Reader competition see page 24

ELECTRONICS WORLD

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OCTOBER 2003 £3.25

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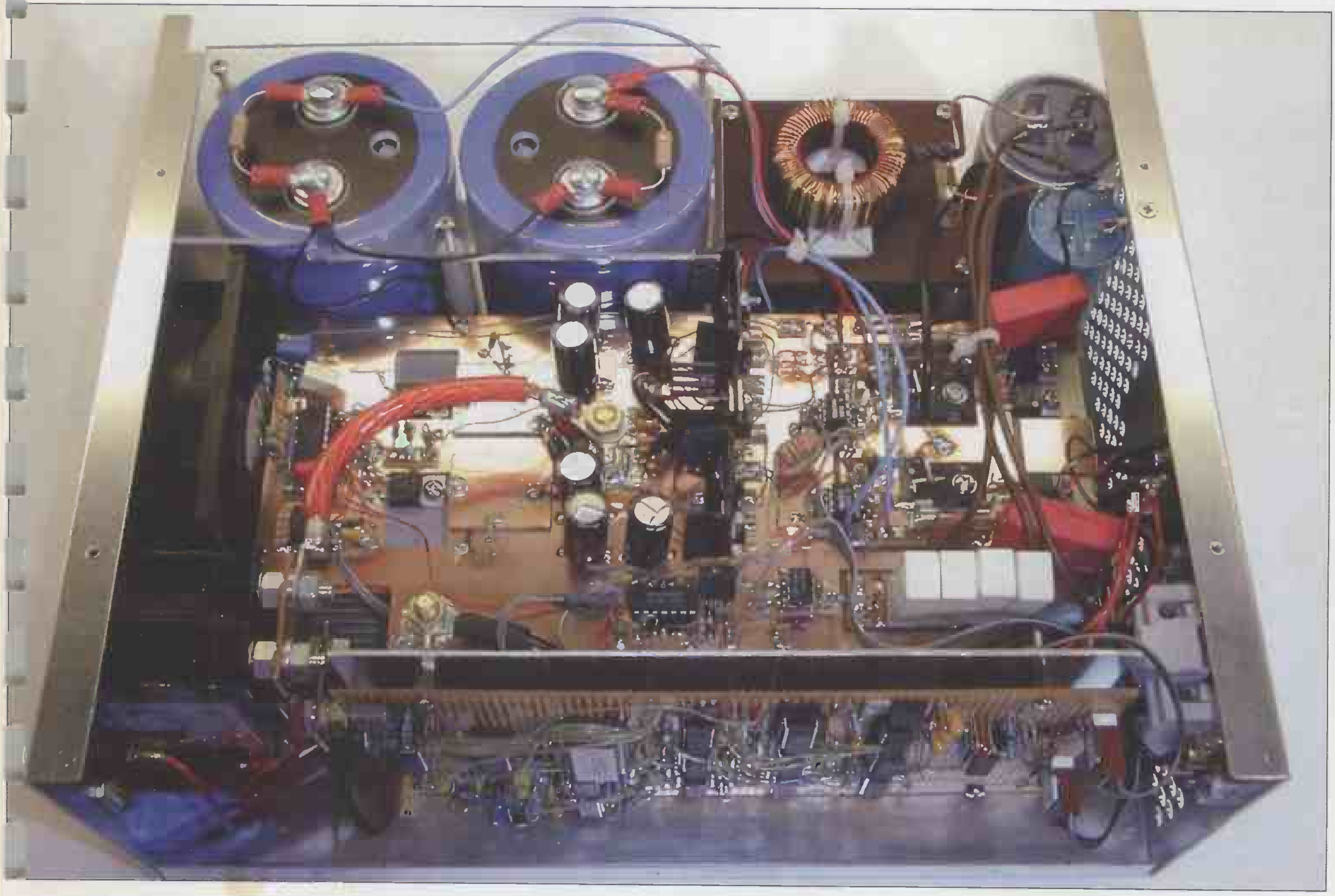

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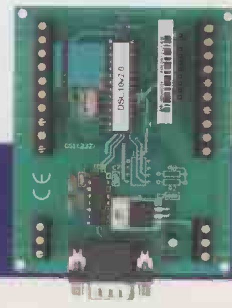

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MISCELLANEOUS

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Tektronix 1735 Waveform Monitor	£1100
Tektronix AM503 - AM503A - AM503B Current Amp's with M/F and probe	from £800
Wayne Kerr 3245 - Precision Inductance Analyser	£1750
Bias unit 3220 and 3225L Cal.Coil available if required.	(P.O.A)
Wayne Kerr 3260A + 3265A Precision Magnetics Analyser with Bias Unit	£5500
W&G PCM-4 PCM Channel measuring set	£3750

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On the turn

I keep getting reports lately that the industry is on the up. Whilst there is bound to be some PR 'hype' – reports are suggesting that recruitment is returning and capital spend in some quarters is definitely up. In a recent report commissioned by the IFC and produced by Booz Allen Hamilton, they reckon 16% of global electronics manufacturing, currently running at $65Bn annually – is performed in emerging countries. More than half of the current emerging market production and 8% of total global production is already in China and $46Bn or 77% of all emerging markets new production growth through 2005 will be in China, causing China to pass Europe as a source of electronics manufacturing. Food for thought.

I'm definitely getting more press releases about new products – that are genuinely new – and just some half-baked 'improvement'. One of which is the subject of our reader competition, courtesy of Observant Electronics. The product description will undoubtedly speak for itself – so good luck, but if you are not so lucky, we'll be running a reader discount offer in subsequent issues.

I've had few negative comments about Nigel Cook's leader in August. Whilst the piece may not have been quite up to the usual 'standard' (although I thought it was perfectly OK) I feel that some of your comments were extremely rude and not befitting of professionals. Rather than create an unholy slanging match, there will be no correspondence published on this subject. If however, you'd like to contribute a leader yourself – I am always happy to listen to any new ideas.

Housekeeping

Yet again, I have to report that EW has moved offices. I suppose it's a sign of the times that publishing is in a mess like the rest of industry and we are 'consolidating' to share resources with other similar magazines in the group. So, we've moved to sunny Swanley in Kent and because the extra journey would have been too much – we've lost Jackie Lowe as well. While we look for a replacement her mail addresses remain active – but I'm doing her job as well – do not hold your breath for a reply – I'm going as fast I can! Thanks to Jackie for all the hard work she's put in over the years both with us (Highbury) and previously at Reed Publications.

I have also decided to make some new email boxes that will enable us to deal with you more efficiently. These new addresses are sprinkled around the magazine – but basically if you need me, use EWeditor, if you have an admin problem like we owe you money or an issue – use EWadmin if you have a letter for publication – use EWletters and finally if you've got a circuit idea – please use EWcircuit. All addresses are @highburybiz.com

There will undoubtedly be hiccoughs while mail of all types gets re-directed (and possibly lost) so please bear with us for a few months as we sort ourselves out.

Phil Reed

New editorial and advertising address

The Highbury Business Communications office previously at Cheam, Surrey has moved to Swanley in Kent. All correspondence intended for the editorial and advertising departments should be addressed to:

Electronics World,
Highbury Business Communications,
Nexus House,
Azalea Drive,
Swanley,
Kent, BR8 8HU

The switchboard phone no. is
01322 660 070
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Fax 01322 616 339

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Fuel cell laptops in 2004

NEC has unveiled a prototype notebook PC with an internal fuel cell.

The fuel cell boasts the world's best output density, claims NEC, of 40mW/cm, achieves an average output of 14W and a maximum output of 24W.

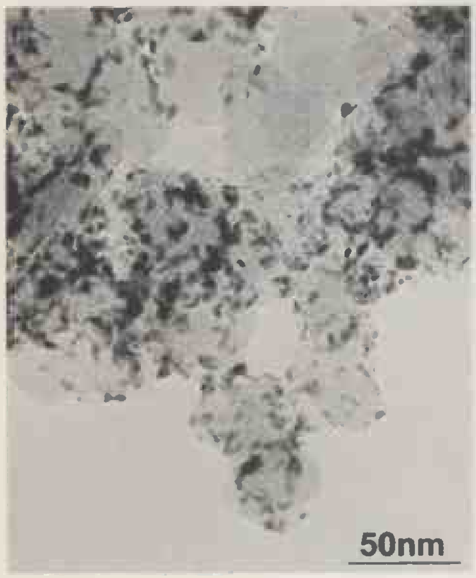
The firm aims to release a notebook PC with a built-in fuel cell on the market by end of 2004, and to make a notebook PC equipped with an internal fuel cell that has a battery life of 40 hours available within two years.

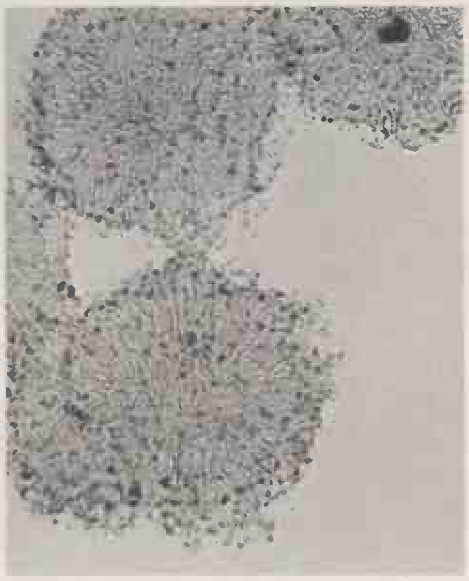
The cell employs carbon nano tubes, in the form of what NEC calls carbon-nano-horns.

"We use the carbon-nano-horn as an electrode covered by a catalyst using white gold. If carbon-nano-horn is used, the catalyst can be attached on the surface of an electrode evenly and in very small particles," said NEC. "If acetylene black is used in the electrode as usual, the catalyst cannot be attached evenly in very small particles."

How the catalyst is attached is the impact on the power of the battery.

"Because of the evenly attached catalyst and the small particles, our battery can generate bigger power with lower cost than other ways," said NEC





The little black specs are white gold catalysts, small and evenly distributed (better) in one case, and clumpy and large in the other.

Fuel cell specification

Average output	14W
Maximum output	24W
Voltage	12V
Weight	900g
Fuel weight	300g (300cc)
Fuel concentration	10% methanol
PC weight incl. cell	2 kg
PC size	288 x 280 x 40mm
Operating time	5 hours aprox.

Academy recognises youth and experience

The Royal Academy of Engineering has elected its youngest ever fellows in Paul Westbury (33) and Professor Colin McInnes (35).

Westbury is a structural engineer at Buro Happold, and was project engineer for the Millennium Dome. McInnes is Professor of space systems engineering at the University of Glasgow with significant work on solar sails and autonomous guidance of space vehicles.

"Britain is privileged to be the home of some of the world's best engineers," said Sir Alec Broers, president of the RAE. "The Academy is working hard to ensure that these people are recognised and that their skills are utilised for the greater good of the country."

Both engineers beat the RAE's previous age record held by Professor Polina Bayvel, of the department of electrical and electronic engineering at University College London, who was 36 on election.

The RAE also elected Gordon Moore, chairman emeritus and co-founder of Intel, as a foreign member.

"Not only have Gordon's pioneering technological efforts revolutionised the design of virtually all modern products, but his business leadership and his philanthropic activities have had an enormously positive impact on society," said Broers.

Keyboard gets virtual

A chipset and optical components that project the image of a keyboard and then tracks a user's fingers has been developed by US firm Canesta.

The technology could be used with mobile phone and other handheld devices in lieu of a physical keyboard.

Three devices make up the system: A scanning projector to create the 'keyboard'; an infra-red LED and optics to create a beam; and a sensor which locates fingers in space, assessing when they touch the virtual keys.

By using IR light the system can work in any ambient light conditions. A wide angle lens in the projector allows the keyboard to be imaged from relatively low angles.

Other applications for the system include intelligent airbags that sense the size and position of a car's occupants, security systems that detect the difference between an intruder and pets, and virtual reality games with no mechanical input.

Silicon circuits top 100GHz

Researchers at Infineon Technologies in Germany have constructed silicon germanium (SiGe) circuits that run at up to 110GHz.

The firm made three devices; a 110GHz dynamic frequency divider, an 86GHz static frequency divider and a 95GHz voltage controlled oscillator.

The dynamic divider has a ratio of two, runs from a 5V supply and draws 62mA. The static design divides by 32, and consumes 180mA at 5V. The VCO circuit operates between 95 and 98GHz with a phase noise of –97dBc/Hz. Output power is –6dBm, said Infineon. At 5V it draws 12mA.

