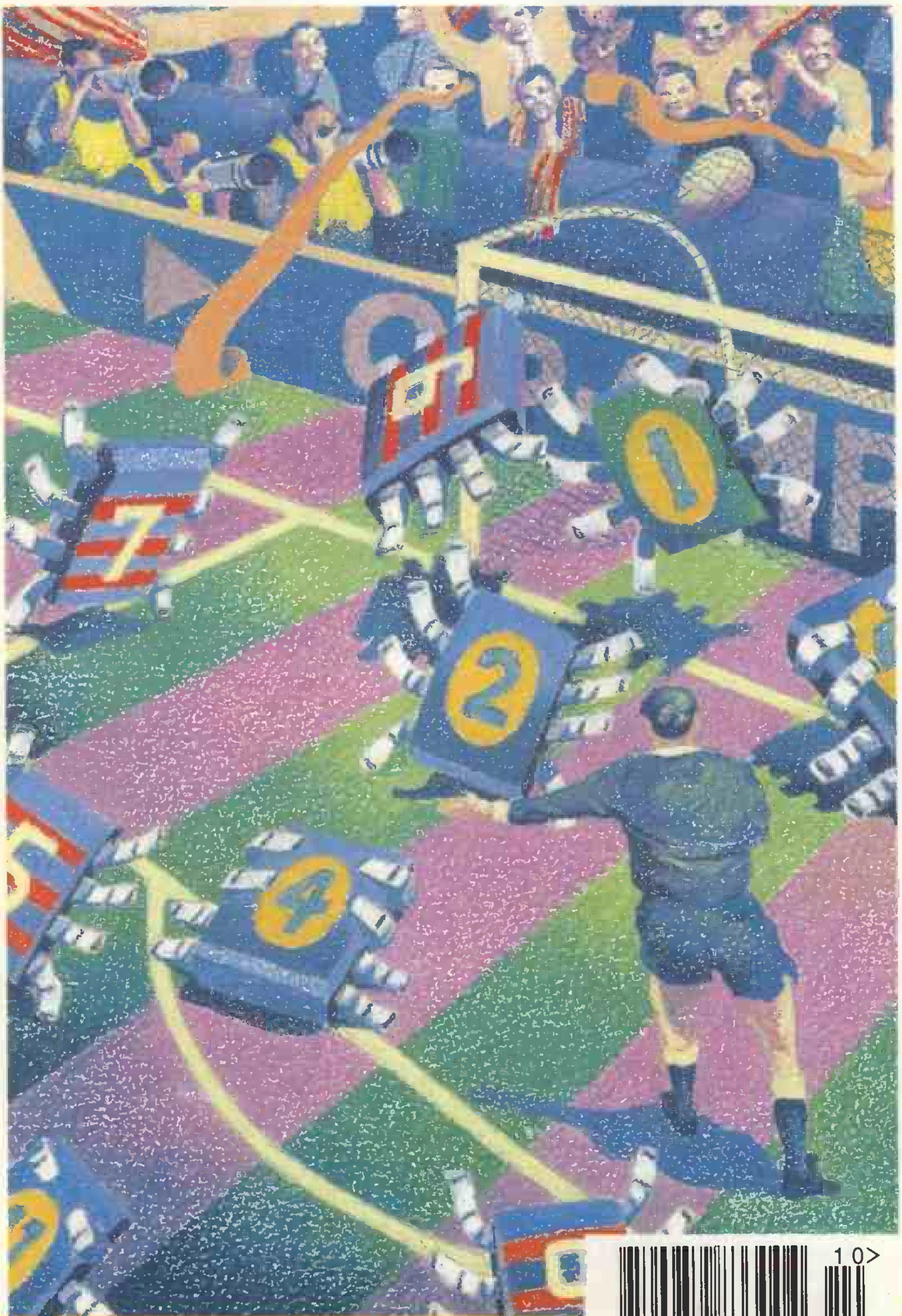
ELECTRONICS WORLD + WIRELESS WORLD

Denmark DKr. 70.00
Germany DM 15.00
Greece Dra.760
Holland Dfl. 14
Italy L. 7300
IR £3.30
Spain Pts. 780
Singapore S$ 12.60
USA $6.70

A REED BUSINESS PUBLICATION
SOR DISTRIBUTION

OCTOBER 1994 £1.95



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Cover: Hashim Akib

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By finding a new application for a high-voltage IC comprising power mosfet drivers and a 555-type timer, you could win one of six mosfet designer's kits worth £1800 in total.

Full details on page 837...

REGULARS

In next month's issue: Making the most of CAD. Owen Bishop takes an in-depth look at CAD from the design engineer's viewpoint, using working circuitry for demonstration.
And John Gregg looks at a new generation of atomically engineered magnetic materials.
There will also be details of our writers' award – the prize for which is a £4000.

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CIRCLE NO. 103 ON REPLY CARD

VISA

Goodbye, goodbye

I have always promised that I would never allow sporting similes to appear in this magazine. After all, it is simply not cricket. But since this is my last editorial for our magazine, I shall make an exception.

How else can you note the performance of Great British Electronics plc other than to compare it with our national cricket or football teams and their past triumphs? I say this because the rumours are insistent that our last major semiconductor maker, GEC Plessey Semiconductors, is about to be sold off to US company Rockwell. It looks like we are about to say goodbye to our last chance to compete in a $100bn worldwide industry, a figure which is expected to rise to $200bn by the turn of the century.

It is worth recalling that just a decade ago, we had five major indigenous semiconductor makers: Ferranti, Plessey, Marconi, STC and Inmos. A similar list for two decades ago would have been double that length. There are a host of complacent voices in the UK Government and the Civil Service who ask if the passing of a national semiconductor industry really matters...

The same voices say that the real value comes from building semiconductors into systems rather than in the making of the devices themselves. They see that wafer lines require astronomic investment, produce irregular returns and don't employ too many people. Contrast this with equipment assembly operations which are undemanding of their bankers and backers while employing any number of redundant miners and shipbuilders.

To anyone who has worked in electronics as I have, who has sat in this office and watched in impotent fury as we have moved down the world technology league table, the seductive voices were wrong, wrong, wrong. I say "were" because, realistically, the time has long gone when we could have turned the situation around. We are now marginalised in an area fundamental to electronics design.

Semiconductor development and manufacture controls absolutely the design of the end-equipment. We mostly define the nature of a complete system at the silicon layout stage. If we wanted to add significant value in system building, we should have retained control of the enabling technology. We didn't, we haven't while our national competitors have and thus the fists of rage. The picture of Arnie Weinstock abdicating totally from any responsibility to our last major semiconductor concern will present a fitting epitaph to the UK's high technology decline.

Returning to the sporting simile, there are two remaining hopes. Root for the Europeans such as Temic, SGS-Thomson, Siemens and Philips in the hope that they play a few away matches in our country; support our junior league of small silicon design houses which have been singularly successful in contrast to the big league. Either way, we must keep playing the game to retain any sort of advanced technology manufacturing base.

I am saying goodbye to Wireless World after six years in the editor's chair to take up a position with another magazine. I have enjoyed my tenure and I would like to thank both readers and contributors for their support and the help which they continue to give to our magazine. I know that my successor, Martin Eccles, currently WW's deputy editor, will continue to develop this venerable journal towards applied electronic design. This has always been my ambition and I believe it to be one that Martin shares. **Frank Ogden**

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ISSN 0959-8332

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

Overseas advertising agents: France and Belgium: Pierre Mussard, 18-20 Place de la Madeleine, Paris 75008. United States of America: Ray Barnes, Reed Business Publishing Ltd, 205 E. 42nd Street, NY 10117.Telephone (212) 867-2080. Tlx 23827.
USA mailing agents: Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. 2nd class postage paid at Rahway NJ Postmaster. Send address changes to above.
Printed by BPCC Magazines (Carlisle) Ltd, Newtown Trading Estate, Carlisle. Cumbria, CA2 7NR
Typeset by Marlin Graphics 2-4 Powerscroft Road, Sidcup, Kent DA14 5DT

EC in the dark over r&d

An admission by the European Commission that it does not know what the r&d plans of Europe's major companies are, or how they relate to the EU's own strategy, is expected to create further argument in the UK cabinet.

The Commission plans to create a register of the research and development strategies of Europe's top 500 industrial companies. The aim is to discover the effectiveness of r&d being funded through its Framework Programme and Eureka.

EC spokesman Michel André said: "European enterprises are responsible for 60 per cent of all r&d undertaken in Europe, but the information we need to evaluate this work is missing."

The admission by the Commission will add weight to the UK government's criticism of the way in which the EC calculates funding for R&D programmes. It will also strengthen the government's view that cutting back UK funding to the Eureka programme was the right one.

A spokesman for the DTI said that such a situation could not arise in the UK. He said: We run a UK r&d Scoreboard, giving a breakdown on UK high tech companies, and how much they spend on r&d as a percentage of sales and profit. We also have an Innovation Unit linking national programmes, such as Link, into the work being undertaken by industry and academia.

Euro screen is clear world leader

A new liquid crystal screen for portable computers and video projectors has the remarkable property of self-healing any defects in manufacture.

Developed by Philips Research Laboratories at Redhill, UK, the display uses active-matrix thin-film diode lcd is easier and cheaper to make than conventional lcds. It also gives brighter pictures.

The screen measures 24cm diagonally and is made as a plug fit replacement for the conventional lcd panels currently supplied by Japanese manufactures. Philips hopes this will encourage firms making portable computers to switch from using Japanese to European displays.

Virtually all screens in today's portable PCs use thin-film transistor technology to control the flow of electricity through the liquid crystal.

The material is sandwiched between two thin glass plates and light-polarising filters. When electricity is passed thorough the lc material it reacts to alter the angle of polarisation of any light which is passing through. The filters then work like crossed sunglasses, to create light areas where the crystal and filter polarisations have matching angles and dark areas where they cross.

In thin-film transistor screens, the glass plates have a thin film of amorphous silicon semiconductor material bonded to their surfaces. This silicon is formed into a mosaic of transistors which control the flow of electricity through pixel spots of the lc material to create a pattern of light and dark which displays the text or graphics image. Philips uses silicon diodes instead of transistors. The diodes are easier to make than transistors and need only two wire connections. They work with the capacitance of the lc material to switch between on and off states.

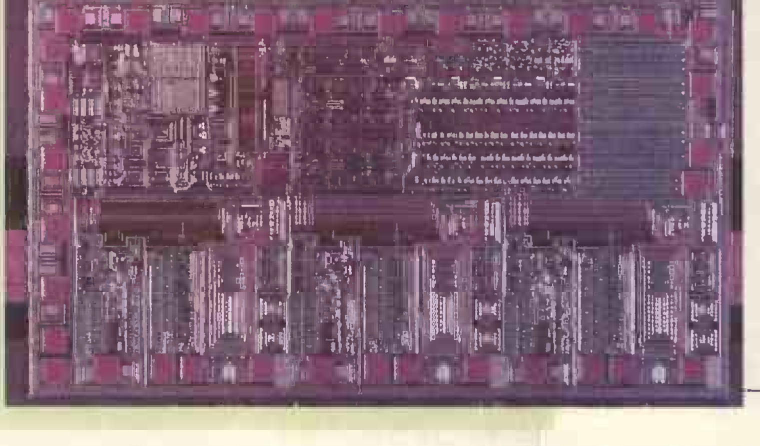
The diodes are made by sandwiching silicon nitride between metal electrodes which are so thin that they are almost transparent. Because the diodes block less of the light which passes through the screen, the picture is around 10% brighter and has crisp contrast with 256 grades of grey. In practice this means that black areas of the image look much blacker than on conventional screens. The diodes also respond rapidly to the switching current so there is no smear on moving objects.

The diodes are easy to mass-produce because alignment of the electrode sandwich is not critical. As long as the two outer electrodes cover at least part of the silicon nitride filling, the diode works as a switch. The filling also has the remarkable ability to self-heal. If it is faulty through a fault in manufacture, and thus passes too much current, its electrical resistance increases to reduce the current.

Philips claims that the diodes are stable in bright light and heat and thus suitable for use in a projector. This means that the new lcd panels can be used in a video projector, where a bright light is shone through a small lcd panel and lens, and onto a large wall screen. Transistor panels break down after a few hours use under these conditions.

Philips began work on the display technology in 1989 and the first manufacturing samples are now rolling off a production line which Philips has jointly constructed with Thomson of France at a factory in Eindhoven, the Netherlands. This was the factory originally built for the ill-fated *Megachip* memory project. **Barry Fox**

Intelligent power dressing: This silicon maze is claimed to be the first dc motor controller chip to bring together high-side drive and low voltage regulator aspects in one surface-mount package. This IC, the Si997CS from Siliconix, is designed for three phase brushless motors and handles between 1-5A of current. Interfacing with microprocessors in photocopiers and printers, the controller will handle over a dozen control and protection functions.



Multilevels boost memory

Intel has revealed that it is developing new flash memory technology which will enable a single storage cell to hold two or more bits of data.

The technology, dubbed *Multilevel Cell*, dramatically increases a chip's storage capacity without needing any more memory cells. Intel says the technique will be applied to memories for cost sensitive mass storage applications such as PCMCIA cards, digital audio and digital photography.

Multilevel Cell technology is simply the ability to write and read four (for 2-bit storage) or more voltage levels on the flash memory's floating gate storage cell. This compares with the conventional two levels for single bit storage and implies the use of three internal reference voltages.

Dr Stefan Lai, Intel's director of flash technology, said the technique has been demonstrated on a 16Mbit die storing two bits per cell. However, Lai also reported Intel engineers have experimented with three bits (eight voltage levels) and even four bits (16 voltage levels) per cell. "Our goal is to provide a 1Gbit flash memory this decade based on four bits per cell and a 256Mbit die," said Dr Lai. The chip would need a 0.35µm process.

Intel's move has interested other flash memory manufacturers. Hitachi engineers in Japan said they were aware of multilevel cell technology but had no plans at present to implement it. They identified tighter process control and slower read times, because the cell has to be compared to a minimum of three reference voltages, as the two principal drawbacks.

Giulio Casagrande, technical manager of SGS-Thomson's flash memory division, said Intel's development made sense. "If you want to address the solid-state storage market, then multilevel cell technology is just one of the alternatives that may provide the breakthrough needed to significantly reduce the cost per bit," said Casagrande. But he questioned the scalability of the approach, suggesting the read voltage would need boosting as Intel progressed to smaller geometry processes.

Dr Lai reports that the main engineering problems to be overcome are designing sufficiently accurate reference voltages and boosting the read currents from the cells. He does not consider the more stringent signal-to-noise constraints (to maintain adequate s/n margin between levels) to be outside the bounds of Intel's existing 0.6mm process even for 16 voltage levels.

"We feel çomfortable at 16 levels even down to a 0.1V margin between levels. The process is the most developed part of the technology," said Dr Lai.

"The read access times are slower but that's not really because we are reading two bits but rather that the sense currents are smaller and take a longer time to charge up the column. So we are looking at more sensitive techniques and are looking at a parallel read operation.

"Reference voltage accuracy is the big issue at the moment. The voltages need to be stable across a range of operating conditions and although there are external references that provide the accuracy, the challenge will be building an internal reference." he said.

Simon Parry, *Electronics Weekly.*

Big TI hopes for the big screen

Texas Instruments looks as though it will lead the market for PALPlus decoders with the launch of Britain's first widescreen TV service using PALPlus technology later this year.

Channel 4 has announced plans to broadcast at least 500 hours of PALPlus 16:9 format material a year starting in October. The £1.5m cost incurred by Channel 4 is being split between the EU and Nokia Consumer Electronics, which will launch a 28in PALPlus tv to coincide with the start of services. The set will cost £1299.

TI is the only company with a highly parallel digital signal processor aimed at pixel processing and suitable for handling the PALPlus decoding in tvs. It could maintain that position until dedicated PALPlus chipsets become available, and they are not expected until next year end at the earliest. "Everyone is using the Texas chip," confirmed Dr Helmut Stein, head of r&d at Nokia Consumer Electronics.

Called the SVP, the TI chip enables three dimensional processing of an entire line of the picture at a time. The chip integrates 960 processing elements in a single-instruction, multiple-data architecture. The number of elements is sufficient for a 16:9 picture format and is quick enough to cope with 16MHz sampling.

A more powerful version of the chip is due at the end of this year.

It will be four times faster, integrate more on-chip program memory and integrate 1024 processing elements – sufficient for 16:9 format pictures using computer-compatible square pixels.

Three German tv stations are already making regular PALPlus broadcasts. Belgium is due to start this year, and Holland and Italy next.

The ultimate train set: Managing the Jubilee line extension of the London Underground could be carried out from a computer terminal. Datel Technology has demonstrated an information management system in a £60m project for the railway. Real-time information from the platforms, signalling and surrounding area will be available from the ten stations on the line.

Video surveillance over the 'phone line

Video surveillance technology developed for the US army by Iterated Systems of Atlanta, Georgia, will soon let unattended cameras in art galleries, car parks and office buildings catch thieves, vandals and terrorists. It relies on fractal compression of moving images and plays a clever trick to send clear moving pictures of any suspicious behaviour down an ordinary 'phone line. Since the system also works over a cellphone link, the camera can be in a field.

The pictures display any distance away on an ordinary personal computer, and are recorded on an ordinary floppy disk. To save money on phone charges the camera automatically dials the telephone number of the display PC only when it registers motion. The same technology can be used for low cost videoconferencing.

Until now videophones have only been able to send very blurred moving pictures down ordinary domestic telephone lines. Thieves are only recognisable if they obligingly remain still and pose close to the camera. Systems which deliver high quality pictures rely either on ISDN digital telephone lines or work only over very short distances.

The US military wanted to put a video camera on a remote-controlled vehicle, and send it into a dangerous area, while relaying high quality moving pictures by low quality radio links. Iterated Systems had already developed fractal technology to compress high quality still pictures into small volumes of digital code. The company made the compressor work fast enough to code moving pictures from a video camera.

Coding a picture by fractal compression reduces the number of digital bits needed to capture good quality by a factor of around thirty. The fractal compressor works by breaking each picture down into component shapes, like pieces of a jigsaw, and then rebuilding the picture by arranging basic shapes which have been previously stored in a library. Conventional picture compression systems work by breaking the picture down into a mosaic of tiny picture points or pixels, and coding each one separately.

Despite the powerful compression achieved with fractals, the number of bits per second needed to display a full screen picture of clear motion is still far too many to send over long distances by a POTS ("plain old telephone service") line. POTS lines, and cellphone links, can reliably carry only around 10 kilobits/second, and at best 20kbits/s.

Iterated Systems gave the US military a clever compromise. At the beginning of the surveillance session, the camera takes around 10s to transmit a high quality still picture of the whole area under surveillance. This picture is frozen on the screen of the display PC. The operator then singles out a small 'window' area of the picture, such as a vulnerable doorway or valuable painting on a wall. The camera now provides a moving image for just that part of the picture.

Because the selected area of the screen can be relatively small, the quality of the moving image is high, but the bit rate within phone line limits. So the overall impression is of a clear picture of the whole area, with an equally clear view of any motion in those areas which need surveillance.

After initial setup, the telephone connection is broken to save money. But as soon as the camera registers any motion in the selected image area it automatically dials the telephone number to connect with the PC, and triggers an alarm to alert the operator.

Alan McKeon, Iterated System's Vice-President of Sales and Marketing, recently demonstrated the system working on ordinary telephone lines between London and Atlanta. Clear moving pictures of an office worker's head and shoulders appear in the middle of an overall, frozen view of the office room.

The compression circuitry will be built into a video camera, along with a telephone modem and auto-dialler, and sell for around £750. All the user then needs is a desktop PC, ordinary phone line or cellphone. Iterated Systems Ltd 0734-880261.

Inmarsat to improve on GPS services

Inmarsat, the international satellite operator, has signalled its intention to compete head-to-head with existing US and Russian global positioning (GPS) satellite services, inviting potential GPS service providers to bid for navigation transponders on its next generation Inmarsat-3 satellites.

Inmarsat claims the new satellites, to be launched at the end of next year or early 1996, will offer a more accurate positioning service than the existing military-owned US Global Positioning System and Russian GLONASS (Global Navigation Satellite System).

In addition, Inmarsat says it will provide an independent "integrity monitoring" service for the existing GPS networks. As well as the navigation signals, the satellites will broadcast an additional signal which corrects errors in the US and Russian services. The move could enable civil airlines to start using GPS for navigation for the first time. Until now airlines have been put off by the unreliability of existing services.

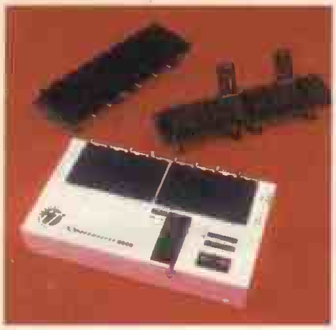
The satellite organisation says Inmarsat 3's correction signals will pinpoint the position of users to within 10m, compared to the 50m currently achievable by using the US service alone. ■

Drive to nowhere: Not in a jam around London but in a driving simulator. Researchers at the University of Iowa have created a virtual highway on the car windscreen. The aim is to collect data for new research into vehicle safety, highway design and driver training.

Three computer-generated images provide the driver with a realistic illusion of moving at speed while also feeling bumps in the road and hearing the wind rushing by. This virtual reality experience includes the sensation of a violent rear end shunt.

Eurofighter scrap

A call that the Eurofighter manufacturing collaboration should be scrapped due to soaring development costs has come from the German opposition Social Democrats. But the German defence .minister Volker Ruhe has defended the multi-billion DM project which is being financed by German, British, Spanish and Italian taxpayers.



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RESEARCH NOTES

Jonathan Campbell

Old masters painted in pixels

The EC programme to put high-resolution copies of Europe's greatest art masterpieces onto CD-rom could take a leap forward with development of a camera that delivers high resolutions, rapidly and from a much smaller unit than previously. The new digital camera has been developed by Lindsay MacDonald of Crosfield Electronics and Reimar Lenz of Munich's Technical University ('An Ultra-High Resolution Digital Camera', *The Journal of Photographic Science*, Vol 42, pp. 49-51) within the European Marc research project.

Monolisa, left, through the eye of a predecessor of a micro-scanning camera designed for digitally archiving works of art, right.

Marc – Methodology for Art Reproduction in Colour, part of the Esprit programme – aims to produce digital representations of fine art for high quality printing or for electronic manipulation and distribution. But current, high resolution digital camera technology is slow, cumbersome and limited to two-dimensions. MacDonald and Lenz believe their camera could change that.

Like previous digital cameras, the Marc camera makes use of micro- and macro-scanning to boost the resolution of conventional ccd arrays.

In micro-scanning, a mask is fitted over a standard low resolution ccd array to make the sampling apertures smaller. Using piezo-electric actuators, the chip can be moved in two dimensions across the image plane allowing partial images to be captured at each sampling point. These can then be assembled in the correct pixel sequence by computer. Colour images are obtained by using a ccd sensor with built-in colour filter stripes and image quality is comparable to 35mm film. The sensors are low-price and the camera is easy to set up, though the piezo-electric actuators must be carefully calibrated and the small aperture means high levels of illumination must be used. But the technique has been used to produce high resolution commercial cameras.

Macro-scanning involves step-moving the camera and lens assembly in front of the scene, by the width and height of the ccd array, to build a complete image by a series of patches. Speed of acquisition is rather slow because the sensor has to be moved about 10mm and must be allowed to settle between adjacent patches. But with sufficient positioning precision, defect-free images can be obtained.

Cameras combining micro- and macro-scanning are already in use, digitising fine art at the National Gallery in London and Neue Pinakothek in Munich.

But the problem is that they are so big, as they need to incorporate a massive rigid frame to make accurate patch movement possible. As a result, pictures must be removed from the gallery to be processed in the laboratory.

The new Marc camera combines micro- and macro-scanning behind a stationary lens so that the need for a large x-y external position mechanism is eliminated. As the image perspective is not altered, the developers say that, with a suitable lens, 3-d objects of arbitrary size can be processed, resolution of the system is limited only by diffraction and the image field size of the lens.

The camera is currently undergoing final debugging before it is used later this to year scan in a series of Flemish masters. During digitisation, each patch needs only 4s to micro-scan with an additional 0.7s to reposition. So a full size image can be acquired in less than five minutes. But the greatest advantage is that the camera can be used on pictures still hanging in position in the gallery. The result is faster processing, no transport problems or extra insurance cover and no worries over humidity damage for what are extremely valuable pictures.

Robodoc with the hip attitude

Hundreds of thousands of people every year undergo surgery to replace hips with artificial implants. But currently in the US, a brave few are being operated on by a rather unusual surgeon – a robot.

The Robodoc surgeon takes over from the human one in cutting the cavity in the thigh bone into which the implant is pushed. Results have shown that the robot's accurate preparation and positioning of the cavity means patients have a better than usual chance of walking properly again.

Trials are still in the early clinical stages. An initial ten-patient single-centre study has been successfully completed and now researchers are in the middle of a 300 patient, multi-centre test.

The robot has been developed by a US team from IBM and Integrated Surgical Systems. In early tests it was used in veterinary clinical trials on dogs needing hip replacement surgery.

The researchers report an order of magnitude improvement in surgical precision compared to manual broaching for cementless hip replacements. Russel Taylor of the IBM TJ Watson Research Center, and colleagues, claim the robot is a step forward in the evolving partnership between humans (surgeons) and machines (computers and robots), a relationship that seeks to complete a task better than either can do alone.

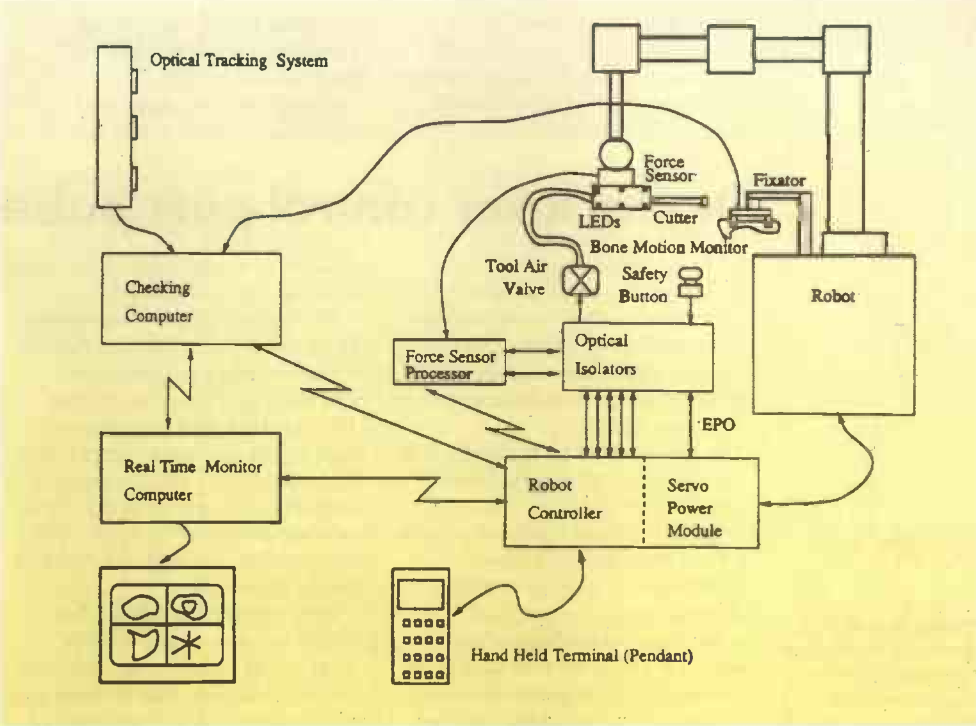
Robots have been used for limited tasks in surgery before. But the group says that the hip replacement application requires ten-times greater accuracy than other uses, while the shapes to be cut are more complex. Safety is more important and the working volume must be much less constrained.

For the patient, preparation for robot surgery begins prior to the operation, with the implanting of three titanium pins through a small skin incision into the thigh bone. A computer tomography scan is then made of the leg, and the pins are located relative to the coordinate system of the CT images. The surgeon selects a hip implant module and determines its position, using the CT data, which is written to disk for use in the operation. During the surgery, the sterilised robot is brought into the operating theatre and the patient data disk loaded. When the patient is ready, his or her thigh bone is fixed rigidly to the robot base and the three titanium pins exposed.

The robot then orientates itself using these pins and computes the transformation from CT coordinates to robot coordinates. Cuts can be made by the robot to produce the desired implant shape at the planned position and orientation relative to the pins.

The human surgeon monitors the robot both visually and by observing a graphical display showing successive cuts. When cutting is complete the thigh-bone is unclamped and the robot is moved out of the way. Plainly, when cutting into a human body, constant position checking and protection against machine failure is vital. But the researchers say there as yet have been no problems.

One of the main needs has been that the human must be in charge at all times, a complex requirement as the surgeon must also trust the system to some extent. Researchers say the system has worked well in surgery and the total surgery time was comparable to manual broaching. A future development could be addition of a head-up display that shows the surgical plan superimposed on the actual patient. It will surely be some time before such robots become part of the scene in normal NHS hospitals. Researchers will need to develop an integral coin slot and pay meter for a start.



Robodoc's participation in a total hip replacement marks the first time in US medical history that robotics technology has been actively employed in an invasive surgical procedure.

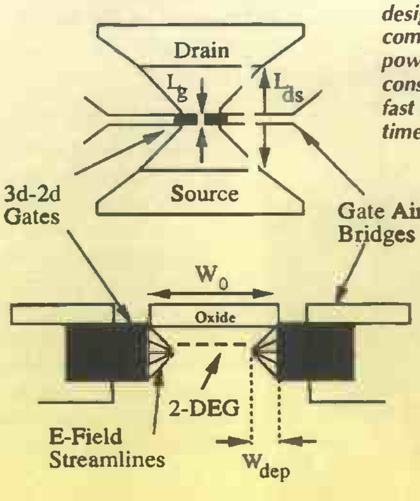
Mesfet redesign cuts power needs

Announcement of a 2d mesfet that makes big cuts in the power consumption/delay product could pave the way for greater vlsi scaling and longer battery life. Conventional fet power needs put an upper limit on vlsi size, while reducing power consumption will detrimentally affect switching speed.

But using a novel design of 2d mesfet, where opposing Schottky side-gates modulate channel width, WCB Peatman and colleagues at the University of Virginia (*IEEE Electron Device Letters*, Vol 15, No 7, pp. 245-247) say they have practically eliminated the narrow channel effect which limits the minimum power consumption in conventional fets.

The power/delay product of a fet is determined by the energy stored in the gate capacitor and is determined by the gate capacitance plus parasitic capacitance, multiplied by the square of gate voltage swing needed to switch between on and off. In the Virginia 2d mesfet, gate design is based on a lateral metal-2d electron gas (2deg) junction, having geometry very different to that of conventional fet devices. Junction capacitance of the 3d-2d Schottky diode is dominated by the direct contact to the 2d electron gas, and the parasitic capacitance is small compared to the junction capacitance.

Transconductance is said to be higher than achievable in conventional 1μm fets suffering



Unusual mesfet design that combines low power consumption with fast switching times.

from the narrow channel effect while threshold voltage and sub-threshold ideality factor are reported to comparable to state-of-the-art hemfets. Gate capacitance is estimated to be 0.8fF/µm per sidegate, or about half that of conventional hemfets.

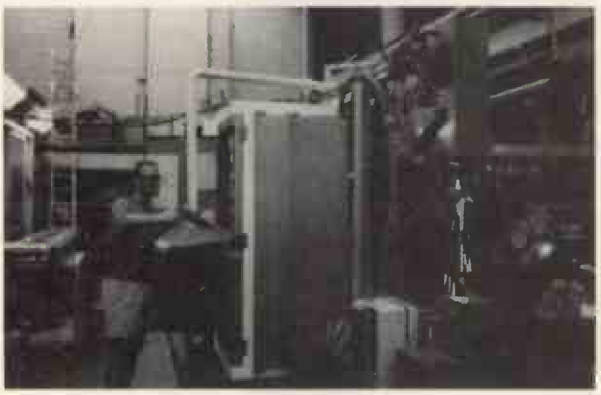
Overall, the device shows a significant reduction in power consumption without loss in speed, and the researchers believe it could have very good prospects for ultra low power circuit applications in the future.

Digital laser control puts pulse-power on site

Heavy section welding – expensive and traditionally carried out in-shop – could become a practical process for use on-site following a breakthrough in welding power obtained from cheaper, lower power lasers.

Researchers led by Stefan Scott at the University of Alberta report (*Applied Physics Letters*, Vol 65, No 3, 1994) development of a multi-kW cw laser demonstrating significantly improved welding properties over conventional technology. At the heart of the system is a CO_2 burst-mode PIE (photoinitiated, impulse-enhanced, electrically-excited) laser. The pie process involves applying dual-polarity 10kV photoionisation impulses coupled with high voltage dc excitation to produce a highly controllable large-volume discharge. A digitally controlled hydrogen thyratron circuit is used to produce the impulses. Penetration with the pie laser is claimed to be 50% better, and the weld profile far superior to normal cw welding at the same average power level.

Alberta researchers squeezing high-power performance from a low power burst mode laser welding X-70 pipeline steel.



Peak optical power is reported to be up to three times greater than cw operation, while the multi-kW average output power is retained. The drawback with conventional high-power cw laser systems is that they need large capital investment and can prove unreliable in constant duty applications. They also allow only average power to be controlled during operation.

Deep penetration welding has tended to focus on peak-power, short-duration, high-frequency laser pulsing. Although such systems give deep penetration, average laser beam power is low. Unfortunately, average power (along with peak intensity) is one of the main factors that makes economical welding possible. But the Alberta researchers say their unit is the first pulsed laser system capable of high peak power, pulse-periodic operation at multi-kW average power levels.

Continuous-wave operation is achieved using low-level digital pulse excitation of a pentode hydrogen thyratron pulser circuit while burst mode is obtained by manipulation of the digital trigger signal. Gating the cw excitation trigger produces basic burst mode operation. The project is built around a 30kW cw pie laser originally designed for application in the oil and gas pipeline industries.

At present the team can weld, using its modified system, 1m diameter pipe and straight sections up to 3.5m in length. The apparatus has also been used to blind-weld heavy plate onto structural members – as found on large oceanic oil tankers and military vessels. This marine application requires penetration of 0.75in to weld through the 0.5in steel plate onto a 0.5in web.

Alberta says the task was completed at an average laser power of only 9kW.

...but will it fly? Helicopter robots able to navigate their own way around an arena, and tracked robots able to manipulate objects on the ground were all busy pumping their servos at the Association for Unmanned Vehicle Systems annual aerial robotics competition. Unfortunately no one machine was yet able to combine both functions – though the AUVS says this day is not far off.

As usual the competition was held at Georgia Tech's Bobby Dodd stadium and Georgia Tech was among the seven different institutions fielding teams. First prize went to the students from the University of Southern California whose 'behaviourial-based approach' impressed the judges.

Mobile phones make for mean streets

No car-bore worth his wheels feels properly dressed without a dashboard full of in-vehicle route-finders and journey information systems. But are we in danger of giving too little attention to the actual mechanics of controlling the car?

A recent Swedish study showed that, despite what we think, we can't even make a hands-free mobile phone call without affecting our driving and, surprisingly, we are most influenced when the road conditions seem the safest.

The research was carried out by Häkan Alm and Lena Nilsson of the Swedish Road and Transport Research Institute. They tested the effects of using a hands-free phone on driver reaction time, lane position, speed level and workload. What they found (*Accident Analysis and Prevention*, Vol 26, No. 4, pp. 441-451), using a driving simulator, was that driving definitely worsened and, against expectations, driver reaction times slowed most when carrying out the easiest tasks. In the hardest conditions, only lateral position was affected. So if the map-reading phone-talking satellite-tracking auto-pilot speeding past at 100mile/h is just wobbling in his lane a little, don't worry. The time to be concerned is when he slows down...

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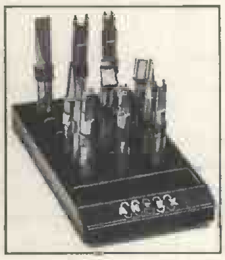

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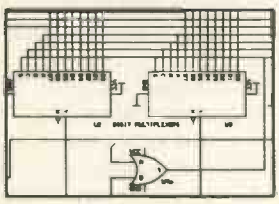

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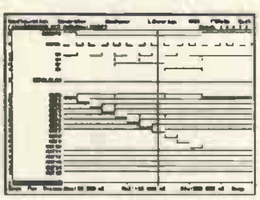

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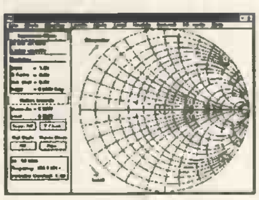

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Voltage regulator chips: switch to linear?

Ben Duncan measures and compares important, rarely documented regulator performance features through three generations of linear ICs and the latest micropower switching types.

An unsuitable choice of regulator can have repercussions that are more catastrophic and far-reaching than others. Even the best data sheets for linear regulators, from companies like Linear Technology and National Semiconductor, do not tell you everything; important graphs are absent and documentation has not progressed in years.

With switching regulators, there is even more to know, and yet less is graphed in proportion. This article charts important ac and transient domain performahce results that are sparsely – and decreasingly – charted by makers.

The line-up

For this evaluation, nine monolithic ICs were chosen, three linear devices and six switchers. For uniformity, all were configured to regulate to +5V. Since there is an increasing tendency to distribute regulation around pcbs[1], some low current parts operating at less than 1A were included, and testing carried out at both 105 and 225mA.

In several cases, both fixed and adjustable voltage versions of a particular regulator are available. Below, an asterisk indicates that the results shown are for an adjustable type.

Beginning with the linear ICs, the *LM340* represents the top grade of classic fixed type, while the *LT317* is a premium example of the adjustable variant that is almost as old. The less well known *LT1129** is a recently introduced low dropout (0.4V), 700mA part. It has a low quiescent current that is claimed not to increase when the regulator is unloaded. **Figures 1-3** show the test configurations.

All the switching regulators were chosen for their low external component requirements, **Fig. 4**. Ignoring the IC and input and output decoupling capacitors C_{in}, C_o,C_{o2}, the *LM2576* uses the fewest – just an inductor and a diode. The *L4962* requires the most, totalling seven, namely two resistors, three capacitors, one inductor and a diode.

Throughout, the switching regulators are differentiated by their current ratings. The first three are in plastic dual-in-line packages. Maxim's *MAX639* is a step-down switching IC rated at up to 225mA. Its current-limiting 'pulse frequency modulated' scheme yields

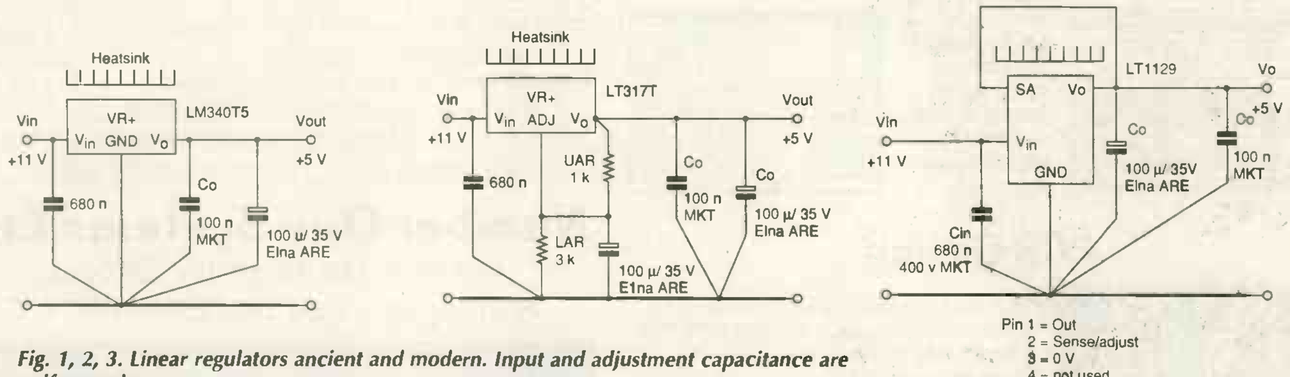

Fig. 1, 2, 3. Linear regulators ancient and modern. Input and adjustment capacitance are uniform values.

high efficiency over a range of loads[†].

Linear Technology's *LTC1174*[*] is a multipurpose switching converter. Here it is configured as the others are made: as a step-down (buck) regulator. Maximum output current of 600mA can be stepped down to 340mA by strapping. The integral switch is mosfet, quiescent current is 130μA and switching frequency is adjustable, up to a higher than average 200kHz.

Within the *LT1176* step-down switching IC is an integral 1.2A bipolar transistor switching at a nominal 100kHz. Response to voltage changes is speeded by using a multiplier in the loop.

SGS-Thomson's *L4962* is a 1.5A step-down switching IC in a Heptawatt seven legged T0220 packaging. It operates at 150kHz. While requiring more parts than others, it includes soft-start. At 50V, the input and differential voltage ratings are higher than any of the preceding ICs.

Finally, National Semiconductor's *LM2576*[*] is a similar category of device in a *Pentawatt* packaging. Operating at up to 63V in its *HV* version, this device is rated at 3A and switches at a fixed 52kHz.

Tests and application

Measurements are focused on graphical performance information readily obtained with a modern lf test set, but scarcely documented by makers, namely:

- Intrinsic noise versus frequency
- Ripple rejection versus frequency
- Spectra of ripple caused by abrupt, repetitive load change.

Switching IC data sheets make efficiency claims, but how often is the efficiency of a linear psu charted ? Being long overdue, a uniform assessment is included here.

Figure 5 shows the test circuit used for noise tests. **Figures 6 to 15** illustrate regulator noise. Getting a clean enough input voltage is the first stumbling block. Loading is stepped to reveal changes that can make a regulator manifest as a current-controlled noise-source. You too may be surprised at the disparity in behaviour patterns – particulary between the switching devices.

Figure 16 shows the ripple rejection test circuit used while **Figs 17 to 25** illustrate the results obtained. Part of the test circuit is a 20V rms audio power amplifier having extended hf response. This can handle a 4Ω load and is used to drive the test network.

A power amplifier is needed because R_{in}, at 47Ω, is seen as a load in shunt with the regulator's input. A higher R_{in} value would reduce loading but also drop the incoming lab supply, when loaded. As a result, it would need to swing above 30V in order to attain the 9V (excluding superimposed AC) required at the regulator input.

Capacitors of 10,000μ+1000μ+1μ in parallel provide resonance-free coupling into the 47Ω, contributing less than –1dB additional deviation from 10Hz to 200kHz. The protective zener was added after the more highly stressed DIL-packaged switchers with marginal voltage ratings were vaporised when loading was removed before reducing the lab supply input.

Figures 26 and 27 illustrate the group spectra caused by abrupt load switching. In test circuit Fig. 26, a mosfet is driven with a 10kHz square wave with roughly equal mark:space ratio. This in turn switches the 22Ω load.

Finally, **Fig. 29** compares efficiency. Each

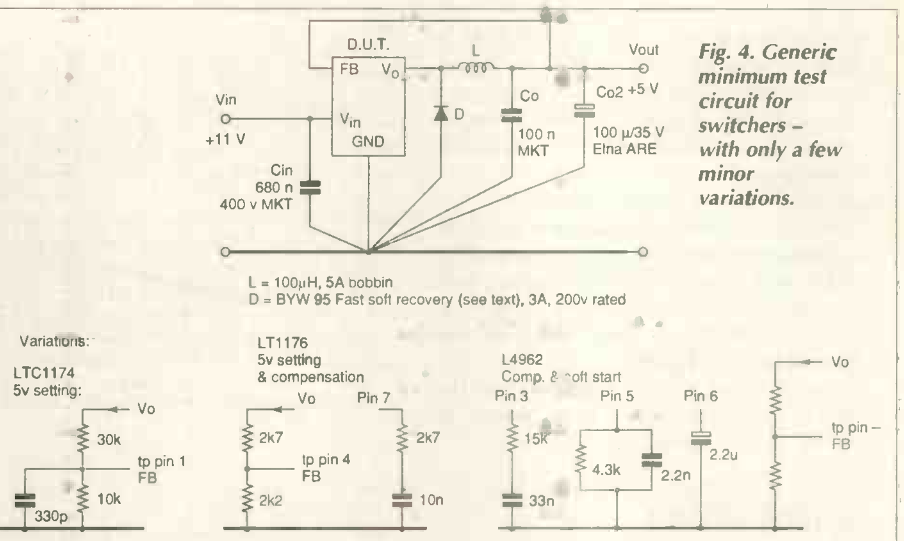

Fig. 4. Generic minimum test circuit for switchers – with only a few minor variations.

