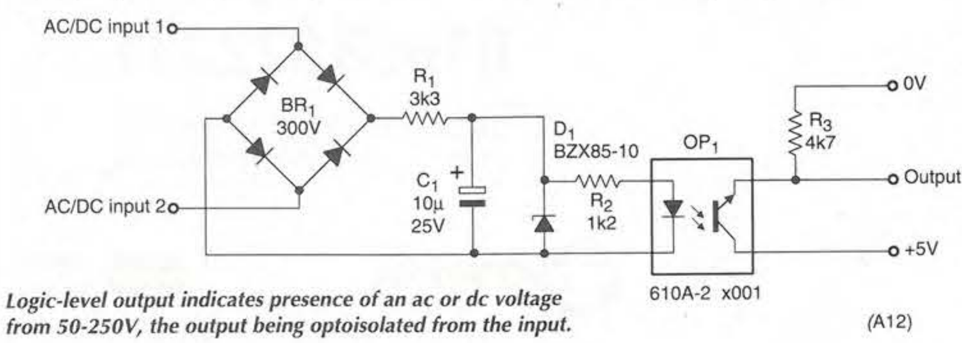
Output is optically isolated from the input.

Since the input goes to a bridge rectifier, it may be ac or dc. Bridge output is dropped across R_1, smoothed and regulated to some extent by the 10V zener. Resistor R_2 limits current to the isolator's led.

When input rises to over 50V, current flows in R_3 and the output reaches 5V. Input voltage limit is about 250V.

The optoisolator shown, a *610A-2 X001*, is a high-voltage type and is recommended.

Keith W Saxon
Deva Medical Electronics
Runcorn



Logic-level output indicates presence of an ac or dc voltage from 50-250V, the output being optoisolated from the input.

Work out the wattage needed for R_1 for maximum expected input – Ed.

Sub-woofer filter extends bass response

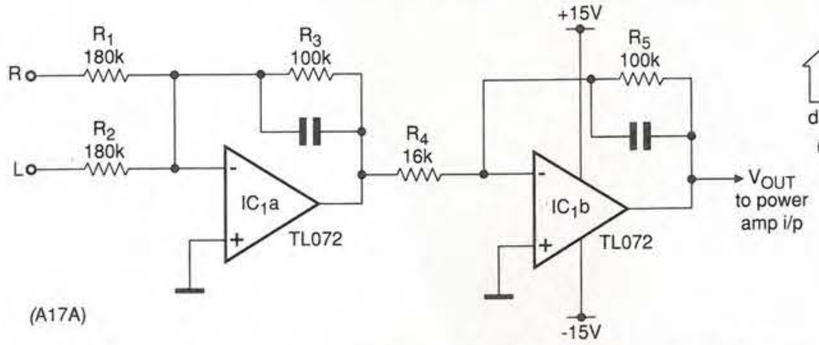
The response of a loudspeaker in a sealed enclosure is that of a second-order high-pass filter, the cut-off frequency and Q being easily determined for any enclosure size.

Such a speaker fed through a filter and power amplifier produces a flat response below its resonant frequency, forming a simple, but effective, sub-woofer.

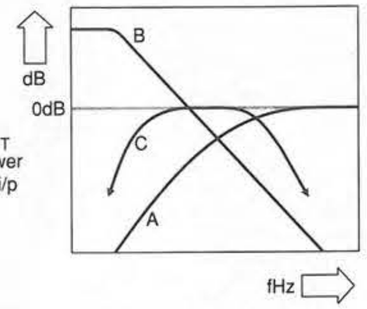
A constant 12dB/octave bass boost, produced by the circuit shown, causes the low-pass system response to become a high-pass one, Q and cut-off frequency being the same as before.

Inputs to the circuit are from left and right speaker outlets, which are added at the virtual earth of the integrating op-amp. This gives a 6dB/octave drop and is followed by a second 6db/octave circuit, the pair providing the 12dB/octave output to feed a power amplifier.

Jeff Macaulay
Chichester
Sussex



(A17A)



A = Enclosure response
B = Filter response
C = Resulting system response

A pair of 6dB/octave stages produce a 12dB/octave bass boost, turning a sealed speaker enclosure into a sub-woofer. Response is shown at C in the output curves.

Mains connection checker

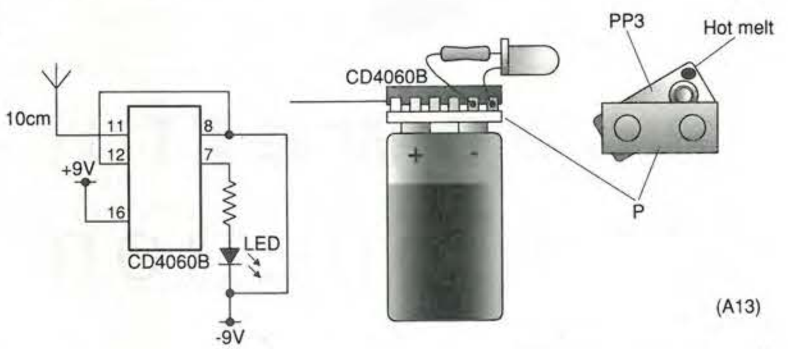
Flashing led shows when a wire is carrying mains voltage.

If you have no meter handy and need to know whether a wire has mains on it, this little device will provide the answer.

An antenna consisting a few inches of stiff wire feeds any signal received to the input of a *CD4060B* 14-bit counter/oscillator, used as a divide-by-16 counter. If the antenna is held near a wire carrying mains voltage, the led flashes at about 3Hz; if the wire is not connected to the mains, the led is either lit or unlit, but does not flash regularly.

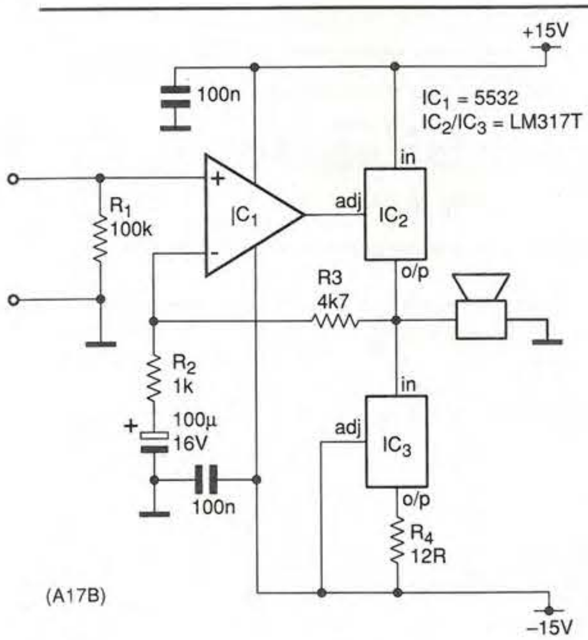
I mounted mine on the top plate removed from an old PP3 battery, one of its press connectors being mated to one of battery studs. To switch the device on, I simply rotate the circuit on its top plate until both connectors make contact. A drop of wax as shown prevents the circuit being switched on accidentally when not in use.

S Arnesen
Oslo
Norway



(A13)

Irregular use of regulators in headphone amplifier



(A17B)

For a few pounds, you can build a headphone amplifier, which gives very good sound quality and which is quite capable of deafening you.

It is a single-ended, direct-coupled Class A amplifier whose main claim to fame is its use of *317T* voltage regulators instead of fets or bjts in the output. Quiescent current for the output stage is set up by IC_3; its adjust pin is taken to negative and the V_{out} pin is 1.25V more positive, current being defined by that voltage divided by R_4.

The other *317T* is the output amplifier, which is in the feedback loop of the op-amp. Its adjust pin produces a constant 50µA sourced by the op-amp to keep it in single-ended operation whatever the signal polarity. Feedback maintains the output voltage and input voltage at the same level, compensating for the 1.25V between IC_2 adjust and output pins.

The regulator ics should be on a heat sink and electrically isolated.

Jeff Macaulay
Chichester
Sussex

Simple, but high-quality headphone amplifier – an unusual use for voltage regulators.

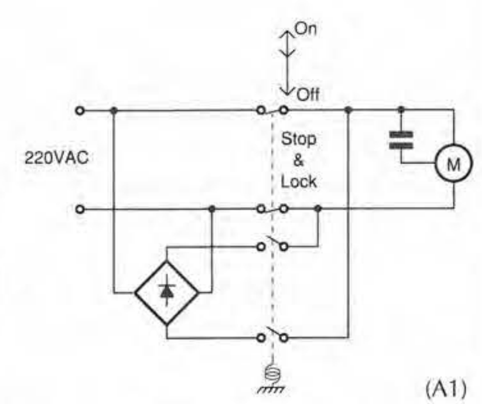
Quick stop for induction motors

Although this method of rapidly stopping and locking induction motors is known, patents covering it have not described this simple variant.

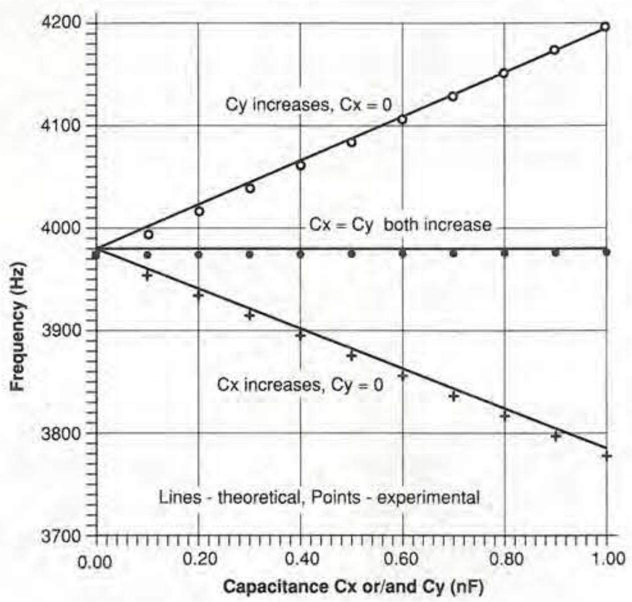
Single-phase and three-phase motors will stop very quickly and be locked if a dc source is connected to the motor. In this arrangement, the dc comes from a bridge across the ac input, which is switched in and out, as is the motor supply.

The possibility of thermal cut-out means that the switch must be released fairly quickly and, if the motor stops too precipitately, a resistor in the dc circuit will slow the action down.

Scott Arnesen
Oslo
Norway



Simpler than usual circuit to stop dead and lock an induction motor by applying dc to it.



Differentially tuned *LC* oscillator

Tuning of the bridge oscillator shown is linearly proportional to the difference in value between the two variable capacitors C_x and C_y, frequency increasing when C_y increases and decreasing when C_x increases.

If $R_2 = R_4$, frequency is determined by the equation,

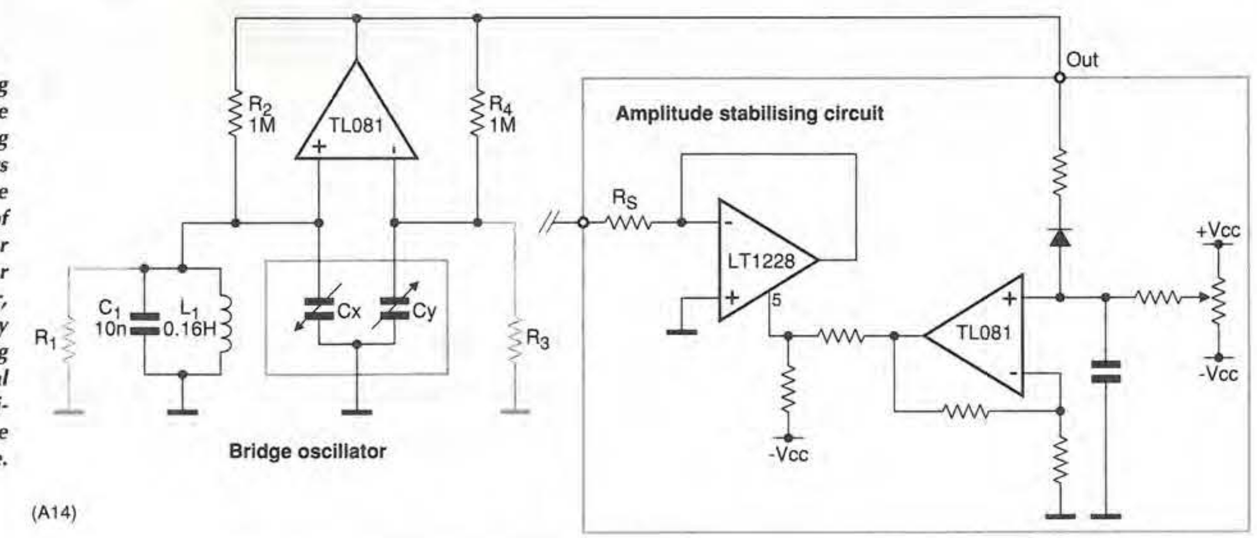
$$f = \frac{1}{2\pi\sqrt{L_1(C_1 + C_x - C_y)}}$$

when $R_3 = R_1$. Resistor R_1 represents parallel losses in the *LC* circuit and R_3 the sum of R_5 and the equivalent resistance of the *LT1228* op-amp in the amplitude stabiliser, which is controlled by rectified output voltage from the oscillator.