A 77GHz radar transceiver is also possible, said the firm. In the long term the firm said it could manufacture devices for 40Gbit/s wireline telecoms, microwave radio links, ultra wideband wireless up to 60GHz and automotive radar.

The basis for the circuits was Infineon's SiGe:C process, which produces transistors with cutoff frequencies of over 200GHz.

IBM has also done significant work in this area. It has produced SiGe transistors with transition frequencies of 350GHz, although it is yet to release details of circuits made with the devices.

In Brief

Baby tags use wideband RF

An RF tagging system that uses ultra wideband (UWB) technology to track documents and patients around hospitals in real time is being developed by two Cambridge companies.

Ubisense, which is developing the local positioning system (LPS) using UWB expertise from technology house Cambridge Consultants, said lost patient records are a major problem. Knowing immediately where records are means fewer patients arriving for surgery without their notes, fewer delays in treatment, and greater patient safety.

Precise applications are still being defined, but Ubisense said tracking the location of babies would be a good trial subject.

Operating at 5.8GHz, a network of sensors distributed throughout a building can locate a credit card-sized tag to within 15cm in real time - an order of magnitude better than that available from wireless LAN or Bluetooth-based systems. Depending on the composition of the hospital's walls, a network should need around 25 sensors for every 1000m^2.

"We're just getting these ready within the NHS," explained David Theriault, CEO of Ubisense. "It just literally has not been done before... We need details about the demand for applications and the cost."

The regulatory position over UWB is still unclear in the UK, although developers' licences are being granted by the Radiocommunications Agency. The FCC in the US is leading the standardisation process for consumer systems.

Broadband meets expectations

A Website devoted to all things broadband has found that ADSL services can reach almost 90 per cent of their rated speed in downloads.

ADSLguide has tested all the UK's 512kbit/s ADSL operators, and found that the fastest, Eclipse Internet, reaches a download speed of 457.2kbit/s and an upload rate of 240.8kbit/s.

All the top ten services, which includes names such as Pipex, BT and Demon, exceed 430kbit/s in download and 225kbit/s in upload.

www.adslguide.org.uk

Amish lead in headlights

The Amish of Pennsylvania, members of a religious sect known for rejecting modern conveniences including cars, are adopting LED headlights for their horse-drawn buggies to avoid being hit by cars at night.

"With conventional lighting systems, buggy owners have to drag a 60 or 70 pound (30kg) battery out of a carriage and lug it over to a charging station after roughly six hours of operation, which typically translates to every two weeks," said Elam Beiler, president of Sunline Solar, a part-Amish-owned alternative energy firm in Pennsylvania. "With LEDs, the buggy battery stays charged for as much as 100 hours of run time, so recharging is required only every six months."

In addition, Beiler predicts that widespread adoption of LEDs in Amish buggies will make it possible to reduce weight of carriage batteries by half in coming years.

Sunline uses a single Lumileds Luxeon red LED in its tail light and eight whites in its headlamp.

"There is some irony in the fact that a culture that resists most of the ingredients of a modern lifestyle is embracing a Silicon Valley technology," said Fran Douros, marketing manager at Lumileds. "In some sense the Amish are even ahead of the technology curve because LED-powered headlamps are available in Amish buggies before they are available in cars."

Amish charging stations are solar powered as they do not use mains electricity.
www.sunlinesolar.com

Insurance costs to soar

The bill for employers' liability insurance could go up by £150m under plans to recover NHS costs in all compensated workplace accidents.

The warning comes from the Engineering Employers' Federation EEF, which added that UK manufacturing will be hit badly by the Health and Social Care Bill.

"The Department of Health expects the employers' liability insurance system to foot the bill but, as this is a compulsory insurance, the costs will simply be spread among all employers in the form of rising premiums," said the EEF's southern regional group.

Moreover, the proposals would unfairly penalise companies with the best safety records, pointed out David Seall, chief executive of EEF South: "Employers who take positive action to improve the health and safety of their employees will subsidise those who do not or who are negligent."

Researchers pave way to completely secure satellite communications

Austrian researchers have used free-space links to successfully transmit a pair of quantum entangled photons to two independent receivers. The work could lead eventually to completely secure satellite-based communciations.

Entangled particles exist in related states that remain undetermined until a measurement is made. Once a measurement is performed on one, the other instantaneously 'knows' what state it is in. In cryptography, the effect allows a key to be generated with complete security between receivers sharing the entanglement.

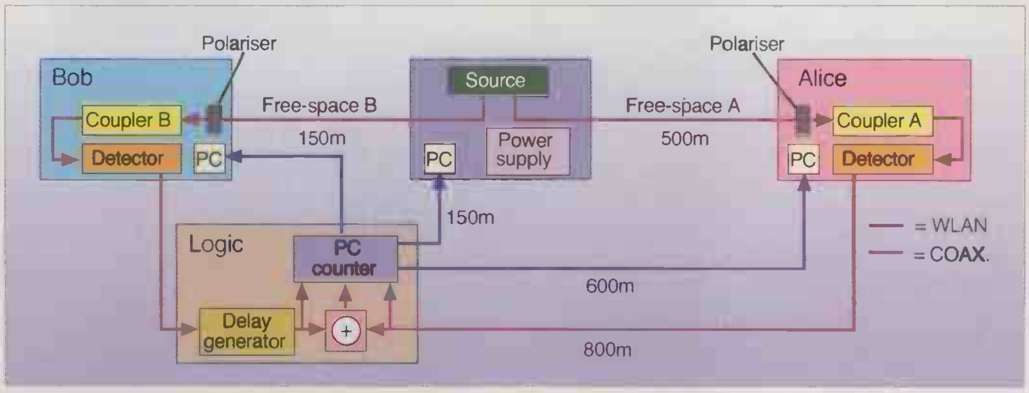
However, entanglement can break down due to interactions with the environment in a process known as decoherence. The scientists, from the Institute for Experimental Physics in Vienna, used free-space links to probe the distances over which decoherence occurs.

Their experiment used receivers spaced 150m and 500m respectively from the entangled photon source, with no clear line of sight between them - and the Danube river in between. The links had an attenuation of around 12dB, which the researchers said is similar to the performance currently available from satellite links, making a satellite-based secure communications system a feasible aim. The research was published in the August 1 issue of the journal Science.

Brian Lowans, team leader in the Single Photon Optical Techniques group at QinetiQ in Malvern, said: "This is a significant achievement and is currently a world record for this technique. It is important because it is currently unknown over what distance entanglement can be maintained before decoherence affects the exchange."

The instantaneous transfer of information between entangled particles when a measurement is made is the so-called 'spooky action at a distance' identified by Einstein, Poldalsky and Rosen in their famous 1935 thought experiment.

QinetiQ itself, with Munich's Ludwig Maximillians University, set a world record of 23.4km for free-space quantum key distribution between two points in the Alps last year. Its system used attenuated laser pulses to achieve single photon exchange.



Antenova, Orange and Queen Mary University team up to increase spectrum efficiency

Cambridge-based antenna company Antenova, with Queen Mary University of London, has won a contract from the Radiocommunications Agency to look into MIMO (multi-input multi-output) antenna arrays for laptops, PDAs and phones.

"This forms part of a wider investigation commissioned by the Radiocommunications Agency into optimising spectrum efficiency to enable future expansion of wireless network systems by using MIMO technology with multiple antennas at the handset and basestations," said Antenova.

Rather than just pick the biggest signal from several antennas, MIMO techniques make use of normally-disruptive multi-path effects and combine the outputs of several receive channels is a constructive way using digital signal processing.

Coding such as variable time delays at multiple transmit antennas assist in signal recovery.

Antenova's specialist is its high dielectric antenna (HDA) technology which makes antennas compact, and less likely to interact than conventional antennas. Queen Mary will be providing a complementary technology known as 'photonic band-gap surfaces', which can enhance isolation between antenna ground planes. Queen Mary will also investigate alternative antenna technologies which could provide both spatial and polarisation diversities for MIMO technology.

Phone company Orange is also involved, providing technical and commercial expertise from an operator's standpoint.

www.antenova.com
www.orange.com
www.qmul.ac.uk
www.radio.gov.uk

Micromachined switch sits on CMOS

STMicroelectronics has released details of a technique to integrate RF switches into circuits fabricated using standard CMOS technology.

"The switch promises to enhance the performance of mobile phones and similar portable terminals where efficient RF switching is required to minimise power consumption," said the company.

The 'Above IC' switch, as it is called, has been developed jointly by ST and long-term research partner CEA-LETI at ST's Crolles1 facility in France. "It meets the four key criteria of high reliability, low power consumption, low actuation voltage and compatibility with SoC fabrication techniques," said ST.

The moving element in the switch consists of a 400 x 50μm beam made from silicon nitride and aluminium, and includes titanium nitride heating resistors and electrostatic holding electrodes. It is fixed at both ends.

Heating with 20mA at 2V bends the beam centre downwards through bi-metallic distortion to meet a gold contact 3 μm below.

Actuation takes around 200μs, an activation energy of 8μJ.

Once the switch has been turned ON, a voltage is applied to the holding electrodes, producing an electrostatic force that retains the beam in position, allowing the heating current to be turned off.

For the first prototypes, the voltage required to achieve electrostatic hold was 15V and with improved stress control for the beam material this is expected to be reduced to 10V.

More than 10^9 switching cycles have been demonstrated, claims ST, without any failure or contact degradation.

Insertion loss is 0.18dB and isolation is 57dB at 2GHz.

ST and CEA-LETI are now working to optimise the electrostatic hold, develop wafer-level packaging and to cut cost by reducing the number of masks required.

Power firm back in business

Custom power supply manufacturer VxI Power has been setup following a management buyout of Bulgin Power Source.

Bulgin Power Source called in the receivers last year, but has now been revived by former managers Grant Ashley and Tim McCann and 14 of the 70 employees made redundant.

"We are hoping to take on more, but it depends on how business grows," said Ashley, a director at VxI.

The firm is aiming its products at the fire protection, security and utilities markets. Typical products include 12/24V AC-DC converters up to several kilowatts.

VxI will also perform small scale contract manufacturing at its Lincoln plant.

UK start-up improves infra-red comms

Optical Antenna Solutions (OAS) of Coventry has produced its first technology demonstrator, an optical device which replaces the lens in optical TO18 packages to improve photodiode performance.

Optical antennas - see the cross section - capture light like normal lenses for rays near the axis. However off-axis rays, which would be wasted by a normal lens, are also directed to the output of an optical antenna by a single internal reflection. This means gain holds up further from the centre axis - out to the critical angle where a single bounce no-longer works.

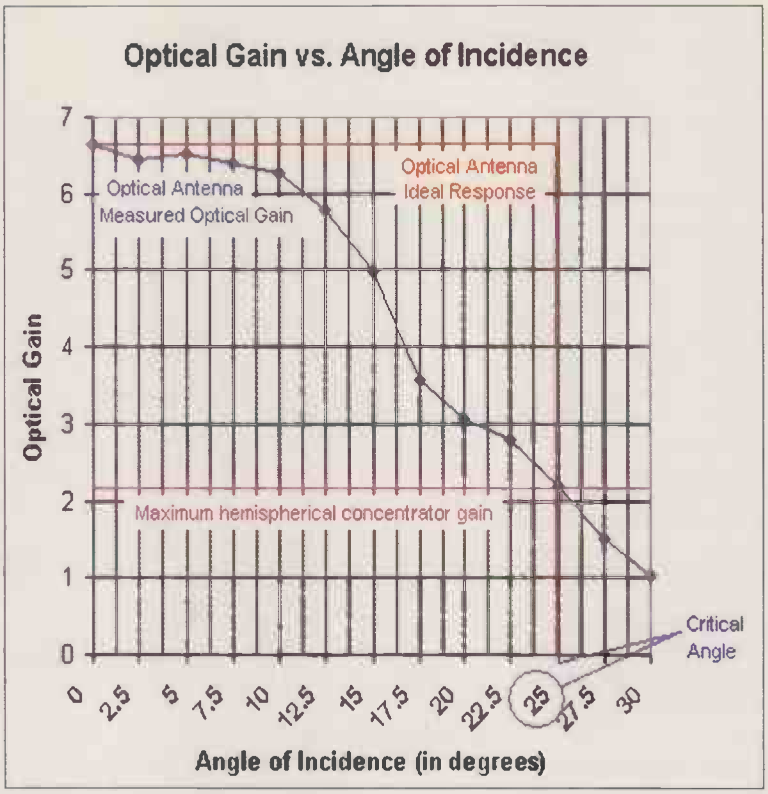
The type of optical antenna the company is developing is more accurately described as a dielectric totally internally reflecting concentrator, or DTIRC, which was developed at Warwick University and which spun-out OAS.