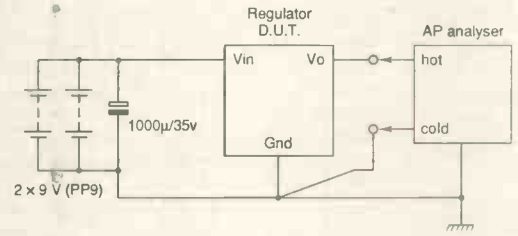

Fig. 5. Noise test network. Zinc–carbon batteries are used as a short–lived, ultra–low noise power supply

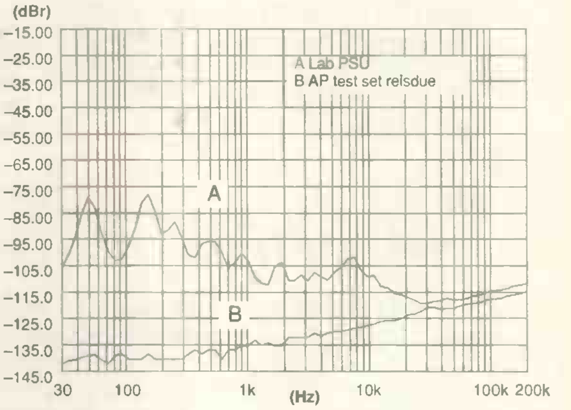

Fig. 6. Ideally, measurement of regulator's output noise should not be affected by incoming noise from the dc (raw) voltage source, but it may be, with all of the regulators having less than infinite and perfectly uniform ripple rejection. The upper plot (A) shows a Thurlby lab supply, considered good enough to test sensitive circuitry, measured at the end of 300mm twisted cabling. It helps to recall that –100dBr is 10μV rms and –60dBr is one millivolt. Mid-band noise can be reduced by decoupling the psu output with low inductance elcaps >1000μF, but below 10kHz any sensible array has little useful effect. The lower curve is the AP residue. A pair of 9V primary batteries paralleled with a 1000μF Elna Low–L capacitor yielded a plot identical to this, and were duly adopted it as a 'virtually noisefree' DC supply for measuring the intrinsic noise.

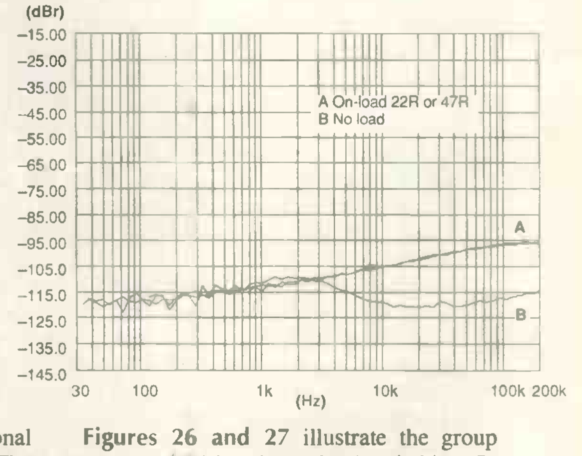

Fig. 7. The 20 year old LM340T-5 has equal second lowest noise, exceeding –95dB in all the tested places. Noise character is smooth. Note how noise is identical with either loading (A) yet how much the unloaded noise droops away above 3kHz (B).

0dBr=1 volt in all graphs

[†] *Maxim's MAX639 was omitted from one of the tests as the initial samples expired readily and a working replacement was not supplied. A process problem may have been the cause.*

regulator was driven at 10V so its burden is about 5V. Loading was kept close to 50% of the rated value using off-the-shelf resistors.

Average (not true rms) input and output currents and voltages were then measured and waste computed from:

$\%\ \text{efficiency}=[(V_o \times I_o)/(V_{in} \times I_{in})] \times 100$

There are surprises. First, efficiencies of the competent switchers converge at around 70%. This is better than for the linear devices, with V_{in} being twice V_o. But remember that if V_{in} is set much closer to V_o, linear efficiency can rise to at least 70% too.

Secondly, the dismal efficiency of the *LTC1174* was confirmed with retests at slightly lower current and after a

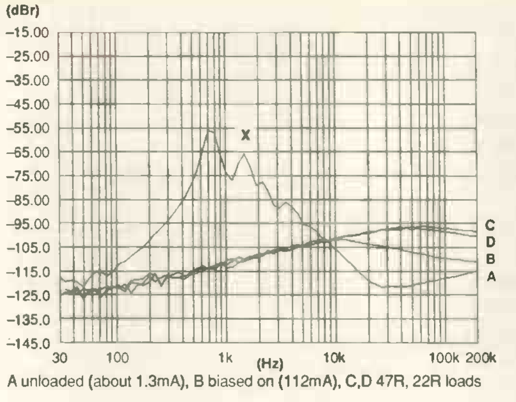

Fig. 8. With an illegally low quiescent current of 1mA, the LT317 'rings' at 850Hz, albeit 55dB down [see ref. 3 for explanation]. From the graph, a strong second harmonic (X) is also evident. A strange tonal-noise modulation effect would occur in unfortunate circumstances where current draw dips to very few mA. When the LT317 is arranged to deliver a minimum of 5 to 7mA, the tone vanishes, and is replaced (middle curve, B) by a smooth white noise that becomes pink noise when viewed from a higher frequency perspective – as noise density is dropping off above 10kHz. At higher currents with the 47Ω and 22Ω loads, audio band noise is unchanged but notice how the hf hinge is shifted up to 65kHz, while noise is about 10dB higher 1.5 octaves either side. Noise character is all-round smooth, as befits a bandgap.

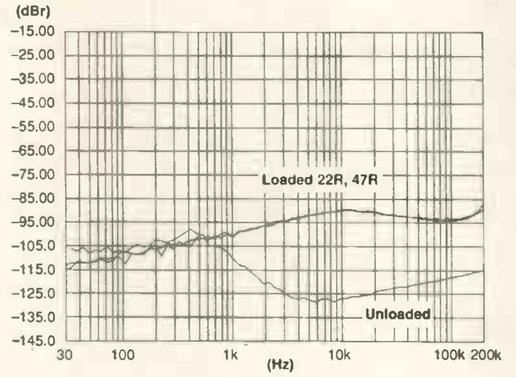

Fig. 9. Listening through the 1/3rd octave sweep, the LT1129 produced the classic, psychologically disturbing noise (sounding in the mid-band when put through speakers like a deathly Antarctic wind) of a crude zener reference. Lower plot shows noise when unloaded. Note same increase in hf noise with loading irrespective whether this is 47Ω or 22Ω, indicating nil noise modulation over this span of current.

Vapid silicon

Both the *MAX639* and *LTC1174* proved instantly destructible by exceeding idiosyncratic voltages (around 11 to 13V) that may not ring alarm bells in analogue design heads. The rejection test network's series resistor R_{in} (**Fig. 16**) causes a voltage drop when running loaded. It is easily compensated for by jacking up the input voltage, but had deathly consequences for these chips when the test load was removed even momentarily. This loss prevented the plotting of the input drive level when in the loaded condition. IC designers should think more clearly before making parts with such arbitrary and low breakdown voltages.

Noise and layout

Even for instrumentation, listening to noise is one of the quickest ways to evaluate its characteristics. One would expect linear regulators to be intrinsically quieter than any switcher. The results show this is mostly true. Excess noise in linear regulator ICs arises mainly from the reference. It has been long established that regulators using plain (cf buried) zener references have a subjectively "gross" noise character. Bandgap-referred (as well as the more modern buried zener-referred) regulators are both measurably quieter and have a smoother, more unobtrusive noise character. Peak noise voltages are up to at least 10 times (20dB) higher than the rms levels plotted. When supplies are bussed-about this may couple into a critical node. **Fig. 6b** amply illustrates why high-end audio perfectionists might dispense with ac mains and the regulators, and opt for cupboards full of car batteries.

All IC regulators demand considered layout. The older fixed types readily oscillate at rf and can even burn out if driven from a distant source without local and quite wideband decoupling. In all test circuits, C_{in} was 680nF low ESL MKT, placed less than 10mm from the IC legs. Most linear regulator ICs also require typically a minimum of 100µF of output decoupling. The *LT1129* is exceptional, being intelligently designed to be stable with C_o of under 10µF. However, above 1kHz, the Z_o and transient response of all this and all other linear regulator ICs employing voltage feedback is increasingly dependent on adequate C_o.

In all test circuits (**Figs. 1 to 4**) the stricter standard of a low ESL 100µF combined with 100nF low ESL+ESR reservoir capacitors which are mandatory for switching regulators was adopted uniformly throughout, so output decoupling has no appreciable part in performance differences. All 100µF elcaps were matched within +2%. Adjustable linear regulators require the sensing resistor to be connected to output pin, but a sample of the output current is not required, and a wrong, non-starred connection degrades Z_o and the transient response.

Turning to switchers[2], we face incisive waveforms with plentiful harmonics. As well as radiating noise, many switchers depend on comparators, and these require robust hysteresis and appropriate filtering so that locally generated noise does not upset the feedback loop(s). As a switching regulator is so easily upset by its own hash, it pays to be kind to it and the environment at the same time, and design the layout for low noise and precision. This is mainly achieved by compact placement, fat star grounds, low inductance, preferably paralleled capacitors of at least two widely spaced values, and the use of adequately rated inductors, preferably toroidal types which radiate least.

With switchers like the *LTC1174* working up to 200kHz and higher, subtler techniques including steps to forestall eddy-currents such as use of Litz wire (plaited conductors) to balance copper losses will be significant. For lowest noise, output (including return) must be taken directly across the output capacitor. **Fig. 19c, Fig.24d and Fig. 25b** are plots demonstrating the results of misconnecting the analyser's cold input just a few inches up the 0V wire back towards the power source, instead of coming off the star ground separately.

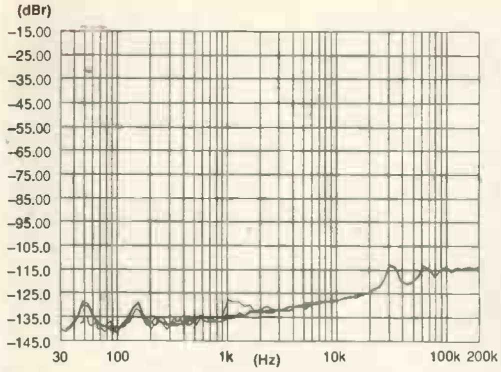

Fig. 10. The prototype of my high power audio regulator [see 4; and discussed in recent issues] was dragged out to show the kind of thing that audiophiles find improves their ability to hear ambient cues and other nuances in recordings: Irrespective of loading, this regulator's noise is indistinguishable from the AP residue in my lab's environment. If anything, the two AP plots are actually the higher of the four here.

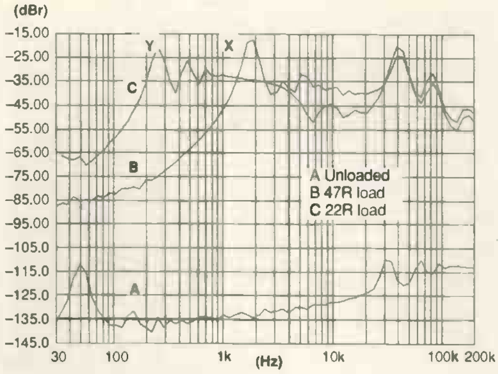

Fig. 11.below The MAX639 is most commendably quiet when unloaded (lowermost), except for 50Hz reception spike (left). Loaded noise is far higher (upper curves). With the 47Ω load, the noise character is rough (like the fixed linears), and curiously includes a 1.8kHz tone (peak at 'X'), as in Fig. 8. With the 22Ω load, switch artifacts are clearly audible; the third octave sweep sounds like a swarm of bees! The bee sound is imparted as the tone spike (at 'X') has shifted down to 250Hz (Y). Overall, noise with this higher loading is unchanged in the decade above 25kHz, is slightly less down to 5kHz, and below 1kHz, 20 to 50dB higher.

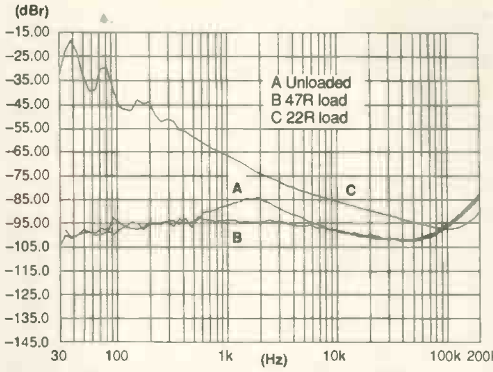

Fig. 12. The LTC 1174 (C for CMOS) switcher is contrastingly noisiest (upper curve) when unloaded. A 2k4 fixed load resistor was added to set a more realistic quiescent 'unloaded' current of 2mA. The peak point indicates an unloaded PRF of 40Hz. Measured noise with 47 and 22 ohm loading is almost the same (A,B). But A's peak suggests a low Q version of Fig. 8's resonant phenomenon, about an octave either side of 1.7kHz.

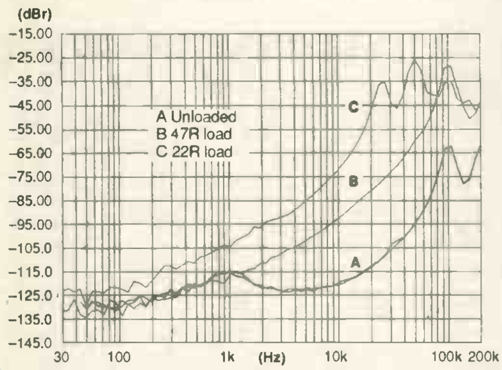

Fig. 13. The LT1176 has commendably low noise below 1kHz, irrespective of load condition. The lowermost curve (A) shows the 100kHz switch frequency, rejected by at least –60dB. The middle and upper curves are for loads of 47Ω and 22Ω respectively. Note that noise increases 20dB for a just over twofold rise in current well away from the maximum current rating of at least 1.2A. The change in noise (ie. noise modulation) might disturb sensitive circuitry drawing discontinuous current. At least noise character is uniformly smooth.

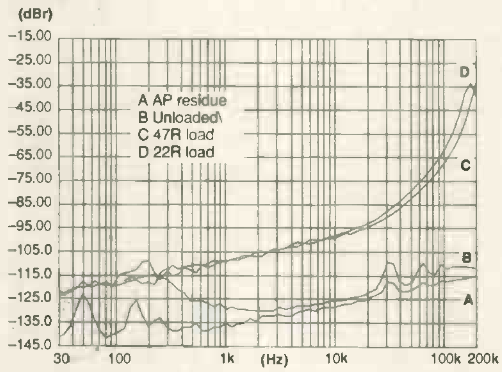

Fig. 14. SGS's L4962 was measured in a later session, so the AP residue was replotted (lowest, A). Compare this to Fig.10. When unloaded, the L4962 is just above the residue and remarkably quiet above 1kHz. Even below 1kHz noise is good for any switcher, at <–105dB. When loaded (47R, curve C), noise rises markedly around the switching fundamental to a –35dB minima. Past this point there is little noise modulation – shown by the negligible change with the 22Ω load (uppermost, D). Noise character is truly excellent for a switcher – as smooth as bandgap linears.

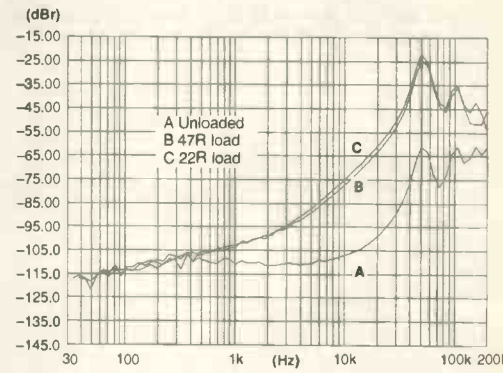

Fig. 15 The LM1576T when unloaded, was as noisy as the L4962 when the latter was loaded. Still, the noise character is similarly smooth. The unloaded plot (A) is quite good in the audio band, but above 20kHz, a train of harmonically related spectra occur, harmonics made visible because this regulator's fundamental is about an octave or two lower than the others at 50kHz. When loaded, the fundamental peaks only 23dB below 1V, and noise is barely changed between the 22Ω and 47Ω loads.

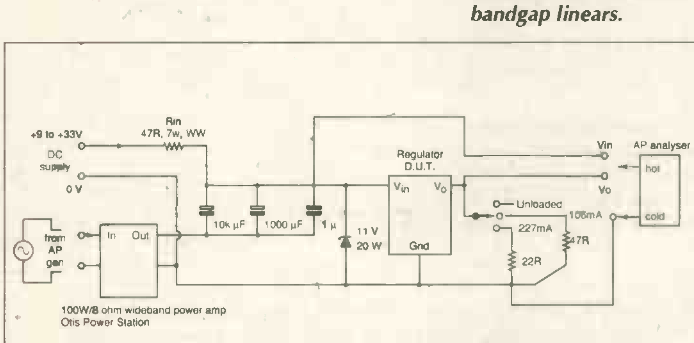

Fig. 16. The Rejection Test Network. Although ripple is fundamentally at power line frequency, one cannot just test at 100/120Hz! The capacitor array, power amplifier and high wattage resistor allows tens to hundreds of milliamperes of incoming DC at 9 volts, to be mixed with 1V rms of sweepable AC from the Audio Precision's generator, with uniform response over 13 octaves (10–200kHz) being preserved at the regulator's input. DC and ac levels have to be finely set to avoid clipping the ac, or overstressing the marginally rated switcher chips; or dropping out on the longest (20Hz) ac peak dip. Any ac appearing at the regulator's output is feedthrough. The amount varies with loading, as open-loop gain is depressed, eg. by beta loss with increasing current.

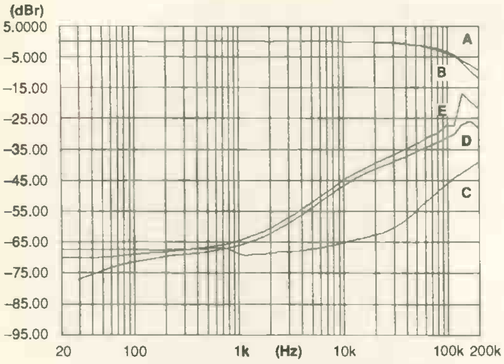

Fig. 17. In this and all the subsequent rejection graphs, the upper plot shows the incoming, 1V swept 20Hz to 200kHz swept test signal as applied to the regulator's input. Much of the rolloff above 30kHz, to –12dB at 200kHz, arises because the drive capability required is borrowed from an audio power amplifier, one of the few with extended ultrasonic response to 200kHz. The incoming signal has been plotted with the regulator's output both unloaded (A) and loaded (B), to confirm that the drive reference only changes slightly around 130–200kHz. Rejection is best when unloaded (C). When loaded with 47 and 22 ohms (D,E) rejection improves slightly below 200Hz but reduces markedly above 1kHz. Note also how rejection barely changes (mainly below 200Hz) with the more than two fold load increase, and how the heavier loading (E) has the best rejection of all below 1kHz, better than the unloaded case.

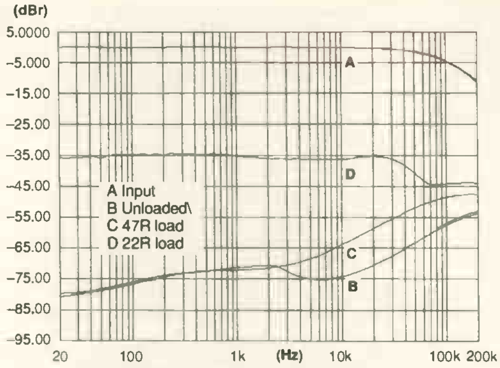

Fig. 18. The LM340 still sets standards, at least at low currents. Whether unloaded (B) or loaded with 47Ω (C), rejection below 3kHz is the same within tolerance and at least –70dB. Above 3kHz, the loaded condition is consistently about 12dB less good but still manages to exceed –45dB. With the heavier (22R) load, rejection is degraded to a less healthy –35dB uniformly with frequency (D). The 30kHz step–down is curious; answers please. Note all three load conditions show identical input levels (A, uppermost), suggesting the input is well buffered at hf.

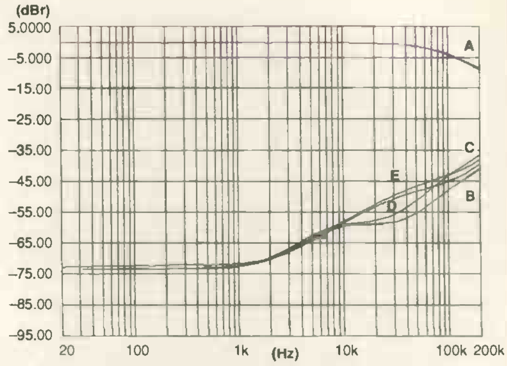

Fig. 19. With the LT317, loaded and unloaded rejection is identical below 10kHz. Above, the 22 and 47Ω load conditions (D,E) are no more than 10dB worse than the best unloaded case (B) at 30kHz, re–converging above. Curve C shows the effect of the unloaded case with a bad ground connection; here the analyser's 'cold' input was coupled to ground several inches back from the output capacitor towards the supply input.

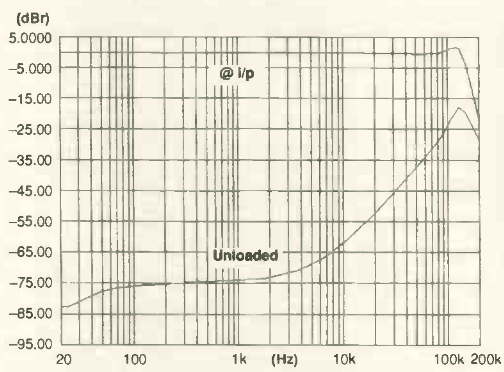

Fig. 21. The MAX639's unloaded rejection is excellent below 3kHz, being better than –70dB. But rejection decay sets in markedly above 10kHz and well before the 125kHz switching frequency. See footnote on p807.

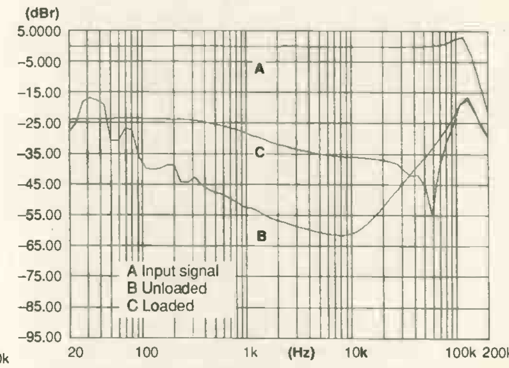

Fig. 22. Unloaded, the LTC1174 has marginally good (>–40dB) rejection between 200Hz and 20kHz, but both modestly loaded (47Ω) and unloaded, it offers a paltry less than –30dB of rejection at line (50/60Hz) and switching (>50kHz) frequencies. "Unloaded" includes the small 2mA bias established earlier (Fig.12). Notice also that the upper, reference plot of the incoming test signal is peaking slightly just above 100kHz, a sign that Z_{in} is affecting the driving amplifier's stability.

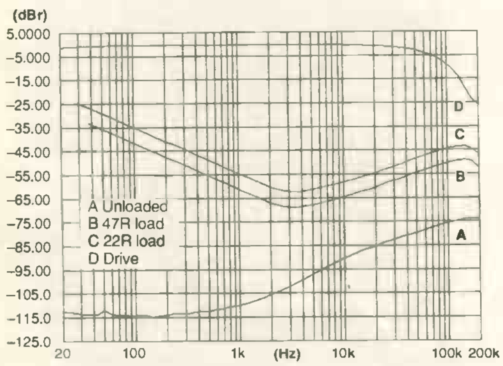

Fig. 20. High power reference regulator for audio. The input shows about 10dB extra attenuation at 200kHz (D), showing imperfect buffering at hf. Unloaded rejection beats all the regulator ICs. Rejection degrades greatly with slight loading, most markedly at LF, where the outcome is least audible. Note the psychoacoustic tuning; after having accepted decay with loading, deepest rejection has been tuned to the ear's most sensitive region, about 3.5kHz. Loading was the same 105/225mA as other regulators. Further up the scale of this regulator's far higher operating capacity (up to 35 Amperes) rejection improves, where again it matters most.

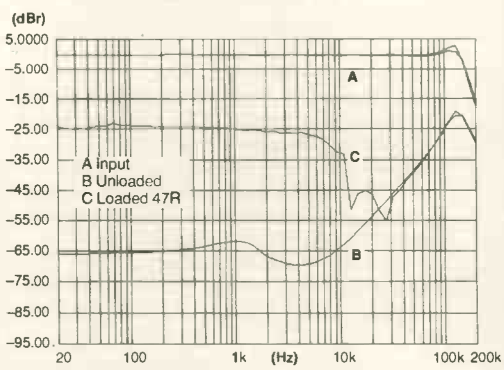

Fig. 23. LT1176 has quite good rejection in the audio band before loading, but with 47R, rejection decays to an unimpressive –25dB at 6kHz. With the input drive (uppermost), note the peaked–up then steeper rolloff at 120kHz, the switching frequency.

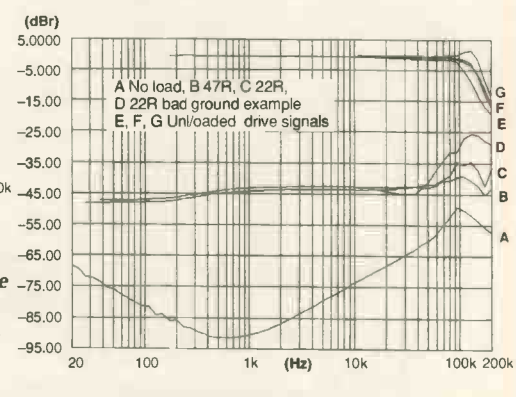

Fig. 24. (Bottom right) Unloaded, the L4962 displays its best rejection around 600Hz, with a negative resonance. When loaded, rejection is a more modest 45dB but is uniform with frequency and load up to 50kHz. Noise at the switching frequency is also at least 15 and up to 30dB lower than any of the other switcher ICs. It is clearly amongst the cleanest switchers. Above 100kHz, the loading effect on the source is slight but more varied than others. Curve D shows the same bad ground connection as in fig.19 (C). Notice how the result is less catastrophic.

cool-down period. Thirdly, you might imagine that the *LT1129* would be slightly more efficient than a 'normal dropout' regulator with emitter follower output. But this is clearly not so for real-world test condition, where V_{in} is higher than it might be.

All the switching regulator ICs were successfully applied by reference to the data sheet alone. And all four switching device makers offer some in-depth advice on critical layout and critical component specification for optimum performance. Linear Technology's *LTC1174* data sheet had the most explicit physical layout recommendations while National's showed the IC internal workings most clearly in relation to the outside world. Physical layout was on five-node analogue *Veroboard* (RS 433 911). Positioning of major critical parts was organised for shortest lead lengths, then repeated within ±3mm for each different device.

Conclusions

Beginning with noise, the benchmarks are set by my own regulator circuit. Similar op-amp-based regulators after Sulzer[5], set even higher or similar standards as do ones yet to be published by Walt Jung[6] which I have had a chance to consider.

For the quietest regulator with the best rejection and lowest output impedance, you need to look at ICs other than those labelled as regulators. On the other hand, while these op-amp based designs are not expensive against performance, they will cost many times more than an IC like the *LM340T*, which is typically under 40p in bulk. The more discrete design will also occupy more space.

Returning to explicit regulator ICs, easily the all-round quietest are *LM340* and *LT317*. In this instance, the latter is slightly quieter below 1kHz, and the former above. With the switchers, *L4962* is the clear leader, staying below –95dB up to 20kHz under the three test conditions. Considering just noise in the audio band, it comes close to equalling the *LM340*. *LM2576* is next best. Such a pattern suggests that switcher noise can be curtailed by using an IC with plenty of reserve current capability.

As for "ripple" (really broadband ac) rejection, the *LM340* exhibits the best figures until it falls apart – at a load current of one fifth of its 1.2A rating (**Fig. 18 curve D**). The *LT317's* rejection hardly varies with load current, but all load conditions share the earlier onset of decaying rejection above 1kHz. High frequency rejection might improve with improved adjustment pin decoupling – not an option with *LM340*.

The *LT1129* is again a slight backwards step from twenty year old technology. Only two of the switchers shows rejection across the range of loading that is remotely acceptable for plumbing around a sensitive analogue circuit. Again, this is SGS's *L4962*, with its near uniform loaded rejection of –45dB, ie. 6mV rms per every 1V rms of ripple sawtooth, with *LM2576* again not far behind. In a real application, the hf (>10kHz) rejection of the switchers will likely be improved markedly if inputs are EMI filtered.

Faced with abrupt load switching, and taking the ripple at fundamental as an indicator of output impedance (Z_o), then the familiar linear duo *LM340* and *LT317* have the lowest Z_o and maintain the overall cleanest supply, but are still perturbed more than 22dB compared with the BDR linear benchmark regulator.

The *LT1129* is by contrast fundamentally worse than even the most perturbed of the switchers, with a closed loop, loaded Z_o that is 6 to 11 times greater than the follower type outputs.

Test conditions

The tested parts were either fixed 5V models or if adjustable, were set to output +5V. As the ICs' maximum rated load current varied from 225mA upwards, and the surprisingly puny PP9 batteries used for noise testing could not support much more than 225mA on–load tests were performed with 47 and 22Ω, 2% CF load resistors, drawing a nominal 105mA and 225mA respectively.

Note: Throughout the following text and graphs, 0dBr = 1 volt rms. –60dBr = 1mV. –120dBr = 1μV. "Unloaded" means, unless stated, that there is only the load of the test equipment – typically under 100μA.

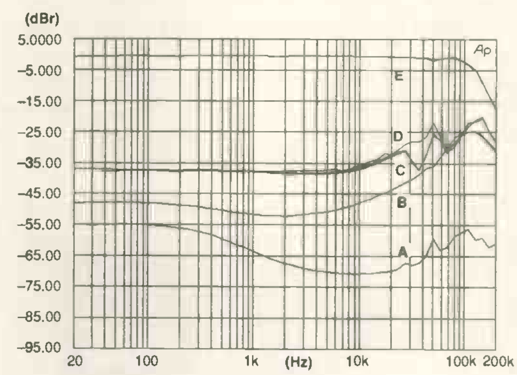

Fig. 25 Unloaded, the LM2576 is quiet and commendably so at ultrasonic frequencies. Again, the outcome of poor grounding practice is shown (B), this time with marked effect. Loaded response (C,D) behaviour is akin to the L4962, only not quite so good by some 7dB.

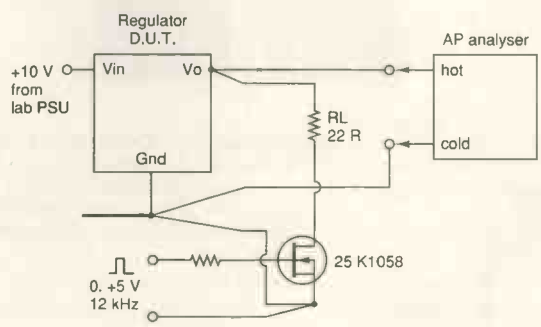

Fig. 26. Abrupt load switching test network. A mosfet switches the 22Ω load in and out at about 12kHz with a consistent (though approximate) 50/50 mark-space ratio. The test set is then set to plot the third–octave spectra above 10kHz to 200kHz. If Z_o were zero, or the feedback were instantaneous and the slew and loop gain infinite, there would be no spectra. The spectral levels give a proportional indication of each regulator's averaged, dynamic output impedance – the product of the static Z_o and transient response.

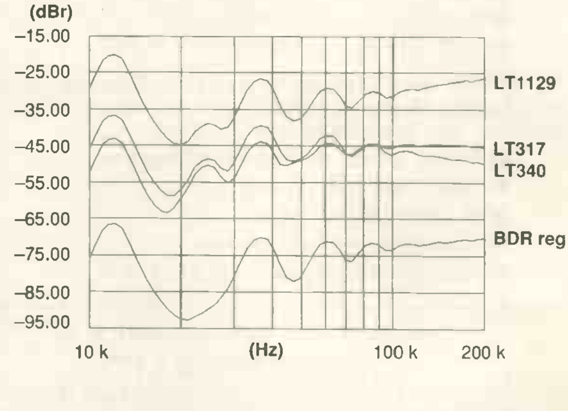

Fig. 27. These abrupt load switching spectra are a new way of looking at a related characteristics simultaneously: Voltage and current step as well as Zo and slew induced perturbations mapped into the frequency domain.
The Audio Precision spectra show how a 12kHz clock (with clean rise and fall of about 1μs) driving a rather abusive 22Ω load would appear on the supply rails at the regulator output. In other words, the results are based on direct, optimum noding. The performance on real pcbs won't get any better and will likely be worse if the clocked supply conductors even remotely share with any other currents.
Of the linears, LT1129 has the highest spectra indicating higher impedance and inferior damping. The bandgap twins, LM340 and LT317 have at least 14dB better performance. LM340 is up to 4dB better than LT317 below 60kHz and vice–versa above 100kHz. My high current audio regulator has the lowest artifacts, up to 30dB below the best regulator ICs. Note how each regulator follows a recognisably similar set of inflexions.

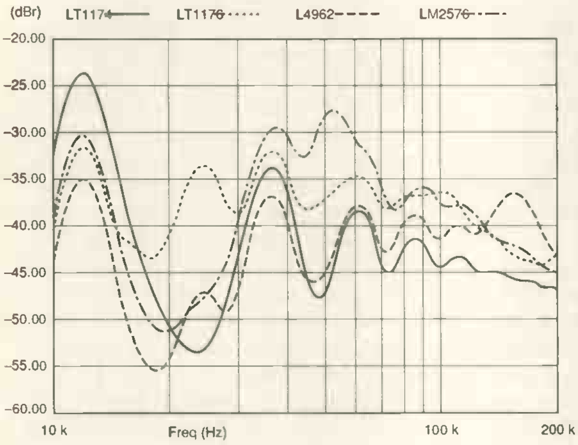

Fig. 28. Turning to the switchers, the inflexion pattern is devoid of correspondence with the exception of the fundamental. At this point (12kHz), the order of performance is similar to the feedthrough ranking, with LTC74 the clear loser (fundamental 24dB below 1V) and L4962 the clear winner. Yet, by 200kHz, the chaotic traces have exactly reversed the positions. Averaging by eye over the bandwidth displayed suggests that (i) the L4962, followed by the LTC1174, will maintain the cleanest rails, (ii) the LT1176's rails will be the noisiest, followed by the LM2576. The MAX639 does not appear; see below.

Efficiency – a score card

Efficiency,	at 50% rated load	
LT 1176	71.5%	576mA
L4962	70.5%	710mA
LM2576	69.0%	1.4A
LT317	52%	710mA
LT1129	49%	333mA
LTC1174	23%	215mA

V_{in} 9.5V ±0.1V; V_o 5V ±0.1V

As for the other switchers, the by now familiar *L4962* is least perturbed at the fundamental, but still near three times (10dB) more than the *LM340*. Note the wide, individualistic swings in ripple above the fundamental of all the switchers other than (oddly) *LTC1174*.

On the basis of these results, regulator IC ac/transient performance has not improved since Linear Technology's first efforts 11 years ago.

Low dropout regulators, while a boon for battery systems, should be used with care in sensitive circuitry powered off-line as their ac (20-200kHz) rejection and output impedance (hence transient line and load regulation) are both inferior to normal dropout parts.

Analogue circuitry employing switching regulators may require considerable rf filtering and stout decoupling. Even if EC or other EMC regulations do not make this mandatory, hash pickup may well affect performance. In this case, the cost of fixing this (in money, space and weight) may exceed the cost of using a linear regulator with a slightly larger heatsink from the outset. In summary – and with the possible exception of applying SGS-Thomson's SMPS chip – *before, while and after switching to switchers, designers must perform regular reality checks!*

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3. E Dietz, *Understanding and reducing noise voltage on three-terminal voltage regulators*, EDN (USA)
4. B Duncan, *PSU regulation boosts audio performance*, EW+WW, Oct '92.
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The author wishes to acknowledge the assistance of Anzac, Linear Technology, Macro Marketing and Maxim. ■

PC
ENGINEERING

From capture to layout

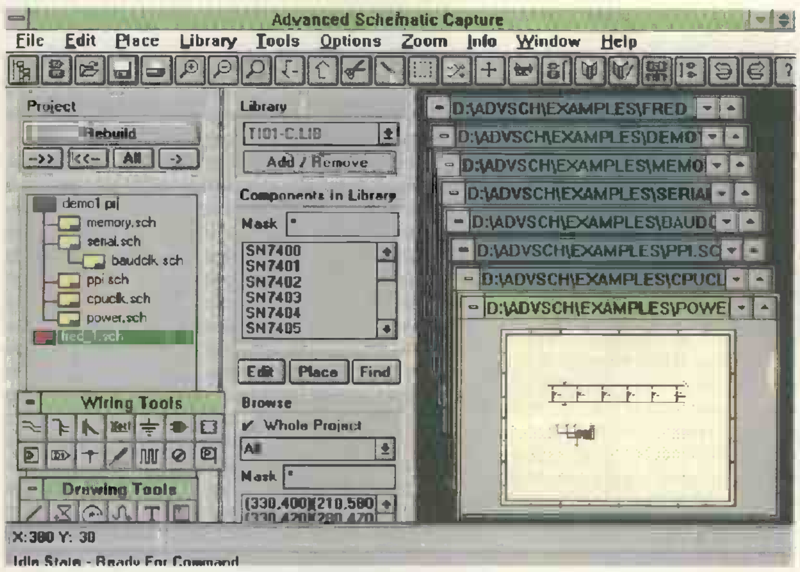
Protel's new pcb design and schematic capture packages both have the prefix 'Advanced', but with so many excellent products on the market how do they compete? John Anderson investigates.

The company known for *Easytrax* and *Autotrax*, Protel, has recently launched *Advanced Schematic 2*. In addition to providing schematic capture, this package executes front-end tasks for *Advanced PCB* – a new pcb design tool covered later in this review.

First launched as *Schematic for Windows*, this product retains the same object orientated editor, but adds many new features. These include library searching, drag and drop editing, and guided wiring.

The software is supplied with professionally produced, comprehensive user, reference and library reference manuals. In addition, there is a strange document entitled the *Environment Guide*, and a software-protection dongle plugging into the parallel port. Installation follows normal Windows procedure, involving a 'set-up' program which unpacks the files and installs its own group in file manager.

On initial start up, you are taken to an interrogation screen where you are prompted for access codes. These are codes that unlock specific features of the software – the review version had an eight-digit code for the schematic capture module.



The Normal operating screen! Note that this picture is captured in 640x480 vga format and the actual editing screen size increases with higher resolution formats.

Editing

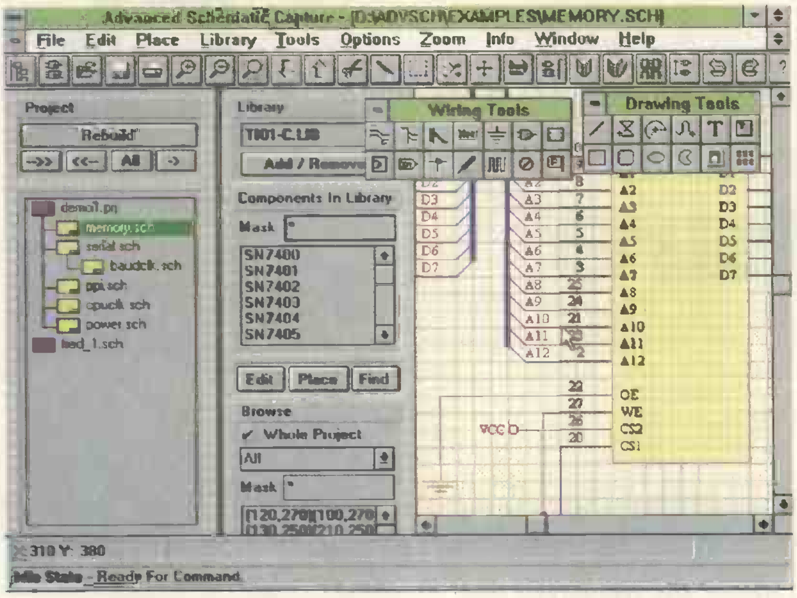
The system requirements state that a minimum screen resolution of 800 by 600 is needed. This type of specification is quite unusual and it was not until I ran the program that the reason for this requirement became apparent. There are two dialogue panels, namely the Component Browser and Project Manager, which in standard vga take over half the editing screen area. These can be turned off, but realistically, because the Component Browser is an essential part of the software operation, it needs to be kept on all the time.

On-screen working space is made even smaller because of two floating toolbars carrying wiring and drawing tools. The distinction between wiring and drawing tools is important, and obvious – except for the icons. In particular, because both line types have similar default colours, it is easy to inadvertently select a drawing rather than the wiring tool, and hence fail to connect up the components electrically.

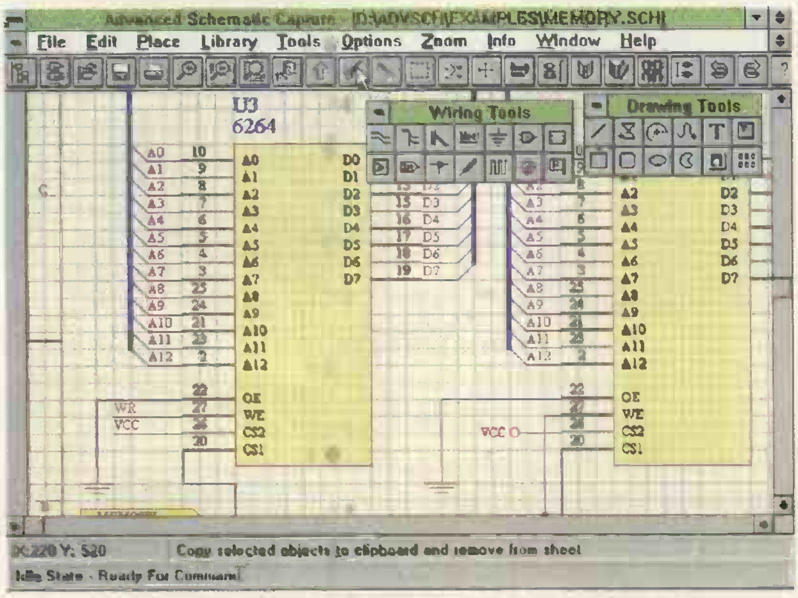
Capture works as follows. After selecting the component browser, the user selects which of the standard libraries to use via the Windows 'add feature' method. Back at the browser, any of the libraries in the selection list can be chosen. Once a library is chosen a list of components from that library is displayed. A specific component may then be selected and moved onto the drawing sheet.

This format of selection and placement seemed to work well, although the method did have weaknesses. In particular, if a multiple element component is selected, for example a 4-by-2-input nand gate, the placement system described above always places the first gate of the four, and you need to undertake a specific subsequent task to edit the gate identifiers or use the toggle part number icon to select any of the remaining three gates.

Editing facilities work well with the Windows clipboard, allowing selected items to be moved to the clipboard and then pasted to a new sheet. If the items are copied to the



Zoom in and the placement grid is clear.



There is much more room if the project manager and component browser are off.

clipboard then pasted back into the same drawing, no prompt or correction is made for the duplicated identifiers. Again using the clipboard, items clipped can be moved to other applications in Windows '.WMF' format – a nice touch for desktop publishing of technical manuals.

Moving around the sheet is somewhat awkward. This is mainly because the automatic panning is rather limited and only becomes available once a editing function is selected. The alternative is to rely on Windows scroll bars.

There are several files that can be generated from the schematic. There is a bill of materials, in tabular and comma separated variable form for spreadsheet or database use. There is also netlist output for interfacing to the pcb cad package, which may be output in any one of over 30 different formats.

Most important is the electrical rules check, or erc. This feature reads through the schematic database, generating a list of rule violations. Examples include multiple components with the same identifier, unconnected lines and floating input pins.

Rules for the erc are set in a user programmable matrix of errors and warnings. As an example, connecting output pins together would normally be an error but leaving input pins unconnected may be considered warning. The designation of pin type is set in the library edit facility.

Project control

Facilities within Advanced Schematic for project control are excellent and intuitive. Any number of sheets in a project hierarchy can be connected, and then whole projects loaded automatically. Each sheet is available at a click, which moves the selected sheet to the top window of a cascaded window stack.

Selecting the Library Editor from the Schematic Editor library menu results in a program running in another window which looks very similar to the schematic editor. However it automatically loads and decodes the currently selected library in the Library Browser.

As well as adding bit maps in any one of a variety of forms from '.WMF' and '.PCX' through to PostScript, the editor provides all the normal vector drawing facilities. The list of standard libraries is impressive, with a total of 76 libraries in all amounting to over 12,000 components taking over 12M-byte of disk space.

Compatibility

A level of compatibility is maintained with the earlier Protel dos based schematic program, with some of the libraries arranged in a similar multiple vendor form.

Files generated by the earlier dos product are loaded by the editor, but a warning is provided that some components are converted from a bitmap form to vector form. In practice this did not seem to cause any problems and existing designs loaded without incident.

Electrical rule checking of the new software is better than that of the dos product. This can result in errors being reported in dos-based schematic designs which had passed the equivalent report facility in the old product.

When the schematic editor and the companion *Advanced PCB* editor are open at the same time, it is possible to cross probe – that is, select a part in the schematic and then jump automatically to the corresponding pcb component. This works in the reverse direction, perhaps suggesting that these two products might have been sold as one.

Back annotation from the pcb to the schematic, sometimes called the 'was-is' function, is supported. Forward annotation, where changes in the schematic are transferred to the pcb, is also provided. In this case, annotation reflects all netlist changes through removal of obsolete tracks and component footprints – dangerous!

Advanced Schematic has routes to interact with other programs. There is, as you might expect, the direct access icon to the *Advanced PCB* product, but direct execution of analogue, digital or mixed signal simulators is also possible. However, these are not supplied with the package and the level of support is little more than that of launching another dos or Windows program.