The graph shows a linear change in frequency for a change in either of the variable capacitors.

Lech Tomawski and Mariusz Slawiec
University of Silesia
Katowice
Poland

Varying either of the two tuning capacitors changes the frequency of this oscillator in a linear manner, frequency being proportional to capacitance difference.



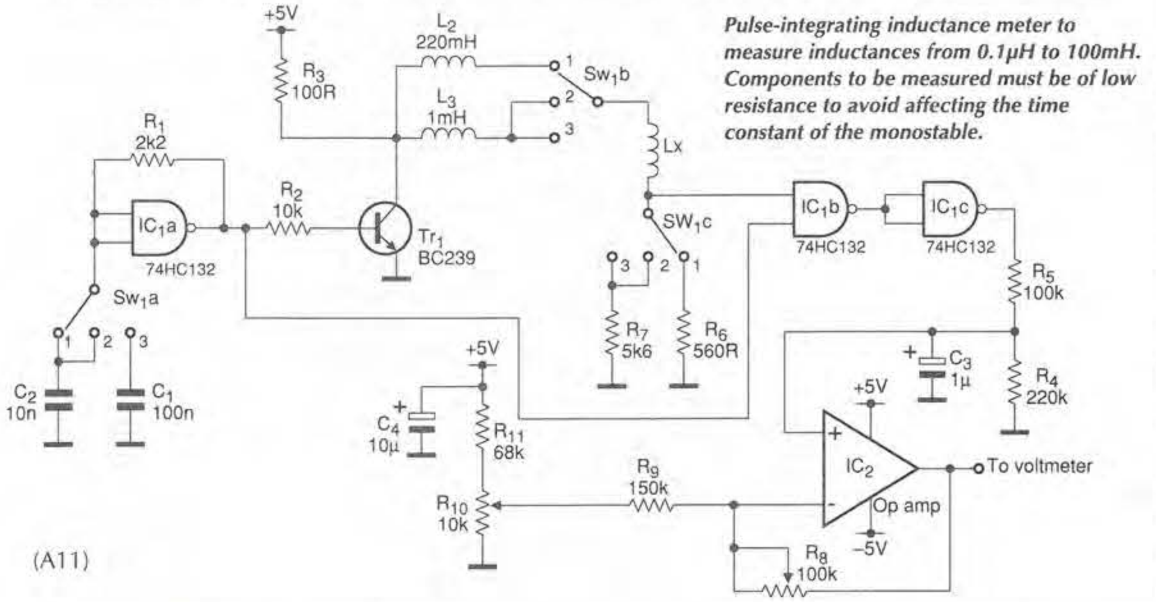
Direct-reading inductance meter

This inductance meter will measure down to 0.1μH and up to 100mH, displaying the result on a digital voltmeter. It uses much the same principle as capacitance meters in which the pulses from a monostable flip-flop are of a length proportional to the capacitance.

The monostable here is U_{1a}, which drives the transistor. When the transistor is cut off by the monostable, the unknown inductor takes current through R_6 or R_7 and R_3. On the next half-cycle, the transistor conducts and the inductor discharges its energy through the resistors to ground, developing a voltage on the input of U_{1b} for a time depending on the total inductance in the circuit and the resistance of R_6, that of the inductor and stray resistances.

Pulses at U_{1c} output are integrated and applied to the output op-amp, which scales the voltage level and sets zero for the digital voltmeter. I use a 4.5-digit meter on the 2V range, the three inductance ranges being 1mH, 10mH and 100mH.

Flavio Fontanelli
Genova
Italy



Pulse-integrating inductance meter to measure inductances from 0.1μH to 100mH. Components to be measured must be of low resistance to avoid affecting the time constant of the monostable.

1-180s timing in two steps

Analogue timer *MC14536* provides intervals that vary by a factor of two for each position of a hexadecimal switch

Intermediate steps are not possible without the addition of a separate divider chain. This circuit interposes steps of two to give, with sixteen positions of the switch – a progression 1, 1.4, 2. 2.8... 180.

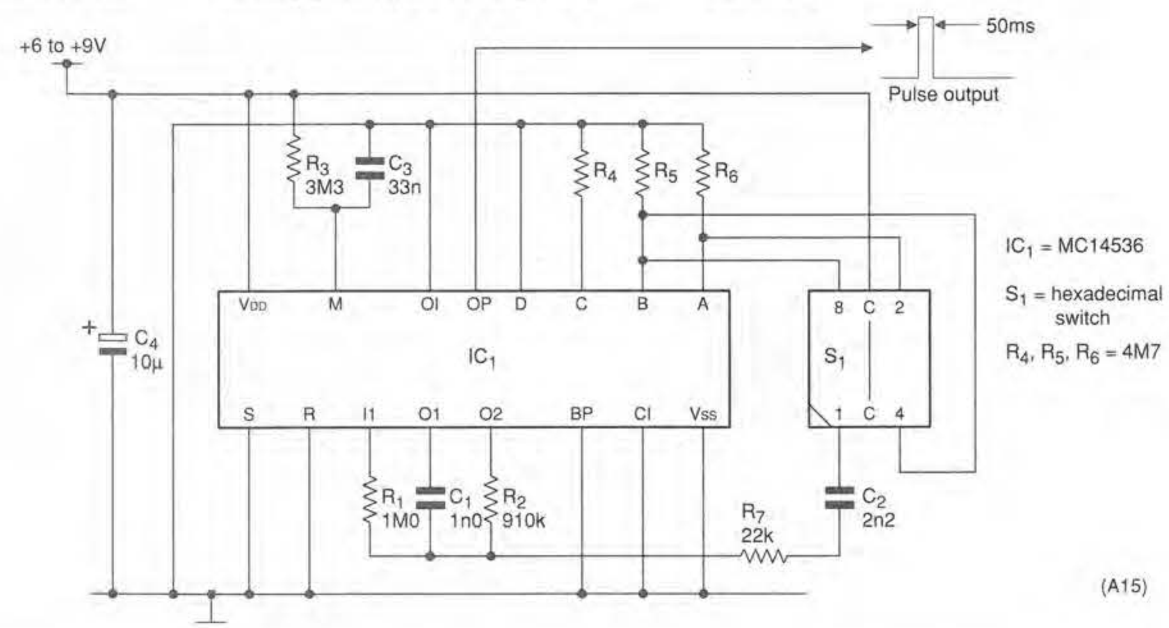
To obtain the intermediate steps, capacitor C_2 is connected by the least significant bit pin of the hexadecimal switch, the capacitor having a value 2.2 times that of the main timing component C_1. Other switch outputs go to the timer programming inputs A, B and C and the switch output *c* to 6V in response to the binary code corresponding to the sixteen switch positions. On even positions of the switch, C_2 is floating, while for odd positions it is connected to 6V and the clock based on $R_{1,2}C_1$ slows down.

The extra capacitor must be larger than the original timing capacitor since the waveform at the bottom of the clock components is attenuated, besides being slowed by C_2; R_7 allows the use of a preferred value for the extra capacitor. Minimum frequency of the timer must be 512Hz to take account of the internal divider between the clock and the monostable output circuit; this falls to 362Hz when C_2 is connected.

Stability is about 1% over a supply range of 6-9V. Pulse output from pin 13 will drive an earpiece or led, but ensure that pickup on C_3 does not cause spurious oscillation.

C J D Catto
Cambridge

Additional timing capacitor connected by the hexadecimal switch adds two steps between the original X2 steps in this MC14536 timer, which produces periods of 1-180s in the 1, 1.4, 2, 2.8... series.



LETTERS

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

Feedback

Having read Mr Ellis' letter in the September issue with attention, I am of course immeasurably pleased at being invited to 're-optimise' my amplifier design.

I agree entirely that the Miller capacitor looks as though it would do a lot more good when wrapped around the output stage as well, because this promises that the crossover distortion could be greatly reduced. Unfortunately, no amount of wishful thinking will alter the fact that this connection is prone to hf instability because of the extra phase-shifts in the output stage. Mr Ellis admits this when he says "I did observe some oscillation..." I have no doubt he did. I would take this capacitor speculation more seriously if someone published some measurements and proved it really did work.

On a model amplifier – small-signal output devices, hence much quicker – I found the effect on crossover spikes disappointingly small.

I fear Mr Ellis has not completely understood the subtle nature of the generic amplifier circuit; the second stage does not and cannot load the input stage. I profoundly disagree with his labelling the second stage as the 'driver' when everyone else calls the devices directly before the output transistors the drivers.

It seems to me there is quite enough confusion about, without getting The Naming of Parts wrong as well. The second stage provides all the voltage gain, so calling it the voltage-amplifier stage seems to me pretty uncontroversial.

The voltage-amplifier stage does not load the input stage, because the input pair is a transconductance amplifier – i.e. voltage-in, current-out. The voltage-amplifier stage on the other hand is a transimpedance stage – current-in, voltage-out – which would ideally have zero input impedance. This is the whole basis of the pole-splitting action that makes this beloved circuit so dependable.

If Mr Ellis had studied my Blameless design a little more closely, he would have seen that the input-pair local feedback measures he recommends with such enthusiasm have already been implemented. Indeed I thought I was one of the main advocates of this way of minimising the Miller capacitance required, though in the way I implemented it I called it 'constant-g_m degeneration'.

If the input emitter resistors are increased significantly beyond 100Ω the noise performance is degraded; life is fill of compromises. In selecting for praise the Lohstroh & Otala amplifier he has chosen a bad example. I know this circuit very well indeed, for, unlikely as it may seem, in 1975 I was entrusted with the task of putting a version of this design into quantity production.

While I no doubt made my share, and more, of mistakes during the development period, the eventual conclusion of us all was that the design was over-complex, inefficent, non-linear, and of very doubtful high-frequency stability. Ironically, in this design, loading of one stage by the next really is an issue.

There are other points I could make, but I fear to little purpose. One question remains; Mr Ellis must realise that his understanding of the subject is imperfect, as it is for everyone in varying degrees. Why is he so confident he is justified in lecturing the rest of us?

Douglas Self
London

Designer workaround

Concerning my article 'Designer power supplies' in the August issue, it has been brought to my attention that the program crashes when the 'figure-of-merit' becomes very high. I am grateful to Ralph Goold of St. Albans for supplying the following work-round.

In the initialisation section, the line $V_2 = V_{peak}$ should be changed to $V_2 = V_{peak} - V_{rec}$.

An alternative, though less elegant, work-round is to force the arcos function to return zero if the parameter value exceeds 1.0.

I hope this has not caused problems for anyone.

W Gray
Farnborough
Hants

Thomson's electron

Your timely article by Tom Ivall in the August 1997 edition of *Electronics World* on J J Thompson's 1897 discovery of the electron and its technical context was interesting and informative.

I am particularly interested in J J Thompson himself, his origins, family and further works (if any) he accomplished. Has any reader any suggestions of previous articles or other material to point the researcher in the right direction?

S Katzen
Magee College Belfast

Singing preamp

A long time, I built a version of John Linsley Hood's simple Class A amplifier, (*WW*, April '69). The output transistors were *2N3055*s mounted directly, i.e. without insulation, on two rather inadequate 'Christmas-tree' heatsinks.

For some reason, while I was driving the amplifier hard with a music signal, with the speakers disconnected – like you do – I found I could still hear a very faint, tinny and distorted version of the music, with a large peak in its spectrum at the frequency at which the heatsinks would 'ting' if flicked with a finger.

The sound loudest near the top surface of the output transistor cans and was not affected by moving the wiring around, nor by holding magnets near the transistors. The only factor that affected it was the load; connecting a 6.8Ω resistor across the output terminals made the sound louder.

I wonder if this could be a thermal effect. Does anyone know?

Chris Bulman
Bedford

Silverskin

I was glad to see Nick Wheeler's sensible and timely article on skin effect in relation to loudspeaker cables, but it may be worth mentioning a small additional point.

In rf applications, where skin effect causes currents to be almost entirely near the surface of a conductor, I believe that the use of silver plating is almost never justified simply because silver is a slightly better conductor than copper. The real advantage is that oxidised silver, though unwelcome in appearance, is a far better conductor than oxidised copper.

Bob Pearson

Remote control enhancement

I have an enhancement to Alex Birkett's circuit 'Two-wire remote control' in *Electronics World*'s June issue.

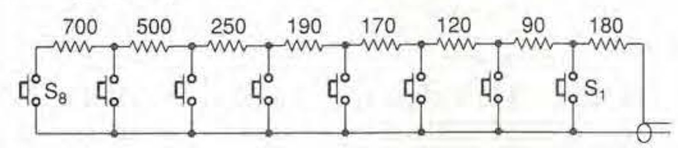
In the original circuit, if two buttons are pressed together, the output corresponds to an unedfined voltage band and is not representative of either button. This voltage depends on the haphazard parallel connection of two resistors.