OAS' partner for the technology demonstrator is German firm Silicon Sensors.

The demonstration DTIRC has a field of view of ±25 degree - see graph - and was swapped into a PD2-6 from Silicon Sensors.

The graph shows the DTIRC performance as well as the maximum performance available from a conventional lens with the same field of view.

Other versions have been produced for low light levels, using Silicon Sensor's series-6 chips, and for high bandwidth applications using high-speed epitaxial series-5 chips.



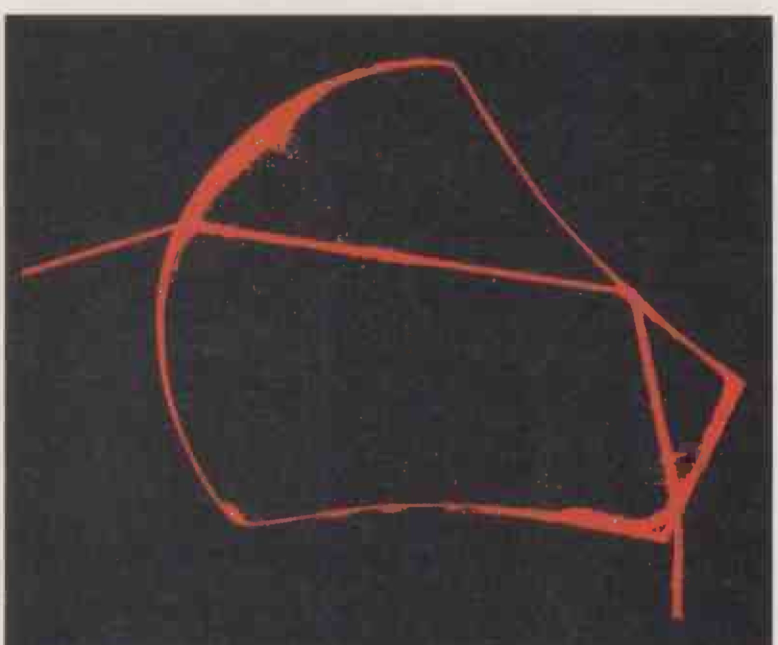
OAS is also working with members of the infra-red communications industry body IrDA to develop improved infra-red devices.

"We expect that opto-electronic components including IrDA devices will go through a period of enhancement to meet the demands and challenges associated with achieving more bandwidth and better signal to noise," said OAS marketing director Alex Clarke.

As DTIRCs can be designed to squeeze light down to a small aperture, they can be designed to reduce the size of photodiode chip needed for a certain field of view. Smaller chips mean less capacitance, and therefore faster operation.

www.opticalantennasolutions.com

University of Warwick postgraduate Roberto Ramirez-Iniguez has earned his PhD for his work developing dielectric totally internally reflecting concentrators, the technology behind Optical Antenna Solutions (OAS).
He previously completed a master's degree in Communications and Real-Time Electronic Systems at Bradford University and started his PhD work there under the supervision of Professor Roger Green.
Green moved to Warwick University to create the Communications and Signal Processing Research Group and Ramirez-Iniguez followed him. Together they founded OAS.



For on-axis light, the antenna acts like a lens. Off-axis light is internally reflected once, and still makes it to the output (right) - where a silicon sensor would collect it.

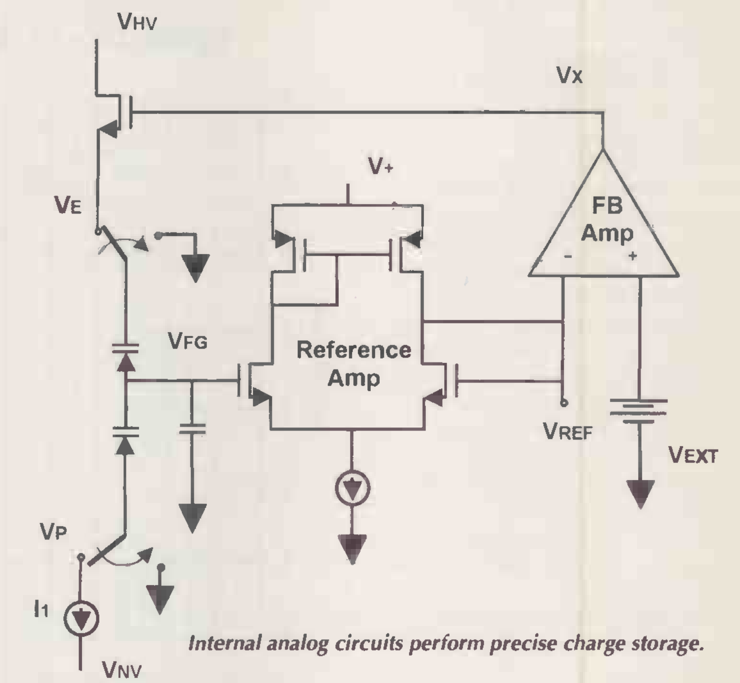
Floating gate puts precision in CMOS voltage reference

Xicor has introduced a floating gate-based precision voltage reference.

Floating gates, mosfet gates that are isolated on all sides, are used in flash and eeprom memory, but are seldom used in analogue applications.

The firm has used a floating gate as a means to fine-tune a CMOS voltage reference, whose output voltage is normally dependant on doping levels and is, as such, untuneable.

The voltage reference is the middle four transistors of the diagram, ladled 'Reference Amp'. Its output is the sum of the threshold voltages of the two lower fets - with the left-hand fet being a depletion mode device which self-biases to have its threshold voltage on its source in this configuration.



Internal analog circuits perform precise charge storage.

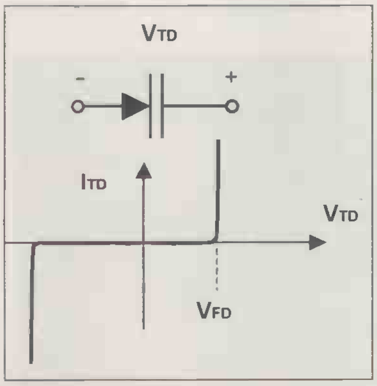
Table 1. Floating-gate voltage reference measured data

Initial accuracy	<0.2mV
Low TC	3ppm/ °C
Low drift	10ppm/1000hr
Low power	500nA at 5V

Charging the floating gate through its isolating oxide - represented by the gate capacitor and two tunnel diodes respectively - sets the threshold voltage and would be a hit-and-miss affair if it wasn't for the rest of the circuit shown.

In it, a current sink draws electrons through the oxide from the bottom of the floating gate while a controllable voltage source feeds electrons through the oxide into the top.

Ideal tunnel diode characteristic.



With a fixed voltage source, there are always a constant number of electrons in the gate, even if some are entering and some are leaving at all times.

The op-amp alters the voltage source until sufficient electrons are on the gate to make the reference output voltage (V_{REF}) equal the calibration voltage (V_{EXT}).

Once this happens, the switches are thrown and the right number of electrons are locked into the floating gate - "without loss for between 10 to 100 years," said the company.

Calibration is actually in two stages, with V_{EXT} slightly adjusted to allow for circuit offsets in the second phase.

Adjusted accuracy is better than 0.2mV.

The right hand lower transistor actually includes a buffer amplifier to reduce the output impedance of the reference.

The on-chip calibration circuit is disabled permanently after use to save power, leaving the overall current consumption at an astonishingly low 500nA.

Temperature coefficient can be inherently good in CMOS voltage references, and Xicor claims to have pulled out all the stops with its latest design to get 5ppm/°C.

The first two devices are the X60008CIS8-50 (±0.5mV) and X60008DIS8-50 (±1.0mV).
www.xicor.com

Xicor reference specification

5V output
±0.5mV initial accuracy
5ppm/°C temperature coefficient
10ppm/1,000Hr stability
$30\mu V_{p-p}$ noise voltage (0.1 to 10Hz)
±10mA output current
500nA supply current
–40 to +85°C operating range

Give me Sunshine

Internet connections running at 54Mbit/s and costing just £10 per month should start soon in the east of England with a service called Sunshine.

The wireless service is aimed primarily at business users, but the operator said it would welcome residential subscribers. It promises no connection fees.

By the end of the year, the service will cover areas within 10km of Bingham, Bottesford, Bourne, Elton, Scalford, South Witham, Stamford, and Uppingham.

The firm behind the high speed access is planning futher rollouts in Cambridgeshire, Essex, Leicestershire, Lincolnshire, Norfolk, North Lincolnshire, Rutland and Suffolk next year. By 2005 it hopes to have national coverage.

High performance 500W mains inverter

Do you think that commercially available power inverters are just not good enough? This design by Paul Bennett should bring many improvements over most commercial designs.

This project is similar to my design for a 400Hz inverter published in February 1999, but has a perhaps more useful output of 240V 50Hz. It is powered from 12V, and has been specifically designed to start and run a full size fridge.

Most refrigerators still use a traditional split phase induction motor, usually sealed inside the compressor casing. These motors have two windings, one for running and one for starting. Starting torque is provided by the starting winding having a different L/R ratio to the run winding and so generating the necessary phase shift. When the motor is up to speed the starting winding is disconnected. This is accomplished by a current relay on the run winding or PTC thermister on the starting winding.

When the motor is at rest there is no back emf in the windings due to no rotor rotation. Current is therefore only limited by the winding inductance and resistance. The starting current is therefore very high and significantly lagging. Any inverter therefore has to support this reactive current long enough to get the motor started. Once running the current is much reduced but remains lagging. At rest all the true power ends up as heat in the winding. Once running this divides mainly into mechanical power output and some remaining winding copper and iron loss. In-between there is a point of maximum power transfer, for a brief moment, as the motor accelerates through the point of maximum torque.

The starting and running conditions can be measured, to gauge what inverter rating is required. A low AC voltage can be applied to the motor, low enough so that the motor doesn't turn. The current can then be measured to give the stalled impedance. When running the inductance can be tuned out to give the lowest current using capacitors. Alternatively many DVMs these days can measure real and apparent power.

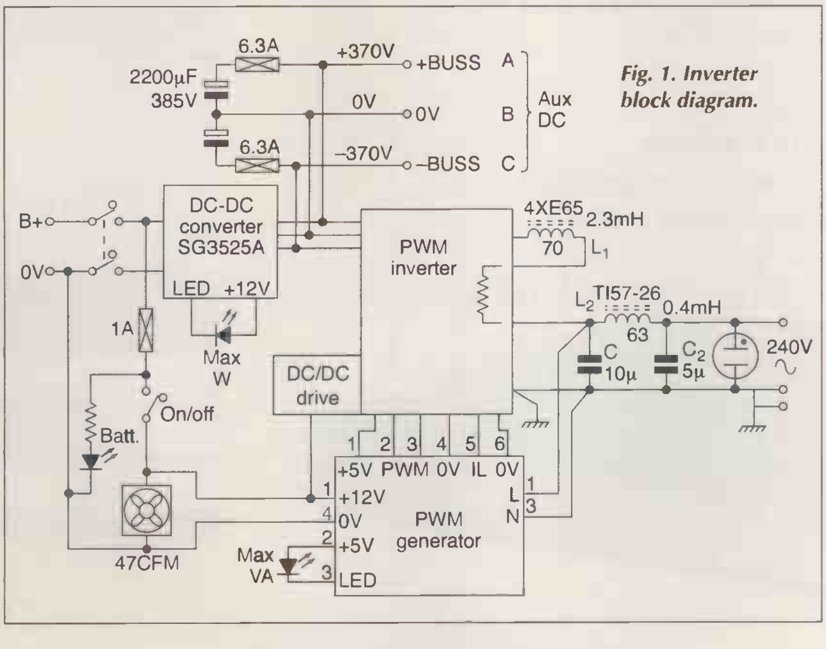


Fig. 1. Inverter block diagram.

Design choice

As we can see from the box on page 15, the typical 120W device requires 1500VA to start. With such a high VA to support during starting, a single stage design was ruled out. A simple 50Hz transformer square wave design would have to pass all the VA between the load and the battery/smoothing capacitors. This would lead to high copper losses unless very heavily engineered. Induction motors have reduced efficiency on square waves as harmonic currents produce torque at harmonic speeds which the rotor obviously cannot follow. PWM sine wave transformer designs are widely used in budget UPS addressing the harmonic problem but not the reactive current.

A two-stage design has been adopted with a DC-to-DC converter followed by a PWM half bridge.