One exception to this is support for the analogue simulator, *S*. The software produces *Spice* compatible net lists, together with an ascii text input facility to add other *Spice* commands to the *Spice* control. These might be, for example, generator frequency or simulation parameters.

Other than *Spice*, the Advanced Schematic outputs for the somewhat obscure *EEsof* and *Touchstone* simulators.

Advanced PCB 2

This is an upgraded version of the Windows-based integrated pcb layout program that Protel released two years ago. In addition to updating the package, Protel has also integrated *Advanced PCB 2* with *Advanced Schematic 2*.

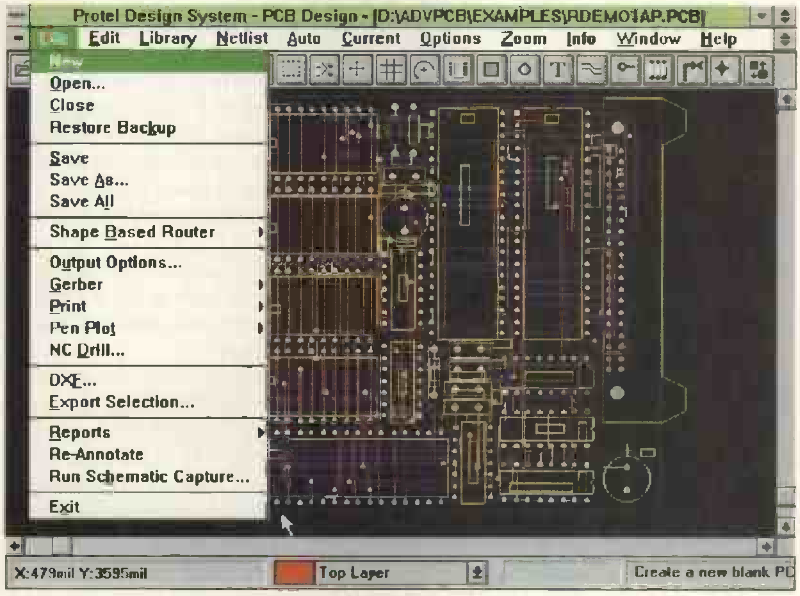
Compared with the earlier Protel for Windows, this new package has 123 new functions and features. Although some of these are little more than corrections of problems with the earlier version, some of them are valuable. There is a split plane feature which allows a net to be assigned to one or more copper planes for example, and a preview-mode display allowing quick scanning between layers, displaying only one at a time.

The package is supplied with comprehensive user and reference manuals, an *Environment Guide* and a parallel port dongle. Installation follows the normal Windows procedure, involving a 'set-up' program which unpacks the files and installs its own group in file manager. It also requires modification of autoexec.bat.

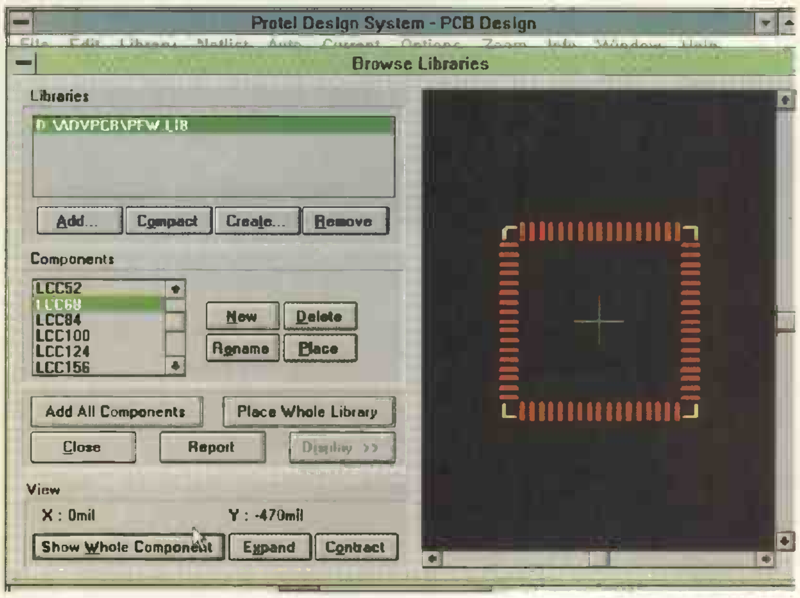
Both manuals are well produced and include a command reference and user guide. However the package is straightforward to learn and use, so the manuals are only of importance should you run into problems. The level of detail in the manuals is commendable – extending even as far as providing the exact format of the pcb-file database. Stored in ascii form, the database could be edited by any word processor.

On initial start up, you are taken to an interrogation screen where you are prompted for access codes. These are codes which unlock specific features of the software. The review version had an eight-digit code for each of four modules, namely '*PCB*', '*Advanced PCB*', '*Advanced Route*' and '*Advanced Place*'. How advanced all these advanced features are remains to be seen.

Dimensional limits on pcb size are 100 by 100in, while positional resolution is 0.001in. Even with toggling between imperial and metric units, the system maintains an accuracy of 0.005in. It supports up to 16 signal layers, 4 internal power planes, 4 mechanical assembly layers, 2 silkscreen overlays, 2 resist masks, 2 paste masks, drill guide, drill drawing, multi-layer and drc error layers – 34 in all!



Advanced PCB's conventional windows editing environment.



Pattern selection from library by name and preview.

Editing

Without further delay, and without reading the manuals, straight into the editor; load a file, Windows style, and start work. Moving around the pcb is done with the mouse, cursor keys or scroll bars. Zoom level is controlled via page up/down keys, the zoom toolbar, which is a window zoom function, or the zoom menu, from which any zoom can be set.

Commands can be executed from Windows pull down menus, but experienced users will find using two key sequence mnemonic hot keys much quicker. Pressing 'PA' for example places an arc. Pressing the Q key at any time toggles between metric and imperial units.

There are hot-key shortcuts for automatic pan and zoom. If you are using auto pan, for example, while dragging a component, holding down the shift key will pan the display at four times the normal rate. When using page-up and page-down to zoom, holding down the shift key causes slow zooming at 0.1 of the normal rate.

Interconnection starts with the pcb netlist generated by schematic capture. This defines the set of pcb footprints and the connectivity between nodes. Once the components are placed, a rat's nest shows what needs to be connected and as the nodes are connected, so the rat's nest is removed. Once all the node connections are complete, the engineering rule check evaluates whether nets are complete and separate from one another.

Component placement is usually the key to a good pcb, making it easy to route and optimising it electrically. The strategy for auto-placement facility in *Advanced PCB* is not specified, but it is usual for optimisation to be based on total net length and real estate use. The facility worked well, if rather slowly, but using local auto-placement, quicker placement of specific areas can be achieved.

The autorouter can be set for a wide variety of routeing strategies and passes, including pre-routes, smd stringer and fanout, memory routes and line probe. A maze router and shape router can be selected. Performance of each of the facilities does depend upon the pcb type and density, but the overriding impressions are of very slow progress combined with excellent final results.

Speed – or the lack of it

For any pcb layout package, it is essential that the redraw speed is fast. When you are trying to visualise how to route or place an item, you do not want to have to wait for the system to redraw. I tried *Advanced PCB* with a logic board of about 25 ICs and the redraw time was about 2-3 seconds on a 33MHz 486DX PC. This was significantly worse than the redraw time using the earlier DOS based *Autotrax* product. This made it clear why the Preview Mode display had been implemented!

Autorouteing and placement are other areas where the speed problem prevails. If it takes an hour to try and route a few dozen routes – and fail to route about half of them – then productivity is certain to be poor.

Advanced PCB retains the same review and report structure as its predecessors. Although functional, some competing products provide on-line design-rule checks which can stop you doing something silly at the time. The design-rule check function was very slow compared with the old DOS based Autotrax – probably taking twice as long to do the same job.

There is a wide variety of component libraries, and as each library may have hundreds of components the new library search facility is very welcome.

The component outline library provides data on pad size and placement for a wide variety of component footprints totalling over 300. There are options available to generate your own library components, or indeed modify those provided. This is all achieved within the program library editor. Each 'component' may then be loaded to the pcb in the same way as any other.

Engineering change order, or eco, is a new Protel function. Following closely the PADS pcb system, it checks the pcb database for changes made during routeing and produces a file which can be read into the schematic.

When the schematic editor and the companion *Advanced PCB* editor are open at the same time, it is possible to cross probe – that is, select a part on the pcb and then jump automatically to the corresponding pcb component on the schematic.

CELL VERSUS SHAPE-BASED ROUTEING

Most autorouters are grid-based routers using a map of grid cells to define every available cell on the pcb. For tight tracking, a small grid size is required and the memory requirements escalate alarmingly. A four layer pcb on a 0.001in grid only 5 by 5in with one byte per cell, for example, requires 100M-byte of memory. Although the day when a standard PC has this much memory may not be too far off, shape-based routeing offers high resolution routeing by only checking pcb objects while routeing.

Protel's *Advanced SB Route* is an optional shape-based autorouter. The benefits of the shape-based router is that it describes the routeing problem more precisely and in much less memory than a grid router. Applications where shape-based routeing will score are off-grid metric placements, fine-pitch smds or staggered pga objects, where high resolution is required.

SYSTEM REQUIREMENTS

Advanced PCB 2:
Windows 3.1 in standard mode with a 286 or better processor and at least 1M-byte ram.
Advanced PCB requires at least 4M-byte of ram with Windows 386 enhanced-mode recommended. Will work with a 386 with maths coprocessor, but a 486 processor is recommended.
MS DOS 5.0 or later.
Advanced PCB needs a minimum display resolution of 800x600. Larger screens are recommended.
Output to a Windows supported pen plotter or printer is provided plus separate direct HPGL output. Gerber format photoplotter output is also possible.

Advanced Schematic 2:
Windows 3.1. 386 processor minimum, 486 and svga video, i.e. 800 by 600, preferable. Larger screens are recommended. System will work with standard vga.
20M-byte hard disk space plus 8M-byte ram.
Output to a Windows supported pen plotter or printer.
Separate direct HPGL support. Gerber format photoplotter output.

SUPPLIER DETAILS AND PRICE

Premier EDA Solutions, 133 Cardiff Road, Reading, Berkshire RG1 8ES. Tel 0734 57 44 44, fax 0734 599 519. Prices for the Protel packages: *Professional PCB* £695, *Advanced PCB* £1250, *Advanced Route* £695, *Advanced Placement* £695 and *Advanced Schematic 2* £695. The productivity pack reviewed is priced at £2795. There is a competitive upgrade scheme plus educational and volume discounts.

Conclusions

Advanced Schematic 2 is a truly excellent electronic cad tool. Although it has some weaknesses, these can be spotted and overcome in the normal use of the product to provide an electrically consistent schematic.

The ability to link to the simulation products is perhaps rather overstated, because it is only the *Spice* interface that seems to be properly supported. Many of the so called 'tool' facilities are just as easy to launch by task switching to program manager – an example of a facile menu option is that of launching the windows clock!

With a full Windows help system, you should be able navigate through the program almost without reference to the documentation.

Fast selection of library components and a really good project orientated sheet hierarchy offer great user productivity, but this can only be achieved using the high resolution screen. A standard vga screen will operate – but is barely usable.

At less than £700 the package is competitively priced, and offers users the opportunity to truly upgrade their dos products to a product with much greater power and functionality. With a high resolution screen and fast processor the functionality and performance is equivalent to the best workstation products.

Although this review was carried out on a 33MHz 486DX PC with 8M-byte of ram, all the functions on the latest version of *Advanced PCB 2* ran very slowly. This is sad, because the software looks good and handles well, but without a 100MHz Pentium, it could bring on a case of severe frustration.

Advanced PCB offers a workstation level of functionality, but without the best PC processor speed and extended memory size it cannot deliver sufficient speed for good user productivity. ■

Audio design leaps forward?

Designers have long recognised the theoretical advantages of combining feed-forward error correction with feedback. But in his design for a feed-forward audio power amp Giovanni Stochino looks to have succeeded in putting theory into practice.

Since its invention by H S Black[1] in the 1920s feed-forward error correction has found practical application in radio frequency and microwave amplifiers[2]. But it has never been used, in Black's form, in audio power amplifiers[3]. The reason is probably the inherent difficulty in accurately and efficiently applying Black's feed-forward principle to audio power amplifiers over the full audio frequency range.

But a newly-developed circuit technique could do just that, and, within specified limits, put Black's true feed-forward principle to work in high power audio amplifiers.

Experimental results demonstrate the effectiveness of the proposed technique, but first , a look at some of the underlying theory.

Feed-forward or feedback?

The general input-output relationship of a power amplifier, before applying correction, can be written as $V_p=V_iG_p+E_p$. G_p is the voltage gain, generally a function of frequency and load impedance, and E_p is the error component that includes the amplifier's non-linear distortion and noise. E_p depends on input voltage and load impedance, and on frequency.

When negative feedback is applied (**Fig. 1**), the input-output relationship of the corrected amplifier becomes $V_o=A_{cl}V_i+E_{fb}$. A_{cl} is the closed loop voltage gain, substantially defined by the feedback network, and E_{fb} is the residual error component after feedback correction.

Analysis shows that distortion can never be completely nulled by negative feedback – though feedback is effective in reducing distortion as long as there is enough gain within the feedback loop.

Feed-forward is based on a different mechanism of error correction. The basic scheme (**Fig. 2**) incorporates a criterion network (α, γ and *SC*) to determine, isolate and extract power amplifier error; an auxiliary amplifier *AA* (low power requirement, low distortion

Fig. 1. Basic elements of a feedback amplifier. Analysis shows that distortion can never be completely nulled with this configuration.

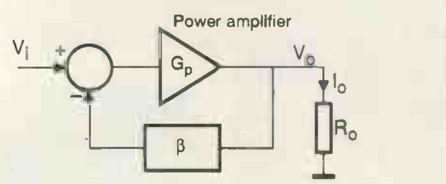
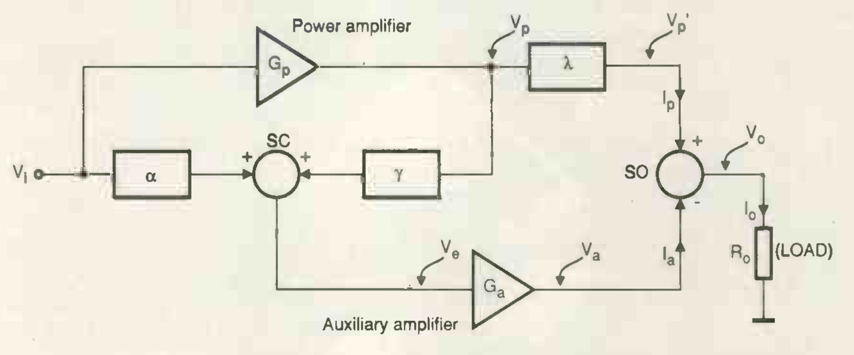


Fig. 2. Principles behind feed-forward. In this configuration, power amplifier error is extracted, determined and isolated via a criterion network comprising α, γ and SC.



and low noise compared with the power amplifier, PA) to provide a buffered copy of E_p; and output summing network *SO*. In *SO* the error component of *PA* and its copy available at the output of *AA* cancel out to provide a distortion-free output voltage on load R_0. Phase-amplitude equaliser network λ is added to the basic scheme to improve the error-correction mechanism at high frequency.

The scheme should include a few delay lines to compensate for amplifier propagation delay and connections. But their influence is negligible in the audio frequency range.

Simple analysis of the diagram gives:

$$V_0=V_p'-V_a$$
$$=V_iG_p\lambda-G_aV_i(\alpha+\gamma G_p)+E_p(\lambda-\gamma G_a)+E_a$$

where E_a is the error component (distortion plus noise) produced by the auxiliary amplifier. Proper operation of the feed-forward technique requires that $E_a<<E_p$ so the effective output error is $E_{ff}=E_a+E_p/\rho$. Term $\rho=1/(\lambda-\gamma G_a)$ can be defined as the distortion rejection factor of the feed-forward amplifier and describes the effectiveness of feed-forward in removing distortion in the power amplifier. E_{ff} reduces to its lowest value of E_a when $\rho=\infty$, that is when $\gamma G_a=\lambda$, and shows the potential of the feed-forward mechanism to completely null distortion E_p in the power amplifier.

The further condition $\gamma G_p=-\alpha$ should be satisfied to nullify the component $V_e^*=V_e(V_i)$ (see panel p. 822 for definition of V_e^*) at the input of the auxiliary amplifier. This would minimise both E_a and power handling requirements for the auxiliary amplifier. The mathematics implies that when $\gamma>0$, G_p and α have opposite signs.

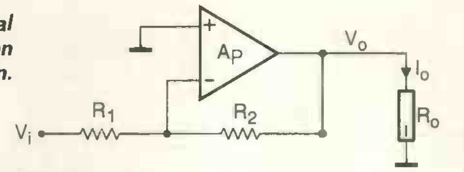


Fig. 3. Simplicity of a practical feedback amplifier configuration with inverting gain.

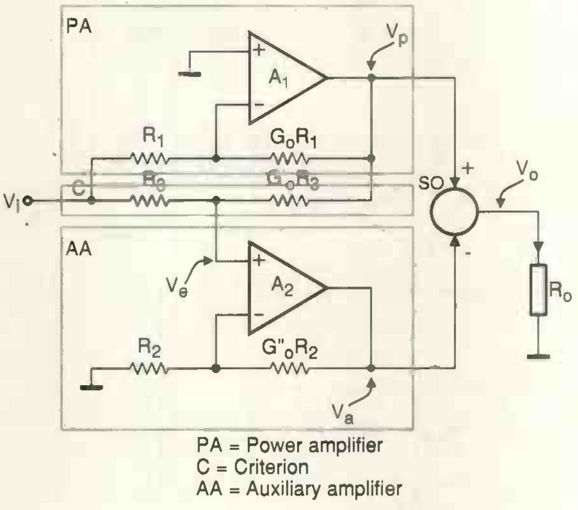


Fig. 4. A practical feed-forward amplifier incorporating feedback amplifiers.

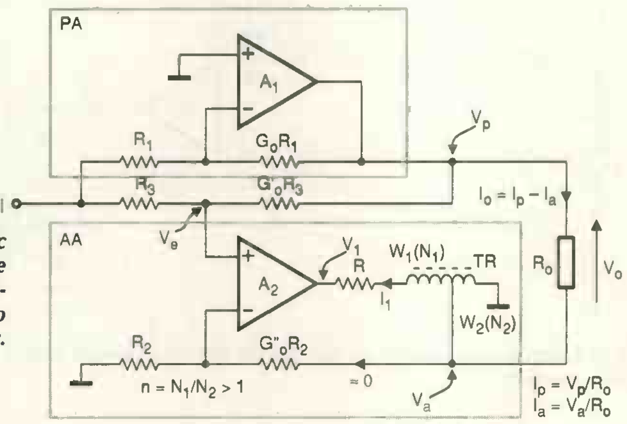


Fig. 5. Basic scheme of the proposed feed-forward audio power amplifier.

Feed-forward more promising?

Distortion E_{fb} of a feedback amplifier can never be nulled, but it can be substantially reduced in the range of frequency and input voltage, where the feedback factor is much greater than 1.

As a technique, it is less effective at the highest frequency of the audio range and in the crossover region of class AB amplifiers, where the feedback factor can be low and deviation from linearity is high[4].

On the other hand, negative feedback amplifier configurations are very simple and require no matching of components (**Fig. 3**).

Feed-forward error correction is much more complex. But better distortion results are possible. In theory, the error of the whole power amplifier can be reduced to that of the auxiliary amplifier alone, even at high frequencies and in the crossover region. The advantage is that the auxiliary amplifier needs to handle only moderate currents and voltages. So it can be designed to provide much lower distortion (for instance it can be operated in class A) than the power amplifier, and very low distortion can be achieved.

Neither feedback nor feed-forward error correction can completely null the output error of a power amplifier. But feed-forward is more promising, virtually nulling distortion of the power amplifier, leaving only the low residual error of the auxiliary amplifier over the load.

Combining feedback and feed-forward

Tight matching of parameters in the feed-forward scheme (Fig. 2) can be achieved, simply and steadily, by using negative feedback. The strategy helps precise definition of gain in both the power amplifier and the auxiliary amplifier – provided the open loop gain of both amplifiers is high in the audio frequency range. So feedback and feed-forward techniques can be profitably combined in a true low-distortion audio power amplifiers. In a practical application (**Fig. 4**), the power amplifier and auxiliary amplifier have the gains defined by their respective feedback networks: $G_p=V_p/V_i\cong-G_0$ and $G_a=V_a/V_e\cong(1+G_0'')$ provided $A_1/G_0>>1$ and $A_2/G_0''>>1$. But there is also $\gamma=1/(1+G_0')$, $\lambda=1$ and $\alpha=G_0'/(1+G_0')$. As a result, $G_0''=G_0$ and $G_0'=G_0$.

The scheme is a practical way of assuring that the fundamental conditions for proper operation of feed-forward technique are always satisfied. But the problem remains in implementing the output summing network – probably the most difficult obstacle in the basic feed-forward error correction scheme.

The simplest and most intuitive way of realising this summing network is where corrective voltage V_a is directly transferred into the load's loop. Voltage V_0 across the load is equal to $V_p-V_a=-G_0V_i+E_p-E_p+E_a=-G_0V+E_a$, which is consistent with feed-forward theory.

But if, in this scheme, the auxiliary amplifier has to sustain the full load current, the assumption that the auxiliary amplifier is a low-power, low-distortion (prospectively class A) amplifier is no longer valid.

As a result, we can not assume that $E_a<<E_p$, and consequently the inherent advantage of the feed-forward technique disappears. This is why the simple feed-forward configuration has never been used in power amplifiers[3].

It also explains why, though the advantages of the feed-forward technique, in conjunction with feedback, are generally recognised, Black's feed-forward error correction technique has found only limited application by audio designers.

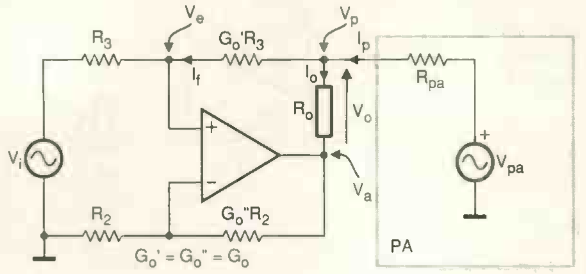


Fig. 6. Non-zero output resistance of PA does not influence the properties of the feed-forward correction mechanism.

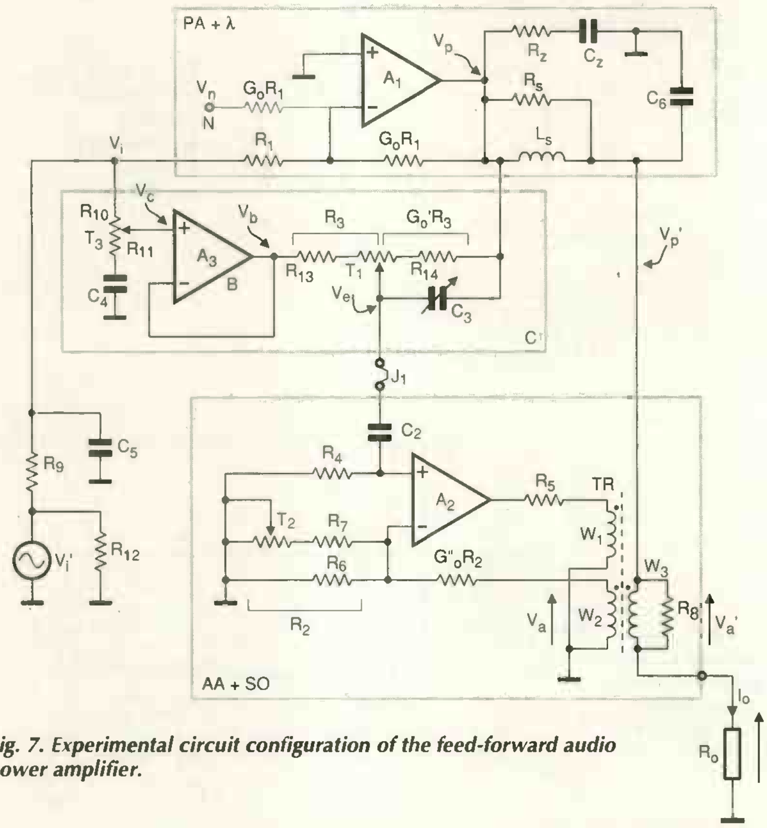


Fig. 7. Experimental circuit configuration of the feed-forward audio power amplifier.

Feed-forward error correction (always intraloop) is sometimes used in audio power amplifiers[5, 6], but Black's basic scheme has yet to be incorporated into audio power amplifier design.

A feed-forward power amplifier that works

We have seen that, in the feed-forward scheme (Figs. 2, 4), the most critical part to be implemented in audio applications is the output summing network (*SO*). Here the power signal V_p coming from the power amplifier and low level corrective signal V_a produced by the auxiliary amplifier have to combine without undesired interaction (ie cross-modulation, frequency instability, gain impairment) to provide a distortion-free output.

What is more, this combination must be performed efficiently without requiring much power from the auxiliary amplifier, and must not be affected by impedance-variations of loads – even loads as difficult as loudspeakers.

In amplifiers for hf use, such problems are less critical. Appropriate networks can be used to implement the output summing function, due mainly to the favourable frequency range and fixed system impedance (50Ω).

But audio applications span an unfavourable frequency range and imply complex and unpredictable load impedances. As a result, circuit techniques commonly used in radio frequency and microwave feed-forward power amplifiers are not practical, and different solutions have to be found.

An effective approach (Fig. 5[7]) has *PA* as the power amplifier to be corrected (usually class AB), and *AA* as the auxiliary amplifier. *AA* should be operated in class A for the best performance and incorporates transformer *TR* in its feedback loop. (Resistor *R* also includes the resistance of winding W_1 and the output resistance of A_2.)

The unique role of *TR* is to provide both the wide-band impedance matching of the auxiliary amplifier to R_0 and the power-efficient means for injecting the corrective signal V_a into the load's loop.

Transformers are usually avoided in solid state audio power amplifiers, as they are expensive, bulky, band-limited and not suited for very low distortion applications. But when used in unconventional ways, as in this case, their unique properties can prove useful.

Putting transformer *TR* in the feedback loop of the auxiliary amplifier has two very important effects. The flux produced in the magnetic core of *TR* by the power component of the load current is automatically annulled by the feedback that forces voltage V_a to be insensitive to power component variation. So no restrictions are imposed on transformer size and core material by the amount of power that the power amplifier transfers into the load. In most cases a small transformer can be used.

Open loop output impedance of the auxiliary amplifier can also be extremely low (a few mΩ) in the full audio frequency range and above. The consequence is that undesired interactions and cross-modulations between power amplifier and auxiliary amplifier, as well as the sensitivity of the auxiliary amplifier to load impedance variations, are strongly reduced. A further benefit is that the primary winding of *TR* is driven, virtually, by a voltage source, since *R* tends to zero. This widens the frequency bandwidth of *TR*, whose practical low frequency corner f_0 turns out to be as low as a few Hz, even if a small ferrite core is used to improve its bandwidth and linearity.

Transformer operational requirements

The function of the transformer – to permit injection of the corrective current into the load without interaction with the main current component – is performed by cancelling the core flux generated by the main current component.

This flux neutralisation is carried out by the coercive action of the auxiliary amplifier's feedback loop and is effective as long as the current and voltage available at the output of A_2 are adequate and the loop gain remains high. The only effective flux in the transformer core is therefore produced by the corrective voltage V_a.

For frequency $f>f_0$, the peak flux density B_p and the peak voltage V_{ap} are linked by $2\pi f B_m S_e N_2$ where S_e is the effective cross-sectional area of the transformer core. So the amount of corrective voltage V_{ap} that can be provided to the load is limited by the core geometry, through S_e, and the core material, through B_s (ie the saturation flux density), since it must always be $B_p \leq B_s$. The amount of available corrective voltage can also be seen to increase in direct proportion to frequency.

As an example, take a toroidal ferrite core with S_e at 100mm² and B_s at 200mT. If N_2 is 20, V_{ap} is 50mV at 20Hz and V_{ap} is 5V at 2kHz. Compared to an output of 100W/8Ω, they represent peak correctable errors of 0.12% and 8.7% respectively.

Performance matches well with that of class AB audio power amplifiers, exhibiting nonlinear distortion that rises with increasing frequency, and extends normally, say, from 0.01% to 1% in the audio frequency range.

Amplifier requirements

Class A operation is mandatory for amplifier A_2 to achieve the lowest distortion with low level error signals. High gain and low noise

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are also recommended, as is low output offset voltage to stop any noticeable direct current flowing into the transformer winding.

Other requirements are wide bandwidth and high slew rate so that the correction capability of the auxiliary amplifier is extended to the highest harmonics of audio signals.

Output voltage and current handling capabilities of the auxiliary amplifier are dictated by the peak values of load current and corrective voltage: $V_1 \cong nV_{ap}$ and $I_1 \cong I_{op}/n$ where $n=N_1/N_2$ is the turns ratio and is>>1. The turns ratio is used to trade-off voltage for current to reach the best level of performance of A_2 with a reasonable power consumption. For example, I_{op}=8A, V_{ap}=0.5V and n=40 gives V_1=20V and I_1=0.2A. Therefore, A_2 can be powered from ±25V and its output stage biased at 0.2A for class A operation.

Only 10W of power is consumed by the auxiliary amplifier – a reasonable and worthy amount if compared with the 256W of undistorted audio power furnished to an 8Ω load.

Auxiliary considerations

An important, yet often overlooked, characteristic of feed-forward schemes based upon Black's principle, is that the voltage across the load is not defined by the power amplifier, but by the auxiliary amplifier only. In other words V_0 is not theoretically dependent on power amplifier parameters (output impedance, gain, linearity etc). Power amplifier output could even be completely uncorrelated with V_i, and power amplifier output impedance could be high and non-linear without affecting the output voltage value ($V_0=-G_0V_i$ in Fig. 5).

We can also deduce, mathematically, that $V_a=V_iG_0+V_p$ and $V_0=V_p-V_iG_0-V_p=-V_iG_0$ to show that output voltage is always equal to the desired value, regardless of the power amplifier output voltage V_p. So any deviation of V_p from its ideal value affects the output of the auxiliary amplifier but not output voltage V_0.

We reach the same conclusion if we take into account the non-zero output impedance of the power amplifier **Fig. 6** – particularly important in the crossover and clipping regions, where comparatively high values of output impedance can be experienced.

Assume $G_0R_3>>R_{pa}$. In this case we have $I_f<<I_p$ so $I_p \cong I_0$ and the node voltages are:

$$V_p=(V_{pa}-V_a)/(R_0+R_{pa})+V_a,$$

and,

$$V_a=V_{pa}+G_0V_i.$$

Solving simultaneously, substituting and assuming in normal operation that $V_{pa}=-G_0V_i+E_p$, we have:

$$V_p=-G_0V_i(1-R_{pa}/R_0)+E_p$$

and

$$V_a=G_0V_iR_{pa}/R_0+E_p$$

You can see that the auxiliary amplifier has to contribute a certain amount of signal voltage to the load in addition to the copy of the error voltage E_p. This amount is proportional to the ratio R_{pa}/R_0.

Power amplifiers with poorly biased class AB output stages can actually have closed loop-output impedances comparable with load impedance, at least in the crossover region. In such cases, the auxiliary amplifier contribution to the output signal voltage in the crossover region can prove significant.

Clearly, the role of forcing the desired voltage across the load is undertaken by the auxiliary amplifier.

In well-designed feed-forward amplifiers, the power amplifier provides the power to the load, while the auxiliary amplifier is limited to providing accuracy and precision only. Nevertheless, should the power amplifier fail to do its job (for instance due to crossover mechanism), the auxiliary amplifier would be forced to provide power as well as precision. Obviously, the auxiliary amplifier is designed to provide only a limited amount of precise corrective voltage.

It is worth noting that the auxiliary amplifier is always stable because even with the worst-case positive feedback factor ($R_{pa}=\infty$),

$$F_p=R_3/[R_3(1+G_0')+R_0]$$

is always lower than the negative feedback factor $F_n=R_2/[R_2(1+G_0'')]$ when $G_0'=G_0''$.

Practical circuit

In a practical circuit implementation (**Fig. 7**), the first-order error due to the finite gain-bandwidth product of the amplifiers can be taken into account and compensated for.

Transformer *TR* is modified to provide error correction to a grounded load and its secondary windings W_2 and W_3 are close-coupled to assure that $V_a'=V_a$. Follower *B* buffers the input voltage source and the phase-amplitude-equaliser (R_{10}, R_{11} and C_4) from the criterion network. R_g and C_5 form an input low-pass filter cutting off input frequencies above 100kHz. Decoupling C_2 avoids undesired dc operation with the auxiliary amplifier. Trimmers T_1, T_2 and T_3 facilitate calibration of the complete amplifier and help achieve the best distortion performance.

In analysing the circuit, the effects of C_2 on the output voltage can be neglected, with C_2 assumed to be ∞.

In the Zobel network (R_z, C_z, R_s and L_s) commonly used at the output of class AB audio power amplifiers, R_s and L_s in conjunction with C_6 implement the λ amplitude-phase equalisation network (Fig. 2).

As well as its normal role of separating the power amplifier from the load at frequencies far above the audio range, this network also limits positive feedback around the auxiliary amplifier at the highest frequencies, improving

Magnetic flux cancellation

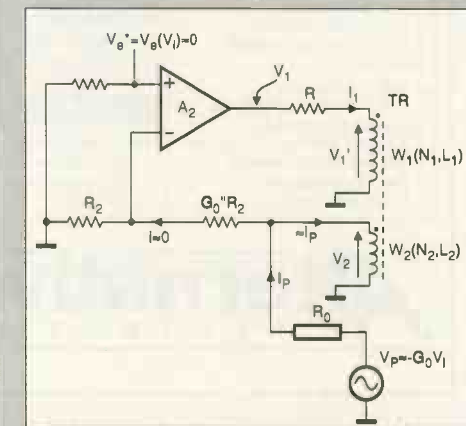
This diagram helps analyse the mechanism of magnetic flux cancellation $\Phi(V_i)$ in the transfomer core, due to main signal current $I_p \approx V_p/R_0 \approx -G_0V_i/R_0$. It is operated by the auxiliary amplifier.

$$\Phi(V_i) \cong V_2(V_i)/(2\pi f N_2)$$

$$V_2(V_i) \cong -G_0V_i\frac{R}{n^2R_0}\cdot\frac{1}{\{1+A_2/[n(1+G_0'')]\}}$$

From this,

$$\lim V_2(V_i) = \Phi(V_i) = 0$$
$$(A_2 \to \infty \text{and/or} R \to 0)$$



Turning Black's feed-forward principle into practice

Load impedance 'seen' by amplifier A_2 is high enough to allow true low distortion operation of the auxiliary amplifier.

Auxiliary amplifier has to process only small error components, and being class A operated, its percent error contribution to load current is extremely low.

There is no appreciable common-mode-induced distortion because component $V_e^* = V_e(V_i)$ is virtually zero.

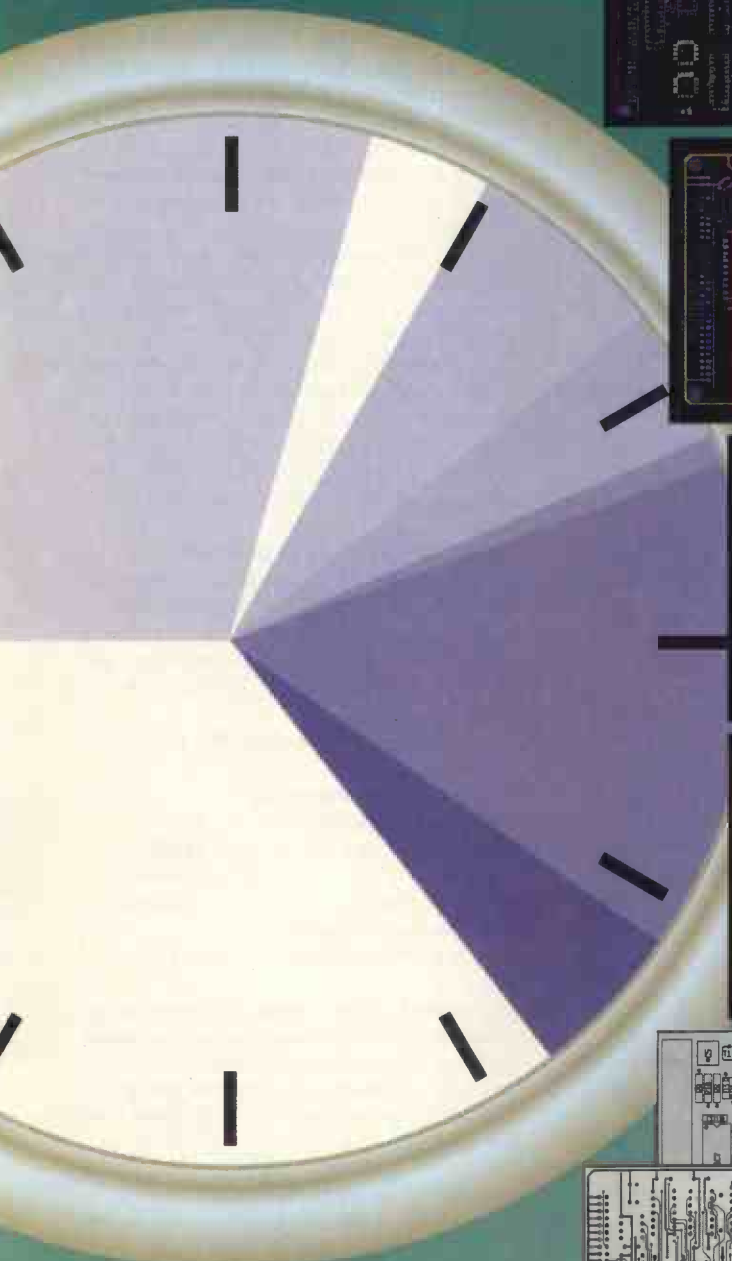
Transformer distortion, if any, is reduced in proportion to the loop gain of the auxiliary amplifier.

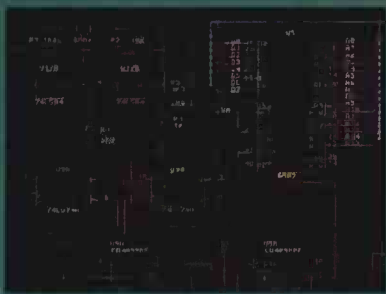
Extremely low open loop output impedance means the auxiliary amplifier's closed loop gain and distortion performance are insensitive to load impedance variations.

Wide bandwidth achievable for the low power auxiliary amplifier also allows a large reduction of the highest harmonics of audio signals.

Error correction technique (Fig. 5) can be applied to all power amplifier circuit configurations, with inverting as well as non-inverting gain.

FROM CONCEPT TO ARTWORK IN 1 DAY





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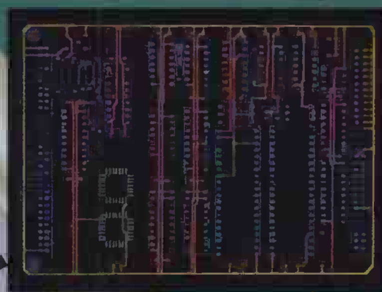

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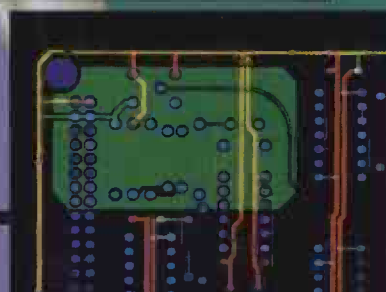

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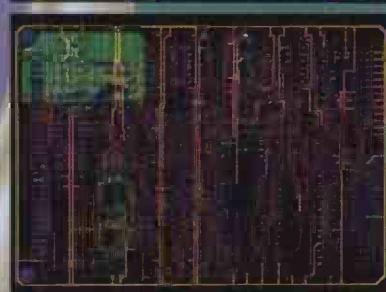

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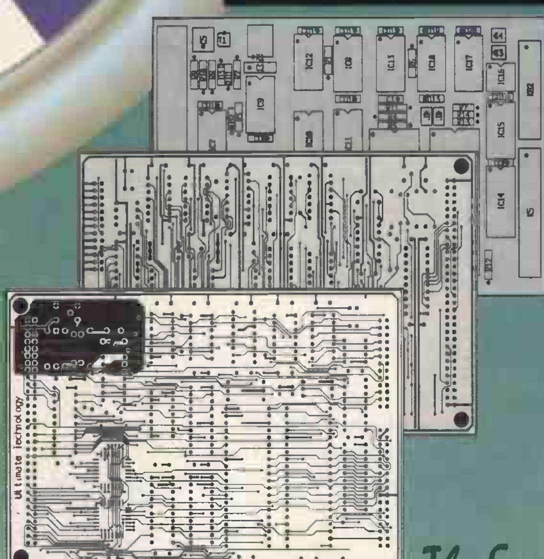

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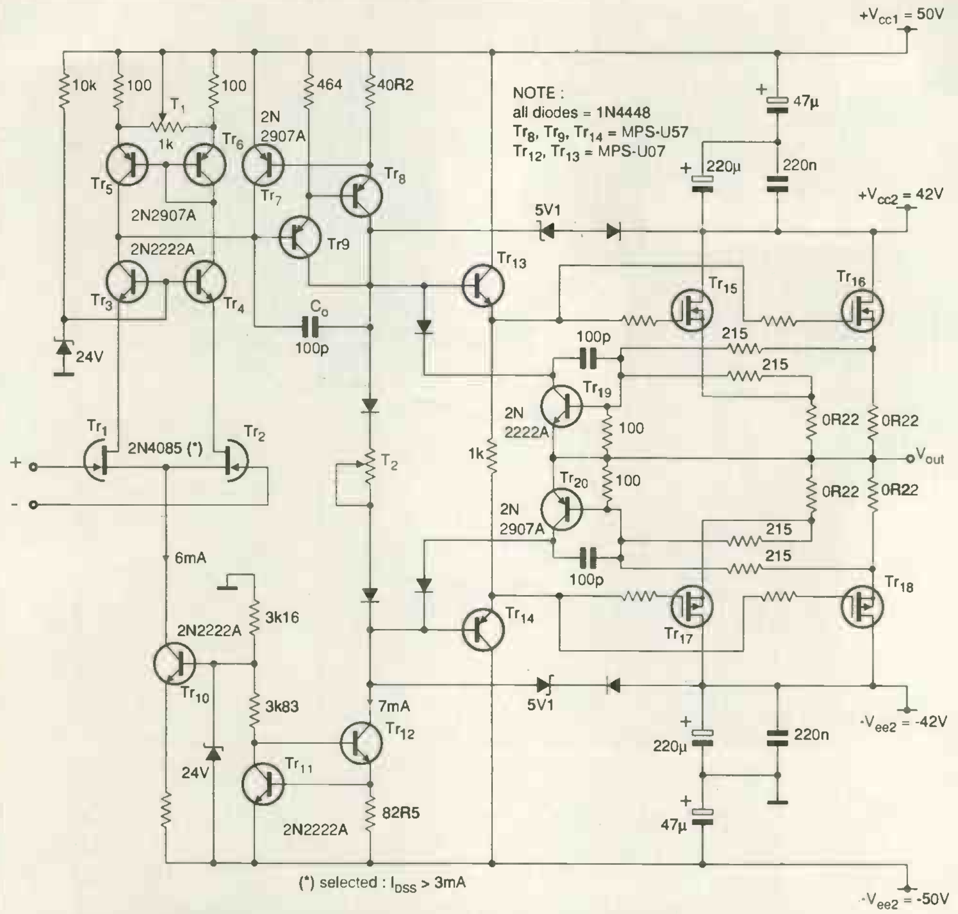

Fig. 8. Complete circuit diagram of the power operational amplifier A_1. Output power into 8Ω is 100W and slew rate is 60V/µs.

frequency stability of the whole amplifier.

Resistor R_8 controls the high frequency output impedance of *AA*. Trimmers T_1 and T_2 permit conditions to be satisfied. C_3 compensates the frequency response of the auxiliary amplifier, and the input capacitance of A_2.

In the analysis, a single pole frequency response is taken for all amplifiers. So, in addition to condition $G_0'=G_0''=G_0$, we shall assume $V_p=[-V_iG_0/(1+s/p_1)]+E_p$. Also,

$$V_b=V_c/(1+s/p_3)\cong V_c=V_i(1+s/z_0)/(1+s/p_0),$$

where p_0 is $1/[C_4(R+R_{11})]$, z_0 is $1/C_4R_{11}$ and V_a is $V_e[(1+G_0)/(1+s/p_2)]+E_a$.

In general we have $p_3>>p_1$ and $p_3>>p_2$ so that the assumption $V_b\cong V_c$ has no appreciable consequences.

As for λ – assuming a first order response of the auxiliary amplifier – we expect a first order low-pass frequency response so that $\lambda\cong 1/(1+s/p_5)$ where $p_5=1/(C_6R_6)$, corresponding to the assumption that $R_5<<sL_s$.