If the parallel ladder of buttons and resistors is replaced by a series resistor chain with buttons to short the 'cold' end of each resistor to ground, as in the diagram, accidentally pressing two buttons together results in an output corresponding to one of the buttons pressed – specifically the one nearer the 'hot' end of the resister chain. This is arguably less confusing to the user.

An additional advantage is that tolerance requirements for the resistors may be relaxed as the effect of variations will tend to cancel out. The values given are for use with Mr Birkett's circuit.

I have come across a similar idea used to reduce the number of connections from the main board to the front panel in a multi-standard monitor. The monitor's designers had used an all-singing all-dancing chip allowing software control of every conceivable parameter. In this case the selected resistor was used to charge a capacitor whose charging time was measured by the control processor,

Chris Bulman
Bedford



Enhancement to the two-wire remote control circuit to make accidentally pressing two keys give more predictable results.

NEW PRODUCTS CLASSIFIED

Please quote *"Electronics World"* when seeking further information

ACTIVE

Linear integrated circuits

Regulator ics. Allegro offers the *A8188S Series* of low drop-out voltage regulators for use in battery-powered equipment. Output is 2.5-3.3V at 250mA peak, the pmos series element allowing a drop-out voltage of 90mV at 60mA. Quiescent current is 55μA and does not increase much near drop-out; sleep current is under 1μA. A bandgap reference provides stabilisation to within 10mV and regulation to between 25mV and 30mV. Both single and dual output types are available. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Stereo amplifiers. National Semiconductor's *LM4752* and *LM4755* are dual audio amplifier ics delivering 11W into 4Ω and 7W into 8Ω from 24V and are intended for use in miniature systems and compact television receivers. Gain is internally set at 34dB and thd is 10%. The *4755* has a built-in mute, while the *4752* needs external components for muting. National Semiconductor GmbH. Tel., 0049 1805 32 7832; fax, 0049 814103515.

Logic

Programmable clock chips. Cypress announces a new version of its *CY7B991*, the *CY7B9911 RoboClock*, which retains the programmable skew facility, but operates at 100MHz. Skew control is selectable between outputs for 700ps to 1.5ns to reduce the effects of crosstalk and emi when high-speed clock signals are distributed around multiprocessing systems. The devices provide zero propagation delay and 50% duty cycles, as required by PCI, are guaranteed. Cypress Semiconductor UK Ltd. Tel., 01707 338888; fax, 01707 338811.

Memory chips

3.3V, 64Mb dram modules. Smart Modular Technologies has announced new 3.3V, 64Mb, 72-pin dram-based modules operating on 5V designed for use in servers and workstations needing 1Gbyte or more. Two problems are thereby solved: voltage translation and a step-down converter built in allow these 3.3V simm devices to work in existing 5V systems; and four times as much memory, compared with existing designs using 16Mb drams, become available without the need to redesign the motherboard. Smart Modular Technologies. Tel., 01908 234030; fax, 01908 234191.

Microprocessors and controllers

New PICs. From Microchip, the new *PIC16C77* range of 5mips eeprom-based, low-power microcontrollers provided with 8K by 14word of program memory, 376byte of data ram and an 8-channel, 8-bit a-to-d converter with sample-and-hold, ±1lsb accuracy and a 16μs acquisition time. Other facilities include a real-time clock, one 8-bit and two 16-bit counter/timers. There are two channels of 160ns resolution for input, compare and pwm. Communications facilities, with 33 pins for i/o, include an 8-bit asynchronous slave port, a synchronous serial port for SPI and I²C, a 6.25Mb/s usart with baud-rate generator, two 8-bit pwm outputs and direct led drive and triac interface. Arrow-Jermyn. Tel., 01234 270027; fax, 01234 214674/791501.

Motors and drivers

Small motor clutches. Densitron's *Wrap-Spring* clutches provide torque transmission up to 8.5Nm in small packages, their low inertia allowing greater acceleration than that obtained using rare-earth servomotors at lower cost. Unidirectional or bidirectional drives in normally-engaged and normally disengaged versions are available. In multi-axis systems, the clutches allow several axes to be started or stopped quickly at a speed synchronised to that of a single drive motor and the ability to engage and disengage under electronic control allows drive to be taken from other sources such as handwheels or petrol engines. Densitron Europe Ltd., 01959 700100; fax, 01959 700300.

Microwave components

Low-noise microwave amplifiers. Narrow-band amplifiers from DBS Microwave operate at frequencies in the 2-22.5GHz range, covering the 16 commonly used bands. Low noise is the feature: a typical example being the *DBS-0910N310*, which has a maximum noise figure of 1dB for a stage gain of 30dB. Gain flatness in its frequency range of 9-10GHz is 0.5dB and output power 10mW, or 100mW for the *N420* version. Variants giving 5W output are available, albeit with slightly higher noise figures. All types contain a voltage regulator to give 12V or 15V and will accept a 20dBm cw input. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Travelling-wave-tube amplifiers. TMD offers the *PTC Series* of travelling-wave-tube amplifiers for microwave instrumentation are smaller and lighter than earlier types. All use TMD's switched-mode psu technique and improved power converter design has allowed lower costs. Three 19-in rack-mount models cover the 1-18GHz frequency range at up to 200W output power – 1kW pulsed – and models working at up to 40GHz at 80W cw can be supplied. Thorn Microwave Devices Ltd. Tel., 0181-573 5555; fax, 0181-569 1839.

Optical devices

White leds. White light-emitting diodes from Sloan AG emit white light with no mixing of primary colours, giving a typical brightness of 400mcd at 20mA and 3.6V. Colour temperature is 8000K and viewing angle 60°. They are rated at 100mW. Roxburgh Electronics Ltd. Tel., 01724 281770; fax, 01724 281650.

Orange leds. Dialight can now offer a range of 3mm led indicators emitting "distinctly orange" light, which seems a fairly defensive way of putting it. These board indicators come in a number of forms: single-led types in a variety of heights, bi-level versions and arrays, tri-level and quad-level led arrays. Wavelength is 600nm, light output 6.5mcd and forward voltage 2.2V at 10mA. Viewing angle is ±30°. Dialight. Tel., 01223 424313; fax, 01223 423493.

Discrete active devices

Efficient GaAs fets. *MGF0900 Series* n-channel, Schottky gate GaAs fets from Mitsubishi are intended for uhf high-power use. Three new types, *0909A, 0910A* and *0911A* are 2.3GHz devices, the *0909A* providing 45% high-power added efficiency when compared to other similar devices with a power gain of 11dB at 20dBm power input at 2.3GHz. Power output is typically 38dBm. The *0910A* is similar but for Class A operation, while the *0911A*, also for Class A, has added efficiency of 40%; at 2.3GHz, output power is 41dBm and power gain 11dB. Mitsubishi Electric UK Ltd. Tel., 01707 276100; fax, 01707 278692.

PASSIVE

Cameras

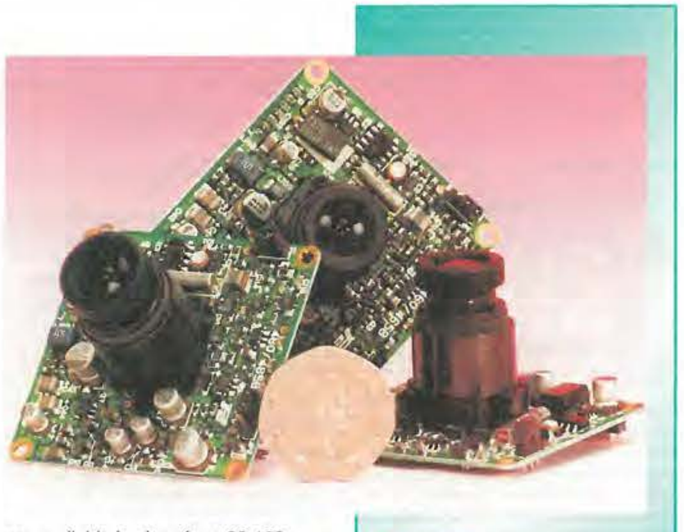
Small ccd camera. Sony offers the *Micro Unit CCD* family of ccd cameras, which measure 18.3 by 7.1mm, complete with timing generator, sample and hold ic, digital signal processing and lens, together with the ccd. Output is conventional Y/C analogue and there is auto white balance and autoexposure. Lenses are 2.9mm or 4mm fixed focus and resolution is 220 tv lines in the centre of the picture. Colour, exposure, white balance and minimum gain are all programmable by external eeprom. Sony Semiconductor Europe. Tel., 01256 478771; fax, 01256 818194.

Passive components

Toroidal transformers. Drake Transformers can offer 1kVA toroidal transformers with dual secondaries of 40, 45, 50, 55 and 120V and dual primaries for supplies of 120 or 240V, 50/60Hz. Outer insulation can be specified as polyester tape wrap or woven tape with a varnish finish. The transformers are mounted using a dished washer with a central bolt or made with the centre filled, with a plain hole or threaded insert. Temperature range is –10°C to 40°C, with a maximum of 65°C temperature rise on full load. Drake will make transformers to customers' own specifications. Drake Transformers Ltd. Tel., 01268 560040; fax, 01268 560261.

Low-esr electrolytics. *Evox Rifa PEH 3XX* snap-in electrolytic capacitors provide high ripple current capability and equivalent series resistance down to 21mΩ at 20°C and 100Hz. The range of values is 150μF and 12000μF at ratings from 40V to 450V. Diameters are 30mm or 35mm and maximum case length 50mm. Campbell Collins Ltd. Tel., 01438 369466; fax, 01438 316465.

Backlighting transformers. Toko's *BLC Series* of miniature inverter transformers are designed for use in cold-cathode fluorescent lamp power supplies for display backlighting in lcd television, camcorders, laptops, etc. Output voltage is typically over 1kV and the range of power available is 0.7W to 6W, with a choice of input and output voltage. Sizes range from 12 by 11 by 5mm for the 0.7W type to 21 by 23 by 6.5mm for a 6W version. Dielectric strength of the independent primary and secondary is over 2.5kV. Cirkit Distribution Ltd. Tel., 01992 444111; fax, 01992 464457.



Mixed-signal ICs

Pwm drivers. M S Kennedy *MSK4220/1* board-mounted motor driver/amplifier modules contain all the drive and control circuitry in the one hybrid chip. Each has a complete H bridge and switching circuitry and the modules are suited to the drive of brush motor speed control and Class D amplification. In the case of the *4220*, one analogue signal is needed to control motor speed and direction or an audio signal for amplification. Both devices are on a ceramic metal-coated substrate for direct mounting on heat sink. Ashwell Electronics Ltd. Tel., 01438 364194; fax, 01438 313461.

Connectors and cabling

Ultra-SCSI connector. Honda has a slim, high-density cable-to-board connector to meet the Ultra-SCSI specification, having two rows of phosphor bronze contacts at 0.8mm spacing. The contacts are underplated with nickel and then plated with palladium and gold. IDC termination is present and the connectors use jackscrews for retention and to ensure shielding. Honda Connectors. Tel., 01793 523388; fax, 01793 521777.

S-m signal/power board connector. *Molex Microfit* surface-mounted board connectors are believed to the first true s-m types to handle both signal and power lines, the 3mm-pitch pins withstanding 1.5kV ac and carrying 250V ac at 5A per contact. A solderable hold-down tab avoids the need to drill holes in the board. Connectors are available with 2-24 contacts in even numbers and there is polarisation and positive latching. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

High-density board connectors. Robinson Nugent's *PAK-5* low-profile, high-density, SMT board-to-board connectors have, as an option, a floating contact in the receptacle to cope with potential torsional effects and up to 0.3mm of discrepancies caused by tolerance build-up. They are available in sizes from 20-100 positions and mated stacking height of 5mm. Contact pitch is 0.5mm. Also available are bellows contacts to allow stacking heights of 3-8mm to resist mating stress and contact resistance of 80mΩ maximum. Robinson Nugent (Europe) Ltd. Tel., 01256 842626; fax, 01256 842673.

Crystals

Ceramic crystals. Total Frequency Control has the *Type CX2* miniature surface-mounted crystal, on 16mm tape and reel for high-volume use. Frequency range is 10MHz to 100MHz, the latter on a fundamental of 40MHz, at a stability of ±10ppm for –10°C to 60°C, load capacitance being 30pF. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie @ tfc.co.uk.

Displays

Bitmap lcd driver. The *NJU6583* driver by New Japan Radio drives 33 by 96 dot graphics displays, containing 3696 bits of display data ram, microprocessor interface circuitry, an instruction decoder and 32 common and one icon common by 96-segment lcd drivers. The display is transferred to display data ram by serial or 8-bit parallel interface. Standby current is 0.05μA and operating voltage 2.4-6V. Young-ECC Electronics. Tel., 01628 810727; fax, 01628 810807.

Filters

900MHz saw filters. Designed for the GSM and AMPS mobile communications market, Fujitsu's *F5CE-D2* high-stopband saw filter measure 3mm square in a surface-mounted LCC package. The filters use a double-mode surface acoustic wave design to provide a stopband attenuation of about 60dB with in-band ripple of under 1dB. Temperature drift is 33ppm/°C. Fujitsu Microelectronics Ltd. Tel., 01628 76100; fax, 01628 781484.