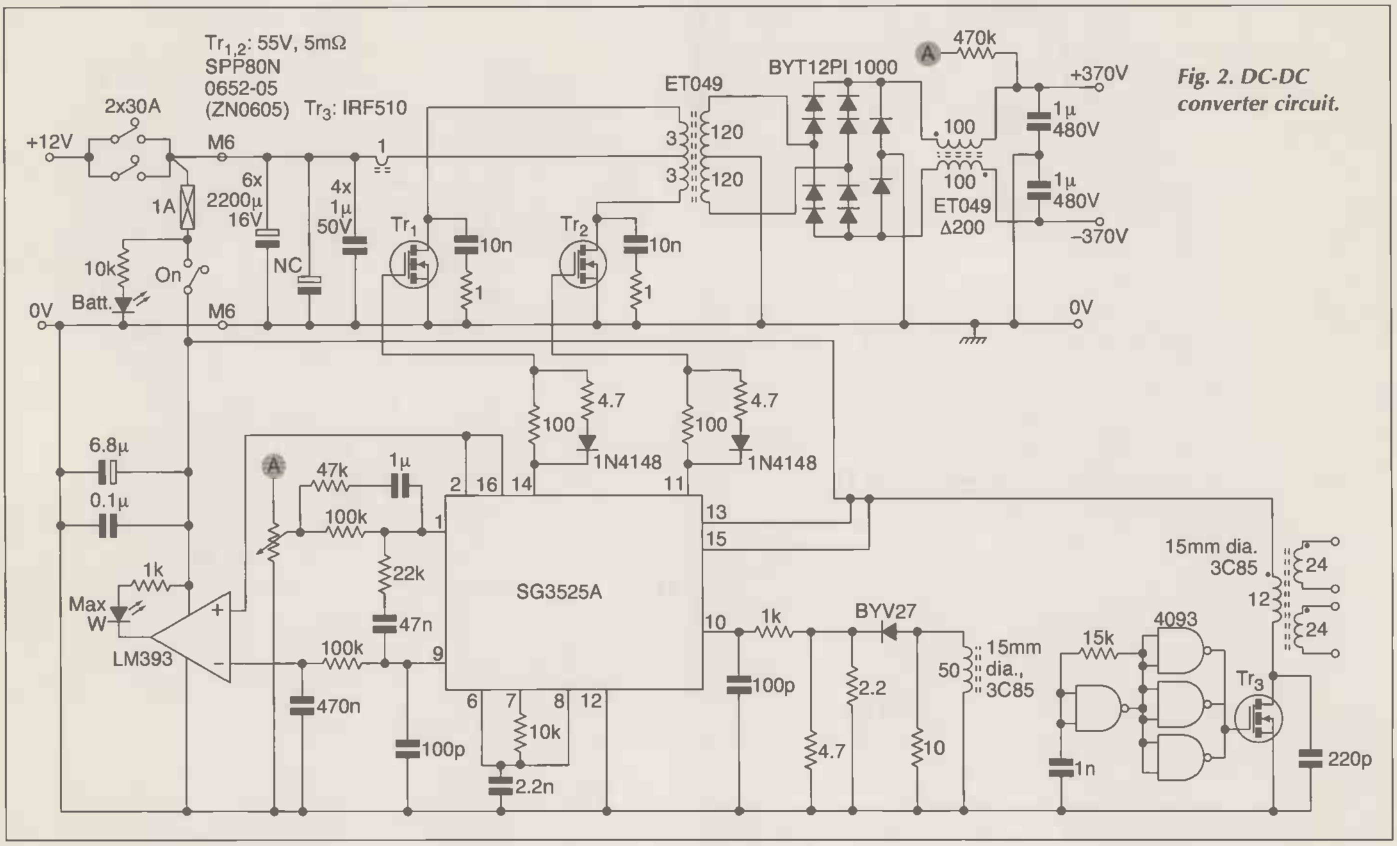


Fig. 2. DC-DC converter circuit.

Large electrolytics, shared between the DC converter output and PWM inverter input store energy to supplement peak demand. Only the PWM section needs to match the fridge peak VA. The DC converter can be scaled to more manageable levels for continuous demand. Generally electrolytic capacitors are more volume efficient at high voltages and avoiding very low impedance levels helps with circuit layout. **Fig. 1** shows the block diagram.

Topology choice

The DC - DC converter is a traditional transformer coupled push pull design with a voltage mode pwm controller. A split secondary and differential choke gives two symmetrical outputs, +/-370V w.r.t. ground. The cross-coupled choke ensures good regulation of the negative rail even though only the positive is sensed. Current limiting is provided by detecting the centre

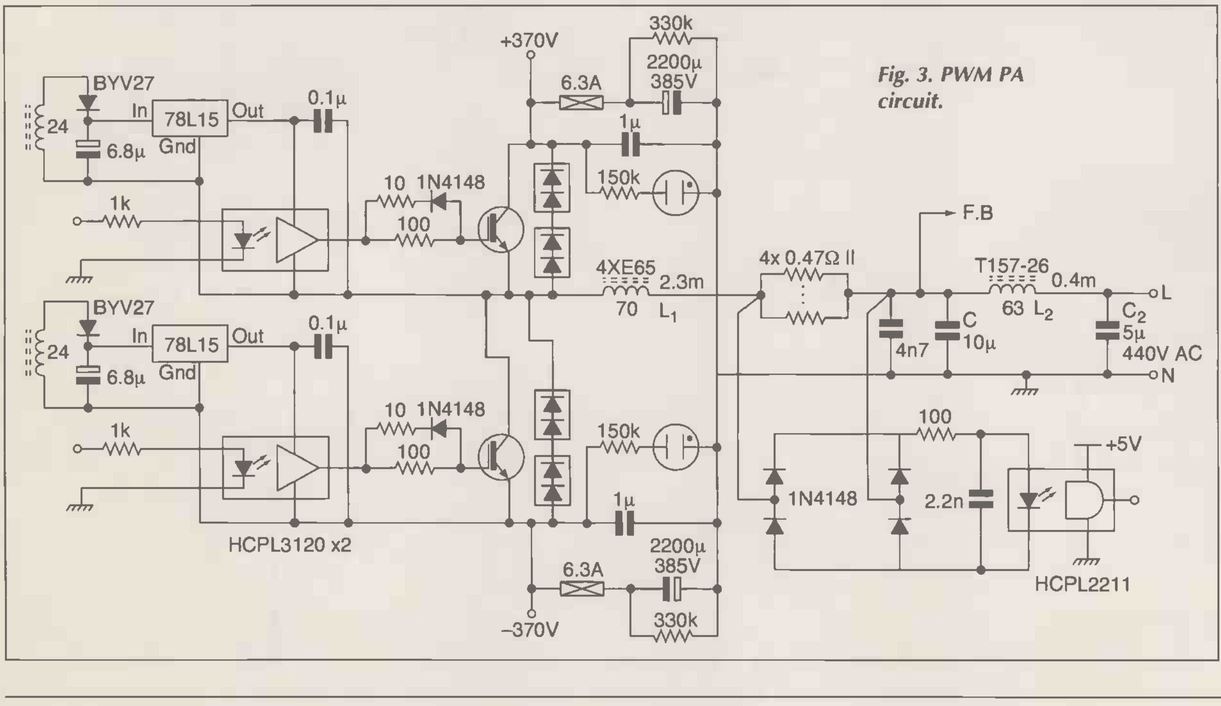


Fig. 3. PWM PA circuit.

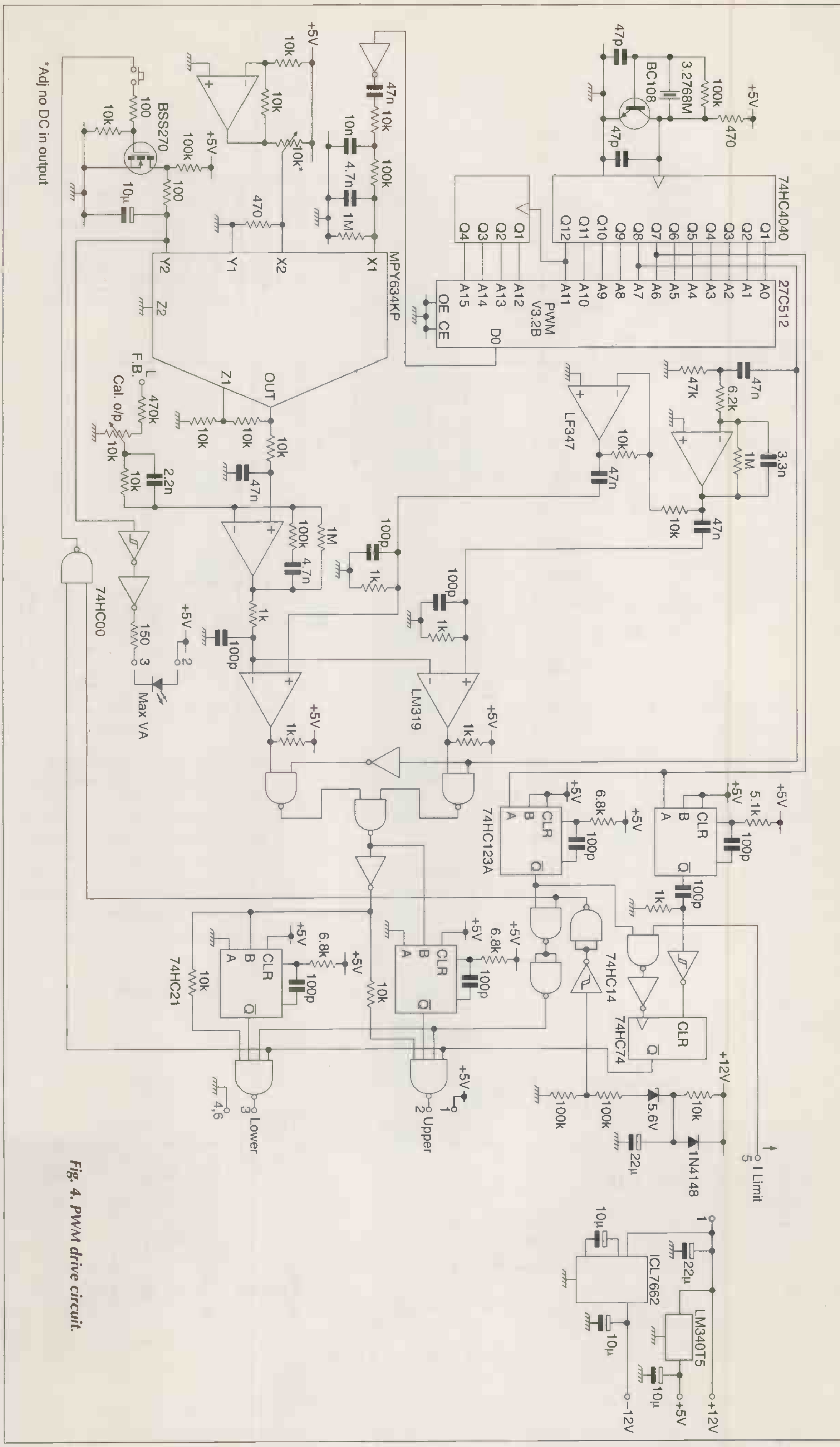


Fig. 4. PWM drive circuit.

tap current with a transformer and terminating the PWM pulse. The current limit effectively determines the true power (W) rating of the over all inverter. **Fig. 2** shows the full circuit.

The author experimented with the possibility of a phase shift resonant design. These offer various advantages such as low loss switching, placing the inductor in the primary circuit and low diode stress. The main disadvantage is high circulating current in the primary leading to poor efficiency at low load. It also requires a full bridge with high side switches.

A half bridge is used for the output. This has to support the full VA, i.e. deliver 6A rms or so without any restriction. The current limit for this therefore determines the overall inverter VA rating. Current is sensed by a fast optocoupler placed in between the main filter inductor L1 and smoothing capacitor C1. In this position, rather than directly at the transistors, avoids the reverse recovery spike. The current at this point is continuous between PWM cycles so a transformer would have to have a frequency response down to 50Hz.

The DC - DC converter control loop needs to be chosen with care. The output load draw has a 50Hz ripple which alternates between the positive and negative buses. The loop is chosen quite slow to average this out, to avoid peak currents in the DC - DC converter which would otherwise reduce efficiency slightly. The slightly increased ripple on the busses is easily coped with by the PWM output control loop with negligible increase in distortion.

Fig. 5. Completed unit external view.

Fig. 6. Completed unit internal view.

Component choice

Fuelled by automotive and portable IT needs, low voltage power MOSFETs have made great strides in recent years. The Infineon SPP80N06S2-05 used here provides 5mΩ at 55V in a To220 package. And only cost a few pounds. The IRF1405 from International Rectifier offers a similar performance.