Regarding the value of the distortion rejection factor, the validity of the above expressions for λ and p_5 is substantially independent on the load impedance value, since V_a' is applied in series with R_0. Then, according to Black's scheme in Fig. 2, applied to our circuit (Fig. 7), the error voltage V_e can be written as:

$$V_e=V_i(\alpha+\gamma G_p)+\gamma E_p$$

where

$$\alpha=[(1+s/z_0)/(1+s/p_0)].[G_0/(1+G_0)(1+s/p_4)],$$

$$\gamma=(1+s/z_3)/[(1+G_0)(1+s/p_4)],$$

$$\gamma G_p=-G_0(1+s/z_3)/[(1+G_0)(1+s/p_1)(1+s/p_4)],$$

$$z_3=1/(G_0R_3C_3)$$

and

$$p_4=(1+G_0)z_3.$$

The condition that $\gamma G_p=-\alpha$ can be met if $p_0=p_1$ and $z_0=z_3$ so that the error V_e reduces to γE_p.

We also want to satisfy the condition $\gamma G_a=\lambda$. Substituting for each term reveals that this condition is true if $z_3=p_2$ and $p_5=p_4$ so that $\rho=1/(\lambda-\gamma G_a)=\infty$. Therefore, the power amplifier error E_p turns out to have been completely removed from the output voltage and the above can be written as:

$$p_5=p_4=(1+G_0)z_3=(+G_0)p_2\cong 2\pi f_{T2},$$

where f_{T2} is the nominal gain-bandwidth product of the auxiliary amplifier.

The interpretation of making z_3 equal to p_2 and $p_5=p_4$ is that zero z_3 is introduced to compensate for the first-order phase-amplitude errors caused by the pole p_2 of the auxiliary amplifier, while the low residual errors due to the collateral pole p_4, associated with block of the criterion network, are counteracted by means of the high frequency pole p_5 of the phase-amplitude equalising network λ.

In other words we can state that all potential limitations of Black's feed-forward error-correction mechanism, due to the dominant pole of both the power and the auxiliary amplifier, have been actually counterbalanced in this practical implementation.

The more accurate expression of λ, written as $\lambda=(1+s/z_s)/(1+s/z_s+s^2/\omega_0{}^2)$ where z_s is R_s/L_s and $\omega_0{}^2$ is $1/L_sC_6$ shows the additional potential of the λ network to compensate for a more realistic second-order frequency response of the auxiliary amplifier, by suitable choice of z_s and ω_0.

This accounts for the high distortion rejection factor (30 to 60dB for frequencies up to 1MHz) that has been measured after calibration on the prototypes (see 'Measurement results') with different load conditions. The end result is that very low distortion figures can be expected – and attained.

An additional property of the feed-forward implementation depicted in Fig. 7, with its floating winding (W_3) able to inject the corrective voltage V_a into the load's loop, is that it easily lends itself to iterative application, reducing output error to extremely low levels.

Power op-amp A_1 in practice

The viability of the error-correction technique discussed so far has been demonstrated by prototypes of an 100W/8Ω audio power amplifier (**Figs. 8, 9 and 10**), assembled and calibrated according to Fig.7 using the theory analysed above.

Power op amp output stage (**Fig. 8**) includes

Main characteristics of A_1

Output power into 8Ω	100W
Output power into 4Ω	160W
Slew rate	±60V/µs
Power bandwidth	≅200kHz
Gain-bandwidth product (measured at 1MHz)	≅11MHz

two pairs of complementary n- and p-channel power mosfets whose quiescent current can be adjusted with trimmer T_2. Different supply rails are used to improve amplifier efficiency.

Output offset voltage can be adjusted with trimmer T_1.

Auxiliary op-amp A_2

Amplifier A_2, Fig. 9, has a mosfet output

Main characteristics – amplifier A_2

Voltage gain	≥10^8
Output voltage range	≅±20V
Output current range	
Class A	≅±200mA
Class AB	±800mA
Slew rate	≅±500V/µs
Gain-bandwidth product (measured at 1MHz)	≅300MHz
Output offset voltage	≅±600µV

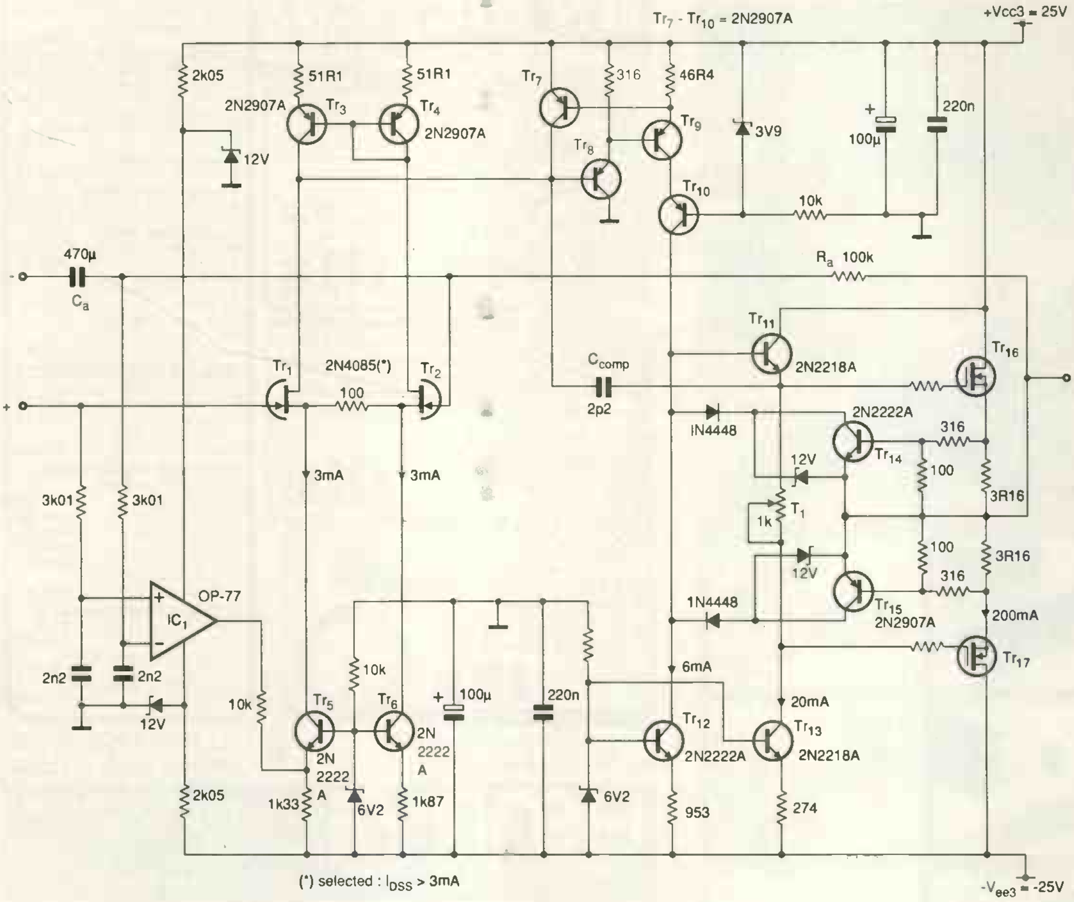

Fig. 9. Power output requirement for the auxiliary operational amplifier, A_2, is lower. In class A, output is ±200mA while in class AB it rises to ±800mA.

stage, biased by adjusting trimmer T_1 at 200mA for class A operation in normal working conditions. But it can operate in class AB operation, when A_2 is forced to sink or source higher currents, due to variations in load impedance or clipping for example.

The circuit is a combination of a high-speed, high dynamic-range amplifiers (Tr_{1-17}) and a precision integrated op-amp IC_1. The main task of IC_1 is, with the help of coupling capacitor C_a and feedback resistor R_a, to keep the offset voltage below a few hundred millivolts and to increase the low frequency open loop gain of the overall amplifier.

IC_1 also helps reduce the low-frequency voltage noise ($1/f$ noise) associated with jfet pair $Tr_{1,2}$.

Main characteristics – buffer

TDH with 1V/600Ω	
from 20Hz to 20kHz	≤0.0005
—3dB small-signal bandwidth	≅15MHz
Voltage noise density	≅1nV/√Hz
Slew rate	≅±20V/µs

Buffering

Buffer *B* of **Fig. 10** consists of an op-amp voltage follower and makes use of a high-performance integrated operational amplifier, featuring low noise and very low distortion.

Transformer

Transformer *TR*'s core is a small toroid – 23mm external diameter, 14mm internal diameter and 7mm high. Its cross-sectional area S_e is approximately 31mm^2. The core material is Ferroxcube-grade 3E2, having a saturation flux density B_s of about 350mT, and a useful linear range of ±200mT. Turns ratio n is 30 (N_1=300, N_2=N_3=10).

Secondary windings W_2 and W_3 are close-coupled with parallel and cross-coupled thick wires and all windings are uniformly wound along the core length. When driving the primary winding with a source resistance of 5Ω, the –3dB small signal bandwidth of the transformer extends from 5Hz to about 13MHz.

Complete auxiliary amplifier and transformer characteristics

Voltage gain	≅18.1
Gain-bandwidth product	≅10MHz
Slew rate	≅±17V/µs
Thd+noise	
@5kHz	≅0.01%
Thd, V_a=0.5V/1Ω	
@50kHz	≅0.05%

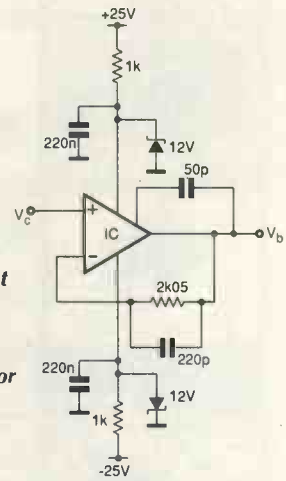

Fig. 10. Buffer B. At frequencies from 20Hz to 20kHz, distortion of this design is less than or equl to 0.0005%.

Amplifier calibration

Amplifier calibration has been performed with a load of 8Ω in parallel with 0.2µF using the following procedure.

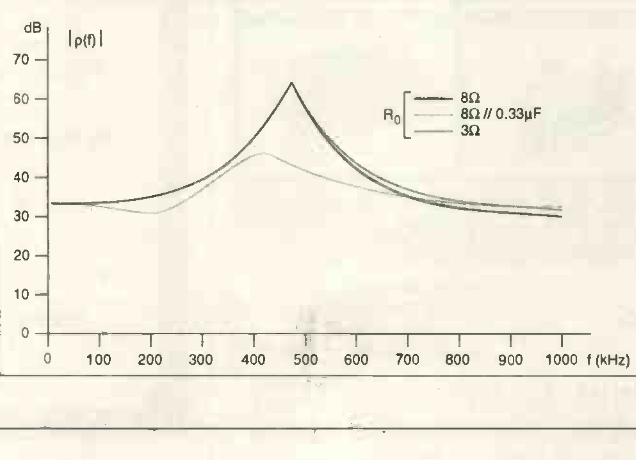

Fig. 11. Magnitude of the distortion rejection factor ρ after calibration.

Step 1: Jumper J_1 is opened to isolate *AA*.

Step 2: Corner frequency of both *PA* and *AA* is measured and recorded ($G_0'' \cong G_0 \cong 18.1$).

$f_{c1}=p_1/2\pi \cong 422\text{kHz}$
$f_{c2}=p_2/2\pi \cong 650\text{kHz}$

Step 3: Nominal value of C_3 is determined with $z_3 = p_2$:

$C_3=1/(G_0R_3p_2) \cong 13.5\text{pF}$.

Step 4: Nominal values of C_4, R_{10} and R_{11} are found by applying $p_0=1/C_4(R_{10}+R_{11})$ and $z_0=1/C_4R_{11}$ and using $p_0=p_1$ and $z_0=z_3$. Since $R_{10}+R_{11}=R_{T3}=2\text{k}\Omega$ we obtain:

$C_4 = 1/[(R_{10}+R_{11})p_1] \cong 188\text{pF}$
$R_{11} = 1/C_4z_3 = 1/C_4p_2 \cong 1.3\text{k}\Omega$.

Step 5: The aim is to meet the condition defined by $\gamma G_a=\lambda$. Signal V_i'=100mV/3kHz is applied to the input and trimmer T_1 is adjusted so that $V_e^* = V_e(V_i)$ reaches a minimum. Then the frequency is increased to 100kHz and trimmer T_3 is adjusted so that $V_e^* = V_e(V_i)$ is again at a minimum.

Step 6: Connect jumper J_1 and repeat step 5.

Step 7: Input of the amplifier is grounded and a forced error signal E_{pn} is produced at the output of *PA* by applying the input voltage $V_n \cong 50\text{mV}$ (the amplitude of V_n must be kept below the limits set by $V_{am} = 2\pi f B_m S_e N_2$, as shown in Fig. 7. Since $E_{pn} = E_n$, this method maximises, in a wide frequency range (up to 1MHz), the distortion rejection factor $\rho = 1/(\lambda - \gamma G_p)$ of the auxiliary amplifier.

Frequency of V_n is first set at 3kHz and trimmer T_2 is adjusted so that the output voltage $V_0(V_n)$ is at a minimum. Then, the frequency is increased to 300kHz and C_3 is adjusted again to have maximum rejection. A network analyser would simplify amplifier calibration, allowing optimization of π in the 1kHz to 1MHz frequency range.

Step 8: Repeat step 5

Measurement results

Figure 11 shows the magnitude of the distortion rejection factor as a function of frequency, achieved for the experimental prototypes which have been calibrated.

We see that π values extending from magnitudes of 30-60dB have been achieved in the wide frequency range 200Hz to 1MHz. Even better results can be expected with more care taken in layout and power distribution design. These values translate into an equivalent degree of reduction of the total harmonic distortion, thd, of the power amplifier, as demonstrated by test results (**Figs. 12 and 13**).

Two significant levels of the total bias current I_{bias} of the power amplifier mosfet output stage are taken into account. The first one, I_{bias}=1mA, representing a very poor biasing level, helps prove the ability of the proposed technique to counter-balance the effects of the comparatively high output impedance of the power amplifier in the crossover region.

In addition, it shows the ability of feed-forward to reject high-order harmonics normally generated by poorly-biased output stages.

The second level, I_{bias}=100mA, is closer to the normal level of biasing of power mosfet output stages and demonstrates the effectiveness of the proposed technique to correct small amounts of distortion.

Results (Figs. 12 and 13) show that the measured improvement ratio of about 30dB is in good agreement with the value of distortion rejection reported in Fig. 11, and gives clear evidence of the effectiveness of the proposed feed-forward technique.

Only the worst case (f=20kHz) thd+noise versus output level (volt peak-to-peak/8Ω load) is reported. All other measurements taken at f<10kHz are, after applying the error correction technique, very close to the instrumentation limits.

Effectiveness of the distortion rejection mechanism with audio programs, has also been simulated by superposing a white-noise voltage at the output of the power amplifier.

A white-noise level of V_n=0.5Vrms was injected at input node *N* while the amplifier was delivering 20Vpk-pk to the load with f at 1kHz. Unfiltered noise appearing across the load was 32dB lower than that measured at the output of *PA* – a high level of rejection in agreement with theoretical expectations.

The final test report refers to the output noise levels of the amplifier, before and after correction. They are 0.79mV and 0.38mV, respectively. ■

Components

These component values were used in the prototype. All resistors have 1% tolerance.

R_1	= 12.5kΩ
G_0R_1	= 226kΩ
G_0	= 18.08
R_2	= 89.6Ω (nom)
$G_0''R_2$	= 1.62kΩ
R_4	= 10kΩ
R_5	= 5Ω
R_6	= 110Ω
R_7	= 332Ω
R_8	= 5.11Ω
R_9	= 1kΩ
R_{10}	= 0.7kΩ (nom)
R_{11}	= 1.3kΩ (nom)
R_{12}	= 4.7kΩ
R_{13}	= 909Ω
R_{14}	= 17.8kΩ
C_2	= 1µf
C_3	= 13.5pF (nom)
C_4	≅ 188pF (nom)
C_6	= 13.5nF
T_1	= 220Ω
T_2	= 1kΩ
T_3	= 2kΩ
L_s	= 1µH
R_s	= 8Ω
R_z	= 10Ω
C_z	= 47nF

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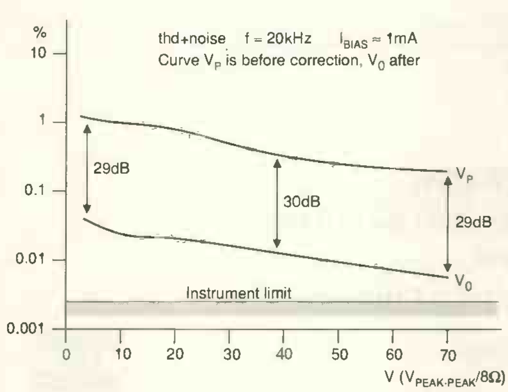


Fig. 12. Total harmonic distortion + noise versus output level. Test conditions: I_{bias} = 1mA, f = 20kHz.

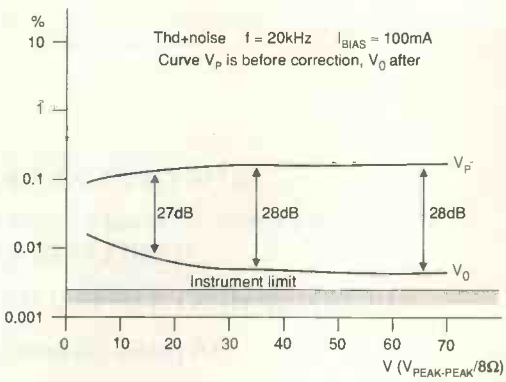


Fig. 13. Total harmonic distortion + noise versus output level. Test conditions: I_{bias}=100mA, f=20kHz.

NEW WAVE MICROWAVES

7: mixers and signal conversion

Microwave signals must be detected and demodulated to reveal their information. Mike Hosking describes components and circuits for this application.

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

So far in this series, solid state devices encountered have been those associated with microwave signal generation and amplification. The complement to this is the extraction of information contained in the signal if any. Information may take the form of carrier frequency and power level together with any amplitude, frequency, phase, pulse or code modulation.

Other things may be important... attenuation, amplitude and pulse modulation and limiting. Phase shifting is also required, together with signal routing, as in multi-throw switches. For instance, phased array radars require electronic control of phase and amplitude to each element for beam shaping and steering; microwave receiver applications generally require input limiting for receiver protection; communication systems often use frequency or channel switching for frequency or time division multiplexing.

Signal detection

Diodes are the mainstay of microwave signal detection. Two forms of circuit are commonly employed: the simple diode detector, **Fig. 1a)**, or down-converter of Fig. **1b)**.

At the higher frequencies, the technology of the diode is different with much attention paid to reducing parasitic reactances associated with packaging and bond wires. Diode types include point contact, a chemically sharpened metallic whisker is brought into physical contact with the semiconductor; the backward diode, a pn junction which operates by tunnelling action. A bipolar transistor or fet junction itself may be used as the non-linear detection element. However, the most common form of microwave diode is the Schottky barrier device.

As operating frequencies increase into the microwave region, charge storage makes the pn junction less effective which limits switching speed. The Schottky diode is a metal-semiconductor junction (**Fig. 2**) made with either Si or GaAs. The junction area is defined by etching a small hole in the oxide passivation layer and depositing a metallic contact (in a choice of metals or alloys) on top of the hole. Bonded contact to the external circuitry is then made to the deposited metal.

With no external voltage applied to the junction, electrons in the n-type layer diffuse into the metal contact, leaving behind a space-charge, or depletion region containing positive charge. This fixed charge tends to inhibit the further flow of electrons until a threshold voltage is reached. This point, the built-in potential difference, is called the barrier height of the Schottky junction. Depending upon choice of metal and semiconductor, this barrier lies typically between 0.3V and 1V. Externally applied voltage beyond this causes the junction to behave as other rectifying junctions.

The key Schottky feature is the fact that the depletion region is highly insulating with virtually no minority carrier current (1pA or less). Thus, the junction stores negligible charge and can switch extremely rapidly between forward and reverse bias making it suitable for high frequency operation.

Simple diode detection finds use in instrumentation: for example, as part of the AGC in signal generators and as a sensor in the measurement of 's' parameters using a scaler network analyzer (**Fig. 3**).

Certain types of radar detection and surveillance receivers also use wideband diode detection for the reception and de-interleaving of complex pulse trains. In addition, as the detec-

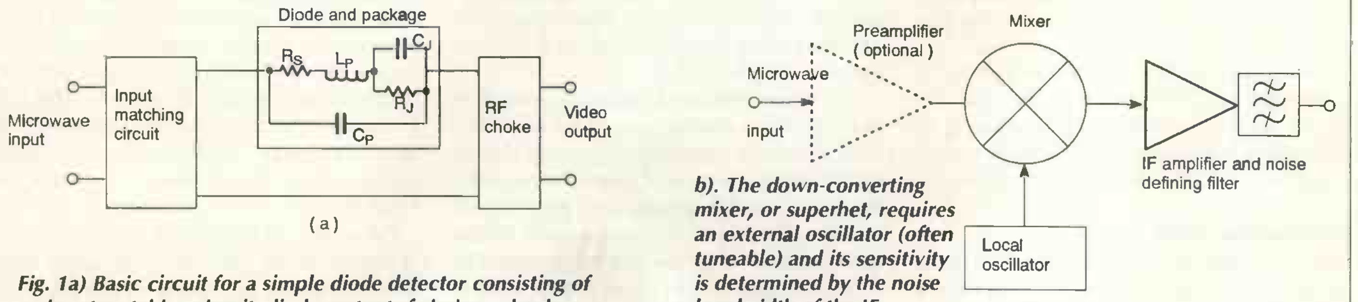


Fig. 1a) Basic circuit for a simple diode detector consisting of an input matching circuit, diode, output rf choke and a dc path for any external bias.

b). The down-converting mixer, or superhet, requires an external oscillator (often tuneable) and its sensitivity is determined by the noise bandwidth of the IF circuitry.

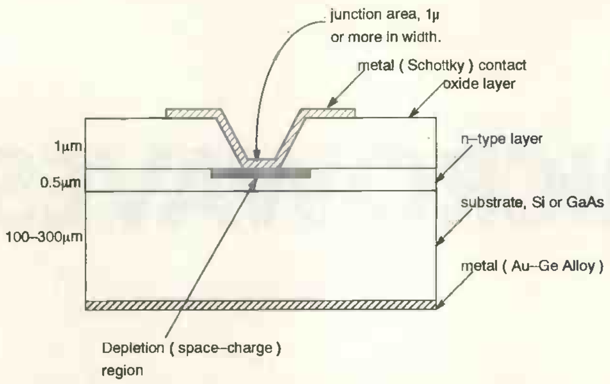


Fig. 2. The Schottky barrier diode is one of the main devices for microwave detectors and mixers, as its virtual absence of minority carriers results in negligible charge storage and fast response.

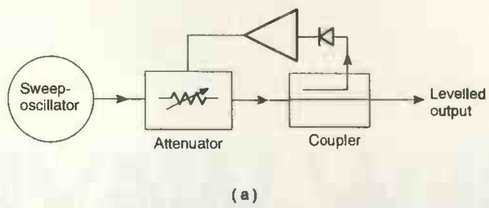
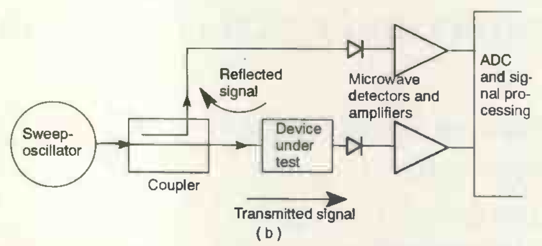


Fig. 3a. One typical application of the diode detector is as part of a levelling loop to sample the amplitude of the main signal so that fluctuations can be reduced.



b). The diode is also found as the wideband sensor in scalar network analysis where the incident, reflected and transmitted signals are used to determine a component's s-parameters.

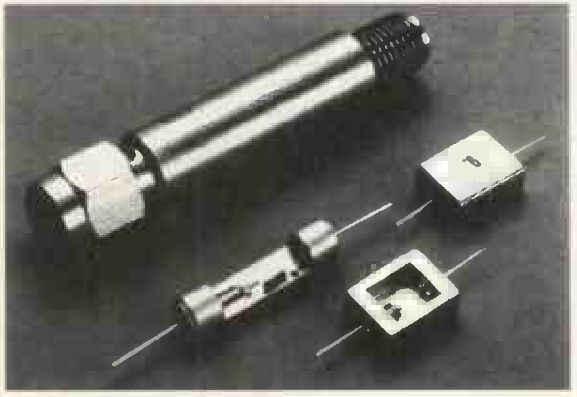


Fig. 4. A collection of diode detector circuit modules available to the microwave designer, already matched and with either standard connectors or with pins for 'drop-in' application. Courtesy of M/A Com Ltd.

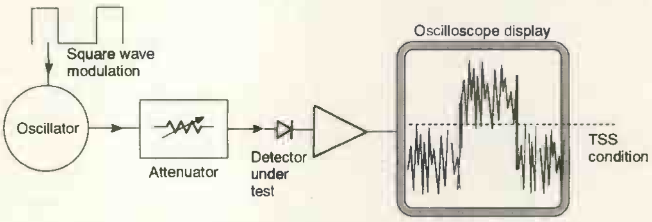


Fig. 5. Equipment used for the measurement of tangential sensitivity, determined by the coincidence of noise peaks on a pulse modulated signal.

tor output voltage is related to the microwave input power, a simple transfer calibration can yield a moderately accurate power meter, suitable for both peak and average power measurement.

Commercially available detectors with integral matching may cover an instantaneous bandwidth of 10MHz to 20GHz. **Figure 4** shows a number of matched and packaged detectors suitable for direct circuit connection or as drop-in modules. A well designed diode and input matching circuit usually requires just an output rf bypass, with possibly a small dc bias and a video amplifier.

Detection sensitivity of the simple diode detector is typically several orders of magnitude less than that of a superheterodyne receiver; The limits are the barrier potential but, more so, the 1/f noise and any dc bias current noise.

From a system point of view, the detector is performing a rectifying function and, thus, the output only contains information relating to the amplitude of the microwave input signal. Frequency and phase information is lost.

Detector specifications

A point of great interest is the minimum input power level which can be detected to provide a useable output. This results in a parameter, unique to microwave applications, called the tangential sensitivity (TSS). For circuits and devices operating at microwave frequencies, there exists a conceptual difference from their lower frequency counterparts in the treatment of voltage and current. These quantities have little physical significance at small wavelengths as they cannot be measured directly. Furthermore, being a function of position within the transmission line circuits, they have no unique value. Instead, circuit analysis is performed using electric and magnetic field distributions and the quantities actually measured are power and impedance. Also, as we saw earlier in the series, the situation is complicated further when a particular circuit element, active or passive, becomes a significant fraction of a wavelength in extent. **Fig. 5** shows the test arrangement for the measurement of TSS: a microwave signal generator, on/off square wave modulation and variable signal amplitude control.

The output of the diode under test is taken to an oscilloscope display via a video amplifier. With no microwave signal present, the display will be just the amplified thermal noise from the test system but, as the signal amplitude is increased from zero, the square wave modulated output from the detector will appear. The TSS is defined as that input power level at which the peak noise level without the signal coincides with the lowest levels on the square wave pulse when the signal is present as indicated in Fig. 5.

Although this is a subjective measurement and operator dependant, it has been found to be repeatable to within about 1dB and is still the method most widely used by diode manufacturers for specifying the low-level sensitivity. The units of TSS are invariably quoted in dB, as a power level with respect to a 1mW reference; i.e. dBm. Values of –50 to –55dBm are typical for good detectors ($<10^{-8}$W). For purposes of calculation consistency, the TSS is usually taken as an input rms signal to noise ratio of 2.5 (4dB) and this should also be referenced to the noise bandwidth and noise figure of the video amplifier.

A further parameter of interest is the actual voltage output from the detector, compared with the microwave power input. This is termed the voltage sensitivity and, while intrinsically having the units of volts per watt, is colloquially quoted in mV/μW. Typical, low power open circuit voltage sensitivities for silicon Schottky barrier and point contact diodes range from about 1mV/μW to 15mV/μW.

The actual achievable sensitivity is dependent on a further parameter: the output resistance of the diode (usually called the video resistance R_v) and its associated loading by any external circuit resistance R_L. For Si

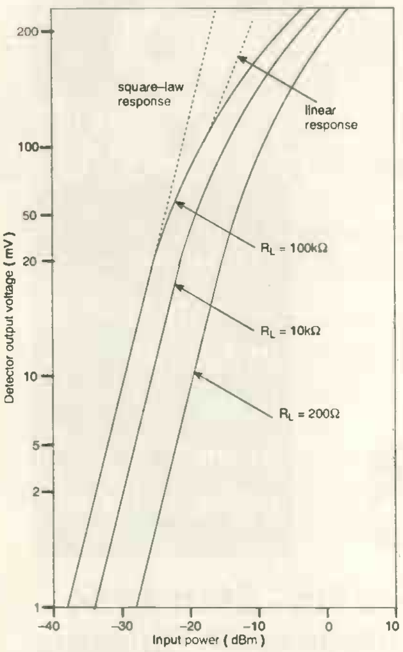


Fig. 6. The diode V/I characteristic passes from a square-law region to linear and into saturation with a forward voltage drop before conduction takes place. In the square-law region, the OUTPUT voltage is directly proportional to the INPUT power.

detector diodes, R_v is typically 1kΩ to several kΩ, although higher values are possible at low microwave frequencies. Thus, the voltage sensitivity will be degraded by the factor and so, for maximum output, R_L must be high.

When detecting pulsed input signals, the fidelity of the pulse output depends upon the time constant set by the output resistance and capacitance. There is, thus, a trade-off which can be made: sacrificing voltage sensitivity with a small value of load resistance (say 50Ω) in order to speed up the rise time.

Input power level also affects the diode parameters and **Fig. 6** illustrates this for several values of load resistance. The main features of the curves are the three regions of different slope. Firstly, at low power levels typically less than a microwatt, the curve follows a square law, so that the output voltage becomes directly proportional to the output power. As the input power level rises, the detection law changes to linear and then starts to saturate. The video resistance and sensitivity properties also change so, for the circuit designer, different matching techniques must be used, depending upon whether the detector is to be optimised for high output, wideband, or flat response.

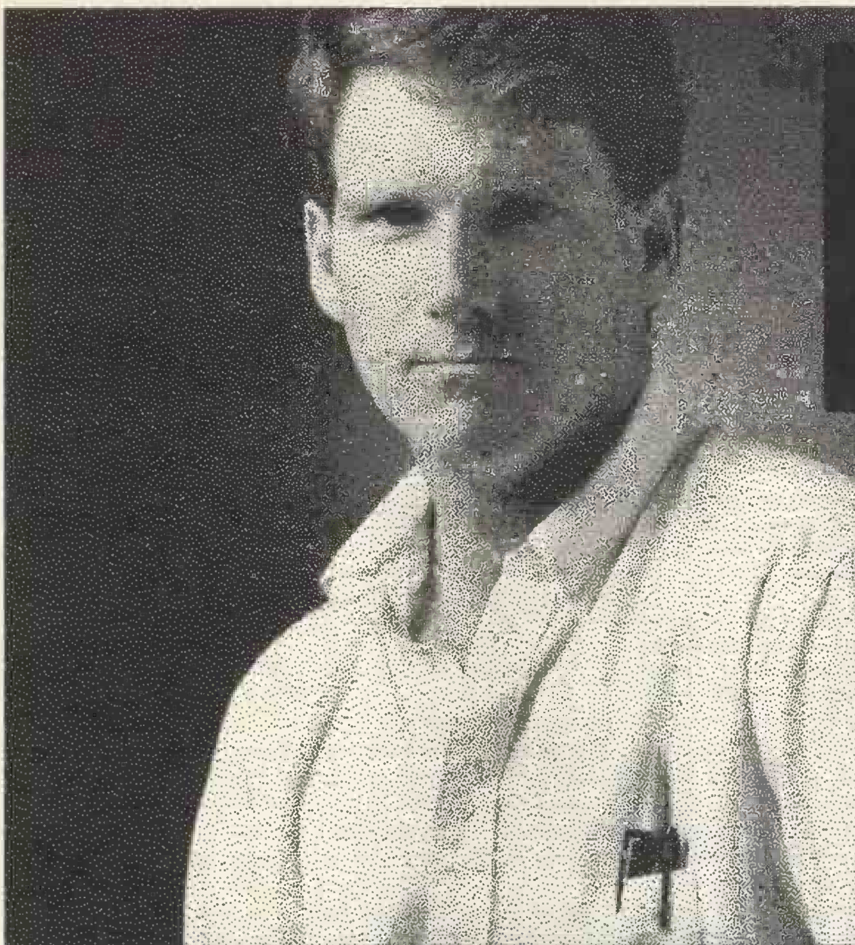
Mixing

Radar and communications systems require the ability to receive signals much lower in amplitude than those which can be detected with a simple diode. A level of one picowatt (10^{-12}W) is not uncommon. For these systems, the process employed uses a separate local oscillator for carrier injection into the mixer circuit. Within the context of this article, the key component is the Schottky barrier diode again, although optimised for the frequency conversion rather than simple detection.

The mixing process combines the microwave input signal with the local oscillator (lo), at the terminals of one or more diodes. The lo signal is normally orders of magnitude larger (about 1mw per diode) than the microwave input and serves to switch the diode(s) between conducting and non-conducting states during each cycle.

This action is analogous to a sampling process and results in the non-linear *V-I* characteristic of the diode generating a theoretically infinite series of harmonics and sidebands of the lo and input signals. In the down-converting mixer, only one of these outputs is required: the difference frequency between the microwave input and the lo (intermediate frequency, IF). Other components generated in the mixing process, shown graphically in **Fig. 7**, are a dc level caused by rectification of the lo signal, together with the sum frequency (or upper sideband) of the two inputs. All of these are rejected by filtering.

Intermediate frequencies of several hundred



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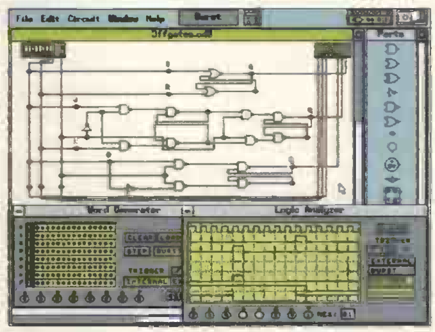
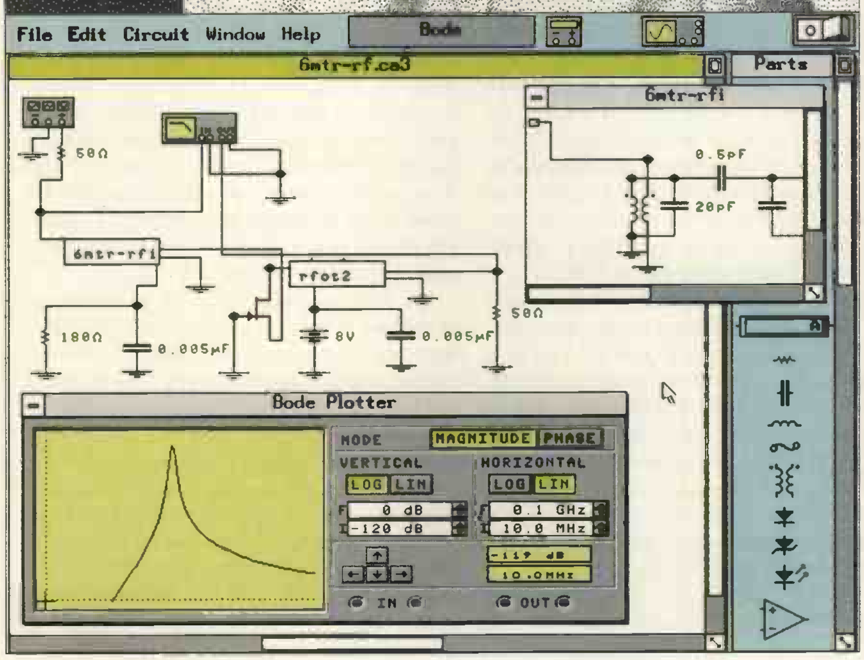


Digital Module: Analyzing a logic circuit.



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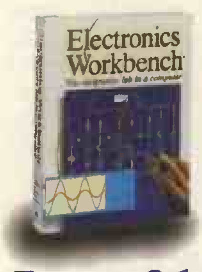

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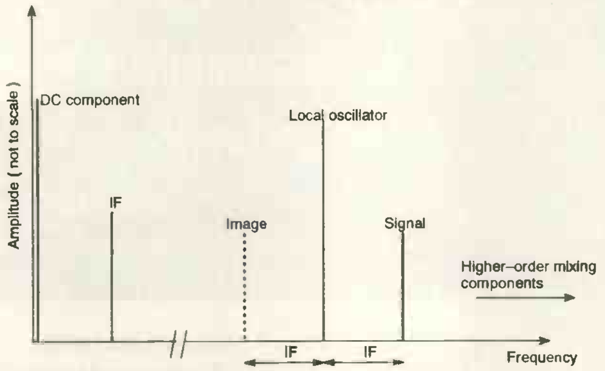


Fig. 7. The spectral components generated during the mixing process consist of a dc level, the IF and image and an infinite series of input signal and lo intermodulation products. Filtering and phase cancellation are used to extract the wanted IF containing the modulated information.

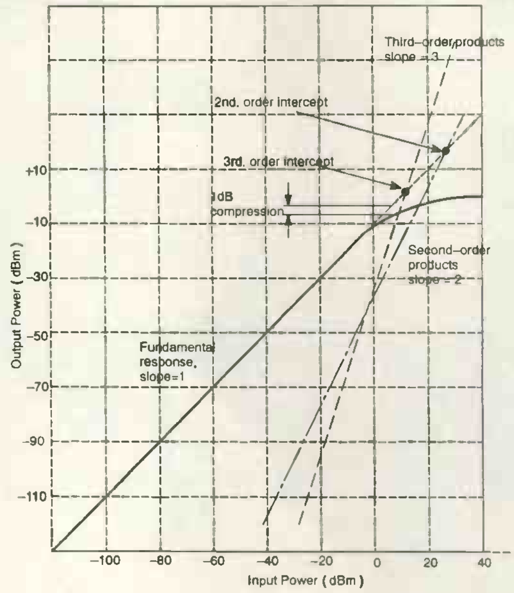


Fig. 8. In determining the performance of a mixer, two key parameters are the 1dB compression point which indicates the onset of saturation and the 3rd. order intercept point which indicates the power level at which intermodulation products will appear.

MHz or less are common in microwave receivers, leading to easier signal processing than would be the case with an input signal at 10's of GHz. Furthermore, all of the information content on the original signal is preserved in the down-conversion process and, knowing the lo frequency, the input carrier frequency can also be deduced.

The switching action overcomes sensitivity problems of the simple detector; The limit for the mixer is set by the efficiency of the conversion process and the thermal noise bandwidth. Other sources of noise, such as lo phase noise, together with the requirement for a finite signal-to-noise ratio, impose further limitations but the overall improvement is, typically, several orders of magnitude over the diode detector.

In very low noise receivers, the lo phase noise is important and so the oscillator can become quite costly. Another point is that whilst the mixer may be broadband in the sense of operating over a wide frequency range, it does not have a wide "instantaneous" bandwidth. This characteristic is fixed by the IF filter response and need only be wide enough to pass the information content of the carrier.Typically, this might be 1-1000MHz and so, if the input frequency is separated from that of the lo by more than this bandwidth, it will not be detected by the mixer. This means that in applications such as surveillance receivers and instrumentation spectrum analyzers, the lo must also be tuneable and often results in a bank of oscillators to cover the full input range. In effect, the mixer receiver sweeps a narrow "window" equal to the noise bandwidth across a much wider RF input band.

Mixer characteristics

Image frequency. The IF produced by an input signal occurs on both sides of the lo. For example, if the lo was at 30GHz and the required IF was 100MHz, then this could be produced by an input signal at either 29.9GHz or 31.1GHz. If the wanted signal frequency was, say, at 31.1GHz then the adjacent or interfering signal at 29.9GHz (called the image) would also produce a simultaneous IF, which could be accepted by the receiver and degrade its performance.

Conversion loss. Conversion loss of the mixer is a measure of the efficiency of the mixing process and is normally quoted in dB. Conversion loss depends on the quality of the impedance match at the RF and IF ports; loss of input power due to the series resistance, R_J and capacitance, C_J of the diode junction. In turn, this loss is also a function of both the lo power level and the ratio of lo frequency to the cut-off frequency of the junction. This cut-off frequency is given by $1/R_JC_J$, which should be as high a value as possible.

Noise figure. Perhaps the most important mixer parameter and related to the conversion loss, is that of the Noise Figure, F. This is a measure of the degradation in signal-to-noise ratio caused by the mixer as the signal undergoes the conversion process and becomes the IF. Noise figure can be related to the conversion loss by the inclusion of a noise temperature, t_m, which takes into account various internal noise mechanisms and the contribution from the IF amplifier, F_{IF}. So,

$$F=Loss(t_m+F_{IF}-1)$$

It has become common (though not universal) practice for manufacturers to quote noise figure assuming a nominal 1.5dB for F_{IF}. Also, t_m is close to unity for many diodes. Thus, the quoted figure must be adjusted for amplifiers of different noise figures. Like many microwave components, the designer has the choice of buying from a wide range of mixers, already packaged and matched, to a quoted noise figure and, in this case, must be careful as to the exact definition implied. Noise performance is usually measured with a wideband noise source and so injects noise power at both signal and image frequencies. This results in a 'double sideband' (dsb) noise figure, whereas it is the 'single sideband' (SSB) figure which is needed. As the DSB output will be twice that of the SSB, the noise figure will appear to be 3dB better... check which is the one quoted.

Dynamic Range and Intermodulation. Most receivers have to operate over a wide range of signal strengths. The lower limit is determined by the inherent noise level, but it is also useful to define an upper limit. In addition, there is no guarantee that a receiver will have only a single wanted frequency at its input; there is a potential problem of spurious responses. Fig. 8 shows the form of a dynamic range plot for a non-linear device relating the input and output power, in this case for a mixer, but applicable in principle to amplifiers as well.

The difference between the absolute values of the two axes is the conversion loss, shown here for convenience as 10dB. The fundamental response curve is that of the IF and it can be seen that this is a linear relationship up to a certain input power, when the mixer starts to saturate. Eventually, at a particular input power level, a 1dB deviation from linear will occur and this is termed the 1dB compression point of the mixer. Linear dynamic range is then defined from the noise level to this point.

If we have the situation of two closely spaced input signals at frequencies f_1 and f_2 (usually called 'tones' in this context) incident simultaneously at the mixer, then the non-linear device characteristic will result in the generation of harmonics, called intermodulation products. These take the form $mf_1 \pm nf_2$, where $m(\neq -1)$ and $n(\neq 1)$ are integers. The order of a harmonic is defined as $m+n$ and the slope of the graph of output power in a particular harmonic against input power is equal to its order. This is shown in Fig. 8 for second and third order intermodulation products, the last of which causes the most concern to mixer users. Its frequency lies close to the wanted IF.

Thus, the better the mixer, the higher the input power level at which this product becomes significant. Manufacturers quote the point at which the third order curve intersects the fundamental as the 'third-order intercept point'. It is, in fact, the point at which the power in the intermodulation product equals the IF power and defines a spurious-free dynamic range for the mixer.

Mixer circuits

Considerable variation exists in the choice of diode mixer design, as specialised circuits can be chosen for particular applications. The number of diodes themselves in a particular mixer can vary from one to eight. Good mixer performance also depends on correctly matched lo and output circuitry. **Fig. 9a** shows a circuit using just one diode, a single-ended mixer. The coupler could be one of the types described earlier in the series, depending on the bandwidth required and the degree of isolation between ports. This last consideration makes wideband operation difficult and there is little scope, other than filtering, to reduce unwanted noise or mixing products. However, the circuit is simple and requires modest local oscillator power, 0dBm or less. A more commonly encountered design is the balanced mixer of **Fig. 9b**. This circuit was illustrated in microstrip form in Part 3. The coupler can be any of the 90° or 180° hybrid types such as branch line or Lange, depending upon the bandwidth. The two diodes of this configuration are connected in reverse polarity: a complete bipolar IF waveform is obtained by summing the diode outputs. This gives a good impedance match at the input ports, together with improved reduction of spurious mixing products. If a 180° hybrid is used then all even harmonics of one of the inputs (the lo, say) can be suppressed.

Another feature is that any am noise on the lo signal appears in antiphase at the two diode outputs and thus tends to cancel at the IF combiner. Harmonic suppression can be extended further using the double balanced circuit of Fig. **9c** to reject all even harmonics of both input signals. Twice the lo power is required, injected via a balun shared with the IF, as there are now four diodes, but the dynamic range is also increased. By extending the basic mixer design still further, it is possible to achieve image rejection without having to attempt what would often be a prohibitively difficult or expensive filtering problem.

Although image and signal inputs produce identical output frequencies, the relative phase of these outputs is different and this can be used as a discriminating factor. Shown in Fig. 9c this type of mixer uses two separate balanced mixers with the signal input routed to each via a 3dB, 90° hybrid coupler; the lo is fed to each at the same phase via a power splitter. A similar, but lower frequency hybrid coupler produces two IF outputs, one of them the upper sideband and the other the lower sideband. It is then only necessary to terminate the output caused by the image.