Cameras

B/w ccd camera. *PEL465* by Premier gives 291000-pixel pictures from a $^1/_3$-in interline ccd sensor. Either a 3.6mm lens or a pinhole 3.7mm f4.5 is provided as standard and there is a range of other lenses to choose from. The camera works in light levels down to 0.5lux, the electronic iris working from $^1/_{50}$s to $^1/_{100000}$s, running on internal or external synchronisation. Video output is 1Vpk-pk at 75Ω and a 12V, 130mW supply is needed. Premier Electronics Ltd. Tel., 01922 634652; fax, 01922 634616.

Chip filters. Mitsubishi's *LT Series* of surface-mounted chip filters are made using a new technique to be produce a steep insertion-loss curve and an insertion loss of –30dB to over 1GHz. They do not need to mounted with any specific orientation and are made in a composite dielectric and ferrite material to allow a number of filter types such as T-section and pi-section versions to be produced. First available are T filters cutting off at 10, 22, 47 or 100MHz, rated at 25V dc, 100mA. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.

Hardware

Rackmount keyboard drawer. Taking only 1U of rack space, the *TR-KBD* keyboard drawer made by Tri-Map is claimed to be simple to fit, since mountings are all at the front. The drawer has ball-bearing slide rails and there is a space for a wrist rest, one version of which has a tracker ball compatible with PS/2 or serial mouse ports. Standard Cherry keyboards fit the drawer, which means that other languages can be used. Standard colour is black to match the rest of Tri-Map's range, but other colours and, indeed, all other characteristics of the drawer, may be otherwise specified. Tri-Map International Ltd. Tel., 01705 424800; fax, 01705 424801.

Miniature power tool. A new cordless power tool from Minicraft, the *MB1037*, is supplied with 25 accessories for cutting, polishing, grinding and drilling in wood, plastics, ceramics, glass and light metals. Operating time is 35 minutes continuously and the charger is included. No-load speed is 9500rev/min, the motor is fan-cooled and there is a keyless chuck. Minicraft. Tel., 01388 420535; fax, 01388 817182.

Flatpack cabinets. Cannon has a new range of 19-in racks and enclosures that come in a flat package and that may be assembled using only 24 bolts. The *2100 Series* includes a wide range of products for data communications and computer hardware. Sizes are 600, 700 and 800mm wide by 600 or 800 deep in heights from 27U to 45U. All panels are included and there is the choice of a steel or glass door. Cannon Technologies Ltd. Tel., 01425 638148; fax, 01425 619276.

Radio communications products

Data communications. Free from Licence requirements, the *TX2/RX2* transmitter and receiver board-mounted modules from RF Solutions provide data comms at 40kb/s over a distance of 300m on open ground or 75m in buildings. Full screening has minimised radiation and the need for external filtering is thereby reduced. Both are in sil-mounting modules, operating on the European 433.92MHz or on the UK 418MHz. The transmitter is a saw-controlled fm type, while the receiver is based on a double-conversion fm superhet, taking 14mA when active. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.

Test and measurement

250Msample/s dso. LeCroy's *9344* digital storage oscilloscope provides 500MHz of analogue bandwidth at 1Gsample/s and 250K points of memory. Four 250Msample/s, 50Kpoint channels may be combined to form two 500Msample/s, 100Kpoint or the one 1Gsample/s configuration. When running at under 200Msample/s, there is 5ns peak detect for fast transients and 'smart' trigger functions to capture infrequent or complex conditions characterised by pattern, interval, dropout, etc. The instrument will carry out over 40 parametric measurements, showing average, low and high values and standard deviation and a pass/fail function tests up to five parameters against selected thresholds, waveform limit test is made against masks defined in the instrument. There is a 9in raster display and a full range of peripheral facilities. LeCroy Ltd. Tel., 01189 344882; fax, 01189 348900.

Oscilloscope calibration. Oscilloscope calibration workstations based on Wavetek's *Model 9500 Calibrator* are introduced in versions with bandwidths of 400MHz, 600MHz and 1.1GHz, upgrading from low to high being provided. All have a new version of the company's pc calibration software, which includes an up-datable library of procedures for commonly available analogue and digital oscilloscopes. All are provided with an Active Head, which applies signals directly to an instrument's BNC input to avoid the distortion and other errors caused by long cables. Active heads will not turn into inactive heads on being dropped, since they have already been dropped and survived the test. The software renders calibration automatic, so that even the most wondrous instruments can be done in ten minutes, reports and certificates being generated. Wavetek Ltd. Tel., 01603 404824; fax, 01603 483670.

Emi test system. Two instruments which have commonly been used together in emission testing are now sold as a package at greatly reduced cost. *ES-Plus* consists of the Rohde & Schwarz *ESPC* 150kHz-1GHz (expandable to 9kHz-2.5GHz) test receiver and the R4131C 10kHz-3.5GHz spectrum analyser by Advantest, the two being accompanied by the R&S ESPC-K1 Windows software. Emc accessories by R&S can be used with both instruments. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Clamp-on power analyser. Hioki's *3166 power Hitester* provides clamp-on measurement and analysis of single or three-phase power lines. An lcd gives instantaneous values of current and voltage, averages and frequency, max/min display and first and second integration displays, which show reactive/apparent power consumption. Options are a floppy-disk drive and harmonic analysis. An RS232 interface is standard. Telonic Instruments Ltd. Tel., 01734 786911; fax, 01734 792338.

Pulse generator. *TGP110* from TTi is a general-purpose pulse generator working at frequencies between 0.1Hz and 10MHz, providing independent control of pulse width, repetition rate and variable pulse delay. Pulses of a fixed width may be generated at any frequency down to a duty cycle of 1 in 10^8, pulse width range being 50ns-5s and period range from 100ns to 10s. Delay is independently adjustable over the same range as the width. As well as continuous pulses, single or multiple pulses may be generated by trigger or gating signals and there is an inverting switch. Square-wave operation is also provided at a frequency set by the period control to provide variable-period edges. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Protection devices

Surface-mounted fuses. Gold-plated contacts on Schurter's s-m *SupraFuses* eliminate corrosion even after storage, so enhancing soldering and general reliability; the fuses also comply with IEC 127-4/2 and UL 248-14 standards. Two types are available: the *Classic*, which is quick acting and the *Chipguard* superfast type to protect fast processors, both use laser-trimmed foil as the fusing element. Dimensions are minute: the Classic measures 4.5 by 2.6 by 1.9mm and handles 0.8A to 4A at up to 125V. *Chipguard* is only slightly larger and carries 1A to 5A at 63V. They are directly solderable or may be fitted with a clip to make them replaceable without soldering. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Literature

Numerical analysis software. National Instruments' *HiQ* interactive problem-solving software is the subject of a new brochure, in which are described functions such as data fitting, numerical integration, data visualisation and report generation. The package will produce 2D and 3D graphs from data and functions and accepts data from Excel and LabView. HiQ is for Windows NT/'95 and Mac. You can obtain an evaluation copy on www.natinst.com/hiq. National Instruments UK. Tel., 01635 572400; fax, 01635 524395.

Power conversion. *Switch on to smart power conversion* is a Celab publication describing the company's power supplies for military and communications use that provide battery backup and charging, management and diagnostics and various protective measures (for 'smart', read 'intelligent'). Celab Ltd. Tel., 01420 477011; fax, 01420 472034.

Emi and thermal interfaces. Product selection software from Chomerics gives guidance on the choice of emi shielding and grounding and thermal interface materials. *inPHorm* uses

Please quote *"Electronics World"* when seeking further information

multiple-choice questions to allow the user to make an application profile, the program then recommending some products, obtained from all Chomerics books, bulletins and brochures, to fill the bill. It faxes or prints data on all the products and the selections are savable and can be added to forms for quotations. *InPHorm* comes on cd or on twelve 3.5in floppies, being compatible with all embodiments of Windows. Parker Hannifin plc, Chomerics Division. Tel., 01628 486030; fax, 01628 476303.

Sensors. Lucas Control Systems can supply a new brochure on its *Schaevitz* sensing products, which include devices to sense and measure pressure, displacement, inertial sensors for slope and tilt and instrumentation. Newly introduced are non-contacting laser sensors and new pressure sensors for difficult environments. Lucas Control Systems Products. Tel., 01753 537622; fax, 01753 823563.

Dsp hardware and software. *Innovative Integration*'s catalogue of products for digital signal processing and development tools is now available. For 1997, the catalogue describes a wide range of products, including the PC32 low-cost, 32-bit supercontroller with a-to-d and d-to-a conversion, the *ADC64* 64-channel, 32-bit, PCIbus data converter board, ViSim/DSP software/hardware for the design of control systems and DASY*Lab* data acquisition software. Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213.

Industrial products. Farnell has a new catalogue of equipment for industrial electronics – separate from the electronic components catalogue. *Monitran* would like to point out that, among the 32000 products, its accelerometers and linear variable displacement transducers with instrumentation are on offer. Farnell Components Ltd. Tel., 0113 2205700; fax, 0113 2891163.

Signal conditioning. IOtech's publication, *The Signal Conditioning and PC-based Data Acquisition Handbook*, is both a primer on the subject and a quick reference guide for experienced engineers. There is an introduction to the use of transducers with pcs by way of data acquisition systems which describes the common interfaces. Discussions on particular requirements for the measurement of strain, temperature and acceleration are included, as are sections on a-to-d conversion, multiplexing, amplification, noise reduction and digital signal conditioning. Scensys Scientific &Engineering Systems Ltd. Tel., 01296 397676; fax, 01296 397878.

Materials

Safety matting. Plastic Extruders offers *Vynastat*, which is a vinyl matting designed to protect electronic equipment and employees from the effects of electrostatic discharge. It is made from semi-conductive pvc, is slip-resistant in the form of an open grid and possesses permanent static dispersal properties. Vynastat comes in 10m rolls, 60cm or 91cm wide. Plastic Extruders Ltd. Tel., 01883 623329; fax, 01883 625506.

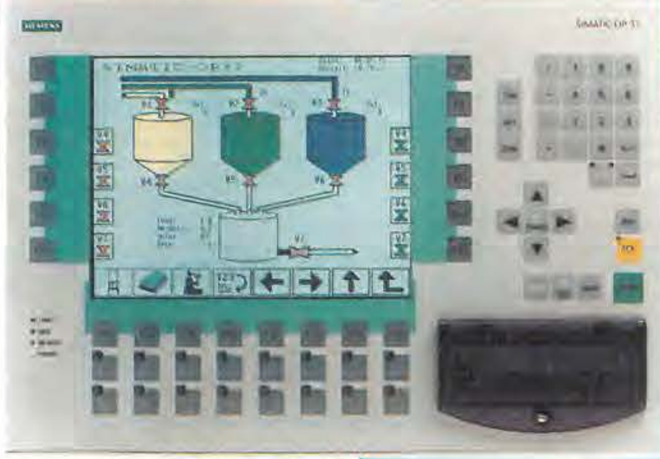
Power supplies

Lithium-ion battery protector. New from Unitrode is the *UCC3911* two-cell lithium-ion battery-pack protector, which has an on-chip fet switch for increased reliability; it copes with short circuits, overcharge and overdischarge. A band-gap voltage reference detects overcharge or overdischarge, opening the fet switch, which is in a feedback loop that controls the fet and allows for pack recovery. In the overcharged state, the loop only allows discharge current to pass the fet and in overcharge, only charging current passes. In the overdischarge state, the chip sleeps until charging current is detected. Battery power drain is 20μA and 3.5μA when shut down. Unitrode (UK) Ltd. Tel., 0181-318 1431; fax, 0181-318 2549.

Lower-cost psus. XP's *NAN55 Series* of universal-input power supplies are supplies in single, dual and triple versions, delivering 55W with forced air cooling, 45W without, from a standard 5 by 3 by 1.2in module. They fully comply with all kinds of safety standard and are protected against overvoltage, shorts and conducted noise to EN55022 level B. Outputs available are 5V, +5 –12V or 5V and ±12V, both available in peak power and regulated versions. XP plc. Tel., 01734 845515; fax, 01734 843423.

1kW supplies. Coutant-Lambda's *RP Series* of power supplies are multiple-output, ac-fed types for largish computer installations such as workstations, file servers and industrial controllers. They are available in ratings of 500W, 750W and 1000W, are CE-marked and may have three, four or five outputs, the main, high-power output being adjustable from 3V to 6V. Other outputs may be 12V, –12V, 3.3V, 5.2V, 24V or 48V. Input is universal and power factor correction is included, and all usual protection is provided. Coutant Lambda Ltd. Tel., 01271 865656; fax, 01271 864894.

15W converters. Power Convertibles' new 15W family of dc-to-dc converters has a wide input range of 4:1 and provides outputs of ±3.3V, ±5\V, ±15V and +12-15V in a variety of combinations. Package size on the highest power-density version, the *WFC15R*, is 1.6 by 2in. An aluminium substrate helps with heat transfer. Power Convertibles Ireland Ltd. Tel., 00353 61 474133; fax, 00353 61 474141.