The output bridge uses two 1200V IGBTs, IR4GPH40U from IR. These are types for switch mode psu applications, i.e. no specified short circuit withstand capability. Devices with this capability usually have inferior switching performance. The device's gain is designed to limit the short circuit current (i.e. come out

Typical measurement from a '120W' fridge freezer

Stalled Z = 39Ω, so stalled current at 240V = 6.2A. Therefore equivalent to 1500VA.
Cold resistance 13.5Ω. Most loss is copper I^2R = 520W. Running current 0.93A.
Approx 8μF could tune this down to a minimum of 0.47A, 113W.
Inverter power and VA is therefore dominated by the starting conditions, requiring approx 1500VA, 500W.

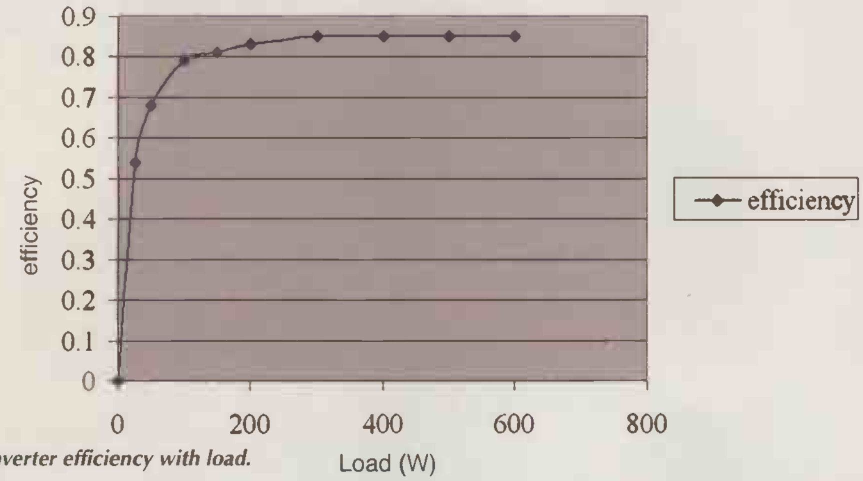


Fig. 7. Inverter efficiency with load.

of saturation), but still needs to be turned off within a few µS, to avoid the destruction by the very high instantaneous power dissipation. In this design current build up under short circuit conditions is limited by L1 to avoid desaturation, and long enough to turn the devices off.

Use of IGBTs allows the necessary fast recovery free wheeling diodes to be connected directly in parallel. Two 600V 8A STTi860 devices from ST are used. These incorporate two 300V die in series. Low voltage diodes have lower reverse recovery charge, and as they are in series, the same charge flows through both diodes. The penalty is higher forward voltage drop, although this also increases for single junction high voltage fast recovery diodes. The low charge reduces the IGBT turn on transient which reduces EMC problems and makes layout less critical.

Output filter

L1 is the main PWM filter choke. Its value needs to be high enough to reduce the 25.6KHz current ripple to an acceptable value. It should be low enough not to attenuate the 50Hz signal significantly and be an acceptable size and shape. It should not saturate under worse case current limit conditions. Four 3C85, E65/32/27 cores were used with a 70 turn winding of 2mm enamelled copper wire. Rubber spacers gave an 'air gap' of approx 2mm for a value of 2.3mH.

C1 is the main output smoothing capacitor. Its value needs to be high enough to pass the remaining 25.6KHz ripple current to ground, minimising the resultant voltage ripple. It also supports some 50Hz inductive reactive current from the load. Any reactive current through C1 not taken by the load circulates through the half bridge, increasing losses, so C1 should not be too large. 10µF is chosen here. Residual 25KHz ripple is approx 1V rms. This is further reduced by L2/C2. C2 also contributes to reactive current circulation. 0.4mH and 5µF are used. **Fig. 3.**

PWM generation

The 50Hz reference signal is itself PWM coded in an EPROM. This is then clocked out and filtered to a high quality sine wave. All of a 512K EPROM clocked at 1.6384MHz stores one 50Hz cycle. Only one bit is used. The other bits could store other output frequencies providing they are an integer multiple of 50Hz. Non-multiples could be stored using an additional bit to reset the EPROM address divider. A reset bit after a count of 436906 approximates to 60.000092Hz. Additional bits could also provide additional phases. The shape of the stored waveform could be changed to more accurately mimic the typically flattop distorted mains. **Fig. 4.**

An error amplifier compares the reference to a sample of the output waveform to generate the required PWM demand. A twin ramp and comparator arrangement is employed for the PWM generator. This fixes one of the PWM edges in time with respect to the master clock. This gives a defined dead time at the end of each PWM cycle to reset the current limit latch. Logic and monostables ensure there is always dead time between conduction of the bridge transistors. A current limit latch turns off both transistors in response to a current limit signal. This is reset during the dead time at the end of each PWM cycle.

Under current limit conditions significant distortion occurs as the threshold is a fixed value and doesn't follow a sine reference. Current limiting is therefore more likely at the peak of the 50Hz cycle. Under partial overload the waveform becomes flat topped. To reduce this distortion the pwm reference sine wave is reduced in response to repeated current limit by a multiplier gain stage.

Packaging

The prototype was built into a simple two-part aluminium case as shown in figures 5 & 6. A 12V 50cfm fan provides cooling. Indicators show battery connected, output on, max W and max VA. The battery indicator uses a high performance green LED. This draws just 1mA – comparable or less to the self discharge of a battery large enough to match the inverter's full output. Unfortunately these LEDs appear only available in clear packages with narrow viewing angle. An old diffused green LED was cut up and glued to the new one, the join being hidden by the chrome bezel.

Performance

The overall true power was checked using lamps. The output half bridge was also checked with lamps and a 1KW fan heater. A variac and transformer providing extra power. The graph shows the overall efficiency as a function of load. The acid test is the fridge, which started and ran fine. The thermostat cycle being a surprisingly long 1-_ hours. Inverter input started off at 143W but this fell to 133W just before the thermostat time out. ■

SAFETY WARNING

This project contains high DC voltages of relatively high energy. Exercise extreme care if recreating this project.

Flat, wideband buffers

Many textbooks cover the emitter-follower or common-collector transistor configuration as a buffer, but they usually don't mention a word about stability. In this article we will show that a buffer circuit can produce gain, and even oscillate under certain source and load conditions. We also look for ways of achieving a flat gain-frequency response. Dewald de Lange elucidates.

The idealised emitter-follower circuit of **Fig. 1.** driven by a source with zero impedance, and not reactively loaded, gives the theoretical transfer function in **Fig. 2.** It shows a flat response beyond 1GHz.

In practice, the source may have some resistance R_G, and stray inductance L_S due to a lead or PCB track; and the load, some capacitance C_L, due to subsequent circuitry – most circuits are capacitive over their normal operating frequency range. When these are added, as in **Fig. 3.** the simple buffer can deliver some unexpected surprises, such as gain or instability at higher frequencies, as a graph of the transfer function shows in **Fig. 4.**

The gain is caused by the transistor's internal base-emitter capacitance C_{be}, and the base transit time – a parameter used in more advanced transistor models, such as the Spice model used in many CAD (Computer Aided Design) software. Both cause a delay from the input to the output.

To understand what causes the gain peak, consider what happens when a step voltage is applied to the input in **Fig. 5.** Internal capacitance c_{be} must first be charged up for v_{be} to increase, and for i_e to increase. The difference between i_e and i_l charges up C_L. When v_{out} reaches the +0.1V target voltage, c_{be} is still charged higher, making v_{be} and i_e larger, and causing v_{out} to overshoot. It takes a finite time for c_{be} to discharge to the point where i_e equals i_l again, and v_{out} settles down to its final value.

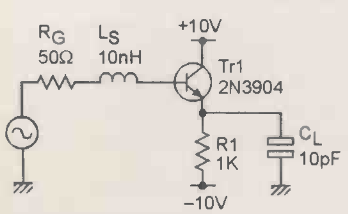


Fig. 3. Buffer with a practical source and load.

By re-arranging the components as in **Fig. 6.** it will be recognised as a Colpitts oscillator. Therefore, given the right source and load impedances, a buffer will oscillate. It is useful to remember this undesirable circuit arrangement when implementing a buffer.

Flattening the response

The purpose of a buffer is to provide a replica of its input signal. Therefore we need to find ways of flattening the gain. This can be achieved by slowing down the rate at which c_{be} is charged, i.e. limiting the bandwidth of the buffer.

For distortionless buffering of non-

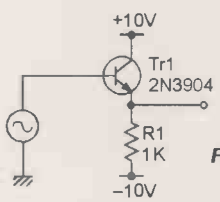


Fig. 1. Simple emitter follower.

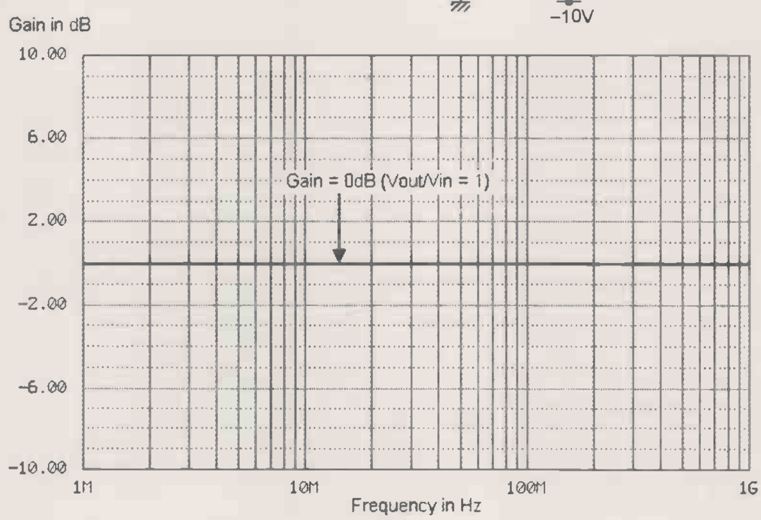


Fig. 2. Response of a simple emitter follower.

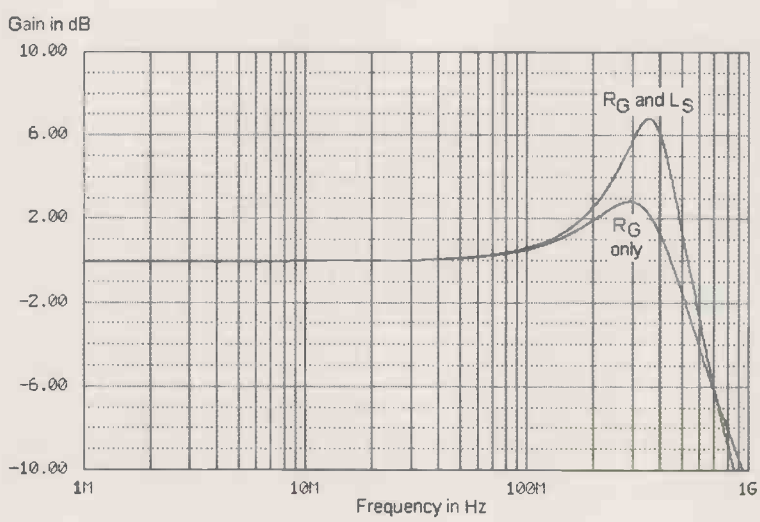


Fig. 4. Buffer response with a practical source and load.

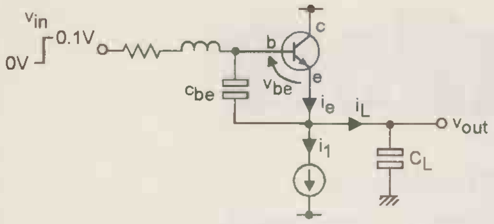


Fig. 5. Effect of internal base-emitter capacitance.

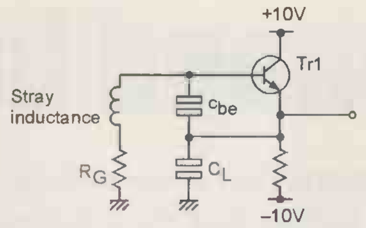


Fig. 6. A buffer can turn into an oscillator.

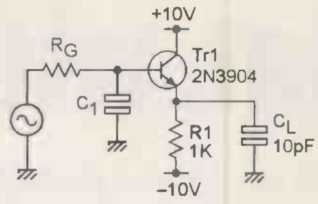


Fig. 7. Bandwidth limiting with RC filter.

Fig. 8. Response of circuit in Figure 7.