Such mixers are complex to build over a wide band, but may deliver an image rejection of 20dB or more. However, even if these mix-

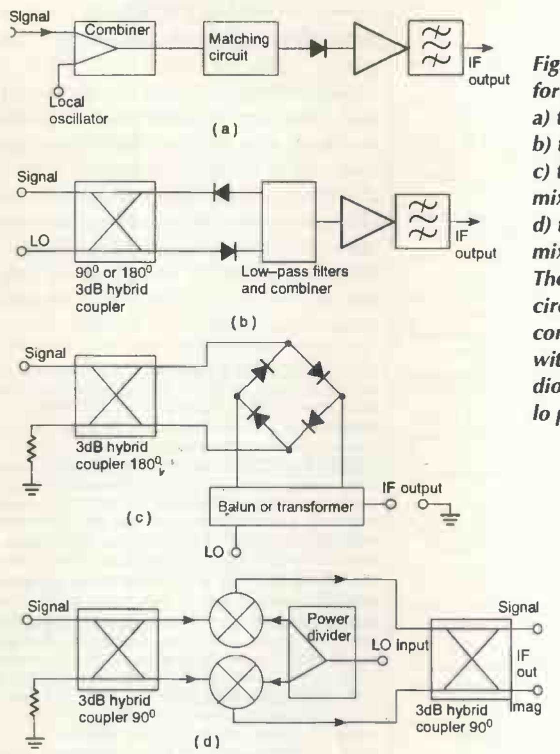


Fig. 9. Schematic circuits for:
a) the single-ended mixer;
b) the balanced mixer;
c) the double balanced mixer;
d) the image rejection mixer.
The complexity of the circuits varies considerably, together with the number of diodes required and the lo power.

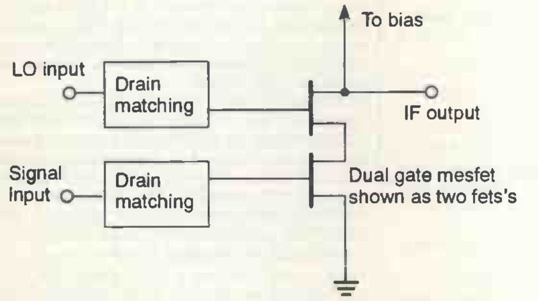


Fig. 10. A dual gate mixer which uses mesfet nonlinearities and features a conversion gain instead of loss. Such devices work into the millimetre wave region.

ers were lossless, there would still be a conversion loss of 3dB because the mixing process generates an equal power second image. By reactively terminating the mixer output at this image frequency, the power can be reflected back onto the diodes in the correct phase to recombine and thus improve the conversion efficiency. The practical improvement obtained is about 1dB.

GaAs mesfets and derivatives may be used as the mixing element. At small drain-source voltages, the mesfet behaves as a voltage controlled linear resistor and thus mixing can take place with both signal and local oscillator signal applied to the gate. Alternatively, the nonlinearity of the transconductance can be used with the signal applied to the gate and the lo applied to the drain. Similar configurations to those above can be realised.

A more elegant technique uses dual-gate transistors to implement the configuration of **Fig. 10**. Such components exhibit a high conversion gain and a lower noise figure than single gate devices; they are also more suitable for integrated circuits and monolithic designs. This is the type of detector often used as the front-end for domestic satellite receivers.

Receiver front end

With the improvements to the noise figure of mesfet amplifiers, it is usual to precede the actual mixer or detector with a low-noise preamplifier. This is because the achievable noise figure from the amplifier can be much less than that from the mixer; so, provided that the amplifier has a reasonable gain, the contribution to the overall noise of the mixer second stage will be small.

For example, if the mixer has a noise figure of 6.5dB and the preamplifier has a noise figure of 2dB with an associated gain of 11dB, then the total noise figure at the input would be 2.12dB and the contribution of the 6.5dB mixer noise figure to that of the overall noise figure is only 0.12dB. There are still considerations, though, of amplifier dynamic range and intermodulation distortion which have to be taken into account in the same way as for the mixer. ■

Next Month. *Superconductors in microwave circuits.*

HIGH voltage versatility

By producing a new and innovative application for this month's free* IC, you could become the owner of one of six mos design kits worth up to £750. Full competition details are on p. 840. Since the IC in question is closely related to a 555 timer with integral 600V power mosfet drivers, the opportunities for new ideas are endless.

International Rectifier's *IR2151* was developed in response to demand for cheaper and more efficient fluorescent lamp ballasts. Energy bills for lighting represent a significant outgoing for many companies, so the pressure for better and cheaper lamps and drivers is great.

Over the past decade or so, electronic ballasts have been gradually replacing traditional passive circuits. Most electronic designs use two power switches in a half-bridge configuration, also known as a totem-pole. The tube circuits comprise *LC* resonant circuits with the lamps across one of the reactances, as shown in **Fig. 1**.

In this circuit the switches are power mosfets driven to conduct alternately by windings on a current transformer. Current in the lamp circuit drives the primary of this transformer, which operates at the resonant frequency of *L* and *C*. Unfortunately, the circuit is not self starting and must be pulsed by the diac con-

*Free *IR2151* oscillator and gate drive IC

The first 500 readers completing the coupon between pages 848 and 849 will receive an *IR2151* fluorescent lamp ballast IC completely free of charge. Additional devices can be bought by contacting Ian Spanswick, Polar Electronics, Cherrycourt Way, Leighton Buzzard, Bedfordshire LU7 8YY. Tel. 0525 377093, fax 853070.

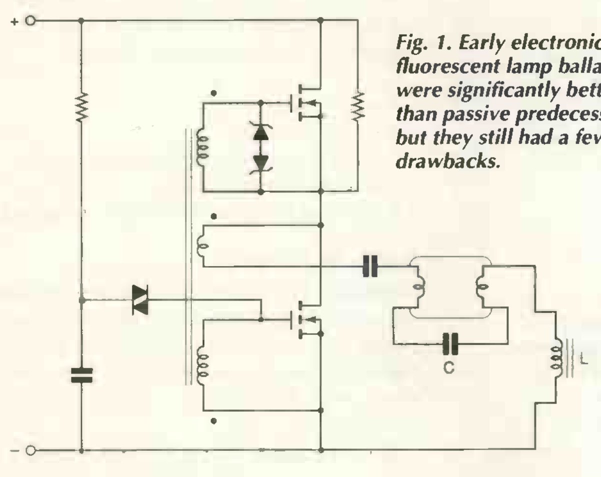


Fig. 1. Early electronic fluorescent lamp ballasts were significantly better than passive predecessors but they still had a few drawbacks.

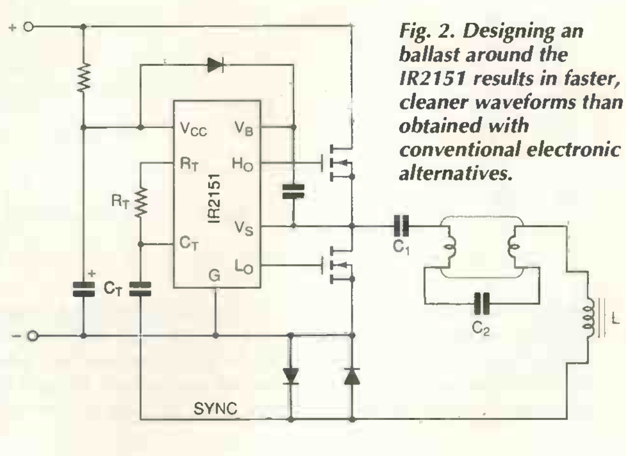


Fig. 2. Designing an ballast around the IR2151 results in faster, cleaner waveforms than obtained with conventional electronic alternatives.

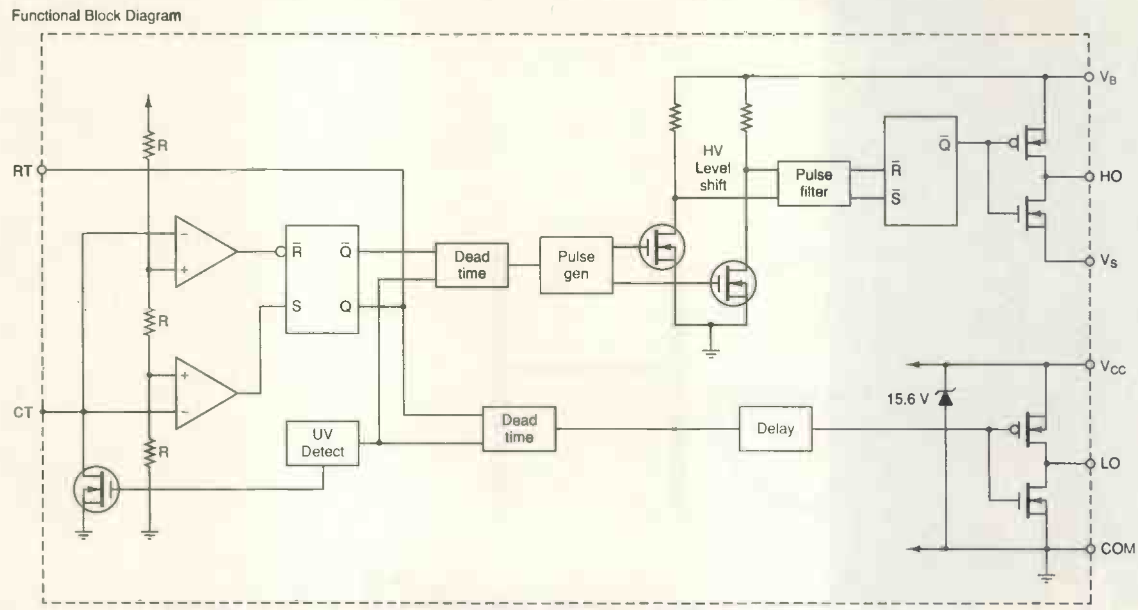


Fig. 3. At the front end of the IR2151 is a timing circuit that is very similar to the 555. Two timing pins are available externally, opening up the possibility for numerous applications other than lamp ballasting.

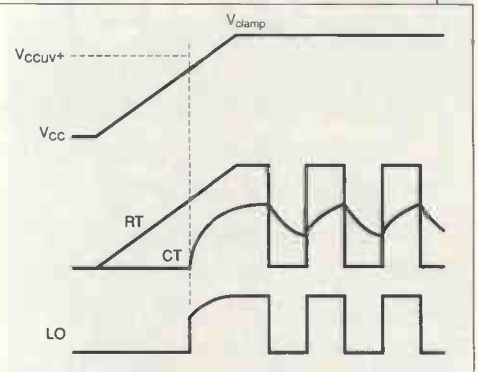


Fig. 4. Timings for the IR2151 when used in a typical self-starting, self-oscillating lamp ballast circuit.

Features of the *IR2151*

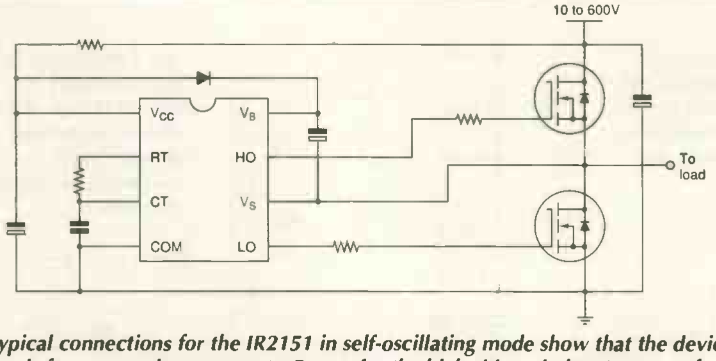
- Floating channel bootstrappable
- Operates to 600V dc
- Tolerant to negative transients
- dV/dt immune
- Undervoltage lockout
- Programmable oscillator frequency
- Matched channel propagation delay
- Low side output in phase with RT pin

The *IR2151* is a high voltage, high speed, self-oscillating power mosfet and IGBT driver with both high side and low side referenced output channels. Output gate drive is 10 to 15V while rise and fall times are 100 and 50ns respectively.

Proprietary high-voltage IC and latch-immune cmos technologies make the device rugged. Its front-end features a programmable oscillator similar to the *555* timer.

Incorporated in the output drivers are a high pulse-current buffer stage and an dead time generator designed for minimum driver cross-conduction. Propagation delays for the two channels are matched to simplify use in 50% duty cycle applications.

The floating channel can be used to drive an n-channel power mosfet or igbt in the high side configuration that operates from a high voltage rail from 10 to 600V.



Typical connections for the IR2151 in self-oscillating mode show that the device needs few external components. Power for the high-side switch gate comes from a bootstrap capacitor of 1μF. This is charged to around 14V whenever V_S is pulled low during low-side power switch conduction. The fast-recovery bootstrap diode blocks dc bus voltage when the high-side switch conducts.

nected to the gate of the lower mosfet.

After initial turn-on of the lower switch, oscillation sustains and a high frequency square wave of between 30 and 80kHz excites the *LC* resonant circuit. Sinusoidal voltage across *C* is magnified by the Q at resonance and develops sufficient amplitude to strike the lamp, which then provides flicker-free illumination.

This basic circuit has been the standard for electronic ballasts for years but is suffers from the following inherent short comings:

– Not self starting
– Poor switch times
– Labour intensive (toroidal transformer)
– Not amenable to dimming
– Expensive to manufacture in large quantity

Inside the *IR2151*

In addition to reducing costs, the International Rectifier *IR2151* removes the drawbacks associated with conventional electronic ballasts. **Figure 2** illustrates the device in a typical configuration. The mosfets shown could be *IRF820* types.

This monolithic power integrated circuit is capable of driving both low and high-side mosfets or igbts from logic level, ground referenced inputs. It provides offset voltage capabilities up to 600V dc, and unlike driver transformers it can provide clean waveforms at any duty cycle between 0 and 99%.

Integrated into the *2151* are a timing circuit, level shifting interfaces and high-voltage mosfet drivers, Fig. 3. Operation of the timing circuit is similar to that of a cmos *555*. As a result, it is possible to define whether the circuit self oscillates, or is synchronised with an external signal. Simply configure the chip's R_T and C_T pins in much the same way as you would those of a conventional *555* timer.

In this type of high-speed mosfet drive circuit, efficiency degrades rapidly if one power mosfet of the pair turns on momentarily before the other turns off. For this reason, the *2151* incorporates a 1μs dead-time generator to help ensure that both power mosfets are never turned on simultaneously. Even when driving power mosfets with 1000pF gate loads, the *2151* is capable of switching on in 100ns and off in 50ns so this safeguard is important.

Propagation delays for the two channels are matched to simplify use in 50% duty cycle applications. When the device is used in self oscillating mode, frequency of oscillation is given by:

$$f = \frac{1}{1.4 \times (RT + 75\Omega) \times CT}$$

Typical timings are shown in **Fig. 4**.

The *IR2155* is intended to be supplied from the rectified ac input voltage and for that reason it was designed for minimum quiescent current. It has a 15V internal shunt regulator so that a single half-watt dropping resistor can be used, assuming 240V ac input. The high voltage rail can be anywhere from 10 to 600V.

Referring again to Fig. 2, note the synchronising capability of the *IR2151* driver. The two back-to-back diodes in series with the lamp circuit are effectively a zero crossing detector for the lamp current. Before the tube strikes, the resonant circuit consists of L, C_1 and C_2 all in series.

Capacitor C_2 has a lower value than C_1 so it naturally operates at a higher ac voltage than C_2. It is this voltage which strikes the lamp.

After the lamp strikes, C_2 is effectively shorted by the lamp voltage drop and frequency of the resonant circuit now depends upon L and C_1. This causes a shift to a lower resonant frequency during normal operation, again synchronised by sensing the zero crossing of the ac current and using the resultant voltage to control the *2151*'s oscillator.

In addition to the quiescent current there are two other components of dc supply current that are a function of the application circuit. One is current due to charging input capacitance of the power switches. The other is current due to charging and discharging junction isolation capacitance of the gate driver.

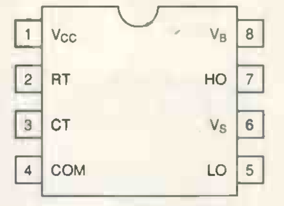
Both components of current are charge related and therefore follow the rule Q=*CV*. To charge and discharge the power switch input capacitances, the required charge is a product of the gate drive voltage and the actual input

IR2151 lead definitions

Pin 2, designated RT, is the oscillator timing resistor input; this resistor normally connects between RT and CT. The signal at this pin is in phase with low-side gate drive output pin LO.

Oscillator timing capacitor input is at pin CT. This capacitor normally connects between CT and the logic and low-side return pin, COM, in order to program the oscillator. Frequency is determined using the equation given in the main article.

High and low-side gate drive outputs are on the HO and LO pins respectively. High-side floating supply feeds via V_B and is returned via V_S. Supply voltage for the low side and logic elements feeds in via the V_{CC} pin. Note that there is an internal zener between this pin and COM so low-impedance supplies above 15V should not be used. Zener voltage is typically 15.6V.



Pin designations for the 8-pin dual-in-line version of the IR2151. The device is also available in small-outline sm packaging.

capacitances and the input power required is directly proportional to the product of charge and frequency and voltage squared:

$$Power = \frac{QV^2}{2} \times f$$

When designing a lamp ballast, follow these pointers. Select the lowest operating frequency consistent with minimising inductor size. Also select the smallest die size for the power switches consistent with low conduction losses. This reduces charge requirements. Usually, dc bus voltage is specified but if there is a choice, use the lowest voltage.

Note that charge is not a function of switching speed. Charge transferred is the same whether the switching speed is 10ns or 10μs.

Because the *IR2151* is designed for off-line supply systems, it contains a zener clamp structure between the chip V_{cc} and the common pin. This diode has a nominal breakdown voltage of 15.6V. Because of the diode, the IC supply voltage is normally derived by forcing current into the supply lead.

Typically, the current is supplied via a high-value resistor connected to the high-voltage supply and decoupling capacitance is connected between V_{cc} and the COM pin. In this way, the internal zener clamp determines the nominal supply voltage. For this reason, the circuit should not be driven by a dc, low-impedance power source with a voltage greater than V_{CLAMP}.

See competition details over page...

Hints and tips on using the *IR2151*

- Never forget – the *IR2151* is a static-sensitive device and should be handled accordingly.
- Limit current into pin 1 to well within the 25mA absolute maximum. Recommended current supplied to this pin is 5mA.
- Do not try to define the voltage at pin 1. This is set internally by a zener.
- Ensure the diode between V_{CC} and V_B has the appropriate reverse recovery capability. Too slow a type and the charge on the bootstrap capacitor could be seriously reduced to the point where efficiency could be affected.
- Remember V_{CT+} at 800V and V_{CT-} at 400V determine the 'toggling' of the S-R flip-flop. Defining the 'charge' and 'discharge' currents enable t_{ON}, and t_{OFF} to be inequalities – thereby determining the 'duty cycle' of the converter (See Fig. 7 of *Application Note AN-995*) Note that 'duty cycle' in this instance refers to inequalities in the t_{ON} of the two channels.
- Observe the waveforms shown in Fig. 4. The apparent 'soft start', determined by the undervoltage detect circuit, might cause problems with transformers because of asymmetry.
- Note that taking V_{cc} low automatically 'disables' both outputs. This facility allows for protection circuits provided the previous clause is taken into account.
- Observe the dV_s/d_t rating in the data sheet. Exceeding this limit could lead to false triggering of the two outputs.
- A continuation of the previous point is the requirement of ensuring pin 6 never goes more than 25V negative with respect to pin 4. Failure to observe this limit could result in false triggering or to failure of the IC.

£1800 COMPETITION

Win one of six design kits

For the six most innovative (and practical) circuits using the IR2151 in any application other than a fluorescent lamp ballast, International Rectifier is giving away a design kit comprising a host of semiconductors complete with storage facilities.

First, second and third best designs will receive prizes valued at £750, £500 and £250 respectively and a further three designs will each receive components and storage facilities to the tune of £100. Components included in the kits have been chosen from across the IR range, which incorporates power mosfets and gate driver ICs.

Since the IR2151 is essentially a 555 timer with high-voltage output, the scope for imagination is wide. Power conversion and motor control are obvious examples, but the chip need not necessarily be used to drive mosfets or igbts.

This competition is open to all EW+WW readers. Entries must reach EW+WW's editorial offices at Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS before December 1 1994. Please mark your envelope clearly with 'IR design competition'.

The best of the designs will be published in EW+WW. Copyright for all submissions will be assigned to International Rectifier.

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Access

VISA

Smaller steps to better performance

Active multipliers find many uses in instrumentation. Ian Hickman explores their use in signal generation as variable attenuators and modulators.

My DDS signal generator, self-designed, covers about 1Hz-320MHz in approximately 1Hz steps. Output is cw only and an external 0-120dB step attenuator provides 10dB steps.

To add amplitude modulation and fine output adjustment, certain constraints had to be considered. For example, though output from the DDS signal generator was about 0dBm, variation over the full range up to 320MHz was nearly 2dB.

Furthermore, several rf amplifier stages would be needed in the rf path, with each contributing further gain variation. So the whole rf path through the modulator/attenuator was enclosed within a levelling loop. Since the leveller output is maintained constant regardless of input- or *load*-variations (within reason), the added advantage is that this represents a zero output impedance point. From there, the load can be supplied via a 50Ω resistor, giving an ideal generator output impedance, independent of the actual output impedance of the last rf amplifier in the chain.

AD834 characteristics

The *AD834* accepts a maximum differential input on both its *X* and *Y* balanced inputs of ±1Vpk-pk, producing at its differential *W* output port a current of ±4mA full scale, according to the relation $W = XY/1V^2$. Output from the DDS signal generator was about 0dBm, or only 630mV pk-pk, so some amplification was indicated to take full advantage of the multiplier's ±1V dynamic range. With frequencies up to 320MHz at the multiplier's input, balanced circuitry is not convenient. Fortunately the device's common mode rejection at both *X* and *Y* input ports is such that in each case one lead can be grounded and the other driven unbalanced (**Fig. 1**).

As recommended in the data sheet when using unbalanced inputs, X_1 and Y_2 are grounded, the inputs being applied to pins 1(Y_1) and 8 (X_2).

Similarly, for convenience, an unbalanced output is taken from W_1, W_2 being returned to the supply, even though this halves the available pk-pk output current to ±2mA. This alternating current is superimposed on a standing 8.5mA (nominal) dc component and is sunk by an open collector output.

The open collector *W* outputs must be operated at a voltage slightly above that on pin 6 (V+) – the manufacturer's recommended method is to insert a resistor in series with the supply to pin 6.

The ac component, flowing in 47Ω load resistor R_6 in parallel with the (nominal) 50Ω input impedance of IC_3, forms the output voltage from the modulator, and is applied to the following amplifier stages.

Note that the 390Ω supply resistor of IC_3 is somewhat lower than the recommended value, so device dissipation will be increased. But this is acceptable for lab use as distinct from a full-temperature-range application.

After further amplification, the signal voltage at the output of IC_4 corresponds to +6dBm when the voltage on the Y_1 input of the modulator is rather less than +1V (or –1V). Thus the level delivered to a matched load at the output is just 0dBm. The rf output of IC_4 is dc restored positive-going by D_1, whose linearity versus signal level is improved somewhat by a hint of

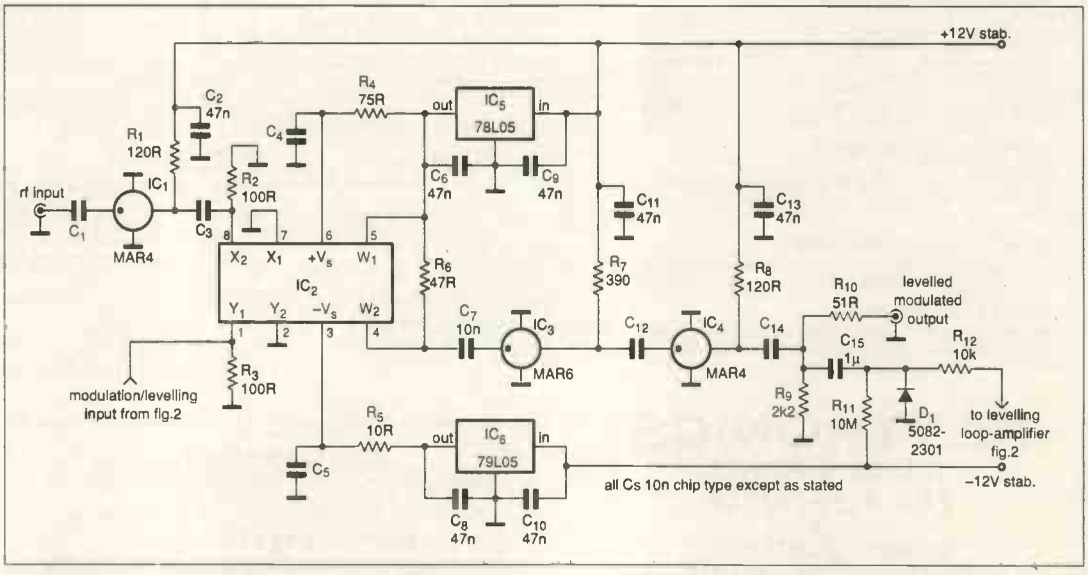


Fig. 1. RF path through the modulator/attenuator/leveller.

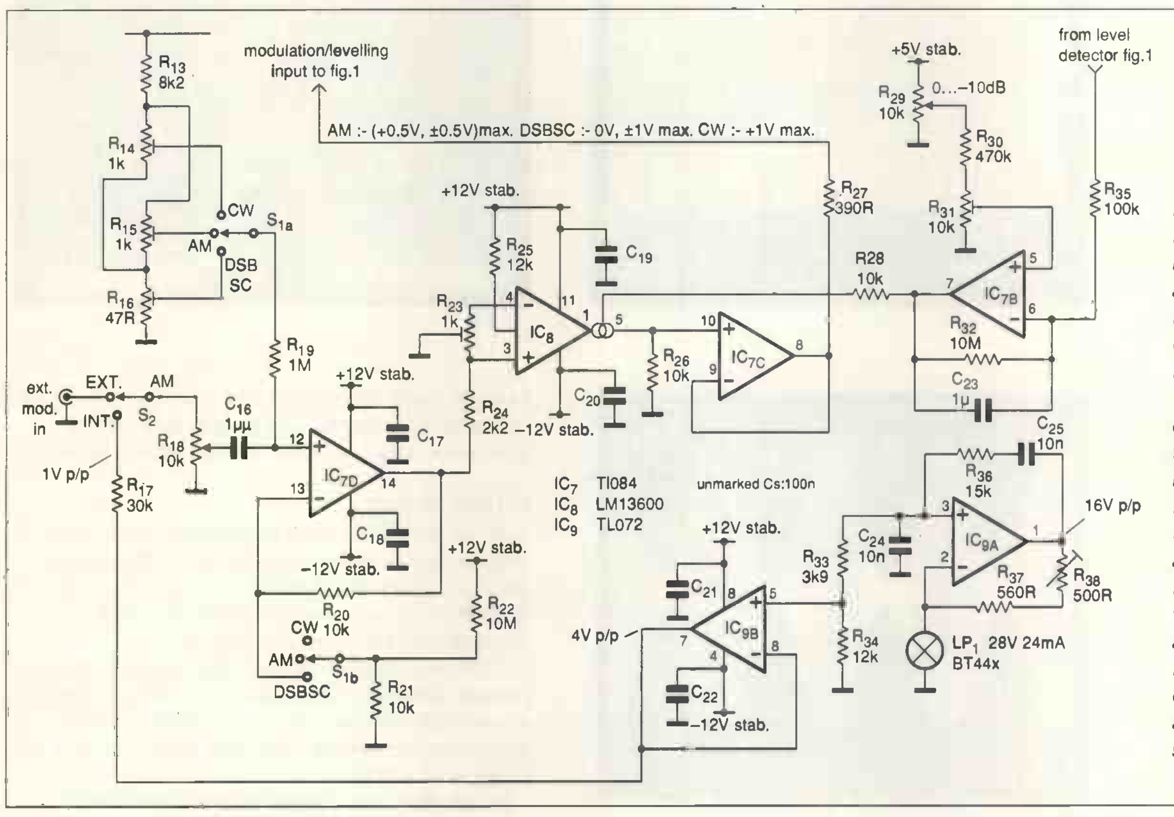


Fig. 2. The remaining modulator/ attenuator/ leveller circuitry, including a 1kHz oscillator for the internal modulation facility (IC_9), the loop amplifier/filter IC_{7B}, the variable gain amplifier (ota IC_8) and the carrier/modulation scaling stage IC_{7D}.

forward bias via R_{11}.

Mean level of the voltage at D_1's cathode is (almost) equal to the peak rf voltage, and it is applied via R_{12} to the modulator/attenuator/levelling loop (**Fig. 2**).

Remaining circuit

The circuitry of Fig. 1 was constructed on a scrap of single sided copper-clad laminate, used as a ground plane. But that of Fig. 2 was built on a piece of 0.1in matrix copper strip board. The level detector output is applied to the loop filter-amplifier, IC_{7B} and associated components.

IC_{7B} controls the gain of the transconductance amplifier stage IC_8 which receives its input from IC_{7D}.

IC_8 is an *LM13600*, of which one half is unused. (An *LM13700* will do too as there is only a minor difference between these two devices. In the *LM13600* the emitter current of the input transistor of the Darlington output buffer is controlled pro-rata with the g_m of the transconductance section, providing improved dynamic range. In the *LM13700*, it is fixed. Since the output buffer is not used in this application, either device will do.)

R_{25} provides the current to operate the *LM13600*'s input linearising diodes.

IC_{7D} produces a dc level which determines the carrier at the output, combining it with an ac signal where modulation is required. For maximum cw output, IC_{7C} (which buffers the ota's output) applies the required voltage, up to +1V, to the modulator's Y_1 input, via R_{27}. This occurs with 0-10dB attenuator control R_{29} set to maximum and with R_{31} suitably adjusted. R_{29} provides an attenuation range of over 10dB, and though its operation is approximately linear rather than logarithmic, it can readily be calibrated with a dB scale. Operation on am is similar, except that the dc level at the modulator's Y_1 input is halved to allow for up to 100% modulation.

Internal 1kHz modulation oscillator IC_9 is included and the modulation depth can be set by R_{18}. R_{18} can be calibrated directly in percent amplitude modulation depth, with the level supplied by the internal modulation oscillator making fully-clockwise equal to 100% am. If the internal modulation oscillator is run with a low output swing at IC_{9A}, such as 4Vpk-pk, setting up is critical and amplitude control poor, due to inadequate drive to the lamp used to stabilise the loop gain. With the arrangement shown, giving 16Vpk-pk, control is tight with little amplitude bounce at switch-on. A 4Vpk-pk output is picked off by IC_{9B}, which is driven from the frequency selective network rather than the output of the maintaining amplifier. IC_{9B}'s output has the advantage of the selectivity of the Wien network – though it amounts to 2.5dB at third harmonic relative to the fundamental, every little is worth having.

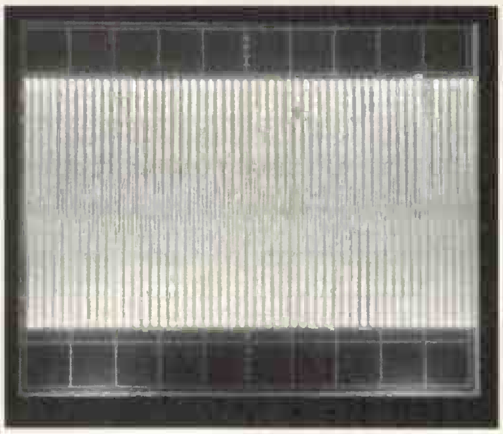
The measured total harmonic distortion at IC_{9B}'s output is 0.01% – almost entirely second harmonic – is presumably due to the IC, as any due to the lamp should be odd order. The result is a little puzzling, as the op-amp spec shows 0.003% thd typical. Nevertheless, the performance is quite creditable for such a cheap, simple circuit.

As an alternative, an external modulation source could be connected, which naturally should not exceed 1Vpk-pk, or the modulation index will exceed 100% when R_{18} is at maximum.

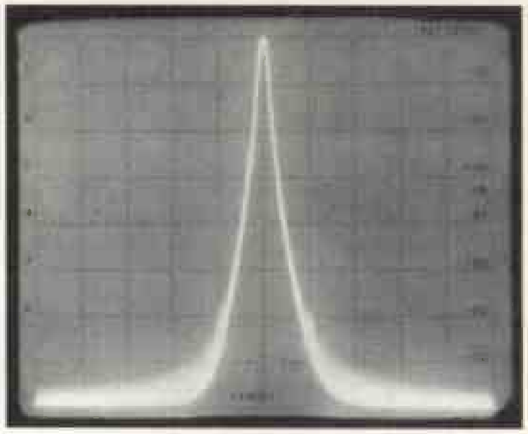
Since it is easily incorporated and could come in useful, a dsbsc (double side-band suppressed carrier) mode is also included.

Switching S1 to dsbsc doubles the gain of IC_{7D} to give a bipolar drive at the modulator's Y_1 modulation

Fig. 3(a). CW output at 10MHz into a 50Ω load (viz a spectrum analyser). Scope settings 0.2V/div, 0.5μs/div.
3(b). Spectrum of (a). Centre frequency 10MHz, 20kHz/div, 3kHz if bandwidth, video filter off, 10dB/div, ref level (top of screen) 0dBm.

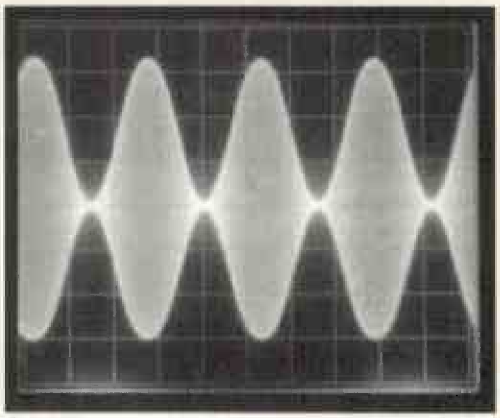


(a)



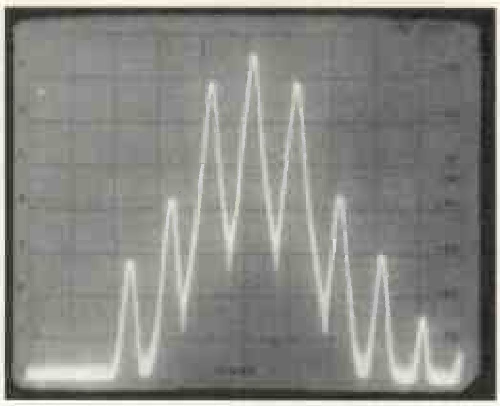
(c)

Fig. 4(a). Output at 10MHz with 100% am at 20kHz. Scope settings 0.2V/div, 0.5μs/div.



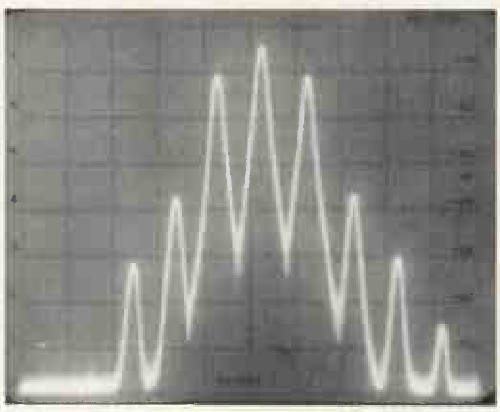
(a)

4(b). Spectrum of (a). Centre frequency 10MHz, 20kHz/div, 3kHz if bandwidth, video filter off, 10dB/div, ref level (top of screen) 0dBm.



(b)

4(c). As (b) but the input rf level reduced by 6dB. Switching to the 1dB division display indicated that the loop compressed a 6dB reduction in input level to a 0.2dB reduction in output level.



(c)

input. R_{16} with R_{22} permits zeroing of any offset at the output of IC_{7D}, and R_{23} can then be adjusted for maximum carrier suppression in the dsbsc output.

Circuit testing

Tests on the circuit, still in breadboard form, show **(Fig. 3a)** the maximum cw output at 10MHz, while the same signal displayed on a spectrum analyser **(Fig. 3b)** indicates a level of 0dBm into 50Ω downstream of the source resistor R_{10}.

Of the two, I believe the spectrum analyser shows the true picture, the scope indicating about 1.1vpk-pk or over 4dB more than this. But you can't believe a *X10* probe, with its 4in earth lead, even at as "low" a frequency as 10MHz.

In fact, the scope trace is included solely for comparison with the am case.

Examining the same 10MHz output, but this time with 100% am and the same scope settings as before, shows that the peak to peak voltage has increased slightly. This is confirmed by the spectrum analyser picture (**Fig. 4b**).

With 100% modulation, had the peak voltage been the same as before, the carrier component in the scope trace would have been exactly –6dBm, ie 6dB below that in Fig. 3b. The fact that it is barely 5dB down indicates that while the levelling loop tries to control the peak level of the output, it is also partly sensitive to the mean rf level. Shortening the time-constant $C_{15}(R_9+R_{12}+R_{35})$ compared to the period of the modulation waveform, would result in mean level control and the carrier level would remain unchanged with modulation depth. But this would limit the highest usable modulating frequency to many octaves below the lowest usable radio frequency – at least in am mode. Restriction-free range for both modulation- and carrier-frequency operation is a priority so the present scheme with (near) peak level control has been retained.

AM and dsbsc modes

At 100% modulation, the second and third harmonic sidebands are only about 26dB and 40dB down respectively **(Fig. 4b)** on the wanted fundamental sidebands. The second harmonic modulation is severe enough to be noticeable (Fig. 4a) as a reduction in amplitude of the negative-going peaks of the envelope relative to the positive. As it is the negative peaks which are sensed by the detector circuit, this largely explains the deviation from true peak-level control exhibited by the levelling loop, mentioned above.

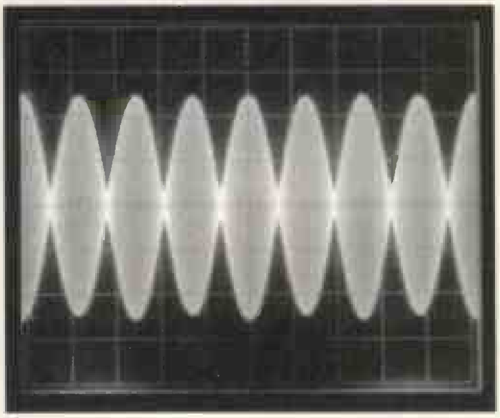
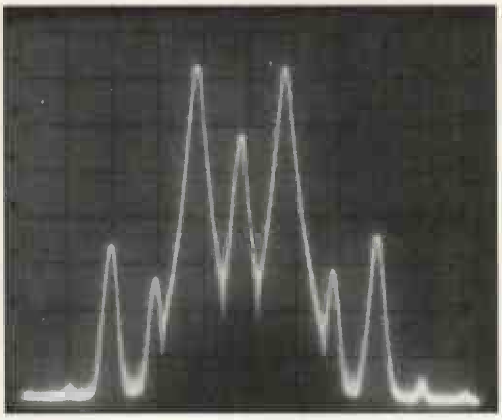

(a)

Fig. 5(a). Output at 200MHz in dsbsc mode with 20kHz modulation. 0.2V/div vertical (but effectively uncalibrated at this frequency, on account of the probe earth lead inductance), 20µs/div horizontal.


(b)

5(b). Spectrum of (a). Spectrum analyser settings as Figure 3(b) except centre frequency is 200MHz.

The levelling loop is really quite effective (**Fig. 4c**), compressing a 6dB reduction in input level to a 0.2dB reduction at the output. Similarity of Figs. 4b and c shows that the second and third order distortion sidebands arise not in IC_1 nor (according to its spec sheet) in IC_2, but in ICs *3* and *4*, both of which are running just a few dBs below their 1dB compression point. An improvement would be to modify the detector circuit to sense the positive peak, or better still use a peak-to-peak detector.

In all modes – cw, am and dsbsc – R_{29} provides the function of a 0-10dB output attenuator, though, for the reasons indicated, with am the modulation should be adjusted for the desired depth before the output level is set.

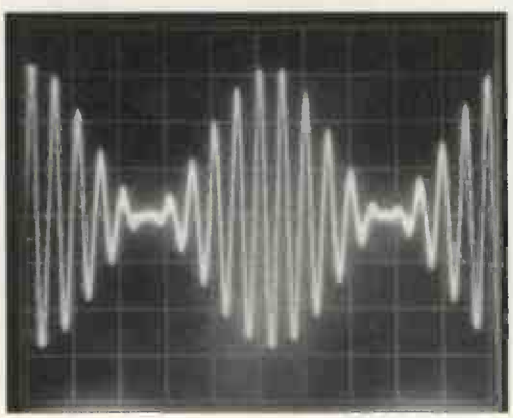


Fig. 6. 10MHz output with 100% amplitude modulation at 1MHz, measured upstream of the 51Ω output resistor R_{10}. 0.2V/div. vertical, 0.2µs/div horizontal.

At 200MHz in the dsbsc mode, the Y_1 input of the modulator is taken both positive and negative, on alternate half cycles of the 20kHz modulating frequency (**Fig. 5a**) So phase of the rf reverses twice per cycle of the modulation. The corresponding spectrum (**Fig. 5b**) shows that the carrier is only 15dB down on the 199.980MHz and 200.020MHz sidebands – despite adjustment of R_{23} for maximum carrier suppression.

The residual carrier is a component in quadrature with that controlled by R_{23} and is not affected by the nulling procedure. It is presumably due to capacitive feedthrough in, or around, IC_2. As expected in this mode, second order distortion sidebands are way down, much lower than third order.

In dsbsc mode, output level is set solely by R_{29} – the "output attenuator". Any external modulation input should be set to 1Vpk-pk and modulation depth control R_{18} to maximum. In this mode, "modulation depth" is meaningless. Whatever the modulation input level, the loop will always try to set the peak output level to that demanded by R_{29}.

Surprising operating frequency

The circuit will operate with input carrier frequencies down to about 1MHz. For lower frequencies, all the capacitors in the rf path, such as C_1 and C_3 etc, should be increased in value. Similarly, the external modulation facilities, while not dc coupled, should work down to a few Hz.

Highest usable modulation frequency is set by the frequency responses of IC_{7C} and IC_{7D}, and IC_8. Response of the modulator's Y input (like its X input) extends up to 500MHz.

An unexpected result was produced by a 10MHz carrier 100% amplitude modulated at 1MHz, monitored at the junction of C_{15} and R_{10}. The modulation envelope seems a very respectable sine wave (**Fig. 6**).

Finding that the *LM13600* worked quite happily at this frequency was no surprise, its typical 3dB bandwidth is 2MHz. But the *TL084* will typically swing only ±2.5V into 2kΩ at 1MHz, even on ±15V rails. So how was IC_{7C} coping on ±12V, with a load of around 500Ω? A quick check with a scope at IC_7 pin eight showed that its output – a sine wave swinging entirely positive from 0V upwards – was distinctly poor. The positive peak was nicely rounded but the negative peak at 0V was distinctly pointed, though this does not show up very well in the scope trace.

You can't get a quart out of a pint pot after all!

Extending the range

For my requirements, the design is adequate and only remains to be recast in a tidier form than the present breadboard.

But clearly the usable modulating frequency range could be greatly extended by substituting more modern, faster, op-amps in place of IC_{7C} and IC_{7D}, and using a faster variable gain amplifier.

To go with current feedback op-amps, the ultimate choice for the variable gain amplifier is obviously yet another *AD834*. AM or dsbsc could then be produced with modulating frequencies up to hundreds of MHz.■

LETTERS

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

Split decision

I think there is scope for confusion in Norm Dye and Helge Granberg's article concerning hybrids (Using RF Transistors: Combined efforts bring power pay-offs, August, 1994). In particular, the rat-race and Wilkinson hybrids are classed as 90°, or quadratic couplers (p. 696).

I thought that a quadratic coupler or hybrid was one which split a single input into two outputs differing in phase (from each other) by 90°; or combined two inputs, differing in phase by 90°, into a single output.

The line coupler and branch line coupler come into this category, but the rat-race and Wilkinson do not. It would have been helpful if Figs. 4, 5 and 6 could have been labelled to show the relative phases of inputs which enable the hybrids to operate with the particular input and output ports. This would have indicated the differences clearly.

In the two diagrams of the hybrid shown in Fig 3, the phase of the output on the right hand side should be –90° not 90°, and in the description of the operation of the rat-race, I think it helps to substitute "combiner" with "splitter" at one point, but I still remain slightly confused.

If any reader wants to know the theoretical bandwidths and other characteristics of these and other hybrids when splitting or combining, I would recommend they refer to 'Hybrid networks and their uses in rf circuits' (*The Radio and Electronic Engineer*, Vol 54, No 11/12. pp.473-489, Nov/Dec 1984.

Dick Manton
Surrey

Listening post

Contrary to Jerry Mead's view (*Letters*, Nov 93) an open mind is not "the most valuable tool in any scientific research project". Of greater importance are a critical mind, lots of hard work and a thorough knowledge of the technology of the relevant fields – usually called "theory" but mostly based on fact.

An open mind is important for researchers to realise that their endeavours may be misdirected, but I feel sure Jerry would not suggest we need to keep an open mind on the phlogiston theory or the concept of a flat earth.