Touch screens. Data entry touch screens from Rafi, using a five-wire technique, are versatile enough to be tailored for individual needs. Operating speed is high, resolution is 4096 by 4096 points, sensitivity 2mm and contact pressure under 100g. The screens are sealed and protected against moisture and dirt and, since they are designed for point-of-sale use, vinegar and ammonia products and cleaners. Fingers, gloved or ungloved and other objects can be used on the screen, which has a light transmission of 75%. All necessary electronics are supplied, including driver software for dos, Windows, Apple Mac, OS/2 and Unix. Rafi (GB) Ltd. Tel., 01737 778660; fax, 01737 778722.

Telemetry receiver. Wood and Douglas has replaced its *TR450* with the *STR450* synthesised telemetry receiver, which is physically and electrically compatible with the earlier model. Improvements include better large-signal handling, an fsk data extractor and either a 1200baud modem or controller options. *STR450* draws less than 11mA from 5.5-14V and meets ETS 300 220 and MPT 1329. Frequency range is 450-470MHz or 350-490MHz to order, the operating frequency being programmed by exchanging a disposable PIC microcontroller or by an internal 2400baud RS232 serial port from a pc. A –80dB local oscillator notch filter is available as an option. Wood and Douglas Ltd. Tel., 0118 9811444; fax, 0118 9811567.

Switches and relays

Pcb relays. *Meisei* board-mounted relays come in the popular switching forms and ratings, are enclosed and have standard 2.54mm pin spacing. *L Series* relays are four-pole, double-throw types switching up to 2A at 100V ac/dc, in an 11.5mm high housing. For dual-in-line mounting, the *M1*, *M3* and *M4* types switch 30W dc or 60VA ac and are meant for telecommunications: *M1* is a dpdt type with 5-48V coils. *M4* relays provide high sensitivity, taking 6.3mA at 48V dc and 29.9mA at 5V dc, while the *M3* versions are rapid-action relays, operating in 2.5ms and releasing in 1ms. Switchtec Electronics Ltd. Tel., 01785 818600; fax, 01785 811900.

Sealed keypanels. *M-TEC* sealed keypanels come in rubber and plastics and may be individually styled in different colours, legends, texture, size, shape, type of window and cut-outs. They panels provide tactile feedback and an audible click, so that bleepers or lamps are unnecessary. Top surface is polycarbonate as standard or polyester for hard use, matt or clear in specific areas, graphics being underneath to preserve them. *M-TEC-R* panels are in silicone rubber and, when used with a bezel, form a sealed sheet with the appearance of individual keys. Hawnt Electronics Ltd. Tel., 0121-784 3355; fax, 0121-783 1657.

Transducers and sensors

Disc thermistors. New miniature disc thermistors from *ESI* are for general use and come in a wide range of values and two sizes. The 3mm type have values between 500Ω and 30kΩ, the 5mm version ranging from

40Ω to 470kΩ, all to within ±5% at 25°C. Operating range is –20°C to 110°C. Ratings are 250mW maximum with a dissipation constant of 3mW/°C over 15s for the 3mm type; 500mW and 7.5mW/°C, 20s for the 5mm version. Anglia. Tel., 01945 474747; fax, 01945 474849.

Small load cells. In a range of sizes from compression types (micro) of 12mm to compression and tension versions (UTC) of 25.4mm in diameter, new load cells from Control Transducers will handle up to 5000kg in 14 ranges. The measuring elements of bonded strain gauges form a 350Ω four-arm Wheatstone bridge taking 10V ac or dc excitation to produce 20mV maximum. Temperature range is –10°C to 65°C. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Linear Hall sensor. Allegro's *A3516* linear Hall sensor ic uses a chopper circuit to cancel offset voltage caused by varying temperatures, among other things. The device is suited to linear and rotary position sensing in industrial, automotive and other trying conditions and produces a rail-to-rail output proportional to the applied magnetic field at a sensitivity of 2.5mV/gauss (gauss are still common in the USA). It is virtually self-contained, possessing a high-gain amplifier, voltage regulator, filter and the offset cancellation circuitry. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Optical shaft encoder. *HD Series Digipots* by Control Transducers are to an open design to keep the cost down, but have the characteristics of enclosed types. Output of this non-contacting design is digital, resolution being 96-2042 lines per rev. Dual, double-sealed ball bearings ensure very low torque and allow speeds of up to 10000rev/min. Current taken is 40mA from 5V and the output is two-channel ttl-compatible for pulse multiplication and direction sensing. Diameter of the smallest is 1in.

COMPUTER

Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Data acquisition

Portable system. *AstroDAQ 2*, introduced by Astro-Med Inc., is a data acquisition system for conditioning, analysing and networking data and is contained in a portable case about the size of a notebook computer. It uses Windows-based software for control and analysis or real-time waveform monitoring and display on a pc. The instrument will monitor data remotely while sending results via a modem to a central pc; twenty 5kHz channels can be recorded, regardless of the number acquired, data being stored on an internal 1Gbyte hard disk. All signal conditioner modules are dsp-based, modules for temperature, bridge, voltage, strain and motion being available. Up to ten units may be controlled by one pc. Astro-Med Inc. Tel., 01628 668836; fax, 01628 664994.

Card reader. Omron's *SGR* family of latching card readers handle magnetic-stripe and/or smart-card reading processes over more than half-a-million passes. The readers use a solenoid-controlled mechanism and red/green leds to show the mode in use, all in the family being configurable to read when the card is pushed or pulled. They support ISO and CP8 chip contacts and are compatible with ISO1, 2 and 3 magnetic stripe standards. A 5V supply is used. Kestronics Ltd. Tel., 01727 812222; fax, 01727 811920.

Development and evaluation

TMP93CM41F starter kit. Toshiba announces the *Topas 900* starter kit for the development and evaluation of the company's 16-bit *TMP93CM41F* microcontroller. The kit contains an evaluation board, Ansi C compiler, assembler and Windows-based debugger, with documentation, software examples and data sheets on cd. With the board comes a serial cable and a power supply and there are connectors for access to mcu signals. Code downloaded to sram on the board can be tested using the source-level debugger, which is a subset of that used with the company's real-time ice. Arrow-Jermyn. Tel., 01234 270027; fax, 01234 214674/791501. ■

STANDBY
keep-alive power supplies

Energy efficiency is today's watchword and, since much equipment is not in continuous use, small, standby power supplies are in demand.

Television receivers, Internet equipment, fax machines and other equipment spending a significant time switched on but somnolent, often use a standby power supply instead of the normal one when in contemplative mode, to reduce power needs and operating cost. Power Integrations Inc. makes the *TOP209* low-cost, switched-mode power supply ic for this purpose which, when compared with discrete switchers, is half the size and weight, uses many fewer components and needs little power itself. An application note on the subject is available.

Figure 1 is the TOP209 functional block diagram and **Fig. 2** shows a typical arrangement of a dc-input flyback circuit using the device. Referring to Fig. 1, Drain is the output and also provides internal bias current on starting, all pins do several things; Control is the input to the error amplifier, carries feedback current for the pwm circuit and connects to the shunt regulator to give internal bias cur-

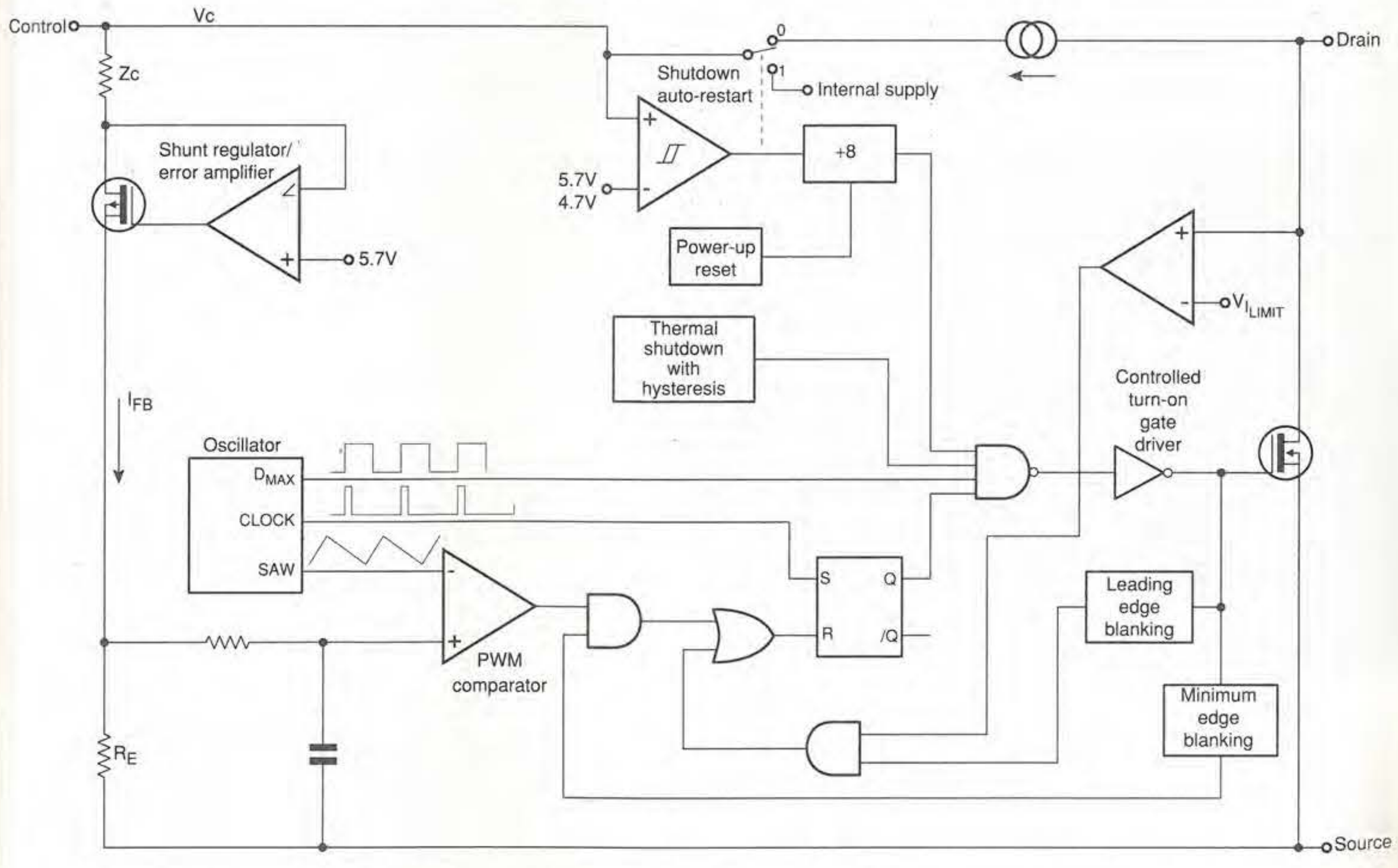


Fig. 1. Functional block diagram of the TOP209 standby power supply chip for output power to 4W.

rent in normal working; the Source pin is common and is actually two pins, normally connected together, one being for the high-voltage return.

Control voltage

Voltage V_c on the control pin supplies all controller and driver circuits and, via an external capacitor shown between control and source in Fig. 2, ensures that gate drive for the output fet is available.

Regulation of this voltage comes in two forms: for start-up and during an overload condition, the regulation is hysteretic in form, as in **Fig. 3**; in normal operation, the shunt regulator amplifier takes over. On start-up, current to supply the internal circuitry and also to charge the external capacitor, C_5 in Fig. 2, is supplied by the high-voltage current source between drain and control pins, shown symbolically in Fig. 1.

At switch-on, V_c reaches 5.7V and the current source turns off, having charged C_5, and the pwm modulator and output stage are turned on. In the circuit shown, the feedback derived from the extra secondary transformer winding keeps C_5 charged and supplies V_c. Normally, the shunt regulator maintains V_c at 5.7V by shunting excess control current through R_E, which determines the error amplifier gain.

In a fault condition in which C_5 discharges to the lower threshold of 4.7V, the current source turns on and recharges C_5; the mosfet output stage turns off, the control circuit being now in a standby condition. A 1V hysteresis in the auto-restart comparator results in the waveform seen in Fig. 3, where the current source is turned on and off, a counter preventing the mosfet coming on again until eight cycles have taken place; the resulting 5% duty cycle limits power dissipation. This whole cycle of events repeats until the fault condition is removed.

Controlled by the voltage across R_E, the pwm modulator drives the mosfet in a duty cycle inversely proportional to the current coming into the control pin, the R_E voltage being compared with the sawtooth from the oscillator to produce the duty cycle ratio, a clock from the oscillator turning on the mosfet via a latch and the modulator turning it off again.