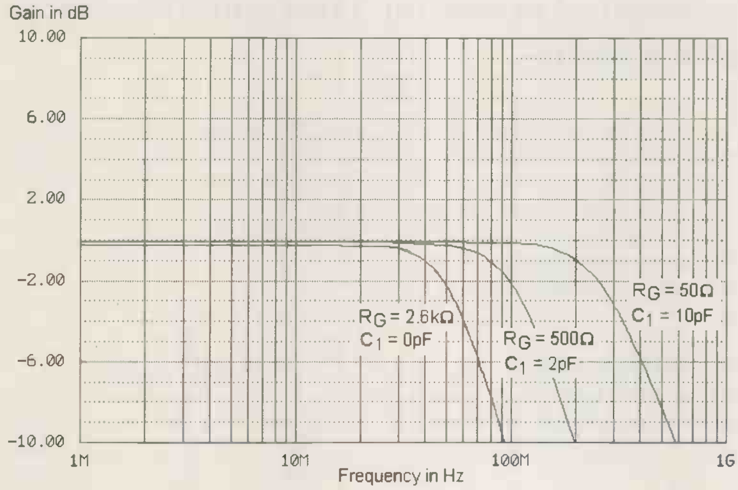


Fig. 9. Circuit in figure 7, with R_G = 500Ω and C1 = 2pF.

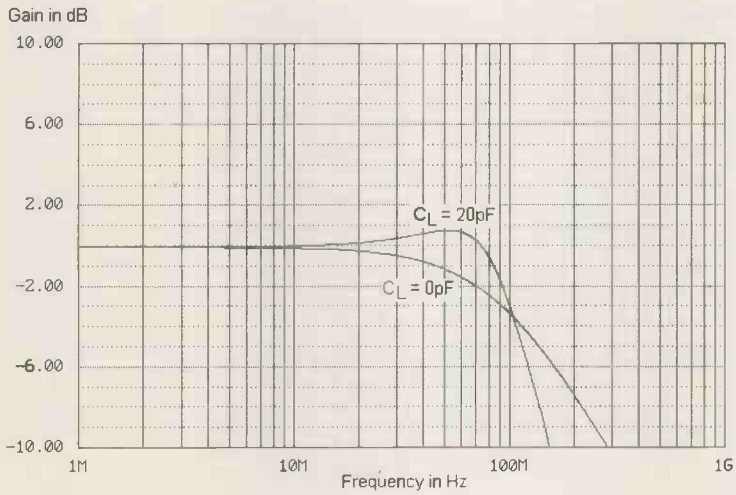
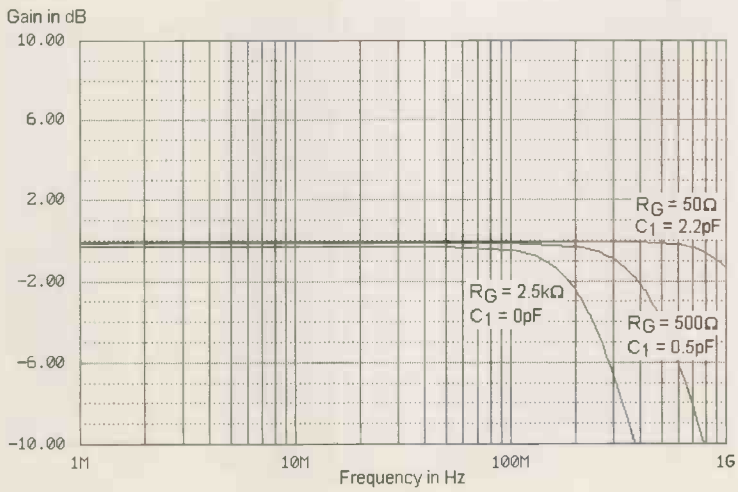


Fig. 10. Response of circuit in Figure 7, with transistor BFG67



sinusoidal waveforms, all frequencies should also be delayed the same – the buffer should have a flat group delay (the change in phase relative to the change in frequency). In this application, it is a small factor compared to the gain.

RC filter

One way of flattening the gain is with a resistor-capacitor low-pass filter on the input, as in **Fig. 7. Figure 8** shows the bandwidth for different combinations of R_G and C_1. These graphs are for a capacitive load C_L of 10pF. Stray inductance is left out for simplicity.

The larger the source resistance R_G, the less the capacitive load C_1 presented to the source, but the lower the bandwidth. A filter capacitor C_1 of 10pF would equal the load capacitance C_L, and may make the buffer redundant, except for partially isolating the source from variations in the load.

A multi-order LC (inductor-capacitor) filter may in principle give a higher cut-off frequency, but achieving a flat gain is more difficult.

Figure. 9 shows the effect of a change in load capacitance, away from the nominal 10pF. This is for R_G = 500Ω and C_1 = 2pF in Fig 7.

The 2N3904 has a current gain bandwidth product f_T of 300MHz, which would limit its performance above 100MHz. What bandwidth could we achieve if we used a microwave transistor such as the BFG67, with a transition frequency f_T of about 8GHz? **Fig. 10** shows the gain for different combinations of R_G and C_1.

RRC filter

The bandwidth can be extended further with a series R_1 and C_1 as in **Fig. 12**. Component values are more critical, and a trimmer capacitor will be required for a really flat response.

Smaller load resistance

If the load has a smaller resistance, the response will also be flatter, as shown in **Fig. 16.** A drawback to the

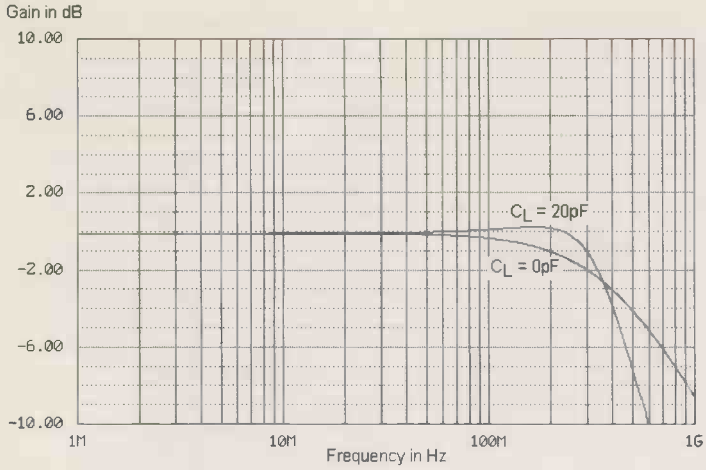


Fig. 11. Effect of load capacitance on microwave transistor.

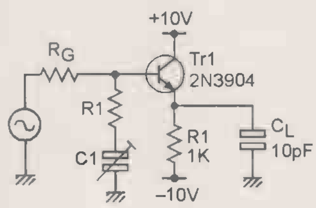


Fig. 12. Bandwidth limiting with RC filter.

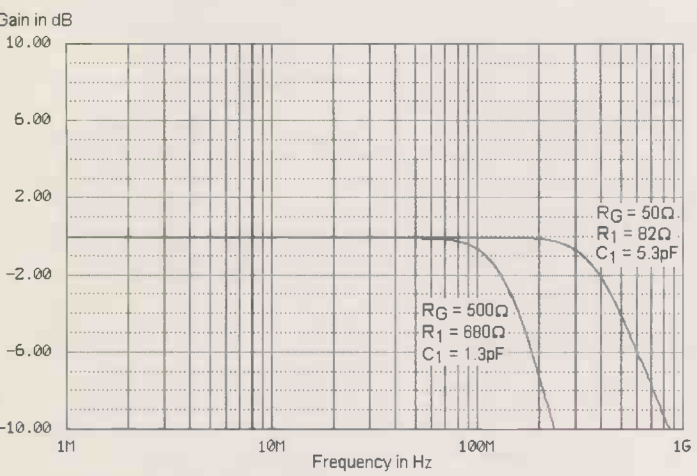


Fig. 13. Response of circuit in Figure 12

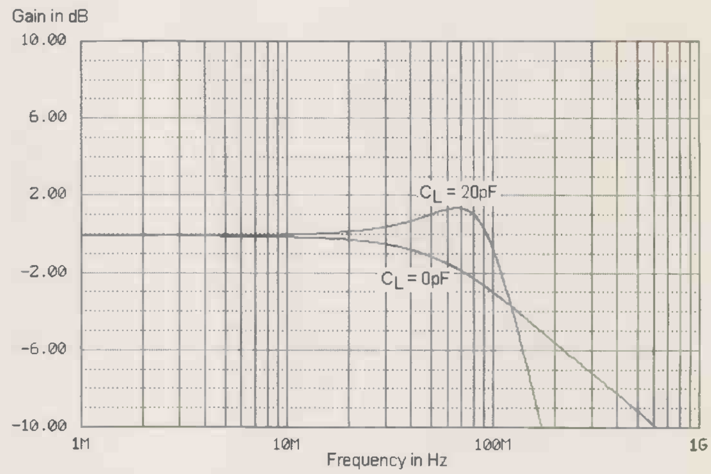


Fig. 14. Effect of change in load capacitance.

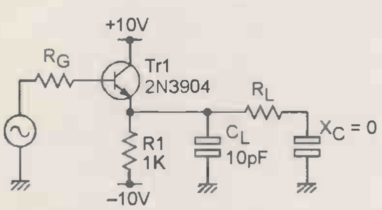


Fig. 15. Adding a load resistor.

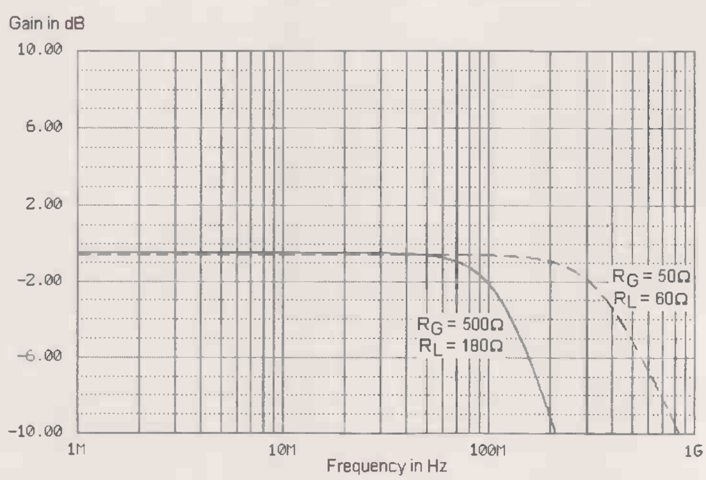


Fig. 16. Response with load resistor.

smaller load resistance is a reduced input resistance.

Other buffer circuits

The complementary compound emitter follower, or feedback pair, of **Fig. 18** reduces the distortion of the single transistor buffer, by keeping v_{be1}, which determines the voltage drop from input to output, constant.

The current i_1 through resistor R2 remains fairly constant, as the voltage across it is set by v_{be2}. If i_1 is chosen at 5 to 10 times the varying base current i_{b2}, it provides a fairly constant current through Tr1 and a constant V_{be1}, when the input voltage and output load current changes.

Since the base current of Tr1 also remains constant, the input resistance is increased.

Due to the extra delay through the second transistor, it will be more prone to instability with a capacitive load, as the response in **Fig. 20** of the circuit in **Fig. 19** shows.

To flatten the gain of this circuit is difficult, even without any load. **Fig. 21** shows one solution for a 10pF load. Gain flatness above 10MHz is dependent on component tolerances.

Large signals

If a buffer is to handle large signals, a current source in the place of an emitter resistor offers several advantages. **Fig. 24** compares the current flows for an output signal that varies from –5V to +5V over a load resistance of 1k, within a ±10V supply. Both circuits draw 20mA at V_{out} = 0V.

In the case of an emitter resistor, the transistor current i_e varies from 5 to 35mA, whilst with a current source, it

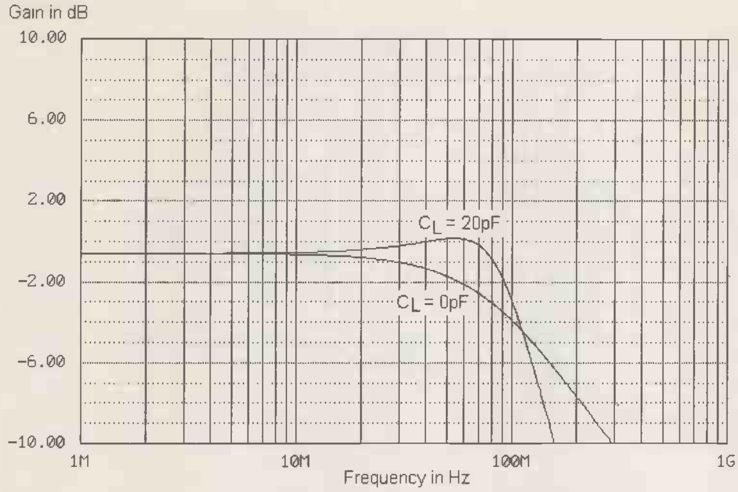


Fig. 17. Effect of change in load capacitance, for circuit in Figure 15 with R_G = 500Ω and R_L = 180Ω.