Jerry says he ranks "developmental listening" (whatever that is, he does not say) "as being as valid in the design of an amplifier as the quest for product safety, long term reliability, applications suitability and an acceptable cost of production". But while safety, reliability, suitability and cost are important engineering constraints in the design of audio amplifiers, they are secondary to output power, sensitivity, frequency response and distortion – all of which are vital to audible performance.

Back in January 1978, in *Wireless World*, Peter Baxandall claimed to have designed an amplifier that "was not listened to at all, but subsequently came top in an independent subjective assessment of many competitive designs from various countries"; and that "Quad... adopt the attitude that if you understand what you are doing thoroughly enough, there is no need for listening tests during the design and development of amplifiers".

Jerry may argue that this is ancient history but in the intervening years I have heard no plausible reason why this should not still be true.

Like Doug Self (*Letters*, January, 1994), I have little confidence in Jerry Mead's experimental procedure which shows scant regard for established testing methods. He would do well to take up some of the recommendations given in Lipshitz and Vandekooy's excellent paper ('The great debate: subjective evaluation', *J Audio Eng Soc*, Vol 29, No 7/8, July/August, 1981).

It is too easy to trick, even experienced, listeners into believing that they hear non-existent artefacts during audio testing. Late last year I took part in a listening test to determine the audible effects of different audio cables. The test was inconclusive, and though no-one demonstrated any real consistency in identifying the correct cable the best results were from one subject who readily admitted he was guessing.

For the test proper we listened to some modern music with which I was unfamiliar. It was spectrally fairly simple in that it had little dynamic nature and few transients, but it sounded sweet enough.

After the test I asked if I could play a CD of my own. The more transient nature of the piece clearly showed that the sound was distorted and was very unpleasant – those who owned the gear probably thought so too, judging by their reaction.

The equipment was based around an ME pre-amp and power amp, designed, I am told, to exhibit fairly high levels of distortion (a few percent). With classical music, it was quite obvious and there was no doubt that a $100 personal cassette player could easily deliver superior performance.

My point is that the perceived performance of any audio equipment depends strongly on the music played. It also shows the futility of increasing the distortion levels in audio equipment to suit the personal tastes of a few, since this can only be done at the cost of reducing the suitability of that equipment for general use.

I have been following the debate over the subjectivists' claims for many years. Perhaps the most annoying aspect is how often they claim to have discovered new 'sonic' artefacts – usually due to poor experimental method – with new theories sought to explain those artefacts while ignoring perfectly valid theories which are already tried and tested.

Such ignorance will continue to attract scorn and their claims are likely to be dismissed with 'facility and derision' for some years to come.

Phil Denniss
University of Sydney
Australia

Radar replication

George Pickworth's account of the coherer is an outstanding example of the way that technical history is enhanced by modelling ancient hardware. I should like to suggest an expansion of his work.

It seems generally accepted that the first working radar was made by Christian Hulsmeyer in 1904. His patent describes a spark transmitter and a conventional coherer mounted one above the other – the decoherer was coupled to an alarm bell which rang when a signal was received. A description of the apparatus appeared in *Wireless World* under the title 'The Telemobilscope'.

There seems little doubt that the device would not work if installed on board ship if for no other reason than that the coherer would almost certainly be reset by vibration (it is interesting to note that the large induction coil was mounted in gimbals to keep the whole assembly vertical; Hulsmeyer's is certainly the first radar to use a stabilised antenna.

A demonstration given in Rotterdam in 1904 seems to have persuaded members of a maritime convention that the Telemobilscope worked and had a range of around 3km, yet examination of the patent specifications casts-considerable doubt on Hulsmeyer's claims.

Replication is the only way to establish if Hulsmeyer deserves his place in the first chapter of so many books published since WWII.

Harold W Shipton
St Louis
USA

Crossover critic

In discussions about the merits of precious-metal loudspeaker cables, I cannot recall anyone raising the issue of crossover networks.

The path between amplifier and listener contains an easily measured performance limiter – the impedance of the crossover network – rising to several ohms. It is likely to exceed the cable loop impedance at only a few hundred hertz either side of the crossover point and the result is frequency-dependent drive-unit damping which significantly affects transient response.

Based on this I have a number of questions:

Why is the 'golden ear' brigade (who appear to detect small changes in cable impedance) not clamouring for systems having the minimum of impedance between amplifier and drive unit?;

why are cable manufacturers not insisting on the removal of filters

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2 x Reed Relay Kits. You get 8 reed switches and 2 coil sets. Order Ref: 148.
12v-0-12v 6va Mains Transformer. PCB mounting. Order Ref: 938.
1 x Big Pull Solenoid. Mains operated. Has ½" pull. Order Ref: 871.
1 x Big Push Solenoid. Mains operated. Has ½" push. Order Ref: 872.
1 x Mini Mono Amp. 3W into 4 ohm speaker or 1W into 8 ohm. Order Ref: 495.
1 x Mini Stereo 1W Amp. Order Ref: 870.
15v DC 150ma PSU. Nicely cased. Order Ref: 942.
1 x In-Flight Stereo Unit is a stereo amp. Has two most useful mini moving coil speakers. Made for BOAC passengers. Order Ref: 29.
1 x 0-1mA Panel Meter. Full vision fact 70mm square. Scaled 0-100. Order Ref: 756.
2 x Lithium Batteries. 2.5V penlight size. Order Ref: 874.
2 x 3m Telephone Leads. With BT flat plug. Ideal for 'phone extensions, fax, etc. Order Ref: 552.
1 x 12V Solenoid. Has good ½" pull or could push if modified. Order Ref: 232.
4 x In-Flex Switches. With neon on/off lights, saves leaving things switched on. Order Ref: 7.
2 x 6V 1A Mains Transformers. Upright mounting with fixing clamps. Order Ref: 9.
2 x Humidity Switches. As the air becomes damper, the membrane stretches and operates a micro switch. Order Ref: 32.
5 x 13A Rocker Switch. Three tags so on/off, or changeover with centre off. Order Ref: 42.
Mini Cassette Motor. 9v. Order Ref: 944.
1 x Suck or Blow-Operated Pressure Switch. Or it can be operated by any low pressure variation such as water level in tanks. Order Ref: 67.
1 x 6V 750mA Power Supply. Nicely cased with mains input and 6V output lead. Order Ref: 103A.
2 x Stripper Boards. Each contains a 400V 2A bridge rectifier and 14 other diodes and rectifiers as well as dozens of condensers, etc. Order Ref: 120.
12 Very Fine Drills. For PCB boards etc. Normal cost about 80p each. Order Ref: 128.
5 x Motors for Model Aeroplanes. Spin to start so needs no switch. Order Ref: 134.
6 x Microphone Inserts. Magnetic 400 ohm, also act as speakers. Order Ref: 139.
6 x Neon Indicators. In panel mounting holders with lens. Order Ref: 180.
1 x In-Flex Simmerstat. Keeps your soldering iron etc always at the ready. Order Ref: 196.
1 x Mains Solenoid. Very powerful as ½" pull, or could push if modified. Order Ref: 199.
1 x Electric Clock. Mains operated. Put this in a box and you need never be late. Order Ref: 211.
4 x 12V Alarms. Makes a noise about as loud as a car horn. All brand new. Order Ref: 221.
2 x (6"x4") Speakers. 16 ohm 5 watts, so can be joined in parallel to make a high wattage column. Order Ref: 243.
1 x Panostat. Controls output of boiling ring from simmer up to boil. Order Ref: 252.
2 x Oblong Push Switches. For bell or chimes, these can switch mains up to 5A so could be foot switch if fitted in pattress. Order Ref: 263.
50 x Mixed Silicon Diodes. Order Ref: 293.
1 x 6 Digit Mains Operated Counter. Standard size but counts in even numbers. Order Ref: 28.
2 x 6V Operated Reed Relays. One normally on, other normally closed. Order Ref: 48.
1 x Cabinet Lock. With two keys. Order Ref: 55.
6½" 8 ohm 5 watt Speaker. Order Ref: 824.
1 x Shaded Pole Mains Motor. ¾" stack, so quite powerful. Order Ref: 85.
2 x 5 Aluminium Fan Blades. Could be fitted to the above motor. Order Ref: 86.
1 x Case. 3½x2¼x1¾ with 13A socket pins. Order Ref: 845.
2 x Cases. 2½x2¼x1¾ with 13A pins. Order Ref: 565.
4 x Luminous Rocker Switches. 10A mains. Order Ref: 793.
4 x Different Standard V3 Micro Switches. Order Ref: 340.
4 x Different Sub Min Micro Switches. Order Ref: 313.

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Infra Red Controller. Made from Thorn TV sets but suitable for other control purposes. Fully built and ready to operate, real bargain, £2. Order Ref 2P304.
Hall Effect. Give positive or negative pulses when magnet passes over. Mounted on small PCB, 2 for £1. Order Ref 1032.
Digital Multi Tester. 30 range, model no 3800, normal price £40, our price £25. Order Ref 25P14. Brand new and guaranteed.
Water Pump with spindle for operation by portable drill, £5. Order Ref 5P240.
Three More Transformers. Order Ref 4P81 is a 12V-0-12V 40W, clamp mounted, price £4 each less 10% for 10 or more. Order Ref 5P236 is a 43V at 2.4A, frame mounted, heavy construction, will withstand considerable overloads, price £5. Order Ref 3P181 is a 12V 3A frame mounting type but without the frames, price £3. We have tested this and find it quite suitable for 50W lamps.
Multi Voltage Auto-Transformer. Could be used to give 350W at 115V for operating regular 115V equipment or it could give this some current at 85V, 120V or 130V. Another use for it is to boost the output from a long line. Could give a 30V or 50V boost up to 300W. Probably has many other uses for its outputs are 85V, 115V, 120V, 130V, 200V, 220V and 240V. A big transformer, price £4. Order Ref 4P79.
If You Use An Invertor to operate radios or TV and similar frequency controlled equipment, then it is advisable to know the frequency of the invertor, otherwise this and/or the equipment it operates can be damaged. We have 100mm square faced panel meters which electronically display the frequency of the supply, providing it is between 45 and 55HZ. Really top class instrument, price £15. Order Ref 15P19.
Mains Klaxon Type Alarm. Very loud output but adjustable. Completely encased, shelf or wall mounting, £5. Order Ref 5P226.
12V 10A Switch Mode Power Supply for only £9.50 and a little bit of work because you have to convert our 135W PSU. Modifications are relatively simple – we supply Instructions. Simply order PSU Ref 9.5P2 and request modification details, price still £9.50.
Speed Controller for 12v DC Motors. Suitable for motors with horse powers up to one third and drawing currents up to 30A. Gives very good control and speed. Uses mosfets and is based on a well tried circuit which appeared in the *Model Engineer* some time ago. The complete kit with case and on/off switch is available, price £18. Order Ref: 18P8.
Fig 8 Flex. Fig. 8 flat white pvc, flexible with .4 sq. mm cores. Ideal for speaker extensions and bell circuits. Also adequately insulated for mains lighting. 50m coil £2. Order Ref: 2P345. 12m coil £1. Order Ref: 1014.
Friedland Underdome Bell. Their ref: 792. A loud ringer but very neat, 3" diameter, complete with wall fixing screws, £5. Order Ref: 5P232.
12v 10amp Switch Mode Power Supply. For only £9.50 and a little bit of work because you have to convert our 135W PSU. Modifications are relatively simple - we supply Instructions. Simply order PSU Ref: 9.5P2 and request modification details. Price still £9.50.
Medicine Cupboard Alarm. Or it could be used to warn when any cupboard door is opened. The light shining on the unit makes the bell ring. Completely built and neatly cased, requires only a battery. £3. Order Ref: 3P155.
Don't Let It Overflow! Be it bath, sink, cellar, sump or any other thing that could flood. This device will tell you when the water has risen to the pre-set level. Adjustable over quite a useful range. Neatly cased for wall mounting, ready to work when battery fitted. £3. Order Ref: 3P156.
Very Powerful Mains Motor. With extra long (2½") shafts extending out each side. Makes it ideal for a reversing arrangement for, as you know, shaded pole motors are not reversible. £3. Order Ref: 3P157.
Solar Panel Bargain. Gives 3v at 200mA. Order Ref: 2P324.

£1 Super Bargain
12V axial fan for only £1, ideal for equipment cooling, brand new, made by West German company. Brushless so virtually everlasting. Needs simple transistor drive circuit, we include diagram. Only £1, Order Ref: 919. When we supply this we will include a list of approximately 800 of our other £1 bargains.

40W-250W Light Dimmers On standard plate to put directly in place of flush switch. Available in colours, green, red, blue and yellow. £2.50, Order Ref. 2.5P9. Or on standard 3x3 cream metal switch plate, £3, Order Ref. 3P174.
45A Double Pole Mains Switch. Mounted on a 6x3½ aluminium plate, beautifully finished in gold, with pilot light. Top quality, made by MEM, £2, Order Ref: 2P316.
Touch Dimmers 40W-250W, no knob to turn, just finger on front plate, will give more, or less light, or off. Silver plate on white background, right size to replace normal switch £5, Order Ref. 5P230.

Motorise that Trolley!
You could with Sinclair C5 ⅓rd hp
12v battery motor
Still available, price £21. Order Ref: 21P1

12/24 DC Solenoid. The construction of this is such that it will push or pull. With 24V this is terrifically powerful but is still quite good at 12V. £1, Order Ref: 877.
Don't Stand Out In The Cold Our 12m telephone extension lead has a flat BT socket one end and flat BT plug other end, £2, Order Ref: 2P338.
20W 5" 4 Ohm Speaker mounted on baffle with front grille, £3, Order Ref: 3P145. Matching 4 ohm 20W tweeter on separate baffle, £1.50, Order Ref: 1.5P9.

LCD 3½ Digit Panel Meter
This is a multi range voltmeter/ammeter using the A-D converter chip 7106 to provide 5 ranges each of volts and amps. Supplied with full data sheet. Special snip price of £12, Order Ref: 12P19.

Telephone Extension Wire 4 core correctly colour coded, intended for permanent extensions, 25m coil, £2, Order Ref. 2P339.
High Power Switch Mode PSU. Normal mains input, 3 outputs: +12V at 4A, +5V at 16A and –12V at ½A. Completely enclosed in plated steel case. Brand new. Our special offer price of £9.50, Order Ref: 9.5P1.
Philips 9" High Resolution Monitor. Black and white in metal frame for easy mounting. Brand new, still in maker's packing, offered at less than price of tube alone, only £15, Order Ref: 15P1.
High Current AC Mains Relay This has a 230v coil and changeover switch rated at 15A with PCB mounting with clear plastic cover. £1, Order Ref. 965.

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Ultra Thin Drills, actually 0.3mm. To buy these regular costs a fortune. However, these are packed in half dozens and the price to you is £1 per pack, Order Ref: 797B.
You Can Stand On It! Made to house GPO telephone equipment, this box is extremely tough and would be ideal for keeping your small tools in, internal size approx. 10½"x4½"x6" high. Complete with carrying strap, price £2, Order Ref: 2P283B.
Ultra Sonic Transducers. Two metal cased units, one transmits, one receives. Built to operate around 40kHz. Price £1.50 the pair, Order Ref: 1.5P/4.
Power Supply with Extras. Mains input is fused and filtered and the 12V DC output is voltage regulated. Intended for high class equipment, this is mounted on a PCB and, also mounted on the board but easily removed, are two 12V relays and Piezo sounder,£3, Order Ref: 3P80B.
Insulation Tester with Multimeter. Internally generates voltages which enable you to read insulation directly in megohms. The multimeter has four ranges, AC/DC volts, 3 ranges DC milliamps, 3 ranges resistance and 5 amp range. These instruments are ex-British Telecom but in very good condition, tested and guaranteed OK, probably cost at least £50, yours for only £7.50 with leads, carrying case £2 extra, Order Ref: 7.5P/4.
This instrument slightly faulty but movement and casing guaranteed okay. Supplied complete with circuit diagram, £3. Order Ref 3P176.
Mains Isolation Transformer. Stops you getting "to earth" shocks. 230V in and 230V out. 150 watt, £7.50, Order Ref: 7.5P/5 and a 250W version is £10, Order Ref: 10P97.
Mains 230V Fan. Best make "PAPST", 4½" square, metal blades, £8, Order Ref: 8P8.
2MW Laser. Helium neon by Philips, full spec. £30, Order Ref: 30P1. Power supply for this in kit form with case is £15, Order Ref: 15P16, or in larger case to house tube as well £18, Order Ref: 18P2. The larger unit, made up, tested and ready to use, complete with laser tube £69, Order Ref: 69P1.
12v 8ohm speaker, only £1.50 and waterproof.
Solar Charger. Holds 4AA nicads and recharges these in 8 hours, in very neat plastic case £6, Order Ref: 6P3.
Ferrite Aerial Rod. 8" long x ⅜" diameter, made by Mullard. Complete with two coils, 2 for £1, Order Ref: 832P.
Air Spaced Trimmer Caps. 2-20pf, ideal for precision tuning UHF circuits, 4 for £1, Order Ref: 818B.
Modem Amstrad FM240 As new condition but customer return, so you may need to fault find, £6, Order Ref. 6P34.
Amstrad Power Unit. 13.5V at 1.9A or 12V at 2A encased and with leads and output plug, normal mains input £6, Order Ref: 6P23.
80W Mains Transformer. Two available, good quality, both with normal primaries and upright mounting, one is 20V 4A, Order Ref: 3P106, the other 40V 2A, Order Ref: 3P107, only £3 each.
Project Box. Size approx. 8"x4"x4½" metal, sprayed grey, louvred ends for ventilation otherwise undrilled. Made for GPO so best quality, only £3 each, Order Ref: 3P74.
Sentinel Component Board Amongst hundred of other parts, this has 15 ICs, all plug in so do not need soldering. Cost well over £100, yours for £4, Order Ref. 4P67.
Sinclair 9V 2.1A Power Supply Made to operate the 138K Spectrum Plus 2, cased with input and output leads. Originally listed at around £15, are brand new, our price is only £3, Order Ref. 3P151.
Experimenting with Valves. Don't spend a fortune on a mains transformer, we can supply one with standard mains input and secs. of 250-0-250V at 75mA and 6.3V at 3A, £5, Order Ref: 5P167.
15W 8 Ohm 8" Speaker & 3" Tweeter. Made for a discontinued high quality music centre, gives real hi-fi and only £4 per pair, Order Ref: 4P57.
Water Pump. Very powerful, mains operated, £10, Order Ref: 10P74.
0-1mA Full Vision Panel Meter. 2¾" square, scaled 0-100 but scale easily removed for re-writing, £1 each, Order Ref: 756.
VU Meter. Illuminate this from behind becomes on/off indicator as well, 1½" square, 75 each, Order Ref: 366.
Amstrad Keyboard Model KB5 This is a most comprehensive keyboard, having over 100 keys including, of course, full numerical and qwerty. Brand new, still in maker's packing, £5, Order Ref. 5P202.
1 RPM Motor. This is only 2W so will not cost much to run. Speed is ideal for revolving mirrors or lights. £2, Order Ref. 2P328.
Unusual Solenoid. Solenoids normally have to be energised to pull in and hold the core, this is a disadvantage where the appliance is left on for most of the time. We now have magnetic solenoids which hold the core until a voltage is applied to release it. £2, Order Ref. 2P327.
Mains Filter. Resin impregnated, nicely cased, pcb mounting. £2, Order Ref. 2P315.
200VA Mains Transformer. Secondary voltages 8v-0-8v. So you could have 16v at 12A or 8v at 25A. Could be ideal for car starter charger, soil heating, spot welding, carbon rod welding or driving high powered amplifiers etc. £15, Order Ref. 15P51.
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CIRCLE NO. 139 ON REPLY CARD

which severely limit any improvement their superbly-constructed products may give?; why are amplifier designers, offering 'damping factors' of thousands measured at the output terminals, not concerned that you are lucky to get a damping factor of five measured at the drive unit?, and why do loudspeaker manufacturers care so little about the transient response of their products that they sacrifice damping for the sake of convention in the placement of filters?

Current technology allows systems to be constructed with one power amplifier for each driver, tailored to the needs of bass, middle, treble units. All crossover networks can be organised at the amplifier inputs, with each drive unit connected direct to an amplifier output. Such systems achieve maximum frequency-independent damping, and incidentally allow complete overload protection for the drivers. So if you really can tell the difference between bell-wire and gold-plated super conductors, then the improvement with this arrangement will be absolutely dazzling.

Wal Hensby
Essex

Historical insight

RL Tufft's reference to speakers driven by moving-iron and balanced-armature (iron) movements (*Letters*) made me wonder if he also remembers the inductor-dynamic-movements, two-iron armatures, producing somewhat better low-frequency response than the other two types.

Back in the late 1920s and early thirties, many of us used to make our own cones, or other diaphragms with suitable frames or mountings for the speaker assembly. We described our interest as high quality sound reproduction, as we did when we progressed to moving coil drives, initially with dc-energised magnets, often at 6V from accumulators or from dc mains supply at 200-230V.

We aspired to owning a PG Voight MC speaker unit – energised in those days by dc – though a very fine example with a permanent magnet came from Ferranti, the Ferranti *M1*. It received a good review in *Wireless World* at that time, with pretty even response up to at least 2000cps.

Ex BBC engineer PR Turner, with a Mr Hartley, produced an interesting unit, a permanent magnet with plastic, brown Bakelite sheet, hand-made into a cone. This worked well, with good low frequency output, in a suitable enclosure, and plenty of hf.

The Hartley Turner MC improved units allowed quite high quality sound to be reproduced from the main medium-wave BBC transmitters. Unfortunately, after dark, a filter was needed to reduce the effect of the accompanying 10kHz whistle caused by the carrier beating with an adjacent transmitter – unless located fairly close to the desired transmitter.

From late 1937 a somewhat better source became available, sound from Alexander Palace and its television signals. The result was good and I must thank the 'magic' of Alan Blumlein at EMI labs, Hayes, for that pleasure.

I am still an enthusiast for high-quality sound reproduction and have my custom-made speakers, making changes from time to time. I keep a pair of Peter Walkes *ELS63*s just for reference.

Finally, I must mention my interest in Douglas Self's work. I often wonder what is his opinion of the Quad 405-2 circuitry. I have modified slightly some that I have and, frankly, find these amplifiers difficult to fault. Again I have another high grade mosfet unit as a comparative assessment.

I moved to Scarborough in 1989 from the south but have not yet encountered anyone who appears to be a contemporary of mine. I look forward to making more contacts.

Harry Dix
Scarborough

Busman's I^{2}C kits

Following publication of 'Busman's guide to I^{2}C' (*EW+WW*, June, pp.479-485), we are offering *EW+WW* readers the *Cameo* development board at the reduced cost of £99 plus vat and delivery (total cash-with-order price including £5 delivery and £18.20 vat is £122.20). The board normally costs £187.41 (inc vat). An information pack giving full details of the board's functions can also be obtained by sending a C4 sae with a £0.57 stamp.

As readers will appreciate, the *Cameo* board allows design and development of 8051 programs, and contains the powerful Philips *80C552* plus monitor prom and up to 32K of user ram. Our offer, which lasts until the end of December, includes a user manual, circuit diagram, and disc with *Cameo WorkBench* comms and example programs. Now for the mistakes. Readers should also note that:

Page 481, Fig. 1, the test for the last data byte sets SDA high for both true and false. There should be an ACK for every byte except the last data byte (Set SDA low ACK if yes).

Page 482, for communications with the Cameo board, the comms program need not be *procomm*, any *RS232* comms package will do.

Page 482, the internal registers on frequency synthesiser *TSA6057* are shown in hex, but the base 16 figures after each data byte should be in subscript to avoid confusion: eg 40_{16} instead of 4016.

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Technical Director TDR Ltd
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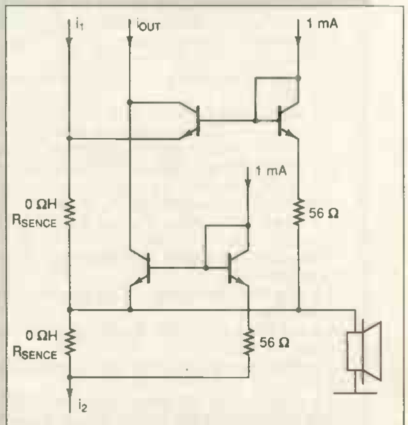
Tel 0666-577464

Discrete behaviour

Douglas Self is correct (*Letters*, August 1994). The circuits described in the references of my previous letter (June) are unsuitable for a discrete amplifier. But I would like to point out that in my original letter I described an output stage of an amplifier which consisted of discrete components and a *CA3046* transistor array. This output stage uses the non-linear common-mode loop technique.

Unfortunately, my letter was too long and this part was not published.

As an alternative, instead of the harmonic mean I have used a different non-linear function which gives similar results but is easier to implement. In the non-linear network (see figure) the output current is approximately proportional to $\exp(-c.i_1)+\exp(-c.i_2)$, where $c = qR_{sense}/kT$ and i_1 and i_2 are the currents through the output transistors. The common-mode loop forces the output current of the network to remain constant.



Non-linear network. The transistors are all part of the CA3046 transistor array.

I would be happy to send Douglas Self the complete amplifier schematic and a short description.

Marcel van de Gevel
Haarlem
The Netherlands

Big science squashes little projects

I agree with R Burfoot (*Letters*, July) that during the past few years, the electronics industry has declined. But it would be grossly unfair to blame the Ministry of Defence. If it had not been for government contracts, inertial navigation, radar and many other major developments might not have taken place.

The tax-payer's money was spent wisely and the national investment has already been returned a thousand-fold.

So what can we do now to halt the decline in electronics?

One answer might be for Government to award contracts for small-but-promising civil projects – studies in robotics, laser technology, artificial intelligence, navigation etc.

But are today's government research funds being wisely spent? The UK, like the USA, has embarked on prestigious but costly research programmes in big science. As a consequence, small industries have been starved of funds.

To take an example from the US, where the government spent $2-billion on an 86km tunnel for the super-conducting super collider only to have the project eventually cancelled by Congress. Surely, there are better ways of spending tax-payers money than seeking a Grand Unified theory or looking for new particles, gravity waves, black holes and dark matter.

The hard fact is that in the short term, big science projects – worthy as they are – are most unlikely to create work or generate wealth. While we wait, our industry is dying.

John Ferguson
Camberley

Fourier dice

Of course R H Pearson (*Letters*, August) is quite right – formally, but in my letter (June) 'Fourier's theory' was loosely used as a collective noun for the many design theories – with and without computer aid – that revolve around simple harmonic vibrations and circular functions.

Although mathematically versatile, these concepts fail to do full justice to the wealth of nuances shaping the signals produced by everyday reality, sound and vision alike.

Hence most reproduction sounds/looks a pronounced artificiality. ■

H G Groenevelt
Rotterdam
Netherlands

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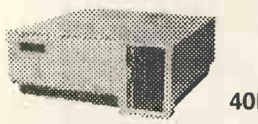

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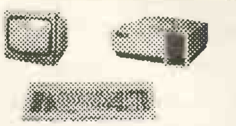

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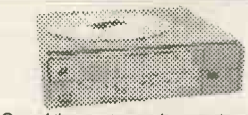

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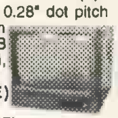

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ACTIVE

Asics

0.6μm PLDs. Altera's new *FLEX 8000A* family of programmable logic devices is a redesign of the earlier *Flex 8000* family, in a new 0.6μm triple-layer metal cmos sram process. There are seven family members, from 2500 to 16000 usable gates, all of them drop-in compatible with the earlier types. Performance improvement over the 8000 family is 75%, from 43MHz to 70MHz. Altera UK Ltd. Tel., 0628 488811; fax, 0628 890078.

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

500ksample/s a-to-d. Sampling at 500k samples per second and using only 75mW from a single 5V supply, Linear's *LTC1278* 12-bit analogue-to-digital converter offers a sinad ratio of 70dB and thd of 74dB at the Nyquist frequency. Integral and differential non-linearity errors are ±1lsb, there are no missing codes over the whole temperature range and drift is 45ppm/°C. With ±5V supplies, the device provides ±2.5V output. A 5mW shutdown feature is included, with rapid wake-up. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

40Msample/s A-to-D. Harris claims the first 10-bit, 40Msample/s analogue-to-digital converter with a simple and reliable pipeline architecture. The *HI5702* uses the company's HBC-10 BiCMOS process to overcome the power v. accuracy compromises of cmos and bipolar solutions and exceeds the performance of any previous device while using about 0.5W less power than its bipolar competitor. The device operates from one 5V supply and offers a maximum integral non-linearity error of 2LSB, with a differential error of 1LSB, digitising a 10MHz 2.5Vpk-pk differential or single-ended input to 10-bit linearity at a minimum 53dB s:n ratio (51dB sinad at 10MHz). An evaluation kit, the *HI5702-EV*, includes clock driver circuitry, a reference voltage generator and a choice of input drive circuitry, together with demo board, sample, data sheet and user's guide. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

Low-power a-to-d converters. Harris's 5V *HI5810* 12-bit sampling a-to-d converter has a 10μs conversion time, sampling at 100,000 samples per second. Analogue input bandwidth is 1MHz and integral linearity is 2lsb over the industrial temperature range. Parallel data outputs are of the three-state bus driver type and there is a selectable choice of resolution. The *HI5813* is a 3-6V type with a track/hold amplifier, 25μs conversion time and 40,000 samples/s. Thame Components Ltd. tel., 0844 261188; fax, 0844 261681.

Discrete active devices

Power transistors. Motorola's *MJ3281A* and *MJ1302A* are PowerBase complementary silicon power transistors for audio, disk head positioning and other high-power linear uses. They are rated at 200V/15A/250W and their f_T is typically 30MHz. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

SM Igbt/Hexfred package. *SMD-CoPacks*, introduced by IR, combine an insulated-gate bipolar transistor and a *Hexfred* fast-recovery epitaxial diode in one surface-mounting package, thereby saving about 40% of the cost of separate devices and up to 70% of the size. First available are *IRGBC20KD2-S/MD2-S*, rated at 10A and 13A respectively. International Rectifier. Tel., 0883 713215; fax, 0883 714234.

Diode arrays. *Rohm* surface-mounted diode arrays contain up to four devices in one package, the range including common-cathode, common-anode and isolated devices. Diode types offered are 0.1-4A Schottky barrier devices, small-signal Schottky types with 0.37V forward drops, 4ns switching arrays and band switching arrays with 1.2pF capacitance at 1MHz and resistance of 0.9Ω at 100MHz. Pin diodes are available in packs of two devices. Flint Distribution. Tel., 0530 510333; fax, 0530 510275.

Linear integrated circuits

Digitally controlled pot. The Xicor *Audio E²POT X9314*, in an 8-pin dip, is a digitally controlled potentiometer with a logarithmic taper to replace the mechanical type in audio circuitry. 'Wiper' position is controlled by asserting chip select, choosing

direction and pulsing the device until the position is reached at one of 32 steps per pulse. Position is then stored in internal memory. Resistance of the X9314 is 10kΩ. Micro Call Ltd. Tel., 0844 261939; fax, 0844 261678.

20MHz function generator. Producing accurate, high-frequency sinusoidal, square, triangular and pulse waveforms with few external components and in response to a 2-digit code, the Maxim's *MAX038* also produces a sync. output. Frequency is controlled by a 2-700μA current and an external capacitor. An external modulating voltage provides pwm or sawtooth waveforms. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

Jfet op-amp for capacitive loads. Linear Technology says its *LT1457* is the first jfet-input op-amp to be optimised for driving large capacitive loads, the dual *C-Load* device being able to handle at least 10nF loads without oscillation. Input offset is 450μV in a plastic dil and 1200μV in SO-8, drifting at 4μV/°C; input bias current is 50pA; voltage noise 13nV/√Hz; and slew rate 4V/μs. Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

900MHz mixer/exciter. Designed as a linear up-converter for American and Japanese digital cellular radio, Motorola's *MRFIC2101* 900MHz transmit mixer and exciter is suitable for analogue cellular and other 900MHz systems such as GSM and ISM. There is a double-balanced mixer and a local-oscillator buffer to reduce LO power and eliminate the need for an external LO balun. The device has a fast power-down control. Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

Laser diodes. Two laser diodes from MPS have power stabilisation, slow start and a heat sink. *CJ51F* (1mW) and *CJ52G* (5mW) continuous diodes are complete with optics and electronics, but are only 22mm long and 12mm in diameter. Power supply needed is 3V. MPS Electronics. Tel., 0702 554171; fax, 0702 553935.

Logic building blocks

Low-voltage logic. Designed expressly for relatively low-performance applications such as palm-tops and point-of-sale equipment, two families of 3.3V low-voltage cmos by TI, *LV-HCMOS* and *LVC*, are said to be equivalent to the 74F series in 5V ttl. The 0.8μm *LVC* family consists of gates and MSI and 8-bit Widebus devices with a standby consumption of 20μA and 7ns propagation delay. Texas Instruments. Tel., 0234 270111; fax, 0234 223459.

Memory chips

Configurable flash memory. AMD's *Am29F400* 4Mbit 5V-only flash memory is user-configurable in 512 by 8 or 256 by 16 form. Eleven sectors of unequal size can be erased

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RF and microwave VCOs. *Vari-L*'s range of voltage-controlled oscillators are meant for use in battery-powered equipment, accepting supplies of 3V, 5V, 12V or 15V. Frequencies covered are 25-50MHz to 3-4.8GHz, phase noise varying between 82dBc/√Hz to 118dBc/√Hz. Packaging includes surface-mount, flatpack, SMA and TO-8. Acal Electronics Ltd. Tel., 0344 727272; fax, 0344 424262.

individually, in multiples or all together. Boot sectors at top or bottom of the address map cope with different microprocessors, the devices having T or B as a suffix. Selected sectors can be protected and embedded algorithms detect and correct erase errors. Advanced Micro Devices (UK) Ltd. Tel., 0483 740440; fax, 0483 756196.

16Mbit dram. Toshiba has a 50ns, 16Mbit dram in a 300mil SOJ package. The *TCS5116400BSJ-50* is based on a 0.5µm process and features hyper-page mode operation. Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

1Mbit srams. One megabit srams by IBM in the *IBM 04XXX* family operate at up to 167MHz in second-level cache applications supporting high-performance microprocessors, with a pipeline access of 4ns or flowthrough of 8ns. Versions with burst mode are available for use with *PowerPC* and Pentium processors. The srams are in 64Kword by 18bit or 32Kword by 36bit form and features include single-clock read/write, self-timed write and low-voltage ttl i/o interfaces. Blue Micro Electronics. Tel., 0604 603310; fax, 0604 603320.

16Mb, 500Mbyte/s drams. NEC's 16Mb and 18Mb *Rambus* dynamic random access memories offer a 2ns access time and a peak data transfer rate of 500Mbyte/s. The *µPD488130* and *µPD488170* 16Mb and 18Mb capacity Rdrams incorporate a Rambus Interface communicating over a byte-wide channel, called the Rambus Channel, to give a transfer rate of one byte in 2ns. If four such channels are used in a system, bandwidth is 2Gbyte/s. Each Rdram has two independent sense amplifier caches to increase data transfer to the arrays and reduce latency. Packaging is a 32-pin vertical SM type. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

Microprocessors and controllers

8-bit risc microcontroller. With 2048 12-bit words of one-time programmable program memory, 73 8-bit bytes of static ram for data and up to 2:1 code compaction referred to non-risc types, Microchip's *PIC16C58A* runs at 20MHz with a 200ns instruction execution time and is claimed to be the fastest available in its class. On-chip peripherals include an 8-bit clock/counter with a programmable prescaler, start-up timer, watchdog timer with *RC* oscillator and 12 i/o lines. Arizona Microchip Technology Ltd. Tel., 0628 850303; fax, 0628 850178.

Fast 8051 controller. While drop-in compatible with the 8051, Dallas's *DS87C520* runs over eight times as fast. It also has 16Kbyte of eprom and 1.2Kbyte of sram. The 8051 core has been redesigned to use only four clocks per cycle instead of twelve, running at 33MHz to give a peak execution cycle of over 8Mips, no change in software or development tools being needed. Power management allows the user to reduce power by 80% by slowing the clock. Dallas Semiconductor Corporation. Tel., 021 782 2959; fax, 021 782 2156.

Low-voltage, 4-bit controller. The *µPD753108* microcontroller, an addition to NEC's 1.8-6V 4-bit range, is provided with a 24 by 4 bit lcd driver. All *75XL* devices have an instruction cycle time of 0.95µs at 1.8V and 0.67µs at 6V, although NEC point out that, since the instruction set is more powerful than that of the earlier *75X* series, fewer instructions are needed for the same tasks. NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

8-bit cmos microcontroller. Zilog's *Z86C04* cmos device is one of the Z8 microcontroller family, with 1Kbyte of rom and 124byte of general-purpose ram and packaged in either an 18-pin dip or 18-pin SOIC. Temperature range is –40°C to 105°C. Power consumption is 50mW and the unit is provided with brown-out protection, fast instruction points of 1.25µs, and stop and halt modes. Fourteen i/o lines are at cmos levels, eleven of them being digital, Schmitt-triggered inputs. Clock speed is 8MHz. Gothic Crellon Ltd. Tel., 0734 788878; fax, 0734 776095.

150MHz processor. As well as reducing the prices of its 100MHz and 133MHz *Orion R4600* processors, IDT has released the 150MHz version, which is claimed to outperform the Pentium at the price of a 486DX. The device has the *Flexbus*, which is a software initialisation mechanism allowing bus interface frequency to be tailored to suit system requirements. Five-volt version are now available, with 3.3V models arriving later in the year. Integrated Device Technology. Tel., 0372 363734; fax, 0372 378851.

Mixed-signal Ics

Engine-knock detection. Harris's *HIP9010* is a mixed-signal device known as an engine knock signal processor, to be used in knock detection subsystems in vehicle ignition control systems. It amplifies and filters the output of a piezoelectric transducer during a short interval about top-dead-centre, so that the signal can be separated from engine noise. Analogue gain and filter characteristics are changed digitally by the system microcontroller to accommodate varying engine conditions. Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

Optical devices

Miniature camera. Henderson has announced a new range of pcb mounting cameras of both the pinhole type and those using the range of interchangeable lenses from 3.6mm to 16mm. Camera units are on a single board measuring 42mm square and are sensitive down to 0.5lux. A range of housings is available, and a remote 12V supply that feeds the unit through a multicore cable. Henderson Security Electronics Ltd. Tel., 9684 274874; fax, 0684 294845.

Laser sensor. A laser photoelectric sensor from Keyence, the *LZ-155* series produces a spot 0.05mm in diameter with a positioning accuracy of 0.005mm horizontally. Detection distance is up to 60mm, at which distance the spot is visible even on a black surface. The device is intended for positioning and counting very small objects, for which a multi-turn potentiometer adjusts sensitivity. Keyence UK Ltd. Tel., 0908 696900; fax, 0908 696777.

Oscillators

Custom crystal oscillators. ACT announces a facility for manufacture of temperature-compensated crystal oscillators with many choices of operating temperature and stability within the limits of –40°C to 85°C and ±5% to ±0.5%. There are six package styles, including hermetically sealed metal. Lead times are down to 20 days. Advanced Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

Power semiconductors

Lamp driver. Microlinear's *ML4874* drives small cold-cathode fluorescent tubes used as backlighting for liquid-crystal displays. The device drives the tubes differentially, taking less power than is the case with single-ended drives and expending less power on stray capacitance in the lcd housing. Efficiency is 95%, obtained by the use of a resonant threshold detection arrangement. Ambar Components Ltd. Tel., 0844 261144; fax, 0844 261789.

TSSOP power mosfets. The new *Litefoot* power mosfets from Siliconix come in n-channel and p-channel form and are small enough to fit on any standard PCMCIA card, being only half the size of others on the market. Power dissipation is 1.5W with no heat sink besides the PCB and breakdown lies between 12V and 30V. On resistance for a single p-channel device is 75mΩ and that for a single n-channel type 50mΩ. Single, dual and complementary devices are available. Siliconix/Temic Marketing. Tel., 0344 485757; fax, 0344 427371.

PASSIVE

Passive components

High power factor capacitors. *Type 6124* and *Type 7124* from Tecate are metallised polyester and metallised polypropylene capacitors intended for use in equipment such as lighting, snubbers and small motors where power factor correction is needed. Standard tolerances are ±5%, ±10% and –5 +100%, dissipation factor being 1% maximum. Both types have an optional thermal cut-off and a bleeder resistor to discharge the capacitor to lower than 50V in a minute. Voltage ratings are 160-250V ac in values of 1-3µF (6124) and 160-500V/1-25µF (7124). Tecate Industries Inc. Tel., 0101 619 448-4811; fax, 0101 619 448-0912.

Transient suppressors. Semtech has transient voltage suppressors in the *SL series* which exhibit only 5pF capacitance. They are designed for use on data lines and handle 300W peak pulse power with a response time of 1ps. Reverse standoff voltages are 5V, 12V, 15V and 24V, breakdown 6-26.7V and maximum clamping voltage at 1A, 9.8-43V. Semtech Ltd. Tel., 0592 773520; fax, 0592 774781.

Tantalum capacitors. Components in AVX's *TAZ* range of high-reliability tantalum capacitors intended for use in medical implantable devices are now reduced in size. The capacitors are qualified to Weibull C failure rates and are available with low leakage current and nine configurations of termination and finishes. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Trimmer capacitors. For those applications not involving high rf power, Jackson's *C824* series of air-spaced trimmers use a low-loss composition front panel and aluminium rotors and stators. Maximum capacitances are 10pF to 100pF and minimum for all types is 5pF, with a linear capacitance/angle relation. Working voltage is 350V. Jackson Brothers Ltd. Tel., 081-681 2754; fax, 081-681 3728.

Dielectric filter. AVX announces the *PDFC* series of dielectric filters meant for use in telecomms, particularly in the DECT sector. Frequency range is 1.8-2GHz, insertion loss 3dB and, for compatibility with the newest equipment, size is 6.5 by 5.5 by 3mm. Filters to provide lower insertion loss and improved stop-band attenuation are available to order. AVX Ltd. Tel., 0252 336868; fax, 0252 346643.

Connectors and cabling

Protected jack connectors. *Murata* announces a series of modular jack connectors with built-in varistors for surge and noise suppression in ISDN terminal equipment. Built-in inductors ensure noise reduction over a wide frequency range. They are rated at 50V dc at 200mA, have a typical impedance 0f 600Ω at 100MHz, a varistor voltage of 250V between line and earth and a –25°C to 60°C operating range. Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

Displays

DVM module. *DMS-40LCD* by Datel Is a series of 4.5-digit LCD meters contained in plastic dips measuring 0.9in by 2.1in by 0.43in, with 0.4in characters. The devices have dual inputs to allow signal input of ±200mV/±2V, ±2V/±20V or ±20V/±200V. Power needed is one 5V line at 2.5mA or 9V at 1.5mA. Backlighting is available. Datel (UK) Ltd. Tel., 0256 880444; fax, 0256 880706.

TV test patterns
Contained In a hand-held case, the *OZAN* television test pattern generator is powered by a 9V battery, although a mains unit is supplied for continuous bench use. It is connected to the rf and line sockets of television receivers and video recorders. Four PAL test patterns – colour bars to the BBC 95% or EBU 100% luminance, grey scale with eight 14.3% steps, cross hatch and red purity – and a 1kHz audio tone are generated, the RF video and audio coming from a 75Ω coaxial socket on channel 36 with audio set to the 6MHz sub-carrier (5.5MHz for other standards). Two 75Ω phono sockets provide composite video at 74Ω and the audio line out at 1kΩ.

Colour lcd. Sharp's *LQ6RA54* is a 5.5in thin-film transistor lcd module in which the pixels are arranged in a stripe rather than in the delta formation of RGB elements, easing the problems of writing graphical information. A black mask reduces internally reflected light by over 80% over previous types and surface reflections are avoided by means of a polariser. Vertical viewing angle Is switchable to either above or below the display centre line. Hero Electronics Ltd. Tel., 0525 4055015; fax, 0525 402383.

Bargraph DMM. Lascar has a large-digit bargraph multimeter that is provided with a 3.5-digit display visible from 10m, even in low light, by virtue of its led backlighting. Connection of the *DMM 977* to the display is by IDC connector. Lascar Electronics Ltd. Tel., 0794 884567; fax, 0794 884616.

Filters

Two-port saw resonators. Saw resonators in *RF Monolithics's RP* and *RS* series, which have nominal phase angles of 180° and 0° respectively, now exhibit a frequency tolerance of ±75kHz. Five of the devices are available for low-power UHF transmitters in applications including the DTI MPT1340 at 418MHz and the pan-European ETS-300-220 at 433.92MHz. Insertion losses are typically 5.7dB and 6.3dB at the two frequencies. Quantelec Ltd. Tel., 0993 776488; fax, 0993 705415.

Instrumentation

EMC probe. An *active near-field probe* by Seaward locates the source of radiated emissions, completing the company's emc test package to the requirements of the EC Directive on Electromagnetic Compatibility. The probe will localise emissions from pcb-mounted components, cables and case apertures and joints in the frequency range 1MHz-1GHz. Although the probe was designed for use with Seaward's spectrum analyser, it is now suitable for most analysers with minor mods. Seaward Electronic Ltd. Tel., 091-586 3511; fax, 091-586 0227.