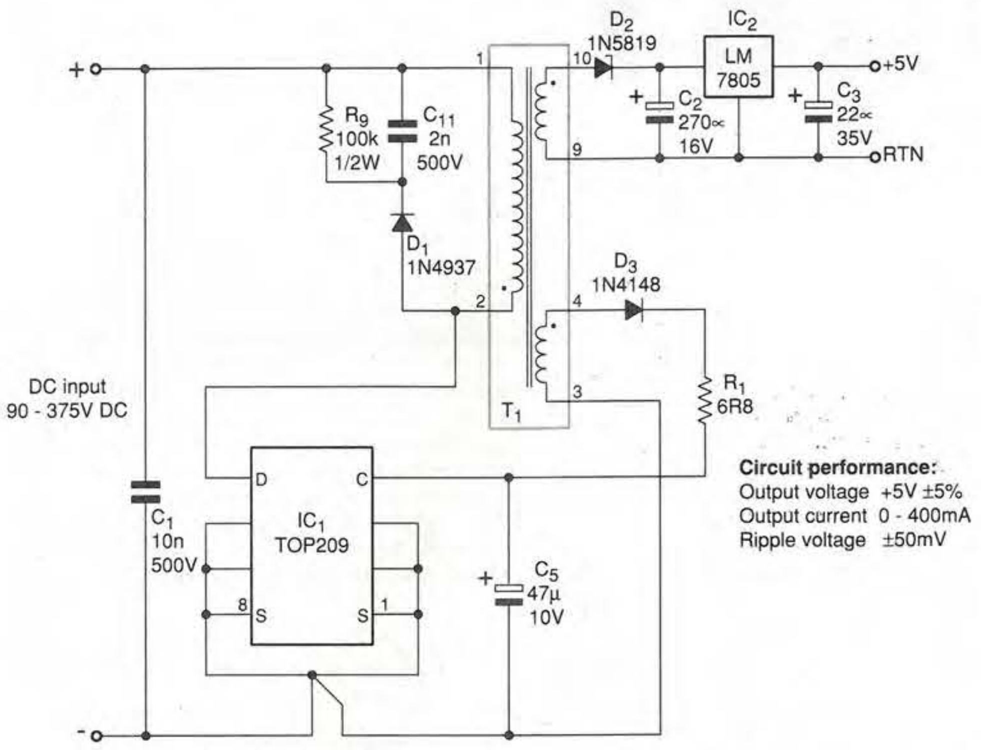


Fig. 2. 2W supply using the TOP209. Using this device, costs are lower and component count, size and weight greatly reduced.

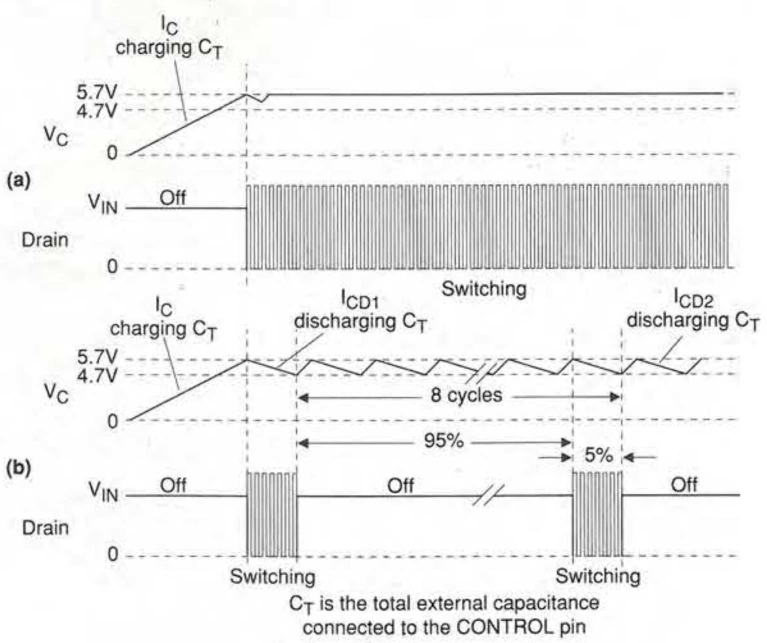


Fig. 3. Waveforms in TOP209 operation. The sawtooth is the auto-restart section, which repeats this operation until a fault condition is removed, power dissipation being thereby reduced.

Protection

If an overvoltage occurs, the high-current pulse into the control pin activates a latch, which turns off the output. Either removing and restoring input power or pulling the control pin below the lower threshold resets this latch and normal operation is resumed. As regards a too high temperature, there is an analogue circuit to act in much the same way as the overvoltage circuit, reset happening in the same way.

Practical circuit

The inexpensive 5V, 2W standby supply in Fig. 2 takes its input from a high-voltage dc source provided by rectified and smoothed 85-265V ac, its output being regulated, even lower cost being obtained by using a transistor regulator or even a zener shunt, albeit at lower output power.

In all cases, the *TOP209* drives the primary with *C*, *R* and diode providing voltage-spike protection and reducing ringing at the drain pin. The other secondary winding provides the feedback to the control pin, charging C_5 to provide bias for the *TOP209*. Capacitor C_5 also filters the voltage to the control pin and sets the auto-restart frequency. ■

Sequoia Technology Ltd., Tekelec House, Back Lane, Spencers Wood, Reading RG7 1PW. Tel., 0118 9258000; fax, 0118 9258020.

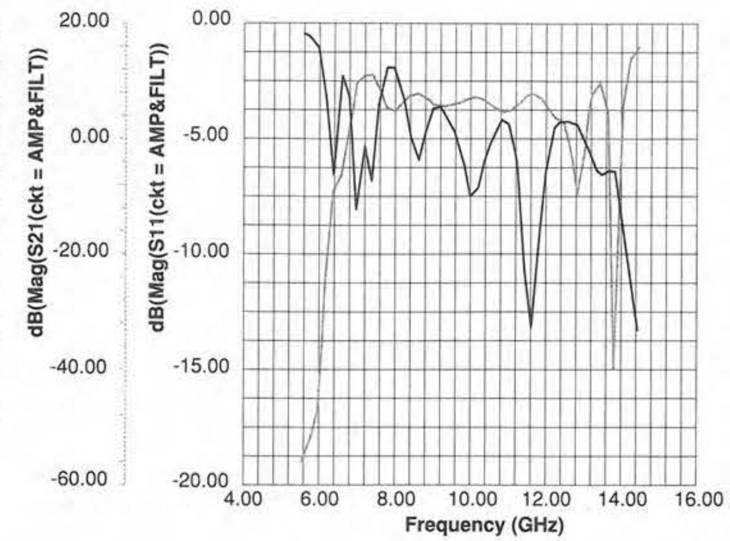


Fig. 11. Simulated performance of combined filter and two-stage distributed amplifier. Grey curve S_{21} is forward-transfer gain while the black one is S_{11}, the input reflection. Note pronounced ripple.

continued from 846

to optimise. The variable block in Fig. 9 shows the values obtained after optimisation, around 0.2mm for both. This is almost impossible to implement in hybrid form.

Figure 10 shows the theoretical response of such a design. Note how the bandwidth has spread compared with the resonant match amplifier. This is particularly evident from the trace for S_{11}.

Finally, **Fig. 11** demonstrates the theoretical effect of putting the filter at the inputs and outputs. Now if I could tidy up that ripple and increase the gain above 20dB, I could sell such an amplifier for about $600 a throw. Then I wouldn't have to scrape a living as a computer nerd any more. Ah well, one can only dream. ■

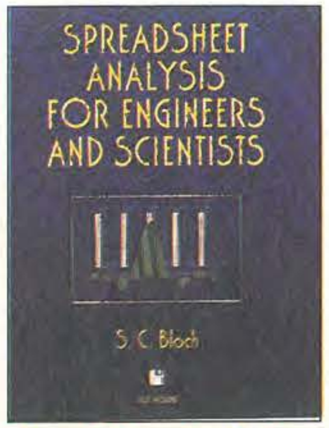
References

1. Thermionic Valve Circuits, WS Percival – British Patent 460 562, 25 January 1937.
2. Ayasli, Y, *et al*, 'A Monolithic GaAs 1-13GHz Travelling-Wave Amplifier,' *IEEE Transcations on Microwave Theory and Techniques*, July 1982.
3. Edwards, T, Foundations for Microstrip Circuit Design, pub. John Wiley
4. Matthaei, G, Young, L, Jones, E, 'Microwave Filters, Impedance Maching Networks, and Coupling Structures,' McGraw Hill.

For more CAD information

Compact Software's Serenade 7, from £14000, Rohde & Schwarz, Ancells Business Park, Fleet, Hampshire GU13 8UZ, tel. 01252 811377, fax. 01252 811447.

Hewlett Packard Series IV, from £18400, HP EESOF, Hewlett Packard, Cain Road, Bracknell, Berkshire RG12 1HN, tel. 01344 366767, fax. 01344 362092.



Spreadsheet Analysis for Engineers and Scientists

With this practical, hands-on guide, engineers and researchers learn, quickly and easily, the latest and most useful electronic spreadsheet methods. Using a variety of interactive techniques, including worksheets, self-test and practical programs on the included disk, *Spreadsheet Analysis for Engineers and Scientists* show you how to harness the power and versatility of spreadsheet programs, including those that contain the fast Fourier transform, complex operations and Bessel functions, and how to customise your own applications.

Includes disk

0471 126837, 336pp

UK £37.50, Europe £39.50, ROW £49.50

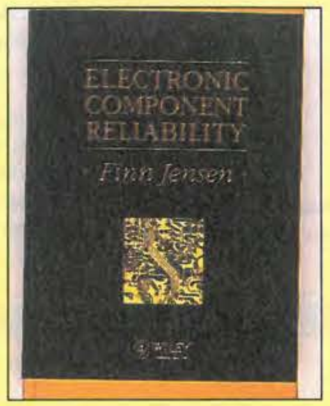
Electronic Component Reliability

Fundamentals, Modelling, Evaluation and Assurance

This text approaches the quality and reliability of electronic components from a unique standpoint. Traditionally the twin subjects of reliability physics and reliability statistics have been treated as seperate entities. Here, the author examines both areas and reveals how components fail and how failures develop over a period of time.

0471 952966, 374pp

UK £50.50, Europe £53.50, ROW £66.50



Fuzzy Logic

Implementations and Applications

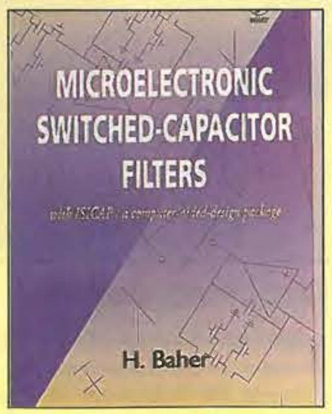
Offering a new perspective on a growing field, this text explores the many hardware implications of fuzzy logic based circuits. As use of AI increases, so the VLSI area of circuits is becoming a growth subject. Opening with an overview of fuzzy sets and fuzzy logic the book moves on to cover a range of non-standard solutions for fuzzy logic VLSI circuits. Future trends, new concepts and ideas are all examined and supported with practical examples from the author's research.

ISBN 0471 950599, approx 346pp,

UK £50.50, Europe £53.50, ROW £66.50

Microelectronic Switched-Capacitor Filters

Switched-capacitor filters and asscoiated MOS integrated circuits are now an established technology finding applications in the telecommunications and instrumentation fields. With unrivalled breadth of coverage, this book surveys the design techniques of an important class of analogue signal processing systems. An accompanying diskette containing a comprehensive computer-aided design package (ISICAP) enables readers to gain a greater depth of understanding of the described techniques. Containing both source code



files and an executable version of the main design package, this alone will be an indispensable tool for many circuit designers.

Includes disk

0471 954047 384pp

UK £75.50, Europe £79.00, ROW £92.00

The I^{2}C Bus

From theory to Practice

With a special emphasis on the I^{2}C Bus, this guidebook through the world of micro controller-managed serial buses presents comprehensive coverage of the theory necessary to design the best possible communications bus for any particular application. The book examines typical industrial and consumer applications and enables the reader to design effectively in a rel-world environment. A disk containing software for the I^{2}C bus is also included.

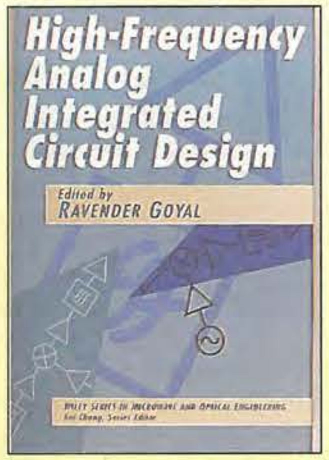
Includes disk

0471 96268 6, 314pp

UK £54.50, Europe £56.50, ROW £65.00

High Frequency Analog Integrated Circuits

As one of the first textbooks to discuss integrated circuit design considerations and



actual designs from the basic concepts, this title provides a solid background in designing basic circuits, advanced circuits and synthesis techniques.