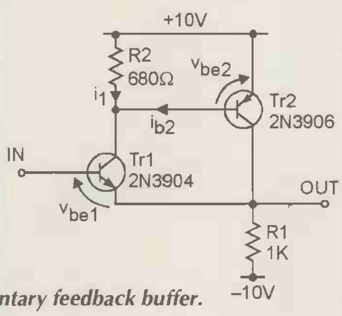


Fig. 18. Complementary feedback buffer.

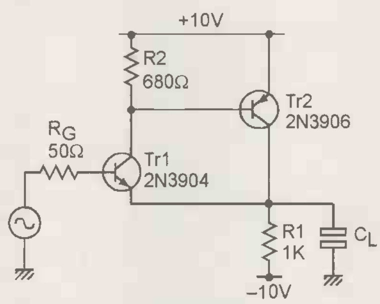


Fig. 19. Complementary feedback pair.

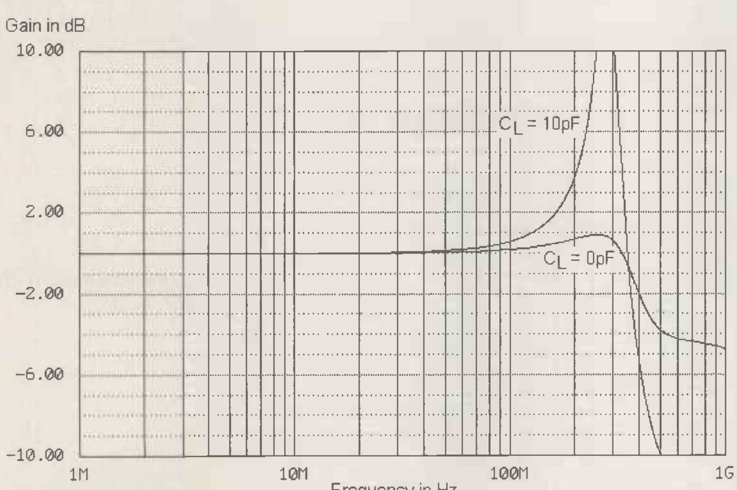


Fig. 20. Response of circuit in Figure 19.

varies from 15 to 25mA. A smaller change in i_e means less change in v_{be}, which reduces the distortion from input to output. Knowing the change in i_e, one can get an idea of the distortion at lower frequencies from a datasheet graph of the base-emitter voltage V_{BE} against collector current I_C.

A current source also allows the buffer to operate close to the supply rails. An emitter resistor limits the negative peak, e.g. in the example given, it is limited to –6.7V.

Current for capacitive load

The maximum current required to drive a capacitive load at a given frequency, can be determined as flows:

From the definition of a capacitor:

$$\text{Capacitance C} = \frac{\text{Electrical charge Q}}{\text{Voltage V}} = \frac{\text{Current I} \times \text{Time t}}{\text{Voltage V}}$$

from which follows the slew rate for a given current and capacitance as:

$$\frac{\Delta V}{\Delta t} = \frac{I}{C}$$

The maximum slew rate for a sinusoidal wave occurs at the zero crossing and is:

Combining the two equations:
Current required I = Vp ω C

Class B buffer

A class-B buffer is better than an emitter-follower with a current source, in that it can provide high source and sink currents when required, whilst drawing a low quiescent current. This makes it particularly suitable for driving a capacitive load at high frequencies. It also has the advantage of symmetrical distortion (positive and negative halves are attenuated the same), and the same DC level from input to output.

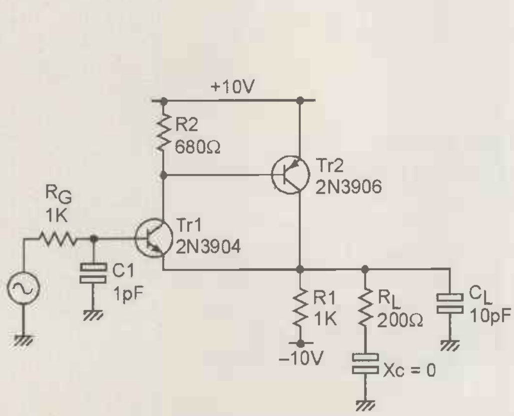


Fig. 21. Complementary feedback pair with flattened gain.

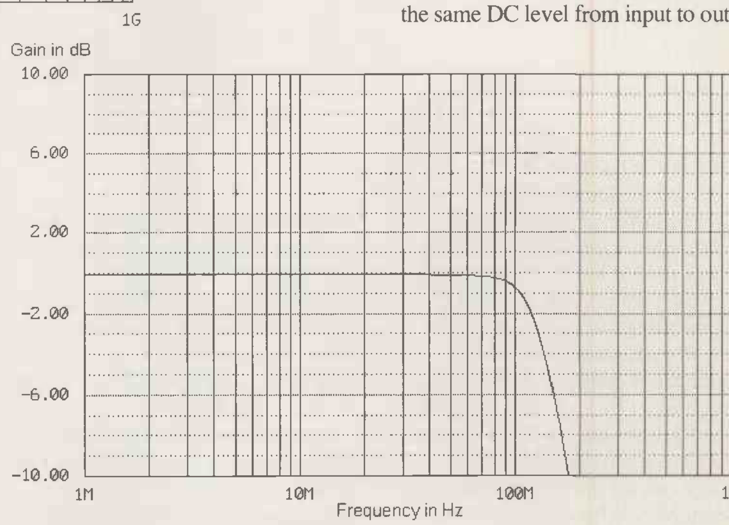


Fig. 22. Response of circuit in Figure 21.

Figure 25 shows one example of such a circuit. The (small signal) frequency response is shown in **Fig. 26.**

Two current sources keep the voltage drop over the diodes constant, and the distortion down. For less critical applications, the current sources can be replaced with resistors, e.g. 4k7 values for large signals, but that would lower the input resistance to half the resistor value, e.g. 2.3kΩ.

Schottky rectifier diodes D2 and D3 are used for a low voltage drop of about 0.2V. R7 and R8 determine the quiescent current through the transistors, and makes the output resistance of the buffer half of their value, namely 7.5Ω.

To reduce distortion, the class B stage is often combined with a differential amplifier, as in **Fig. 28.** The delay through four transistors does however severely limit the bandwidth, when stabilising components are added. Tr1 and Tr2 form a differential amplifier, D1 and Tr3 mirror the current from Tr1, which drives class B stage Tr4 and Tr5. The output is fed back to the base of Tr2. For gain, it is fed back via a resistive divider. ■

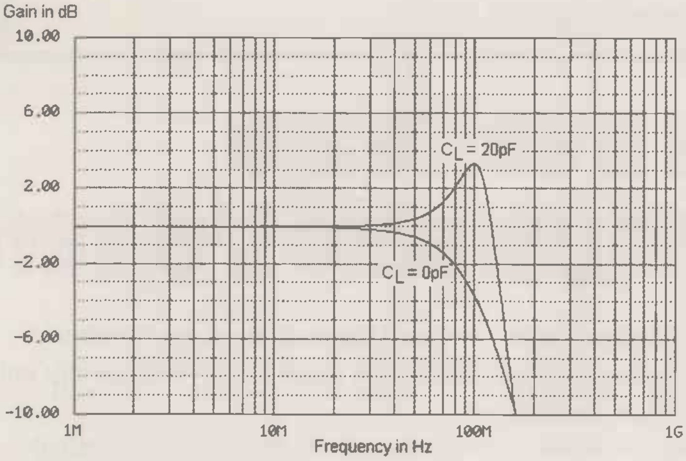


Fig. 23. Change in load capacitance.

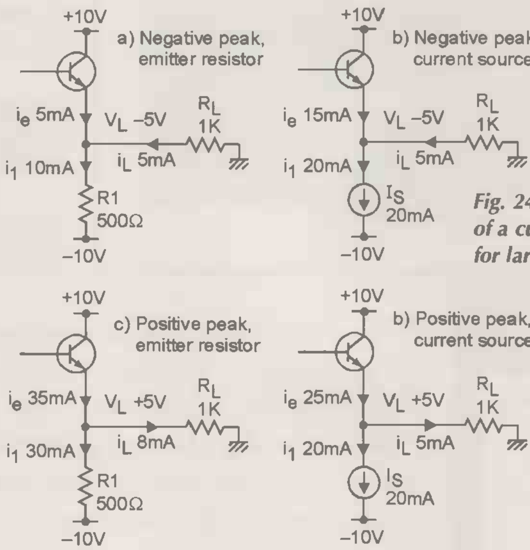


Fig. 24. Advantage of a current source for large signals.

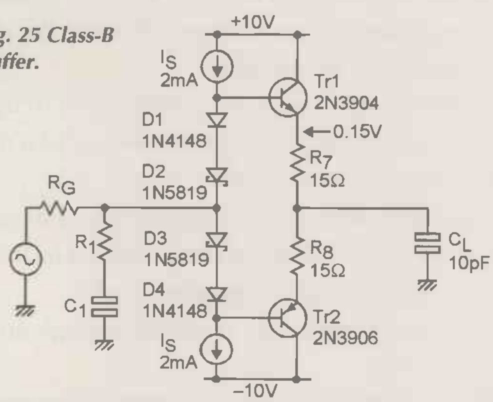


Fig. 25 Class-B buffer.

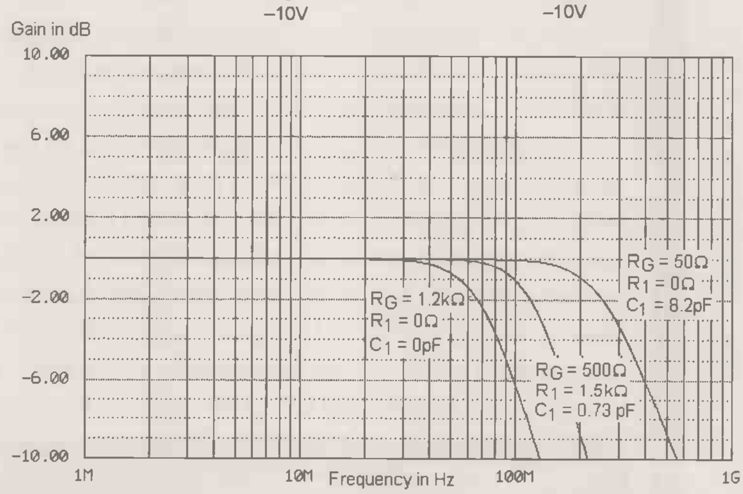


Fig. 26. Class-B buffer response.

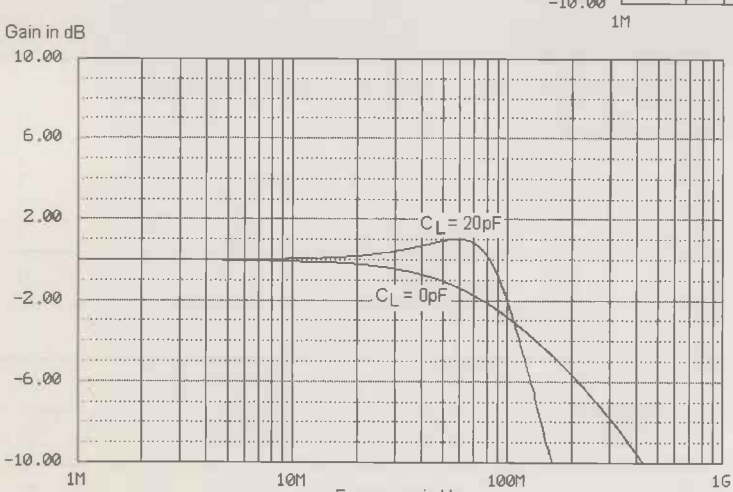


Fig. 27. Effect of load capacitance, with R_G = 500Ω, R1 = 1k5 and C1 = 0.73pF.