Simm tester. ITM has introduced the *TA1011 Test Head* for the *Excel 1000* bench-top production memory tester from TMI Inc., which carries out 100% testing of virtually all memory modules, including 30-pin and 72-pin simms with page-mode up to 40 bits wide and organised in configurations from 64K by 9 to 16M by 9; It also handles 2K and 4K refresh-type drams. The system is controlled by a PC AT and a bus interface for the PC is supplied. Instrumentation Test & Measurement Ltd. Tel., 0202 872771; fax, 0202 871052.

Microwave

Sweepers/synthesisers. Giga-Tronics offers the *GT 9000* microwave synthesiser and the *GT 9000S* synthesised microwave sweeper. They are improved versions of earlier instruments, offering a 2-20GHz range. Phase noise of the 9000S at 2GHz is –95dBc/√Hz at 10kHz offset. output power +13dBm from 10MHz to 20GHz and harmonics at 6dBm are less than –65dBc. Pulse mod. is standard and AM, FM and scan mod. are options. The 9000S is the same, but with analogue and digital frequency and power sweeping. Sematron UK Ltd. Tel., 0734 819970; fax, 0734 819786.

100MHz logic analyser.

The *TA320S* self-contained logic analyser by TTi has 32 channels and a 100MHz acquisition rate. Display is by supertwist lcd capable of

Eight-channel dso
Two instruments in the *Yokogawa DL5100* series of digital storage oscilloscopes, *DL5180* and *DL5140* offer 1Gsample/s sampling and 500MHz bandwidth on all channels, with a 4Kword/channel memory and 8-bit resolution. The display is a 640 by 480 dot colour tft lcd, an interesting feature being the colour accumulation, in which pixels vary their colour according to the number of times they are written. A history memory acquires and reads out up to 120 sets of waveform data and either measured or saved to disk in the internal 3.5in drive. All 120 scans can be viewed simultaneously. The instruments incorporate an Intel i960 32-bit risc processor for automatic measurements and computation, up to 19 standard parameters being measured automatically. Martron Instruments Ltd. Tel., 0494 459200; fax, 0494 535002.

displaying high-res. graphics or 40-column text and control and data entry is by a combination of soft keys and an alphanumeric keypad. There is also an RS-232 interface to transfer data to and from a computer. The instrument has thirty-two data channels at up to 25MHz and eight at 100MHz for asynchronous working. Events down to 5ns in length can be captured. Optional dissassembler pods, each with its own internal software, support a range of popular microprocessors. State and timing displays are selectable and data can be grouped under user-defined names. Search-and-compare facilities are provided, as is non-volatile storage of acquisitions and set-ups. Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

RF counter. *HP53181A* by H-P is one of the company's lower-cost instruments for frequency measurement to 225MHz or up to 3GHz with optional second channels. Period measurement is performed quickly and there is a limit-testing feature with an analogue display mode for pass/fail testing. An HP-IB port provides SCPI(1) compatible programming and an RS-232 talk-only interface gives printer control or data transfer to a computer at more than 200 formatted measurements per second. Hewlett-Packard Ltd. Tel., 0344 362277; fax, 0344 362269.

1GHz spectrum analysers. Two spectrum analysers by Promax, the *AE-366* and *AE-566*, cover the 1-1000MHz frequency range (1750MHz with an optional converter) and cost £2094 and £2800 respectively, the 566 having its own tracking generator and a normalising function to reduce errors caused by connections. Display dynamic range is 70dB and measuring range 15-130dBμV, the display being log. or linear in the vertical direction. Best resolution bandwidth is selected automatically. Alban Electronics Ltd. Tel., 0727 832266; fax, 0727 810546.

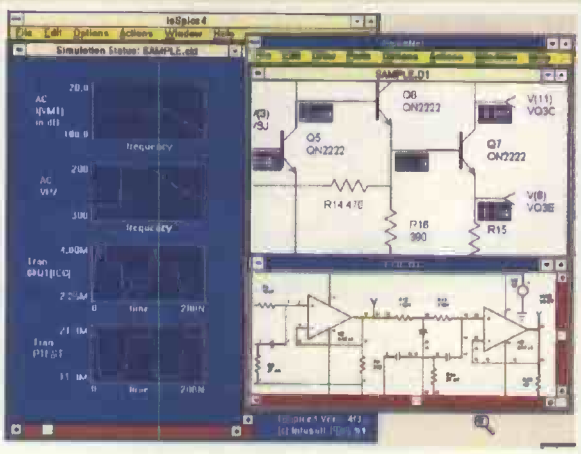
ELF field meter. Over 5-200Hz, Holaday's hand-held *HI-3627* three-axis magnetic field meter measures 0.2mG to 20G, which makes it very suitable for power frequency field measurement. Outputs from three orthogonal sensing coils combine vectorially, the result being indicated by an analogue meter which has a recorder output. Batteries are rechargeable. Holaday Industries. Tel., 0628 478155; fax, 0628 476871.

Literature

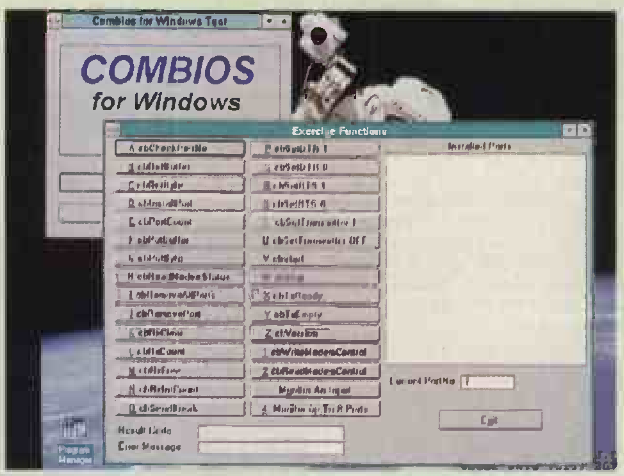
RF/wireless communication. Anglia's *RF/Wireless Communication Components Designer's Data Book* is now available, covering a large range of components. At the end of the book, three articles describe the use of ICs for digital links, attenuator and amplifier Ics for digital systems and lna/mixer ICs. Anglia Microwaves Ltd. Tel., 0277 630000; fax, 0277 631111.

PC instrumentation. Intelligent Instrumentation, a Burr-Brown company, has published the 7th edition of *Handbook of Personal Computer Instrumentation.* Topics covered include signal conditioning, wiring, shielding and data acquisition, together with notes on techniques and applications. The range of II's hardware and software is described. The handbook is free to 'qualified individuals'. Intelligent Instrumentation. Tel., 0923 896989; fax 0923 896671.

Visual C for embedded processors. Since Intel has abandoned embedded C compilers and assemblers, Hitex has produced a guide to the use of the Microsoft C8 Compiler and Visual Workbench for use with embedded processors. It shows how to integrate emulator debuggers into the Workbench to allow rapid swapping between editing and debugging, in addition to many programming tricks needed to address peripherals such as real-time clocks. *Embedding Microsoft C* is available free. Hitex (UK) Ltd. Tel., 0203 692066; fax, 0203 692131.



Low-cost Spice. *ICAP/4Lite* by Intusoft is a low-cost analogue and mixed-signal circuit simulator, based on the company's professional version of Spice. Instead of providing all the traditional Spice facilities and limiting the size of circuit – a common method of producing a cheaper version – Intusoft has allowed unlimited circuit size and reduced the available facilities. The software performs analyses of frequency response, DC conditions, transients and temperature. It also has a schematic entry program to produce a complete Spice netlist, both compatible with the Intusoft professional version. Spice simulation is based on the 32-bit IsSpice4, which gives a real-time waveform display with interactive component changes, printed or displayed inside IsSpice4 or in a reduced version of IntuScope, a graphical analysis program. The ICAP/4Lite package includes schematic entry, IsSpice4 simulator and IntuScope, with a library of over 500 parts. Technology Sources Ltd. Tel., 0638 561460; fax, 0638 561721.



COMBIOS for Windows. *COMBIOS for Windows* provides all the functions found in COMBIOS for DOS, allowing users to develop applications to implement driver buffered serial comms on up to 64 ports. Features include serial ports at any i/o address, all input channels buffered up to 40Kbyte/port, the provision of standard baud rates and data formats and a GUI. It supports any serial RS 232 comms port using the 8250 or 16450 uart and any RS 422/485 port that will enable the transmitter using the uart out 1 line. The facility is independent of language and is not TSR. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 0273 570215.

Strain-gauge selector. In 52 pages, the *HBM* catalogue provides comprehensive data on a range of strain gauges and advice on their use in stress analysis and in various types of transducer. Components and materials for use with the gauges are described in an accessories section. HBM United Kingdom Ltd. Tel., 081-420 7170; fax, 081-420 7336.

Power supplies

Switching controller. The *MAX1771* step-up switching controller from Maxim provides 90% efficiency over a 30mA-2A load, by virtue of its current-limited, pulse-frequency-modulated control, which also takes only 100μA current from the supply. Switching frequency is 300kHz and an n-channel mosfet switch takes loads up to 24W. From inputs of 2-16.5V, output is preset to 12V, adjustable by two resistors. Maxim Integrated Products UK Ltd. Tel., 0734 845255; fax, 0734 843863.

PSU approval. Gardners's *LCS40* switched-mode power supply has been awarded the Industry Approvals EN 60950, UL 1950 and CSA 22.2. It is the first in a new family meant for the medium-volume European market, providing any (reasonable) combination of inputs and outputs at a fair cost. Inputs are 85-265V ac and 120-370V dc and operating frequency is 47-440Hz. Gardners Ltd. Tel., 0202 482284; fax, 0202 470805.

Triple psu. The *Calex Model 3.15.1000* low-noise linear power supply provides ±15V at 100mA to drive amplifier and data conversion circuits and 5V at 1A for logic. Output noise is typically 2mV rms and there is overvoltage and short-circuit protection. Input-to-output isolation is 1500V rms. The supply is for mounting on a pcb, measuring 3.5 by 2.5 by 1.5in. Calex Electronics Ltd. Tel., 0525 373178; fax, 0525 851319.

Dc-to-dc converters. New low-profile versions of the *Cosel* Z series of converters, the *ZU series* are available from XP in 15W and 25W form, total package height being 8.5mm. In common with the standard Z units, the ZU models have input ranges of 9-18V, 18-36V and 37-72V, 500V input-to-output isolation and short-circuit protection, but also have output trim and remote on/off. Stabilisation is 0.1% typical, regulation 1% max. for single output types and noise and ripple 40mV typical. XP plc. Tel., 0734 845515; fax, 0734 843423.

Mobile AC. When plugged into the cigarette lighter socket of, for example, a car, Powerline's *PAC1400* provides up to 140W of continuous ac power or up to 200W for about five minutes; 400W start-up surge can be given for 100ms. This output is sufficient to power a laptop computer or to recharge power packs. The car battery is fully protected. Powerline Electronics Ltd. Tel., 0734 868567; fax, 0734 755172.

Transient protection. A wide range of *Claude Soule* modules to protect power circuits and data lines against transients is now available from Europa. They are DIN rail-mounted, leaded or boxed and protect against rfi, voltage surges or both. Three series, *8748*, *8777* and *8776* protect peripherals such as strain gauges in industrial equipment, coaxial video lines and antenna feeders respectively. Europa Components & Equipment plc. Tel., 081-953 2379; fax, 081-207 6646.

Bendy battery. Ultralife's new *U3VF-X* primary lithium cell is pliable, under 1mm thick and available in virtually any shape. Five versions cover the 70mAh-2500mAh range, weight being 1g-67g. After 10 years, 80% of capacity remains. The cells are suitable for building-in or as stand-alone units when encased in a plastic jacket or hard case. Suvicon Ltd. Tel., 021 643 6888; fax, 021 643 2011.

Radio communications products

Antenna switching relays. Among their other functions, Teledyne's *TO-5* and *Centigrid* relays are suitable for switching between built-in cellular telephone antennas and car antennas. Teledyne claims its TO-5 device to be the smallest and most reliable sealed relay available. The *Centigrid* type is an industrial subminiature, hermetically sealed armature relay. Teledyne Electronic Technologies. Tel., 081-571 9596; fax,081-571 9637.

RF power amplifiers. *Models 604L* and *607L* from ENI cover the frequency range 500kHz-1GHz and 800kHz-1GHz, with linear outputs of 4W and 7W and gains of 40dB and 43dB respectively. They can cope with any load vswr, from open-circuit to short-circuit, without damage. Holaday Industries. Tel., 0628 478155; fax, 0628 476871.

Coaxial switch. The *Toesel Model TS 360-00* is a fail-safe spdt coaxial switch consuming 220mA at 28V. It is a break-before-make type and has position-indicator contacts rated at 60V/350mA maximum, 4V/10mA minimum. Higher power is optional; a special dielectric material allows the switch to handle 1kW at 1GHz, against 200W at 1GHz for the standard type. Switching time is 100ms and life is about a million operations. Anglia Microwaves Ltd. Tel., 0277 630000; fax, 0277 631111.

Transducers and sensors

Hostile-media pressure transducer. For use in wet and corrosive media, the *Sensit p-192* pressure transducer offers pressure ranges of 1-40 bar gauge reference at a sensitivity of 4mV/bar at full pressure and a maximum of ±1% of full-range error from all causes. Offset voltage is 1mV. Kynmore Engineering Co. Ltd. Tel., 071 405 6060; fax, 071 405 2040.

Angular measurement. The Cline Labs *Angular Measurement System* is a battery-powered standalone system needing no external power or extra electronics. It consists of a gravity-referenced clinometer, digital readout and a cable to connect the two. Angular range is ±60° or ±19.9° to a resolution of 0.1° or 0.01° with linearity varying between ±0.1°, 1% of angle and monotonicity, depending on the range of angles being measured. Frequency response is 0.5Hz. Kynmore Engineering Co. Ltd. Tel., 071 405 6060; fax, 071 405 2040.

COMPUTER

Computer peripherals

Magneto-optical storage. With a 1.3Gbyte capacity and average seek time of under 40ms, Sony's *RMO-S570* magneto-optical drive is meant for digital photography and other data-intensive application. Recording density is not constant over the whole disk, but increases on the outer tracks, the increases occurring in zones; inside each zone the density is constant. Maximum data transfer rate is 2Mbyte/s and a 1Mbyte buffer memory improves performance by reducing mechanical movement. Sony Computer Peripherals & Components. Tel., 0932 816000; fax, 0932 817001.

PCMCIA mass storage. *Solid-State File Cards* by IBM form an effective alternative to hard disks in portable computers. They are in PCMCIA Type 1 and Type 2 form, both with a PCMCIA-ATA interface. 3.3mm thick types have capacities of 3Mb, 5Mb, 10Mb and 20Mb, while the 5mm thick Type 2 has either 30Mb or 40Mb. The cards use a single 5V supply at less power than disks and are not, of course, subject to the relatively long access time of disks. An integral controller and dram buffers eliminate the need for flash memory blocks to be erased before new data can be stored. Blue Micro Electronics. Tel., 0604 603310; fax, 0604 603320.

Software

Data acquisition for Windows. Version 4.1 of The *Windmill* data acquisition software suite for Windows now supports Network DDE in Windows for Workgroups, allowing other Windows applications on other workstations to use data collected by *Windmill*. *Windmill* charting and logging modules are now controllable by other applications supporting DDE, for example by Visual Basic programs. Data acquisition from plug-in cards, bench-top units or other sources is at the rate of 50/s down to 1/hour. Windmill Software Ltd. Tel., 061 833 2782; fax, 061 833 219 ■

Multi i/o via the serial port

Providing 64 lines, this i/o interface compensates for a relatively slow serial link to the host PC by having its own 68000 family microprocessor. J. N. Ellis describes how the interface has a range of uses from switching a led to managing a control system.

Several i/o designs taking advantage of the microcomputer's easy-to-use serial port have appeared, one as recently as June[1]. Many provide two eight-bit parallel ports and involve a dedicated parallel-interface chip.

This RS232 interface differs in that it provides up to eight, 8-bit ports, each programmable as an input or output, from one serial port. It incorporates a high-performance microprocessor, which makes operating it easy.

Text commands can be used directly from a terminal emulator to provide interactive control. Alternatively the interface may be programmed via a programming language on a host PC. In this case, the same text commands can be used, provided that the programming language allows access to the host PC serial port.

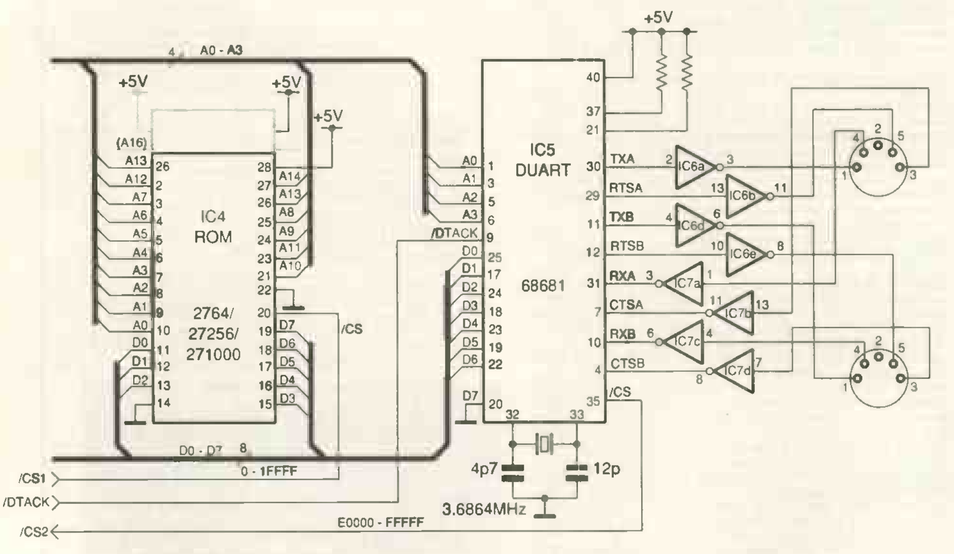
Since this unit operates via a standard RS232 link, virtually any type of PC with a serial port can be used. A basic operating system is available in eprom. An additional benefit of using a microprocessor is that programs – compiled 68000 machine code – could also be downloaded into ram. Alternatively, they could be programmed into rom, and used to operate the ports autonomously. This extends the scope of the interface to use as a programmable controller.

Almost 32K-byte of space, provided by a 256K-bit ram, is available to store small routines. A 1M-bit device ram providing 128K-byte could be used. In fact, decoding circuitry described will drive the full 1M-byte address space of the *MC68008* in 128K-byte segments.

In principle, additional ram could be added using 1M-bit chips up to 512K-byte, but some consideration to the circuitry would be needed. While 32K is regarded as tiny these days, it is adequate for many machine-code programs.

One application for which this interface is eminently suitable is eprom programming. The multiple ports allow for two or three address ports, catering for 16 or 24-bit addressing. Another one or two ports can transfer data in 8 or 16-bit widths, and further port can provide program and verification control signals.

Large eproms can be programmed with this unit, but the RS232 handshaking routines will need careful design to prevent loss of data. Little extra hardware is required. A zero insertion-force socket and jumper pins for selecting the half-dozen or so non-standard pin-outs between different size eproms are needed, together with a selection of logic-controllable programming voltages, for example 12.5, 21 or 24V, or 16V for pals.



Interfacing details

This is a straightforward microprocessor-based design incorporating the often overlooked *MC68008*, **Fig. 1**. The device is an eight-bit external-bus version of the *68000*. It is able to provide much more powerful control capabilities than the *6502* or *Z80*, which are still often used in controllers.

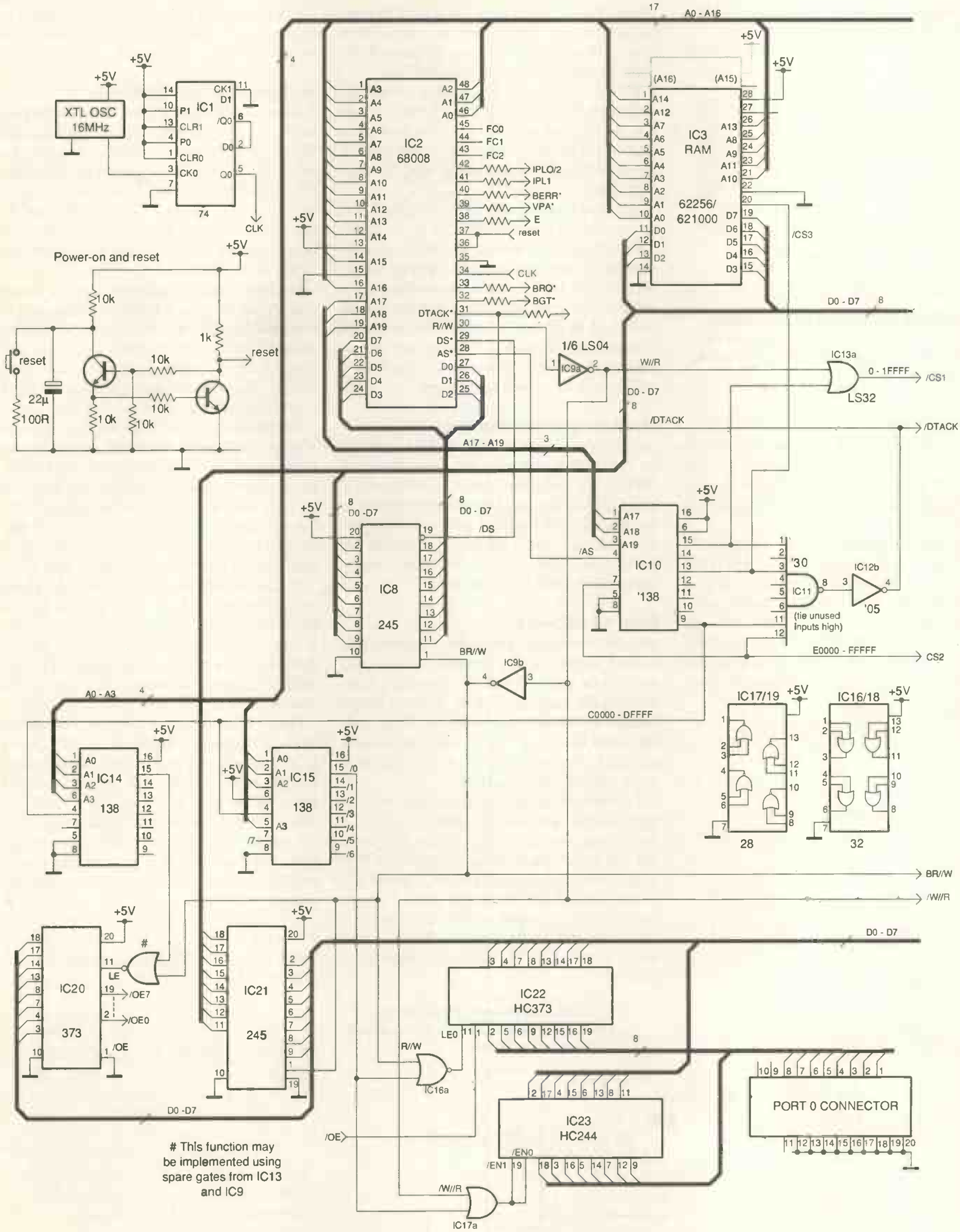
A clock signal drives the processor at 8MHz, corresponding to 2MHz system clocking. A *'138* decoder chip provides eight, 128K-byte address spaces which are filled

Fig. 1. At its most basic, this i/o interface operates 64 i/o lines from PC initiated commands communicated via RS232. Having its own 68000 family processor however, the card can become a versatile programmable controller capable of autonomous i/o.

from zero upwards by a rom, at 40000_{16} and up by ram, at $C0000_{16}$ and up by the i/o ports, and $E0000_{16}$ and up by the universal asynchronous receiver transmitter chip – a *68681*.

Additional ram can be added at unused spaces 60000_{16}-$BFFFF_{16}$, but further address and data buffering may be needed. Static ram keeps the design simple and avoids introducing wait states. The memory map summary is:

Address	Use
00000-1FFFF	rom
20000-2FFFF	rom expansion
40000-5FFFF	ram
60000-BFFFF	ram expansion
C0000-C0007	i/o ports
C0008	control byte
E0000-E001F	uart



7 identical i/o ports to port 0 but driven by /OE1....../OE7 from chip and /O1....../O7 from chip

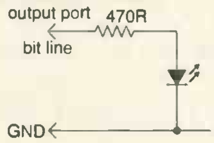


Fig. 2. In hardware terms, connecting an eight-bit d-to-a converter like the ZN428 to the interface involves little more than linking pins to i/o lines.

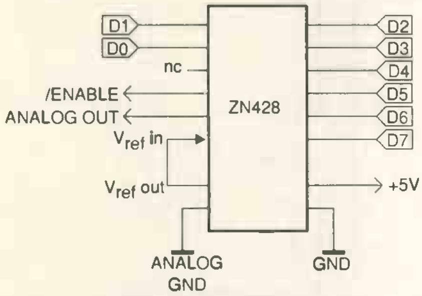
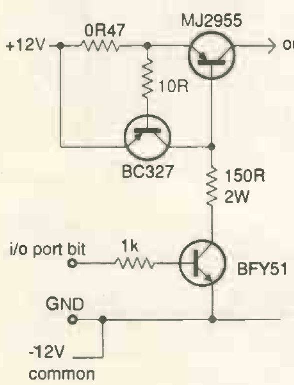


Fig. 3. Capable of delivering up to an amp at 12V, this high-side power switch incorporates current limiting for protecting the circuit in the event of an overload.

Memory map

Two RS232 terminals are provided by the *68681*. Each RS232 socket is driven by *1488* and *1489* serial interface chips. To avoid wasting pins, DIN sockets are used as opposed to the usual 25 pin D type.

Each eight-bit, parallel i/o port comprises an HC373 latch with an LS244 buffer. Decoding these chips is performed using the R/-W line. A read activates the selected 244 buffer, while a write activates the selected latch, allowing data written to an output port to be read back. This configuration was chosen in preference to other parallel i/o ICs for three reasons – software setting up is minimal, drive current is greater, and additional timers and control signals are unnecessary.



To allow each port to be used as an input or output, a ninth decoded HC373 latch controls each output enable pin on the eight i/o port latches. A control byte is written to a 'ninth' port address to select input or output functions on each i/o port. The bit number in this control byte corresponds to the port number of the eight i/o ports: bit 0 controls port 0, etc.

Writing a zero in a bit location allows the corresponding port to become a latched output. Writing a one to that bit turns the output latches off enabling that port to become an input, although inputs can be read from an output port.

It is not a good idea to 'force' an output port to be driven by something else as an input. Should this be essential, it can be accomplished by inserting an open-collector buffer between the latch and the input chip with a suitable pull-up resistor. If any port is required only to be an input or an output, the redundant chip need not be used. The control port is output-only, so if the control byte is needed, it will have to be copied to memory.

Hardware considerations

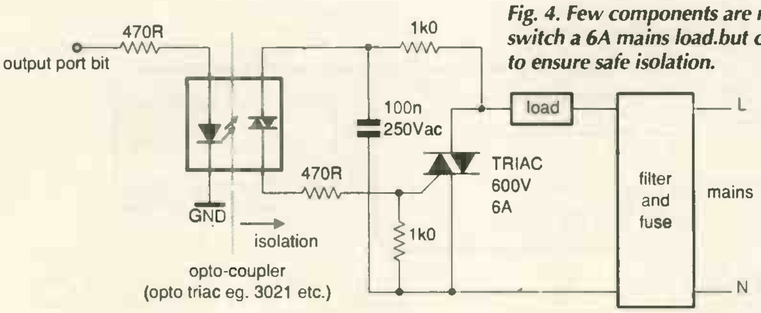
Initially, the bit rate is set to 9600baud. It can be changed by software but only through a 68000-code program. Power supplies of 5V and ±12V are needed. I considered whether to use RS232 interface chips with on-board voltage generators, but I rejected this idea as the devices are expensive. In addition, many applications need ±12V anyhow.

Control software

An eprom-resident controller program will receive simple ascii commands to read and write to the ports. It will also provide a rudimentary file handling system, in which the i/o routine is a separate file called io. This must be started by typing 'run io' from the terminal emulator after switching on, or sent by a program running on the host PC.

To write to a port, the command W*x,y* is typed on the terminal emulator, or sent by a controlling program. Value *x* is the port number and *y* the data in hexadecimal form. The first command is usually to port 9, to set the output status of the other eight ports. Thus, to set ports 0-3 as outputs and 4-7 as inputs, the first command would be w8,f0. Port numbers are counted from zero and the command is not case sensitive.

Fig. 4. Few components are needed to switch a 6A mains load.but care is needed to ensure safe isolation.



To read from a port, the command R<x> is typed. Data is returned using hexadecimal ascii text of the form (*x*)=*y*. A menu is available, command M, with help, H, and quit, Q, to quit the i/o routine. Once quit, the minimal operating system software is in the main command mode to receive, send or run programs. It can even provide a list (dir) of programs in memory. To re-run the i/o program, just resend the command 'run io'.

Rudimentary file transfer

Although non-standard, the file-transfer protocol is reasonably simple. Command 'Re(ceive) <filename>' initiates receiving of a file, which should be given a filename. Filenames can be up to 32 characters, and can be anything, including spaces. The module takes text until new line characters carriage-return/line-feed, or 32 bytes – whichever is first – as the filename.

Once the filename has been received, binary data should be sent. A break should be sent to complete the file transfer. Terminal emulators usually carry out these tasks, but if not, you could write a routine to execute the 'send break'. There is no error checking and file lengths are arbitrary.

File transmission from the interface requires a 'send <filename>' command, at which point the name of the file in memory should be supplied. Transmission is initiated when the receiver indicates it is ready by issuing 'OK' through the serial link. This is to stop transmission until the host PC is ready to receive. It may have to be programmed to do this.

Programs sent as a file are run by typing 'run <filename>'. Unless control software is needed to operate the ports at speed, it is likely that normal operation through an RS232 interface via a program on the host PC will suffice. Control software must be 68000 machine code and written as PIC.

Figure 2 shows how to connect a digital-to-analogue converter. **Figure 3** is a 12V power switch and **Fig.** 4 a mains-power switch using an opto-isolator. Logic for the i/o latches is cmos, rather than ttl, to provide a full 0-5V swing. This simplifies additional circuitry.

Examples of further drive and input circuits that could be used with this interface were published in the article mentioned in the reference. ■

Reference

Teliki, W., *Applied i/o Design for the PC*, , *EW&WW*, June 1994, p. 452.

Control software in eprom

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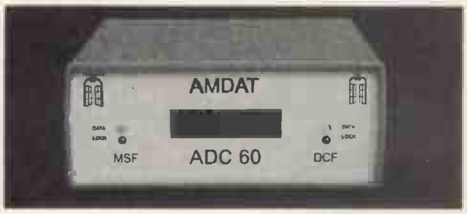

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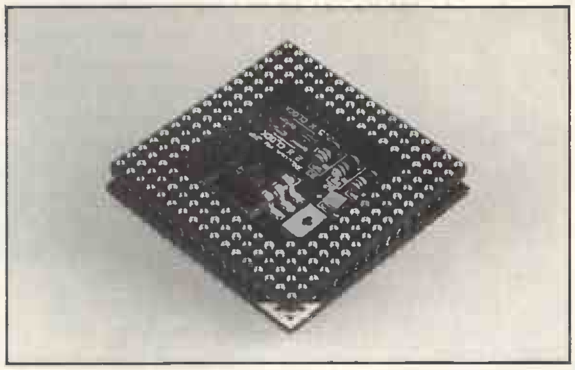

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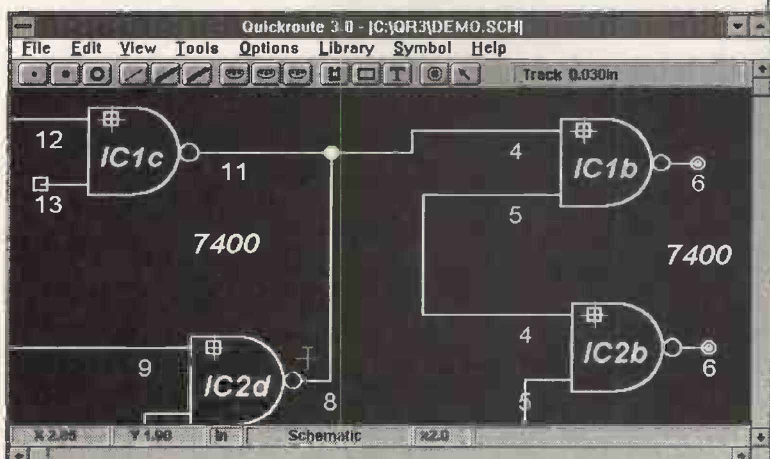

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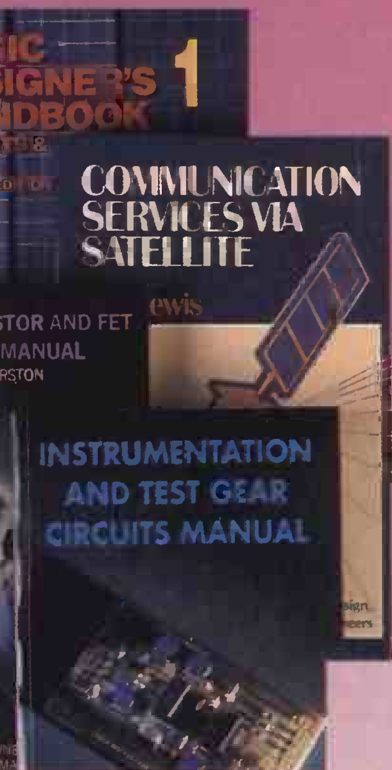

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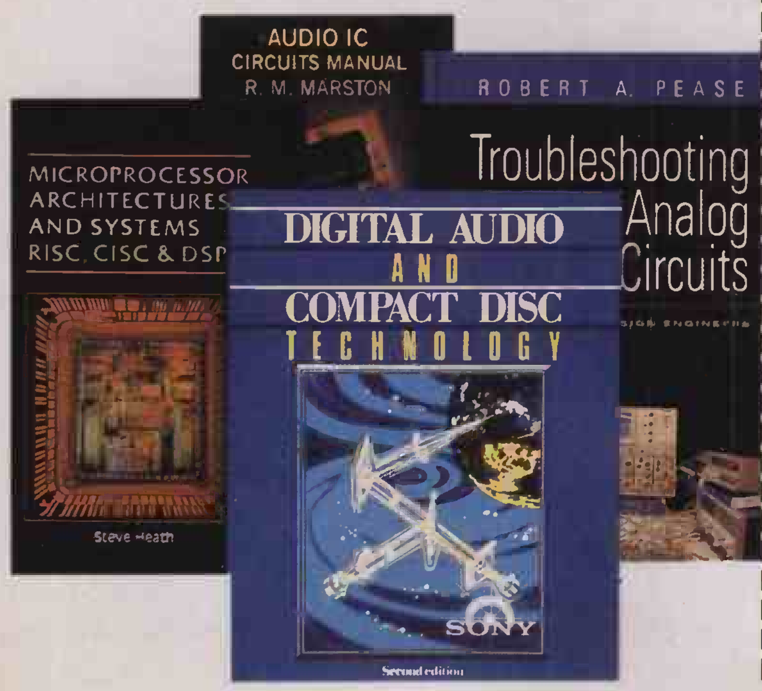

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Wideband op-amp delivers broadcast-quality video

At PAL and NTSC frequencies, the *OPA628* op-amp has a differential gain error of 0.015% and a differential phase error of 0.015° when driving a back-terminated 75Ω cable.

As the device data sheet describes, these specifications are made possible using a classical op-amp architecture involving true differential and fully symmetrical inputs. Separated power supply pins for the input and output stages also eliminate the effects of package and wire-bond parasitic effects.

In performance terms, the device has other interesting features. These include unity-gain stability with a bandwidth to 160MHz, 90dB spurious-free dynamic range and a 2nV/√Hz noise figure. The two-tone third-order intercept is 60dB and gain is flat within 0.1dB to 30MHz.

Both Spice models and evaluation boards exist for the *OPA628*. The data sheet carries in-depth details of the device's performance together with discussions on many aspects including driving capacitive loads, thermal considerations, input protection and pcb layout. The three application circuits shown here are included in the note but there is no further specific information on their operation.

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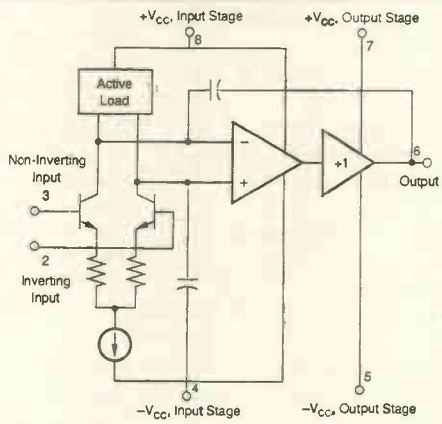
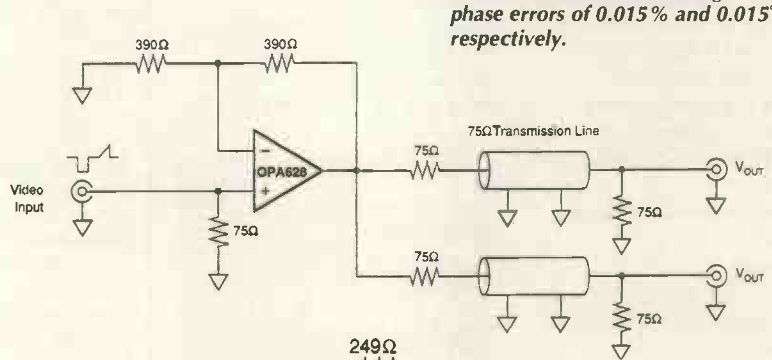
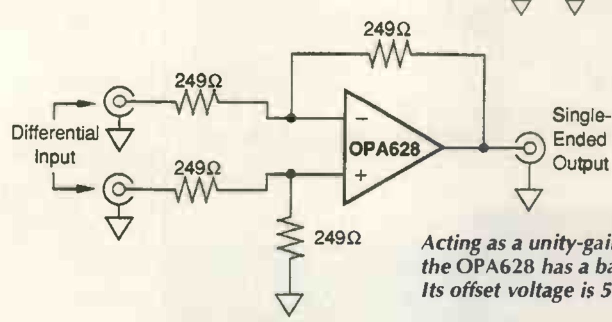


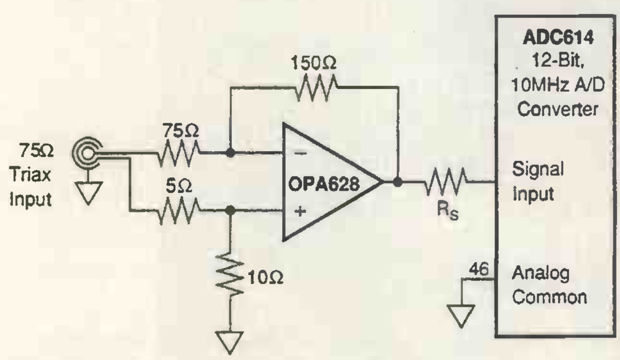
To maximise performance at high speed, the OPA628 has four of its eight pins devoted to power supply – two for the input stage and two for the output. The device operates from ±5V rails.



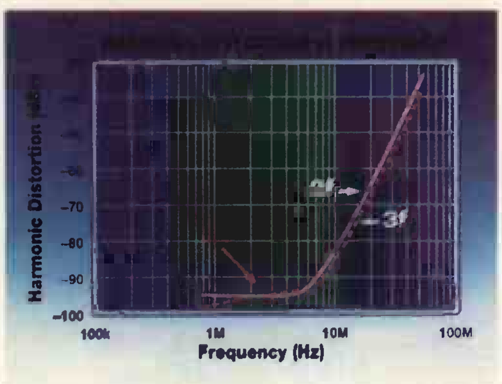
This video distribution amplifier has a high output drive capability of 6V pk-pk into 75Ω. Driving video into a 150Ω load, i.e. 75Ω back terminated, the 628 exhibits differential gain and phase errors of 0.015% and 0.015° respectively.



Acting as a unity-gain difference amplifier, the OPA628 has a bandwidth of 160MHz. Its offset voltage is 500V.



When driving a high-speed a-to-d converter at a gain of two, the OPA628 introduces a linearity error of less than 1lsb in a 15-bit conversion.



At unity gain, the 2f curve is roughly the same as the 3f curve shown here. Increasing gain to 2V/V causes the 2f curve to rise but the 3f curve shape and position remain virtually unchanged.

Function generators use analogue trigonometric synthesiser

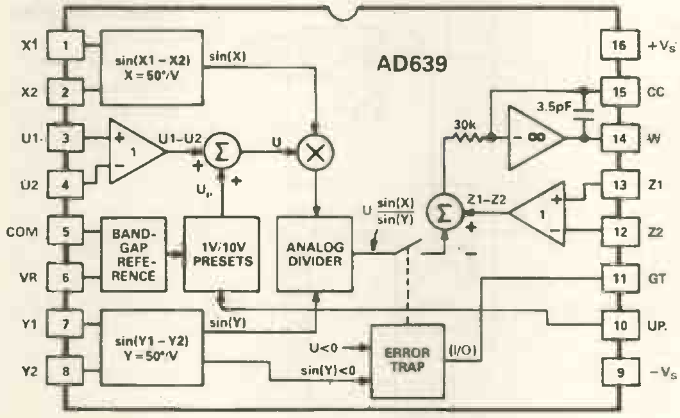
Via pin-strapping, the *AD639* function generator provides all the standard trigonometric functions and their inverses. According to the device data sheet, its law conformance and total harmonic distortion surpass figures previously attained using analogue shaping techniques. Also in the data sheet are a number of application circuits. Two of them are described here, namely a gated function generator and a four-quadrant sine multiplier.

Compared with using rom look-up tables and d-to-a conversion, the device is also faster; in sine mode, bandwidth is typically 1.5MHz. Unlike other function synthesis circuits, the *AD639* provides a smooth and continuous sine conformance over a range of –500° to +500°. When generating a sine wave, law conformance is within 0.02% and distortion levels of –74dB are attainable for triwave to sine wave conversion.

The device generates a basic function representing the ratio of a pair of independent sines:

$$W = U\frac{\sin(x_1 - x_2)}{\sin(y_1 - y_2)}$$

Differential angle arguments are proportional to the input voltages X and Y scaled by 50°/V. Using the 1.8V on-board reference any of the angular inputs can be preset to 90°. This provides the means to set up a fixed numerator or denominator (sin 90°=1) or to convert either sine function to a cosine (cosθ=sin(90°–θ)). Using the ratio of sines, all trigonometric functions can be generated.

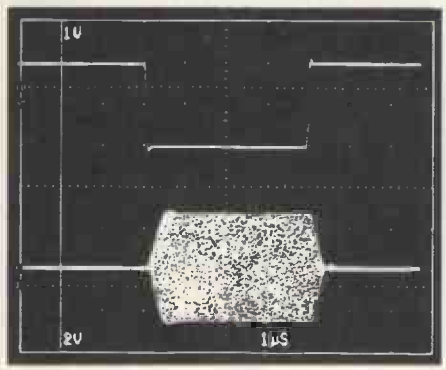


Capable of generating trigonometrical functions including sin, cos, tan, cosec, sec, cot, arcsin, arccos and arctan, the AD639 can produce a sine wave with 0.02% law conformance.