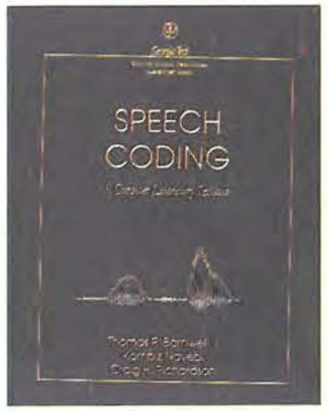
0471 530433 424pp

UK £80.00, Europe £83.00, ROW £95.00

Speech Coding

A Computing Laboratoty Textbook

This is one of the first lab manuals with software dedicated exclusively to speech processing and coding. It takes advantage of the development of the personal computer by making this technology accessible to a wider audience The manual and Dos based software together create a user-friendly digital signal processing lab which allows the user to



perform a wide variety of speech coding and speech processing experiments. The text presents and explains a set of basic speech coders analytically and in terms of the specific parameters controlling each coder. The manual leads the student through the experimental process of understanding how speech coders work and sound via over 70 exercises and projects. The class-tested menu-driven, Dos-based software can be operated by students with little or no training.

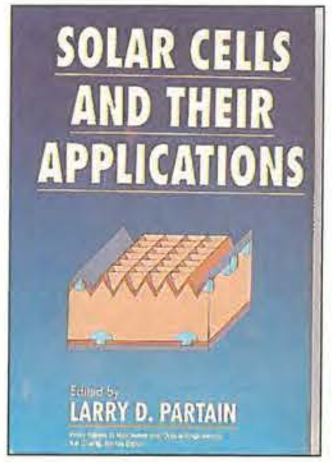
Includes disks

0471 516929, 194pp,

UK £28.95, Europe £30.95, ROW £36.95

Solar Cells and their Applications

The past decade has witnessed numerous important breakthroughs in solar cell technology, many of which have occurred in just the past few years. Far cheaper to produce and maintain, exhibiting a longer lifetime, and considerably more efficient than ever before, solar cells are, at last, in a position to compete with traditional technologies for both small and large-scale energy conversion applications. Including contributions from some of the world's leading experts in the field, this book reports the most important recent advances in solar cell technology. From in-depth discussions of breakthroughs in cell, module, and system technologies to a probing look at important environmental, health, and ssfety issues in the photovoltaic industry, it covers a broad range of topics of vital interest to solar cell researches, power systems designers, and all those with professional interest in current and future capabilities of this important technology. Offers a detailed look at cutting-edge solar technology from an international team



of researchers.

Covers silicon, GaAs, InP, CdTe, a-Si:H, CuInSe$_{29}$ and GaSb solar cells, cells, concentrators, multijunction cell configurations, space cells, and more.

Describes a wide range of applications – from space cells to terrestrial systems

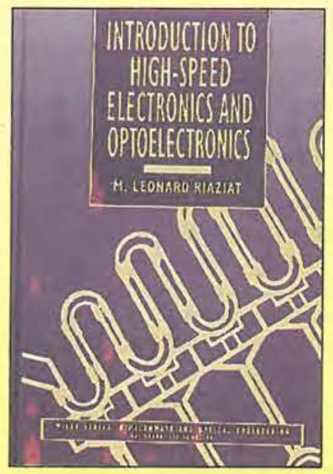
Provides an informal look ahead at the future of solar cell technology.

0471 574201, 596pp,

UK £71.50, Europe £75.00, ROW £92.00

Introduction to High-Speed Electronics and Optoelectronics

Lasers, fibre optics, and high-speed optical systems share many concepts with microwave devices. Furthermore, semiconductor-based optoelectronics and microwave integrated circuits share evolving process technologies. It is only natural, therefore, that students of optoelectronics be introduced to high-speed concepts in a unified manner. This highly practical intensive introduction enables electrical engineers, applied physicists, and students to develop and identify tools for understanding, analysis, design, and characterisation of high speed components. Broad in scope, this unique



text/reference examines the complementary nature of electronics and optics and emphasizes high-speed technology in which the two fields are less differentiated. Beginning with an overview that develops a perspective and appreciation of analog high-speed technology in general, the book goes on to cover devices and circuits used at microwave and millimeter-wave frequencies, optical components, and opto-electronic integrated circuits and subsystems. Particular attention is paid to applications in the area of high levels of interest in this area and because many of the concepts are applicable in other fields. The book concludes with important coverage of the often-overlooked area of measurement and characterization of high-speed devices. Fully referenced and supplemented with hundreds of helpful illustrations, *Introduction to High-Speed Electronics and Optoelectronics* is equally useful as a professional reference or a textbook for senior undergraduate and first-year graduate courses.

0471 015822, 312pp,

UK £65.00, Europe £67.00, ROW £77.00

Risc Systems and Applications

Professor Daniel Tabak has completely revised and updated his two previous books on Reduced Instruction Set Computer architecture to produce this new book, RISC Systems and Applications. The text is a unique, concentrated, detailed description of the architecture and implementation of most recent high-performance RISC systems, such as DEC Alpha AXP21164, IBM/Motorola/Apple PowerPC 620, Sun Microsystems and Texas Instruments UltraSPARC and SuperSPARC, MIPS technologies R10000, Intel i860 XP, Motorola MC88110, Hewlett-Packard PA-7100/8000 and the transputer. It also includes details of pioneering devices such as Berkley's RISC II and Stanford's MIPS and multiprocessor, real-time and workstation systems.

ISBN 0863 801889, 452pp,

UK £50.50, Europe £54.00, ROW £67.00

All prices are fully inclusive of packing and delivery

Return to Jackie Lowe, Room L333, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Please supply the following titles:

Qty	Title or ISBN	Price

** All prices on these pages include delivery and package **

Total________

Name

Address

Postcode Telephone

Method of payment *(please circle)*

Access/Mastercard/Visa/Cheque/PO

Cheques should be made payable to Reed Business Publishing

Credit card no

Card expiry date

Signed

Please allow up to 28 days for delivery

Understanding CAN

Svetlana Josifovska looks at CAN – the network bus standard originally developed for automotive uses that is now finding its way into all sorts of applications.

As industrial and embedded control systems become ever more complex, control-system designers increasingly adopt a distributed architecture for system layout.

Rather than having one, large processing centre with sensor inputs and actuator outputs, distributed control systems provide a means of breaking down the complexity of the tasks involved in system control. These individual control systems each concentrate on their own task, but as the need for higher levels of sophistication gains pace, so does the need to pass information and data between these centres, or nodes.

One versatile distributed architecture communications system that can be used on a macro or micro scale is controller area network, or CAN. A serial bus system, CAN is useful as an embedded communication system for microcontrollers, and as an open communication system for intelligent devices. This makes it a very attractive bus.

Being a real-time serial bus system with multi-master capabilities – allowing several CAN nodes to request the bus simultaneously – this bus system is well suited for networking 'intelligent' devices as well as actuators with sensors within a machine or plant. It is suitable for difficult and harsh electrical environments, where a high degree of real-time operation and ease of use are needed – and at low cost.

Not only in automotive areas

Originally, CAN was developed for use in vehicles. Today, it is increasingly used in industrial, building automation and other applications. It is already used as a standard in some areas, in addition to the automotive industry, although mainly in Europe. Mercedes-Benz is using CAN in its S Class series of vehicles for fast transmissions of up to 1Mbit/s.

And the more widely used CAN is, the more likely it is to appear in unexpected areas such as agricultural, nautical, instrumentation, medical, textile machines.

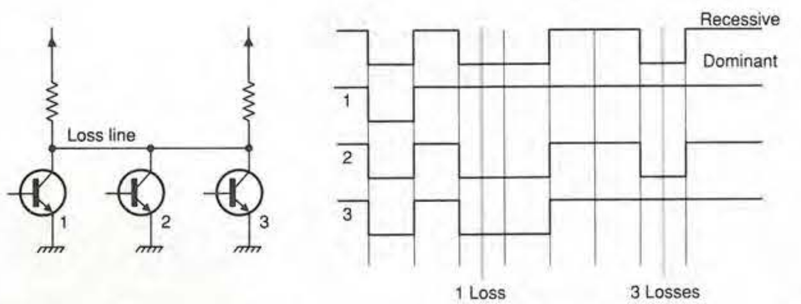
In safety-critical applications, integrity of information exchange is paramount. In the field of medical engineering, CAN is typically selected since it meets the stringent safety requirements. Similar problems are faced by manufacturers of other equipment with very high safety or reliability requirements. Examples of such equipment are robots, lifts and transportation systems.

The relative simplicity of the CAN protocol means that very little cost and effort is needed on training and the CAN chips interfaces make applications programming relatively simple. Low-cost CAN controller chips, allowing simple connections to microcontrollers, have been available since the last decade.

To spring 1996, over 10 million CAN nodes had been installed. Today there are more than 25 CAN protocol controller chips from over ten manufacturers. And as the CAN protocol steadily gains in popularity the CAN chip's availability is guaranteed.

Reliability issues

One of CAN's features is high transmission reliability. In CAN networks, instead of addressing subscribers or nodes, the transmitted messages are prioritised. A transmitter sends a message to all CAN nodes and each



As several nodes can require to send messages on the CAN bus simultaneously, conflicts can arise. These can be resolved by non-destructive bitwise arbitration: each node's identifier can be in one of two states: dominant – logic 0 – or recessive – logic 1. During the arbitration, the dominant state overwrites the recessive state. So the 'losing' nodes become message receivers and do not attempt re. transmission until the bus is available again.

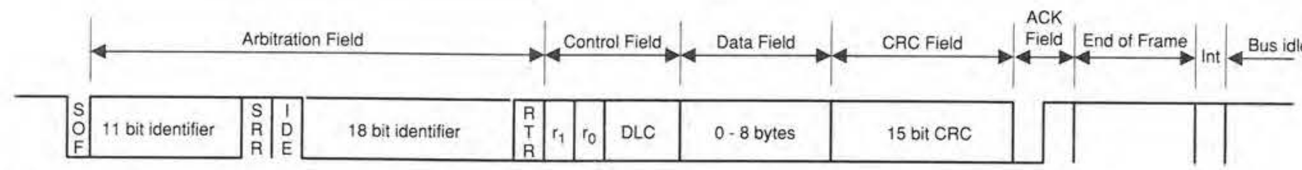


Fig. 1. Format of an extended CAN message specification comprises seven main fields.

decides on whether to process the message or not, based on the identifier received. The identifier also determines the priority of the message as it competes for bus access.

Such a content-oriented addressing scheme allows a high degree of system and configuration flexibility. As the data transmission protocol does not require physical destination addresses for individual components, it is particularly suitable for modular electronics and permits multiple reception – i.e. broadcasting and multicasting.

This addressing scheme also makes CAN suitable for synchronising distributed processes. In such applications, measurements needed as information by several controllers can be transmitted via the network in such a way that it is unnecessary for each controller to have its own sensor. Suitably, nodes can always be added to an existing CAN network without making any hardware or software changes to the others.

Each CAN message may have from 0 to 8 bytes of user information. Longer data information can also be transmitted by using segmentation. The maximum transmission rate specified so far on a CAN network is 1Mbit/s over lengths of up to 40m. For longer distances the data rate is reduced. For example, for distances of up to 500m, the specified data speed is up to 125kbit/s and for transmissions of up to 1km the data rate is around 50kbit/s.

When data is transmitted by CAN, no nodes are addressed. Instead, the content of the message, which can be a parameter such as revolutions-per-minute or engine temperature, is designated by a unique identifier. The identifier specifies the content and the priority of the message. This is important for bus allocation when several nodes are competing for bus access.

If the cpu of a given node wishes to send a message to one or more nodes, it passes the data to be transmitted and its identifiers to the assigned CAN chip. Known as 'making ready', this is all the cpu needs to do to initiate data exchange.

The message is formed and transmitted by the CAN chip. As soon as the CAN chip receives the bus allocation – a condition called 'send message' – all other nodes on the CAN network become receivers of this message 'receive Message'.

Having received the message correctly, each node in the CAN network performs an acceptance test to determine whether the data received is relevant to that node, an operation called 'select'. If the data is of significance for the node concerned it is then processed – called 'accept' – otherwise it is ignored.

Error handling

When an error is detected, the CAN controller registers it, as well as which node it is coming from. It then evaluates it statistically in order to take appropriate measures. This could lead to the disconnecting of the CAN node that has produced the error.

The CAN protocol supports two message frame formats – standard or Version 2.0A – and extended CAN or Version 2.0B, **Fig. 1**. The only difference between the two is in the length of the identifier.

In the standard format the length of the identifier is 11 bits and in the extended format the length is 29 bits. The extended format CAN messages with 29-bit identifier was introduced after the American-based Society of Automotive Engineers, SAE, 'Truck and Bus' subcommittee decided to standardise signals and messages in addition to the data transmission protocols for a variety of data rates. Standardisation of this kind turned out to be easier to implement when a longer identification field is available.

For a standard message, the transmission message frame consists of seven main fields. A message in the standard format begins with the 'start of frame' bit. It is followed by the 'arbitration field', which contains the identifier and the 'remote-transmission request' bit that indicates whether it is a data frame or a request frame.

The 'control field' contains the identifier extension bit that indicates whether the format is standard or extended, a bit reserved for

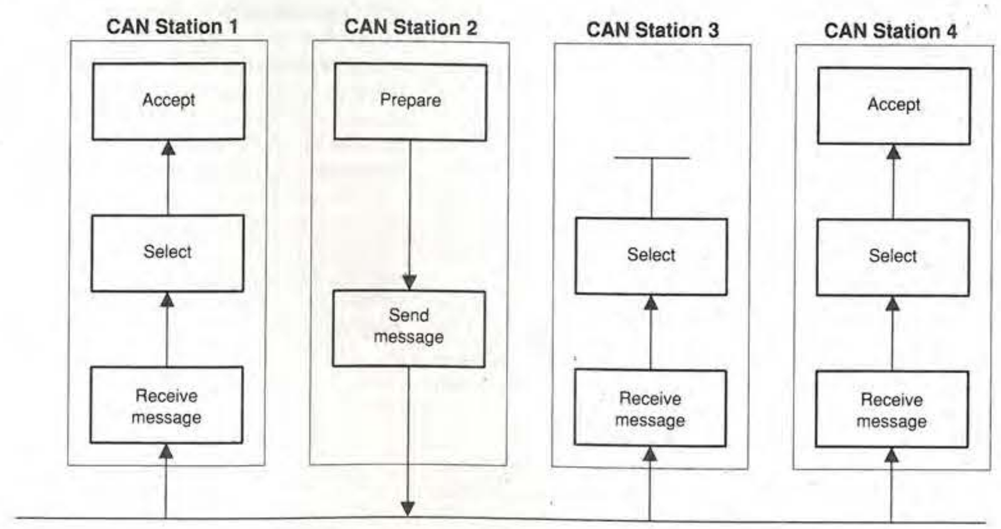
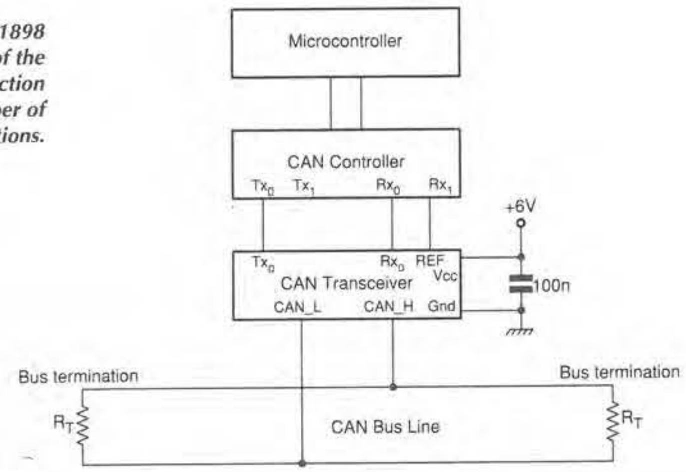


Fig. 2. Each CAN node carries out broadcast transmission and acceptance filtering.

The ISO 11898 implementation of the physical CAN connection is just one of a number of options.



future extensions and a count of the data bytes in the data field.

Length of the 'data field' can vary from 0 to 8 bytes. It is followed by the cyclic redundancy check 'CRC field', which is used as a frame security check for detecting bit errors. The acknowledgement field, or 'ACK field', comprises of a one-bit ACK slot and one-recessive bit ACK delimiter.

The bit in the ACK slot is sent as a recessive bit and is overwritten as a dominant bit by those receivers which have at this time received the data correctly, typically dubbed positive acknowledgement. Correct messages are acknowledged by the receivers regardless of the result of the acceptance test.

The end of the message is indicated by 'end of frame'. 'Intermission' is the minimum number of bit periods separating consecutive messages. If no other nodes demand to access the bus then the bus remains idle.

Extended format messages

In extended format transmission messages the 29-bit identifier consists of the existing 11-bit identifier (base ID) and an 18-bit extension, also known as the ID extension.

Distinction between standard format and extended format is made using the IDE bit, the so-called identifier extension bit. This is transmitted as dominant in the case of a frame in standard format. For frames in extended format it is recessive.

The RTR bit is transmitted dominant or recessive depending on whether data is being transmitted or whether a specific message is being requested from a node. In place of the RTR bit in standard format the substitute remote request, or SRR, bit is transmitted for frames with extended identification.

The SRR bit is always transmitted as recessive, to ensure that in the case of arbitration the standard frame always has priority bus allocation over an extended frame when both messages have the same base identifier.

Unlike the standard format, in the extended format the IDE bit is followed by the 18-bit identification extension, the RTR bit and a reserved bit (r1). All following fields are identical with the fields in standard format. Conformity between the two formats is ensured by the fact that the CAN controllers which support the extended format can also communicate in standard format.

The two formats can easily coexist on one bus. However their relevant messages are prioritised differently. The standard CAN message version always has priority over the message in extended format. This way, collisions of messages during requesting bus access are avoided.

Extended format CAN controllers can also send and receive messages in standard format. On the other hand, CAN controllers which only cover the standard format, or Version 2.0A, can transmit only standard format messages on that network. Otherwise, the extended format messages will be misunderstood.

There are CAN controllers however which can recognise extended messages even though they only support standard format. In these cases the extended format will ignore those messages. As a result, it has been marked as the Version 2.0B passive format.

Error signalling

The CAN protocol signals errors that occur on the network rather than use acknowledgement messages as in other bus systems. For error detection, the CAN protocol has three mechanisms at the 'message' level and two mechanisms at the 'bit' level.

At the message level, there are the cyclic-redundancy check, frame check and ACK errors. At the 'bit' level, errors mechanisms can be one of two types: 'monitoring' or 'bit stuffing.'

Cyclic redundancy checking safeguards the information in the frame by adding redundant check bits at the transmission end. At the receiver end, these bits are re-computed and tested against the received bits. If they do not agree there has been a CRC error.

Frame checking verifies the structure of the transmitted frame by checking the bit fields against the fixed format and the frame size. Errors detected by frame checks are designated 'format errors'.

Each CAN node can transmit or observe bus signals, **Fig. 2**. That way they can easily detect the difference between the bits sent and bits received, which indicates an error. This is 'monitoring'. With 'bit stuffing' the sender inserts into the bit stream a bit which can be removed by the receiver. This permits reliable detection of all global errors and errors local to the transmitter.

When errors are discovered by at least one node by means of the above mechanisms, the current transmission is aborted by sending an 'error flag'. This prevents other nodes from accepting the message which ensures the consistency of data throughout the network.

After transmission of an erroneous message has been aborted, the sender automatically re-attempts transmission – an action entitled automatic repeat request. There may again be competition for bus allocation. As a rule, re-transmission will begin within 23 bit periods after error detection and in special cases the system recovery time is 31 bit periods.

Lock-ups

Although this method of identifying and isolating a defective node proves effective and efficient, it may occasionally lead to all messages – including correct ones – to be aborted. This can block the bus system if no measures for self-monitoring are taken. The CAN protocol therefore provides a mechanism for distinguishing sporadic errors from permanent errors and localising node failures.

This is done by statistically assessing the node error situations. The aim of this is to recognise a node's defects and if need be enter an operating mode where the rest of the CAN network is not negatively affected. This may go as far as the node switching itself off to prevent the abortion of correct messages that have been misinterpreted as incorrect.

On the other hand, errors can go undetected. However the statistical average of that happening on a CAN network operating at a data rate of 1Mbit/s with an average bus capacity of 50% is one undetected error in every thousand years. ■

CLASSIFIED

TEL 0181 652 3620 FAX 0181 652 8956

ARTICLES WANTED

VALVES and CRTs AVAILABLE

ONE MILLION VALVES stocked for Audio, Receiving, Transmitting & RF Heating. Rare brands such as Mullard & GEC available. Also MAGENTRONS, KLYSTRONS, CRTs and SOCKETS.

Large stocks of Russian and Sovtek items.

Please ask for our free catalogues of valves or CRTs.

VALVES, etc, WANTED

Most types considered but especially KT88 (£60), PX4/PX25 (£60), KT66 (£38), KT77 (£18), EL34 (£15), EL37 (£15), ECC83 (£3). Valves must be UK manufacture to achieve prices mentioned. Also various valve-era equipment e.g. Garrard 301, (up to .£80). Ask for a free copy of our wanted List.

BILLINGTON EXPORT LTD., Billingshurst, Sussex RH14 9EZ
Tel: 01403 784961 Fax: 01403 783519

VISITORS STRICTLY BY APPOINTMENT. **MINIMUM ORDER £50 plus VAT**

TOP PRICES PAID

For all your valves, tubes, semi conductors and IC's.

Langrex Supplies Limited
1 Mayo Road, Croydon
Surrey CR0 2QP
TEL: 0181-684 1166
FAX: 0181-684 3056

RF DESIGN SERVICES

All aspects of RF hardware development considered from concept to production.

WATERBEACH ELECTRONICS

TEL: 01223 862550
FAX: 01223 440853

Test & Measurement Instruments

A chance to profit from your own ideas !

An established U.K. manufacturer of test & measurement instruments is looking for new product concepts.

They would like to hear from design engineers with innovative product ideas that could lead to joint developments on either a fee or royalty basis.

In the first instance, please write giving an outline of your product concept along with details of your own experience and capabilities.

Detailed disclosure, should the idea be pursued, would be subject to an intellectual property agreement protecting your rights.

Write to: Box 111, Reed Business Information, Room L329, Quadrant House, The Quadrant, Sutton, Surrey. SM2 5AS

PRECISION SHEET METAL SERVICES

We specialise in the manufacture of small sheet metal parts for the electronic industry; chassis, screening cans, cases etc., sizes under 30cm.
SMALL/MEDIUM QUANTITIES. QUICK TURNAROUND
Please write, fax or phone for quotation

HARRISON ELECTRONIC SERVICES
Century Way, March, Cambs. PE15 8QW
Tel/Fax: (01354) 651289

MEGGAR INSULATION TESTER, up to 2500v, hand cranked, excellent condition, leather case, accessories and leads: £300 ono. Tel: 01793 724151.

WANTED – any AVO model 8 with hopelessly damaged movement. Tel: 01326 312901.

FOR SALE: complete volume of Electronics World/Wireless World from 1945-1983. Plus selection of technical books. For further information please tel: 01920 870289.

WANTED – W.W.II. German and S.O.E. clandestine sets, receivers. Also incomplete sets. 0Z8RO Rag Otterstad, Hosterkobvej 10 DK-3460 Birkerod, Denmark. Fax: ++45 44681514.

WANTED – Old cipher equipment and "spy radio sets". 0Z8RO Rag Otterstad, Hosterkobvej 10 DK-3460 Birkerod, Denmark. Fax: ++45 44681514.

FOR SALE – HP8505 network analyzer. Price: Two pre-war TVs. Jac Janssen, Hogeham 117D, NL-5104JD Dongen, Netherlands. Fax: +31 13 4624 684. Tel: + 31 16 2318 158.

FLUKE 867B graphical multi meter, five months old, with leads and manual: £550 ono. Tel: 01482 815625.

TEKTRONIX 145 PAL Genlock Test Signal Generator, 148 Insertion Test Signal Generator, EV4020 Vectorscope, EV4040 Waveform Monitor. All manuals. Tel: 01383 419282.

TWO TEKTRONIX 561A oscilloscopes. Trolley, manuals, plug ins. Good working order. £60 pair. Details tel: 01273 553505.

PLEASE MENTION
ELECTRONICS WORLD
WHEN REPLYING TO ADVERTISEMENTS

ARTICLES FOR SALE

PCB CAMERA £49.00+VAT
B/W MODULES

32mm. 3.5 lens or pinhole £49.00 + vat each
42mm • lens with audio add £10 + vat

(UK CARRIAGE/PACK/INSURANCE £4 UP TO 10 PIECES)

DISCOUNT FOR QUANTITIES
SEE OUR WEB SITE
OPEN 6 DAYS A WEEK FOR CALLERS
MAIL-ORDER OR EMAIL

HENRY'S 404 EDGWARE ROAD LONDON W12 1ED

Tel: 0171 258 1831 Fax: 0171 724 0322
email: sales@henrys.demon.co.uk internet address: http://www.henrys.co.uk

ADVERTISERS PLEASE NOTE

For all your future enquiries on advertising rates Please contact Joannah Cox on

Tel: 0181-652 3620
Fax: 0181-652 8956

ADVERTISERS INDEX

Anchor	816 and 836	Pico Technology	849
Card Professionals	849	Quickroute	836
CMS	823	Ralfe	880
Colomor	839	RD Research	853
Conford	799	Stewart of Reading	839 and 783
Cooke	859	Seetrax	830
Dataman	OBC	Surrey	830
Display Electronics	862	Tandem Technology	IFC
Equinox	IBC	Telford	811
Hart	868	Telnet	813
IOSIS	816	Temwell	799
John's Radio	800	Texas	799
JPG	830	Those Engineers	877
Labcenter	805	Tie Pie	811
M & B Radio	847	Ultimate	803
Number One	829	Weir	855
Olson	794	Wood & Douglas	839

ELECTRONIC UPDATE

Contact Joannah Cox on 0181-652 3620

A regular advertising feature enabling readers to obtain more information on companies' products or services.