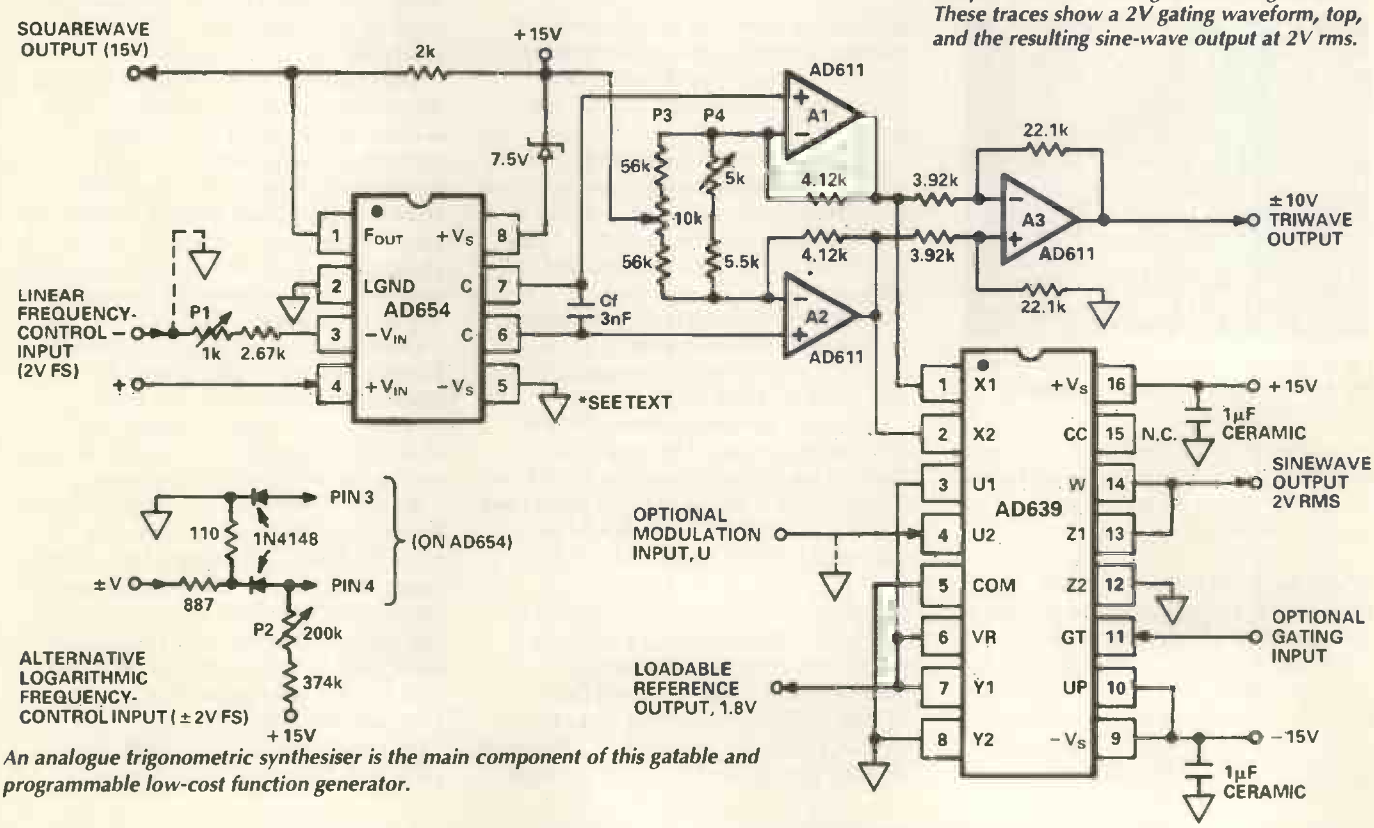
Amplitude of the function is proportional to a voltage U, which is the sum of an external differential voltage (U_1–U_2) and an optional internal preset voltage, U_p. Control pin UP selects a 0V, 1V or 10V laser-trimmed preset amplitude which may be used alone (U_1–U_2=0) or internally added to the U_1–U_2 analogue input.

At the output, a further differential voltage Z can be added to the ratio of sines to obtain the offset trigonometric functions versine (1–cosθ), coversine (1–sinθ) and exsecant (1–secθ). A gating input is available enabling or disabling the analogue output. This pin also acts as an error flag output in situations where a combination of inputs will cause the output to saturate or to be undefined.

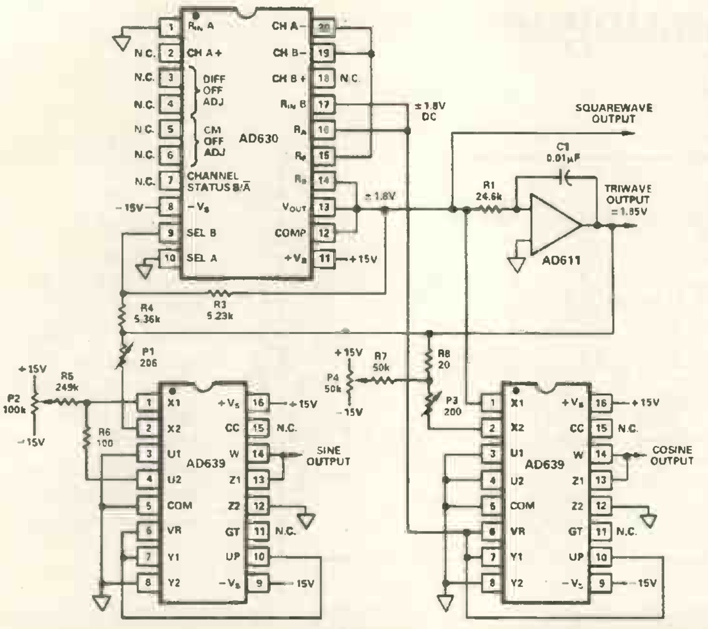
In the inverse modes, the argument can be



Output of the function generator is gatable. These traces show a 2V gating waveform, top, and the resulting sine-wave output at 2V rms.



An analogue trigonometric synthesiser is the main component of this gatable and programmable low-cost function generator.



Designing a quadrature oscillator around the AD 639 trigonometric function generator avoids many of the problems associated with integrator/sign-inverter combinations.

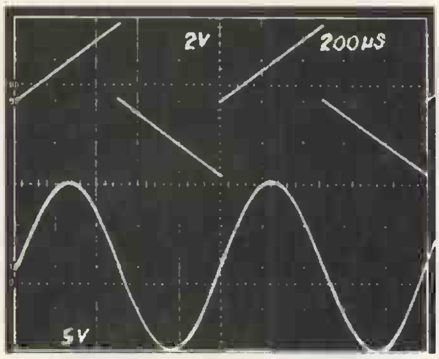


Fig. 1. Top waveform is the difference voltage between the triwave and squarewave. Resulting output is shown in the bottom trace.

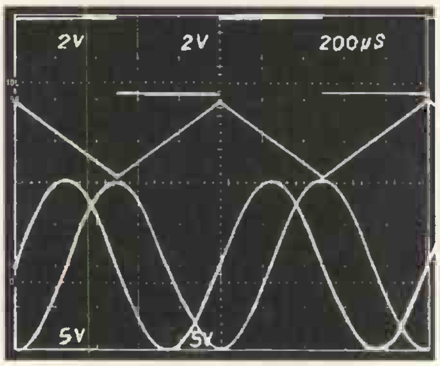


Fig. 2. Timing relationships between all outputs of the quadrature oscillator.

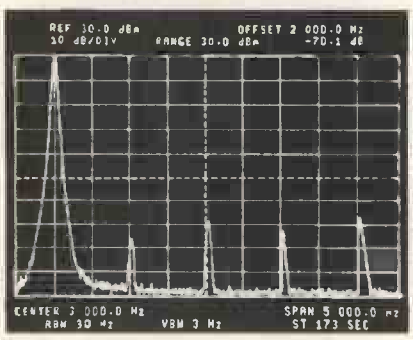


Fig. 3. Spectrum of cosine output at 1kHz for the AD639-based quadrature oscillator.

the ratio of two input signals. This allows the user to compute the phase angle between the real and imaginary components of a signal using the arctangent mode.

Wide-range waveform generator. This is an inexpensive signal generator, providing voltage control of frequency from 20Hz to 20kHz and a preset sine amplitude of 2.8V (within 0.1dB of 2V rms). This output may be further modulated by an input of up to ±2.8V to input U2, or gated off by an input of +1.5V or more to input GT; the oscillograph shows the gated response. If required, a further input can be summed into Z2. The sine output can be set to 10V amplitude by connecting UP to VR and grounding U1.

An *AD654* is used to generate the triwave which appear across timing capacitor C_f, and is buffered, amplified and level-shifted by A_1 and A_2. Using a spectrum analyser, P_3 and P_4 are adjusted to minimise even- and odd-harmonic distortion, respectively.

The triwave linearity is not good enough to realise the inherent capabilities of the *AD639*, but total harmonic distortion is in the –50dB to –60dB range.

Op-amp A_3 provides further gain for a ±10V triwave output. The square-wave output is taken directly from the *AD654* and is unbuffered. It swings between ground and +15V; if pins 2 and 5 of the *AD654* are connected to –15V, this output is 30V pk-pk.

Scaling with the linear input (shown) is 10kHz/V, calibrated using P_1. Frequency can be controlled manually using a potentiometer and the V output of the *AD639*, P_1 has sufficient trim range to provide a full-scale frequency of 20kHz with the 1.8V peak input. The alternative input scheme provides a log-sweep response with an approximate scaling of 10^VkHz (where V is in volts). The range is now from about 10Hz to 100kHz; the frequency should be set to 1kHz with V=0, using P_2. Frequency is now sensitive to variations in both temperature and the +15V supply, but stability will be adequate for many applications.

Because of the exceptionally wide angular range of the numerator function of the *AD639*, it is possible to generate sinewave outputs with 2, 3, 4 or 5 times the triwave frequency using cosine mode for even multiples of the sine mode for odd multiples.

For example, to multiply the output frequency by 3, use the sine function with the X input driven to 5.4V (±270°). Distortion remains low; all harmonics are typically under –50dB, even for the frequency quintupling mode.

Sine/cosine oscillators. Quadrature oscillators generate a pair of sinusoidal outputs displaced by 90°, and are invariably based on a state-variable loop comprising two integrators and a sign-inverter. This approach however needs additional circuitry to control the amplitude of the oscillation. In addition, a trade-off arises between the settling-time of this control circuitry and the distortion level, which is particularly troublesome at low frequencies. Amplitude balance of the two outputs depends on the matching of two time-constants and two tracking analogue multipliers or multiplying d-to-a converters are needed if the frequency is to be programmable.

These problems are avoided using a function-shaping technique based on a triwave oscillator. Only one time-constant is required, so its frequency is more easily controlled.

Amplitude control is eliminated by using the scheme shown. The two outputs have accurate amplitudes of 10V, without the need for an external reference source. Alternatively, they can be individually controlled by external voltages, without any effect on frequency. Variable-amplitude sine and cosine outputs can be added using the Z-input to provide continuously-variable phase control of the output.

The triwave oscillator has an *AD630*. This device alternates the sign of the 1.8V reference from one of the *AD639*s to generate a square-wave output of ±1.8V amplitude. An integrator, formed by R_1, C_1 and the op-amp, generates the triwave.

Amplitude of the triwave is determined by the ratio of R_3 to R_4, and is nominally ±1.845V. This is 2.5% higher than needed at inputs of the *639*s, providing the adjustment range needed minimise distortion. In many applications, all adjustments can be eliminated. To do this, make R_2=R_4=5kΩ, omit $P_{2,4}$, $R_{5,7}$, and replace $P_{1,3}$, and $R_{6,8}$ with short circuits.

Frequency is nominally $^1/_4C_1R_1$, and is 1kHz with component values shown. A variety of methods may be used provide external control of frequency, including the use of another *AD630* in series with R_1, or a multiplying d-to-a converter.

Sine output is generated using the triwave directly. Potentiometers P_1 and P_2 should be adjusted using a spectrum analyser for minimum odd-order and even-order harmonics, respectively. The cosine is generated by using the *difference* between the triwave the square-wave, as shown in the upper wave form of **Fig. 1**. This composite voltage first generates a sine-function over range 0 to +180°, then over the range 0 to –180°, to produce the function shown in the lower wave form, which can be seen to be 90° out of phase with the triwave.

The complete set of wave forms available from this generator are shown in **Fig. 2**. Potentiometers P_3 and P_4 are adjusted for minimum odd-order and even-order cosine harmonics, respectively. **Fig. 3** shows the cosine spectrum for a well-adjusted circuit. Due to the finite transition time back to the baseline in the drive voltage to the cosine generator, a brief spike occurs at the zero-crossing of this output.

Frequency components will be beyond the bandwidth of the output amplifier in the *AD639*, and the energy contained in these spikes will not generally be troublesome. They may be further reduced, if necessary, by adding a capacitor between pins 14 and 15, to roll off the *AD639* output response.

Analog Devices, Station Avenue, Walton-on-Thames, Surrey KT12 1PF. Tel. 0932 253320, fax. 0932 247401.

Switching with igbts reduces lamp ballast size

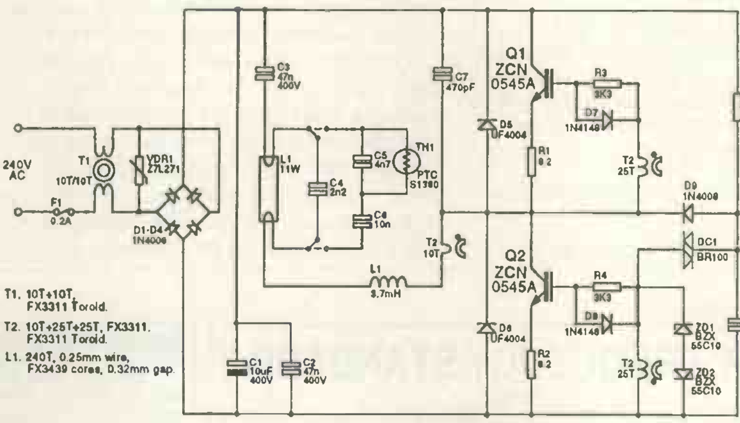
Nearly all insulated gate bipolar transistors, igbts, are high-power devices, but there is a pair of medium-power, low-cost devices in TO-92-style packaging. These are the n-channel *ZCN0545* and the p-channel *ZCP0545*.

The circuit is an 11W off-line fluorescent lamp ballast using two *ZCN0545A* igbts. Efficiency of the circuit is such that it allows the TO92-format E-line igbts to replace the TO220/TO126 bipolar or mosfets commonly used in this application. This both lowers component costs and gives a reduction in circuit size – critical in integral lamp/ballast designs.

The 300ns turns-off capability of the *ZCN0545A* would allow operation at up to 100kHz but the working frequency of the design was set at 40kHz to minimise losses and hf interference.

By controlling the phase of the current flowing in the igbts so that cross-conduction does not occur, switching losses have been virtually eliminated. Also the low effective $R_{DS(on)}$ of the *ZCN0545A* keeps conduction losses to around 60mW in each device.

The first curves below show the voltage and current waveforms of the igbts. Curves in the lower graph show an expanded view of the critical turn-off behaviour of the *ZCN/ZCP0545A* pair. Note in particular that the drain current falls to zero before the



Since the ZCN igbts are efficient, they can replace TO220/126 bipolar or mos transistors and provide savings in both cost and circuit volume.

Inside the igbt

This relatively new type of transistor has a mosfet input device followed by a bipolar amplifier. The high input resistance is ideal for direct drive from microcontrollers. In addition, igbts have a low $R_{DS(on)}$. For a given chip size and BV_{DSS}, the on resistance of an igbt is less than 10% that of a standard high-voltage mosfet at high current.

Like a bipolar darlington, the igbt needs a drain-source voltage of 0.7V before current flows. If the drain-source terminals are reverse biased, the drain-source diode of the input mosfet cannot conduct since the base-emitter junction of the output bipolar transistor is in series. In many applications this provides a very useful reverse blocking capability.

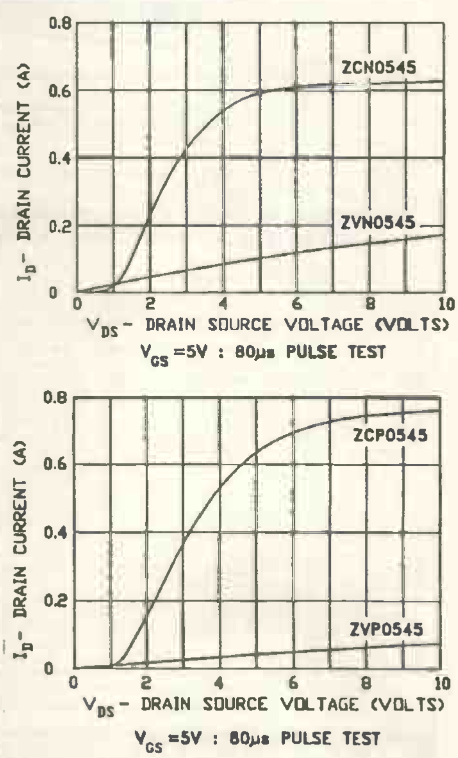
Switching speed is dominated by the characteristics of the bipolar transistor, which can be optimised for either speed or saturation voltage. The *ZCN0545* and *ZCP0545* are designed to be very fast at switching on – in less than 20ns – and their off time is less than 300ns. This makes them suitable for switching applications up to 100kHz.

Since the structure of igbts includes an scr, they have a drain current which, if exceeded, will cause the device to latch up. Latchup can lead to device destruction in some applications. Consequently, the pulsed drain current rating of the igbt should not be exceeded. This rating is temperature sensitive, falling as temperature increases.

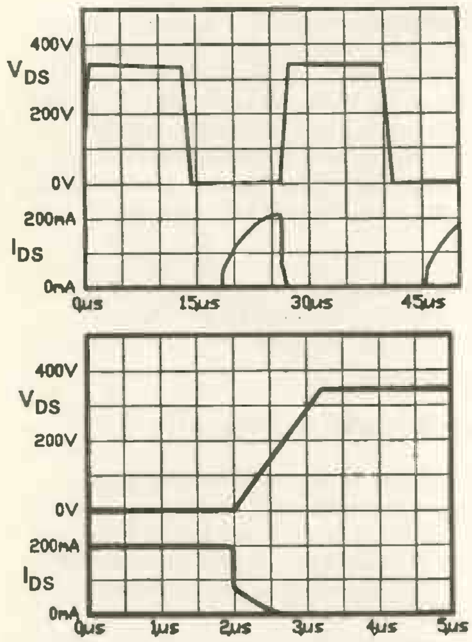
Equivalent $R_{DS(on)}$ of an igbt on the other hand does not change significantly with temperature. Standard mosfet resistances double as temperature is raised from ambient to the device upper limit.

Output characteristics of the *ZCN0545* and *ZCP0545* igbts are illustrated on the right. These curves show typical I_D versus V_{DS} for a 5V logic level gate drive.

To indicate the improvement the igbt structure gives over standard mosfets, graphs of the typical performances of two mosfets with an identical chip size, the *ZVN0545* and *ZVP0545*, have been plotted for comparison.

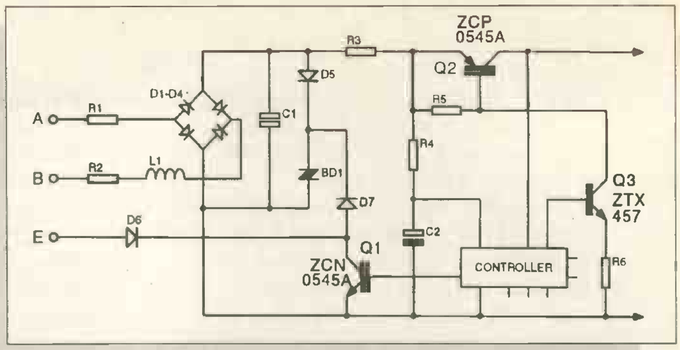


In each graph, the top curve shows characteristics of a ZC type medium-power igbt while the lower curve illustrates a similarly-sized mosfet. For a given chip area and voltage rating, on resistance of the igbt is less than 10% that of a mosfet. Upper graph is n-channel, lower p-channel.



Voltage and current waveforms for the fluorescent lighting ballast circuit, left, and an expanded view of critical turn-off behaviour of the igbt used. Because drain current falls to zero before drain voltage rises significantly, switching losses are low.

Benefits of igbts in telephone hook switches



To withstand normal telephone operating voltages and lightning-induced transients, transistors with breakdown voltages of 250-400V are needed for telephone hook switches, diallers, etc. Normal currents can rise to about 150mA, or much higher on transients.

This 'feature-phone' interface shows the lower igbt being used as an earth recall switch. It provides as high an input impedance as the often used mosfet but lower on-voltage at high supply currents.

Electronic hook switching is provided by the upper p-channel igbt. Its controlled gate drive limits drain current during transients. Expensive p-channel mosfets or npn/pnp bipolar pairs are normally used.

Suppressors used must not operate below 200V. Devices selected normally allow 270V worst-case peaks during transients. Having a drain-source breakdown rating of 450V, the igbt shown simplifies design. Transient protection is aided by D_5 and BD_1.

drain voltage rises significantly, giving low switching losses.

Gate drive for the igbts come from a current transformer connected in series with the ballast inductor. This transformer controls the switching frequency of the circuit and zener diodes $ZD_{1,2}$ set gate drive voltage for both igbts. A diac is used to give an initial kick to start the circuit and the transformer T_1 and vdr control line borne transients and interference.

Two strike circuits can be used. The simplest – and lowest cost – is to use a single capacitor which gives the circuit an instant start characteristic. However this has the disadvantage that the lamp strikes before the heaters heat up fully, leading to tube-end blackening and some reduction of tube life when switched on and off frequently.

Using a two capacitor/ptc-thermistor starter combination improves matters. At turn-on, a 10nF capacitor forces a high heater current to flow until the series connected ptc warms. Resistance of the ptc increases rapidly, causing the voltage across the tube to rise until the tube strikes. Since the tube strikes only after its heaters reach working temperature, life is extended. However this starter option is more expensive and gives a noticeable turn-on delay of around 1-2 seconds.

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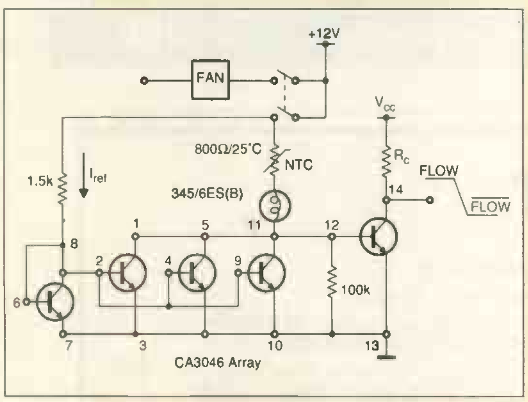
One chip air-flow monitor

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

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

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

John A Haase
Fort Collins
Colorado
USA



Air-flow monitor provides a two-level indication.

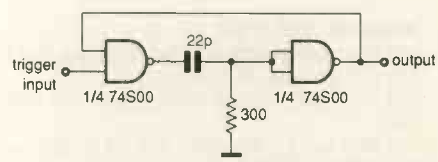
Monostable flip-flop pulses down to 10ns

Since there is no monostable member of the *74S* series of ttl logic, a monoshot comparable in speed with the rest of the family must be made from gates. This circuit produces pulses less than 10ns wide and with 2.5ns transitions.

The falling edge of a 7ns pulse to the first gate triggers the circuit to give the 10ns output when R=300Ω and C=22pF. This is rather better than the performance of ecl monoshots, which give a minimum output pulse width of 10ns.

The two spare gates could be used to invert input and output or in a second circuit.

I K and S R Kaul
Bhabha Atomic Research Centre
Bombay
India



Spare pair of two-input Nand gates performs the function of the monostable flip-flop missing from the 74S ttl series.

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Guitar fuzz box uses radio chip

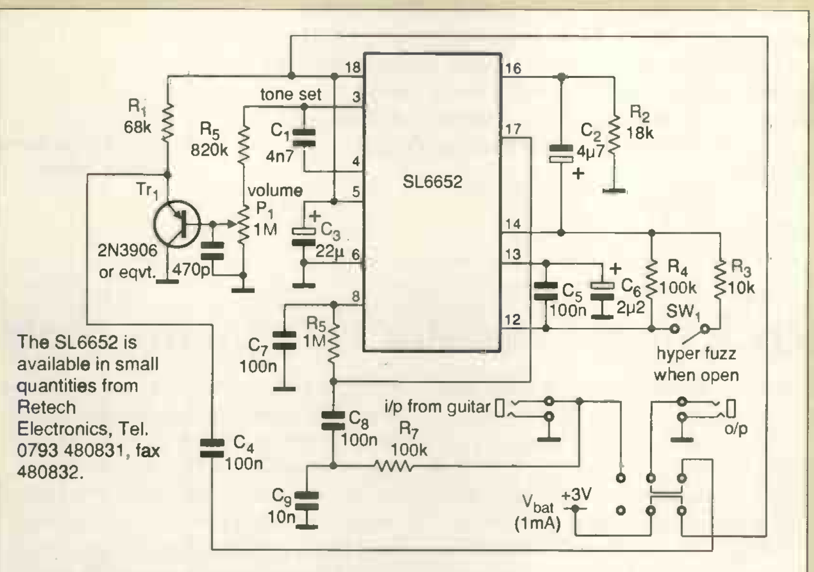
Fuzz boxes to produce the 'heavy metal' sound from guitars rely on limiting the input signal to generate odd harmonics. GEC Plessey's *SL6652* is a low-power IF/AF circuit meant for fm cellular radio, naturally containing a good limiter.

The circuit shown needed no special screening or layout and gave good results without any decoupling problems, the only initial drawback being a harshness in the sound. Resistor R_7 and C_9 solved that problem and two professional guitarists have approved the results. In this application, the *SL6652* draws about 1mA so two AA batteries last a long time.

To use the circuit, turn the fuzz box volume control to minimum and slowly increase the volume, while strumming the guitar, until the correct drive level for the amplifier is obtained. The guitar volume control now functions as a 'sustain' control.

As regards the hyper-fuzz switch – try it and see!

Dave Mapleston and Steve Newton
GEC Plessey Semiconductors



Limiter section of FM radio chip produces 'heavy metal' sound from a guitar.

Triggered sawtooth generator from a phase-locked-loop IC

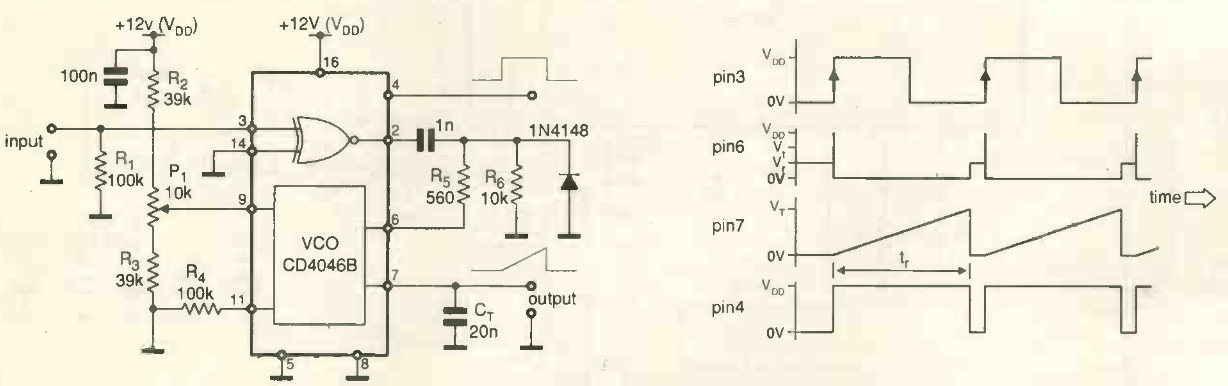
One phase-locked loop IC generates triggered linear sawtooth waveforms, referred to ground, of constant amplitude and positive-going. Alternative methods involve several ICs and multiple power supplies.

In the absence of a trigger pulse, the vco section of the pll holds pin 7 at ground, a current I_c appearing at pin 6, determined by the value of R_4 and the voltage at pin 9. This current through $R_{5,6}$ sets the voltage on pin 6 lower than V_T, the transfer voltage of the vco inverters.

Trigger pulses are buffered by the ex-or phase comparator and increase the pin 6 voltage to change the state of the vco flip-flop, in which condition pin 6 is now grounded and I_c now appears at pin 7, where it charges C_T. When the charging ramp on pin 7 reaches V_T, the flip-flop again changes state and the capacitor discharges into pin 7. The circuit is now stable until the next trigger pulse. The vco output at pin 4 goes high during the ramp.

Ramp duration is around 1ms for these values, but can be set to last from a few microseconds to several seconds by varying C_T, R_4 and the voltage on pin 9.

M S Nagaraj
ISRO Satellite Centre
Bangalore
India



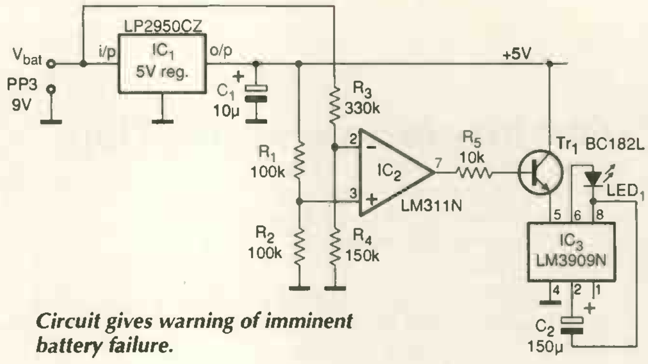
Single-IC, wide-range triggered sawtooth generator produces a linear, ground-referred ramp from microseconds to seconds in duration. Timing is shown on the right.

Low battery-voltage indicator

In circumstances in which battery failure might lead to loss of data – for example, in field data logging – this device will warn of impending doom by means of a flashing led.

Regulator IC_1 powers the circuit from the 9V battery, drawing a very low quiescent current, and supplies a reference voltage via $R_{1,2}$ to the comparator. If the battery-derived input to the comparator falls to the threshold voltage set by $R_{3,4}$, 8.15V with these values, Tr_1 turns on and enables the led flasher oscillator IC_3, its flash rate being set by C_2 (2.3Hz in this case). Changing R_3 to a 1MΩ variable component allows any battery voltage to be monitored. Circuit current consumption is 1.5mA and 2.5mA when the led flashes; micropower devices would reduce this considerably.

Kamru Miah
CSL
Slough



Circuit gives warning of imminent battery failure.

Bench filter evaluator with tuning control

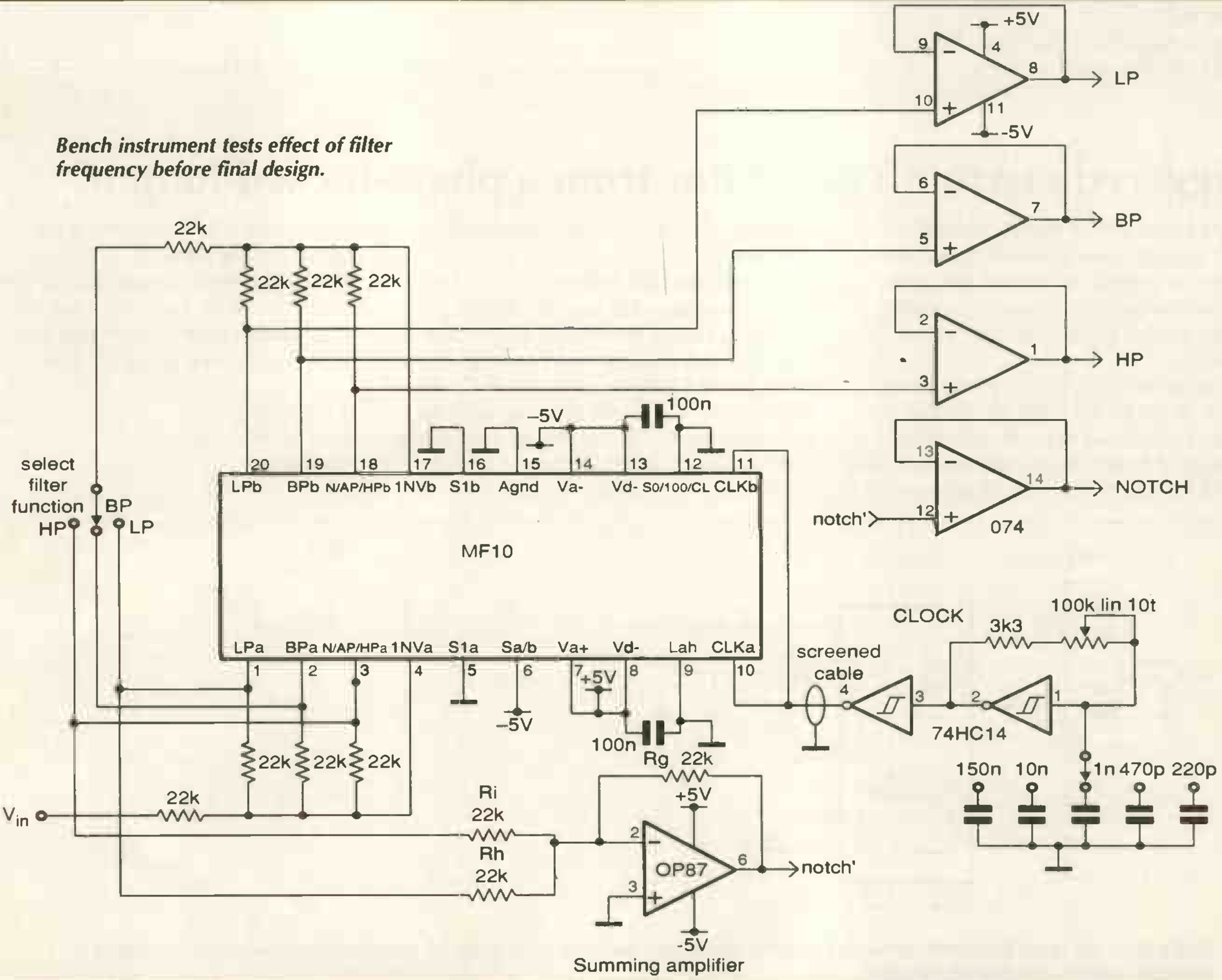
Cascading the two halves of a National Semiconductor *MF10* dual cmos switched-capacitor filter IC makes a bench instrument to evaluate the effects of varying the frequency of a prototype filter section before committing yourself to a final design.

This instrument is effectively a state-variable filter giving the characteristics of low-pass, band-pass and high-pass types with 80dB/decade slopes. A summing amplifier combines high and low-pass sections to give a 40dB/decade notch filter. Clock frequency is variable from 0.83Hz to 14.7kHz in five ranges and it would be simple to drive pins 10 and 11 directly with an external clock, via an internal/external switch.

Centre or break frequency is $f_o = f_{clk}/100$, which is also the notch frequency. Outputs are buffered by the four op-amps.

P J Hale
University of Humberside
Hull

Bench instrument tests effect of filter frequency before final design.



Automatic gain-adjusting bridge amplifier

When a measuring bridge is near balance, amplifier gain must be high to cope with the small bridge signal. When out of balance, however, the large gain is unnecessary and could lead to instability, so that a dynamic setting of gain is the ideal.

In **Fig.1**, the diodes in the feedback loop of a bridge difference amplifier increase their resistance at low signal levels, increasing the amplifier gain. If feedback resistance is $R+\delta R$, δR being the change in diode resistance, then amplifier gain is expressed as $1+3\delta R/4R$. A practical circuit using the *INA 105*, with $R = 25\text{k}\Omega$, is shown in **Fig. 2**.

The circuit in **Fig. 3** uses the same approach to increase the low-signal gain of the classic three-op-amp instrumentation amplifier, in this case an *AD524*. ■

Kamil Kraus
Rokycany
Czech Republic

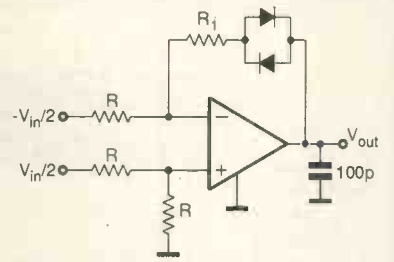


Fig.1. At low input levels, the feedback diodes have higher resistance, increasing the gain of a bridge amplifier near balance.

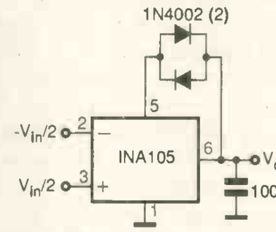


Fig.2. Practical circuit using the approach of Fig.1

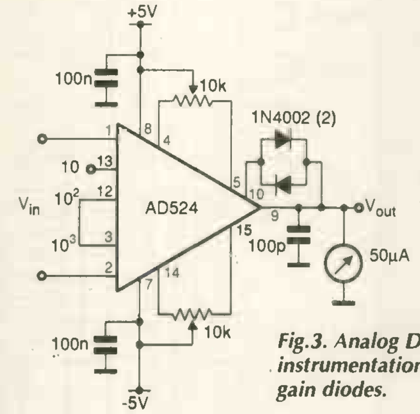


Fig.3. Analog Devices's AD524 three-op-amp instrumentation amplifier with the dynamic-gain diodes.

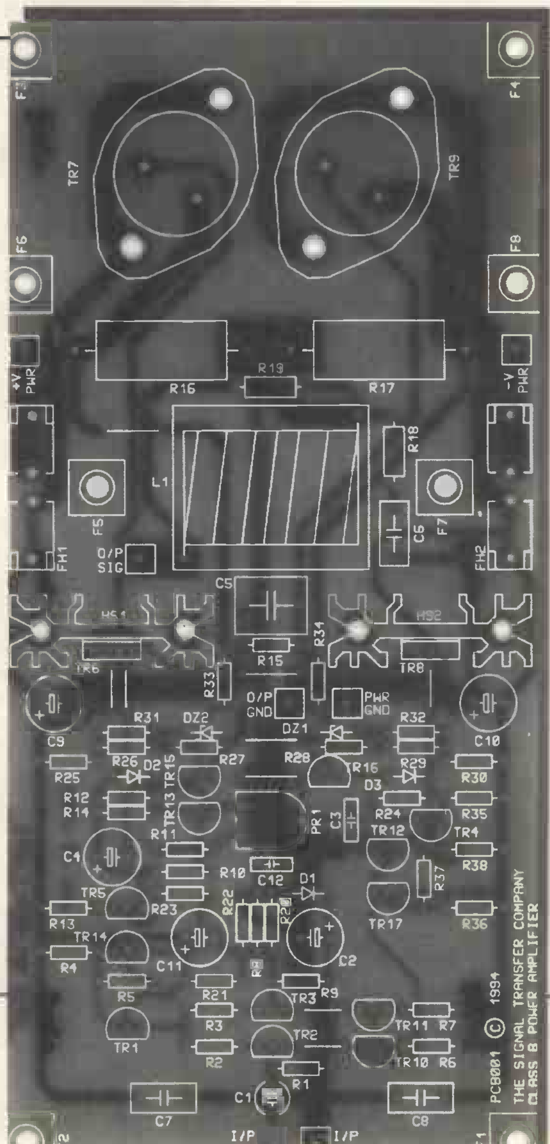
PCBs for Douglas Self's power amplifier series

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

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

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

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



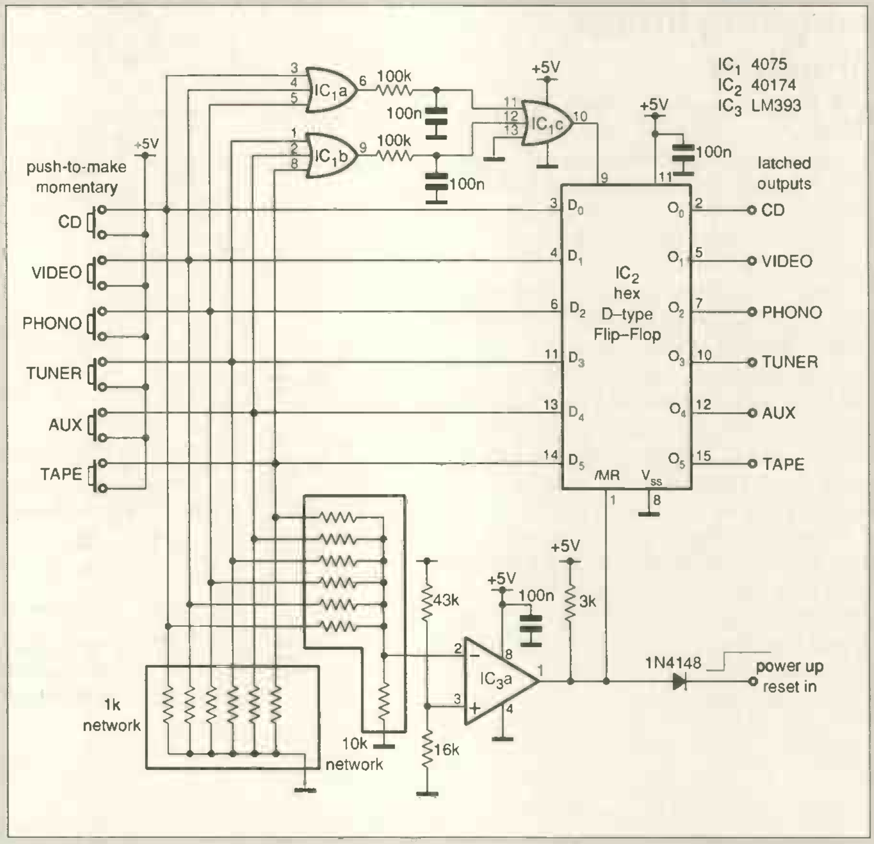
A better feeling about channel selection

The row of mechanically ganged push-button switches sometimes used for channel selection in audio amplifiers and television receivers is effective, but lacking in the feel of quality. This circuit arrangement uses momentary-action, light-touch switches without complicated circuitry.

Pressing any switch causes IC_1 to emit a clock pulse to IC_2 and latch it, the relevant output from IC_2 going high. However, if more than one switch is pressed, the output of IC_3 is low to inhibit all outputs and act as a mute.

The outputs could be buffered and used to drive relays, perhaps also illuminating indicator leds to confirm the selection.

A P Scrimgeour
London N4



Momentary-contact, unganged switches replace the mechanically ganged type to provide a lighter touch.

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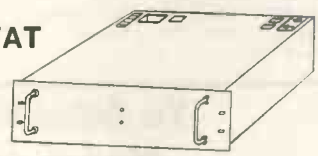

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2019A as above, improved spec	£1950

2022D synthesized signal generator	£1950
2438 520MHz universal counter/timer	£200
2828A/2829 digital simulator/analyser	£1000
2955 mobile radio test set	£3000
2955A + 2960 cellular adapters . . . various configurations . . .	call
6059A signal source 12-18GHz	£750
6460/6420 power meter 10MHz – 12.4GHz 0.3uW – 10mW	£350
893B audio power meter	£350
OA2805A PCM regenerator test set	£750
TF2910/4 non-linear distortion (video) analyser	£1000
TF2914A TV insertion signal analyser	£1250
TF2910 TV interval timer	£500

TEST EQUIPMENT

ANRITSU ME518A pcm portable error rate test set	£2500
BRUEL & KJAER 2511 vibration meter set/1621 filter	£2250
BRUEL & KJAER 2307 level recorder	£1000
BRUEL & KJAER 2317 portable level recorder	£1850
BRUEL & KJAER 1618 band pass filter	£750
BRUEL & KJAER 3513 portable vibration analyser	£3500
BRUEL & KJAER 2515 vibration analyser	£5000
AVO RM215L-2 insulation & breakdown tester	£650
DATRON 1065 digital multimeter	£750
DRANETZ 626 mains disturbance analyser/2 x PA-6001	£750
DRANETZ 606-3 line disturbance analyser	£275
FLUKE 1720A TEEE-instrument controller	£500

SPECIAL THIS MONTH ONLY: HP8673C

synthesised signal generators covering from 50MHz-18.6GHz. AM/FM/PM. Full HPIB control. Supplied in 'as new' condition with current HP calibration. List is over £59,000. We have two to sell this month only, for £9,950 each.

RACAL 9082 synthesized AM/FM sig' gen' 5-520MHz	£600
RACAL 9300 RMS voltmeter – 80dB to +50db	£325
RACAL V-STORE 16, 16-chan instrumentation recorder	£9000
RACAL-DANA 9302 RF milli-voltmeter 1.5GHz	£450
RACAL-DANA 9303 level meter, digital	£1000
ROBERTS & ARMSTRONG t/o-cable end-cut measure unit	£500
TEKTRONIX J16 digital photometer	£250
TEKTRONIX 1503C/03/04/05/06 TDR cable tester	£3250
WAYNE KERR 3245 precision inductance analyser	£3000
WAYNE KERR B905 automatic precision bridge	£950

HEWLETT PACKARD

331A distortion meter	£200
339A distortion meter	£1500
3400A voltmeter 10Hz-10MHz	£250
3335A synthesizer/level generator with option 01	£2500
3336A level generator	£2000
3552A transmission test set	£1000
3586A selective level meter	£2500
415E swr meter	£350
4274A multi-frequency (100Hz-100kHz) LCR component meter	£4000
4275A multi-frequency LCR component meter	£5000
432A/478A microwave power meter 10MHz-10GHz	£400
432A/R486A uwave power meter 26.5-40GHz (waveguide)	£600
5370B universal time-interval counter	£2500
6253A dual power supply 0-20V 0-3A twice	£225
6825A bipolar power supply/amp –20 to +20vdc 0-1A	£350
70300A tracking generator plug-in unit	£2000
70907A external mixer for 70000-ser spectrum analyser	£1750
7035B X-Y single pen analogue chart recorder	£350
8011A pulse generator 0.1Hz-20MHz	£500
8112A pulse generator	£2500
816A slotted line 1.8-18GHz with carriage 809C & 447B	£500
8350B sweep generator main-frame	£2250
8405A vector voltmeter, voltage & phase to 1000MHz	£950
8620C sweep generator, many plug-in units available	£call
8671A synthesized signal generator 2-6.2GHz	£2500

URGENTLY REQUIRED – HIGHEST CASH PRICES PAID FOR 8566B, 8753A/B/C, Spectrum/network analysers, 8590/1/2/3 portable analysers. Please call us if you have high-end capital equipment being under-utilised.

PLEASE NOTE: ALL OUR EQUIPMENT IS NOW OPERATION-VERIFICATION TESTED BEFORE DESPATCH BY INDEPENDENT LABORATORY

We would be pleased to handle all grades of calibration or NAMAS certification by same laboratory at cost price. All items covered by our 90-day parts and labour guarantee and 7-day 'Right to Refuse' (money back) warranty.

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ELECTRONIC UPDATE

Contact Malcolm Wells on 081-652 3620

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The LabWindows/CVI Brochure, from National Instruments, explains how users can build instrumentation applications on Windows PCs and Sun SPARCstations using the ANSI C programming language and LabWindows/CVI data acquisition, analysis, and presentation libraries.

NATIONAL INSTRUMENTS
For further information
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