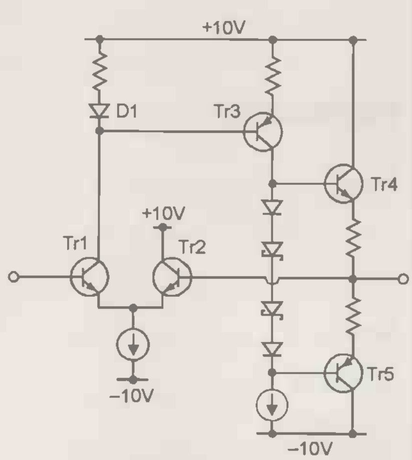
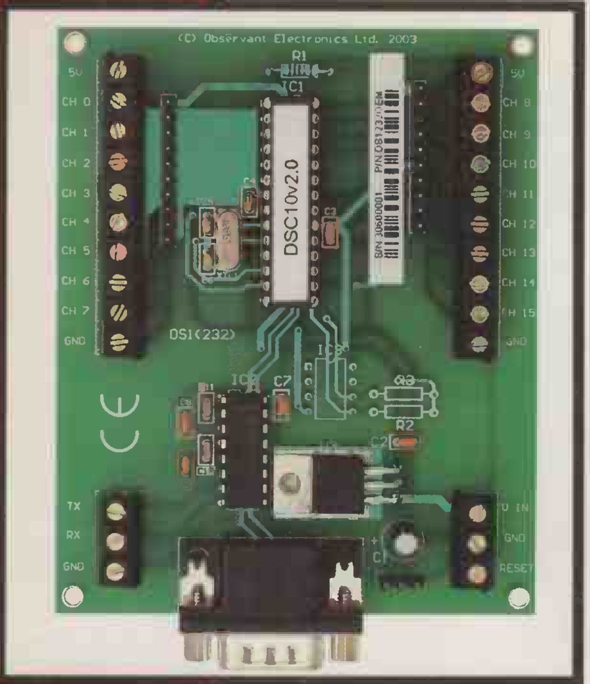


Fig. 28. Class B stage with feedback.



Want to win it? Simply answer the questions in the box below and send it together with your name and address and your entry will go into the hat. Post your answer to Electronics World, Highbury Business Communications, Nexus House, Azalea Drive, Swanley, Kent, BR8 8HU
Please read the competition rules first though.

This competition is sponsored by Observant Electronics
www.ObservantWorld.com

Win a software configurable I/O card

By combining software-configurable channels with a compact instruction set, the Observant Electronics DataStation family of software-configurable, mixed signal I/O products brings unique flexibility to control and data acquisition applications.

The new product overcomes the two main problems of traditional serial I/O solutions: i) limited choice of channel configurations and ii) low sample rates.

Delivering a 12 bit sample rate of greater than 500 samples a second – a 50 times improvement over many other solutions – DataStation is suitable for applications as diverse as machine control, industrial automation, environmental monitoring and robotics. Making it possible for designers to meet the needs of virtually any application with just one software compatible family of products.

DataStation's 16 software-configurable channels can be configured to give any combination of up to 6 analogue inputs (with selectable 8 or 12 bit resolution), 2 PWM outputs, 16 digital inputs and 16 digital outputs; giving a total of 391 different I/O configurations from a single product.

The DataStation family is available as an OEM product and in an IP65 rated enclosure and will soon be extended to include an IP40 rated enclosure plus RS-485 and CAN-bus interfaces in addition to the current RS-232 interface. The winner will receive an Observant Electronics DataStation plus development board and cable (normally worth £119 + VAT). For those readers unlucky enough not to win there will be a special offer running from the November issue.

Competition rules

No purchase is necessary. Strictly one entry per household. Competition closing date 20th November 2003. No entries will be accepted after that date. The draw will take place within the next working week following the 20th November. and the winner will be announced as soon as possible thereafter. The prize is not negotiable. No correspondence will be entered into regarding this competition. No employee of Highbury Business Communications or Observant Electronics may enter the competition.

THE QUESTION

1: How many different I/O configurations are possible with the DataStation?
2: If you win the product, what use will you put it to?

THE ANSWER is

1:____________________ 2:____________________

Name____________________

Address____________________

Entries only accepted on an original or photo copy of this form

NEWPRODUCTS

Please quote *Electronics World* when seeking further information

Digitally programmed capacitor

Designed to replace variable capacitors, Xicor has introduced the X90100 digitally programmable capacitor, which can be used to tune the frequency response of electronic systems up to 400MHz. The device can be set to 1 of 32 discrete capacitor steps ranging from 7.5pF to 14.5pF in 0.20pF increments. Once the desired capacitor value is selected via an up/down interface it is stored in the on-chip EEPROM.
Xicor
www.xicor.com
Tel: +44(0) 1993 700544

High power transistor cuts saturation voltage

Rohm's latest family of small transistors are available in high-power surface mount packages and feature reductions in saturation voltage V^{CE}(sat) of up to 90 per cent when compared with previous technologies, according to the supplier. The low V^{CE}(sat) transistors are designed to offer low energy consumption characteristics in switching and interface circuits without sacrificing current handling capabilities. Package options include 3-pin TSMT3 devices with power ratings of 0.5W, and 6-pin TSMT6 devices rated at up to 1.25W. These packages both have dimensions of 2.9 x 1.6 x 1.0mm. Available in both NPN and PNP versions, the surface mount parts are suitable for power management, load switch applications, and DC-DC conversion in portable applications such as mobile phones, PDAs and camcorders. Collector currents for the transistors range from 1A to 6A, while voltage ratings are either 12V or 30V depending on device chosen. In addition, the company can supply further devices in three smaller packages of SC-70 SC-75A and VMT3 measuring just 1.2 x 0.8mm. Here the I^C capability ranges from 0.5A to 1.5A.
Rohm
www.rohm.com
Tel: +44(0) 1908 282666

Hall-effect sensor with snug-fitting back-biasing magnet

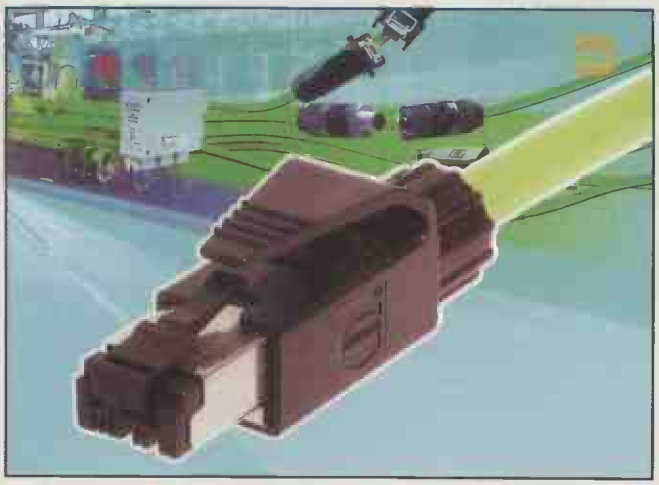
Allegro MicroSystems has developed a packaged Hall-effect sensor IC with a back-biasing magnet designed to achieve high air-gap performance. The SG package allows the magnet to sit closer to the IC than it does in conventional packaging, allowing it to work with a higher air gap and smaller magnet. The resultant package fits into the tight spacing dictated by automotive applications such as ABS systems, said the company. The package is manufactured with a single-step moulding operation, which according to the supplier eliminates the possibility of voids caused by air entrapment during the potting process and results in improved heat dissipation. The lead configuration is suitable for PCB surface mounting and attachment of a bypass capacitor. The spacing provided between the two leads allows an axial-leaded capacitor to be welded across them. Attachment to the sensor can also be made with a lead frame to preclude the use of a costly PCB.
Allegro
www.allegromicro.com
Tel: +44(0) 33 450 512 359

Aluminium instrument housings in all shapes and sizes

Hammond Electronics has added an additional three sizes to its recently introduced 1455 family of extruded aluminium instrument cases, designed to house PCBs mounted horizontally into internal slots in the body of the case or as an enclosure for any small electronic or pneumatic instrument. The recently introduced sizes are the 1455J160 at 160 x 78 x 27mm, the 1455L160 at 160 x 103 x 30.5mm. The two larger sizes accept a standard or extended depth Eurocard. The body of the enclosure is in either clear or black anodised finish.
Hammond Electronics
www.hammondmfg.com
Tel: +44(0) 1256 812812

Ethernet connector

Harting has introduced a range of connectors for industrial Ethernet systems. The range of Ethernet products being offered covers all the relevant international standards being developed on both sides of the Atlantic by organisations such as Profibus Nutzer Organisation, Open Devicenet Vendor Association, International Electrotechnical Commission and Industrial Automation Open Networking Alliance. For industrial Ethernet applications, the range includes the RJ industrial family which offers data-only types in IP67 and IP20 variants and a hybrid (data and power) version to IP67. Also available are Ethernet compatible shielded M12 connectors to IP67 and rectangular HanBrid 3A-RJ45 connectors to IP65. RJ Industrial connectors are the only RJ45 based connectors that can accept 22 AWG gauge conductors, claimed the firm.
Harting
www.harting.co.uk
Tel: +44(0) 1604 766686



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Power supplies for all weathers

Designed for environmentally demanding applications such as mobile radio base stations and roadside electronic cabinets, the BBC series of 200 to 400W AC-DC switching power supplies from XP is a range of rugged units that need no derating over a –20 to +70°C ambient temperature range. The power supplies use a 6mm thick aluminium baseplate to provide effective conduction cooling. All heat dissipating components are attached to the baseplate which in turn is attached to the equipment enclosure. The single output units comply with international EMC and ETSI standards and the active power sharing feature allows simple parallel operation where higher output currents are needed. A version capable of operating at down to –40°C is also available and the power supplies can be conformally coated for use in environments with high humidity. The BCC family consists of 12 standard power supplies with universal AC inputs and outputs from 3.3V DC (50A) to 28V DC (14.5A). Output load regulation is 1.5 per cent for output voltages up to 7.5V DC and 1 per cent above this.

XP
www.xppic.com
Tel: +44(0) 118 984 5515

PCI/LPC embedded modules

Diamond Point is offering the X-board from Kontron, which is a family of high density PCI and LPC-compatible embedded PC modules. According to the supplier, it is intended to close the gap between the previous DIMM-PC and ETX embedded PC modules. The sub-miniature credit card size (68 x 49 x 8mm) modules integrate a host of interfaces including USB, LAN, and serial in addition to sound, LCD and CRT. The omission of legacy interfaces such as PS/2, floppy and parallel enables the use of ARM and or other Risc processor architectures as well as X86. Powered by a single 3.3V supply, the first X-Board CPU has been developed based on a 266MHz National Geode processor. It comes complete with integrated LCD digital, CRT analogue and TV-composite graphics interfaces, three USB ports, 10/100Mbit Ethernet, AC97 digital audio and PCI/LPC bus expansion.

Diamond Point
Tel: +44(0) 1634 722390

Low ESR cylindrical SMT electrolytics

A series of cylindrical surface mount electrolytic capacitors from NIC Eurotech are designed to provide low ESR (equivalent series resistance) performance and long load life ratings. Capacitance values are between 3.3μF and 6,800μF with voltage ratings of 6.3V to 100V DC. ESR values at 20°C and 100kHz are down to 0.0033Ω. The load life rating at 105°C for 105°C for 4x10mm diameter parts is 2,000 hours, whilst for 12.5 x 18mm parts this increases to 5,000 hours. With an operating temperature range of –55 to +105°C, the capacitors called Nack can be used in electronic equipment designed for use in harsh environments. All parts are designed for automatic mounting and reflow soldering.

NIC Eurotech
www.niccomp.com
Tel: +44(0) 1634 722390

High frequency PLLs for WLANs

National Semiconductor is offering its first family of high frequency phase lock loop (PLL) devices. Aimed at applications operating at over 3GHz as well as wireless LANs and 5.8GHz cordless phones, the first products are the LMX2430, LMX2433 and LMX2434 PLLs. The devices feature a normalised phase noise contribution of –219.0dBc/Hz, which will increase radio sensitivity, extend base station range and improve voice quality, said the supplier. The LMX2434PLL has an operation frequency of 5.0GHz RF with 2.5GHz IF, which can eliminate

AC-DC supplies support world standards

Lambda has introduced a series of AC-DC power supplies that accept a continuous universal input from 85 to 265V AC (47-63Hz) and as such will support world standards. The supplies use soft-start circuitry, have an integral class B EMI filter and provide over voltage and over current protection. Other features include an output adjustment potentiometer (±10 per cent), integral power good LED indicator and screw terminals for input and output connections. Four models are available with nominal output power ratings of 50, 75, 100 and 150W. Each model can be supplied with a choice of 5, 12 or 24V output. The company will extend the SWS series soon to offer more output voltage selections between 3V and 48V. CE marked in accordance with the Low Voltage Directive, the new SWS series carries UL, CSA EN60950 and EN50178 safety approvals. All models conform to EN61000-3-2, the SWS100 and SWS150 being fitted with an integral active PFC filter. Other approvals include EN55011-B, FCC-B and EN-61000-4. Operating temperature range is –10 to +60°C

Lambda
www.lambda-gb.com
Tel: +44(0) 1271 856666

PROTEUS

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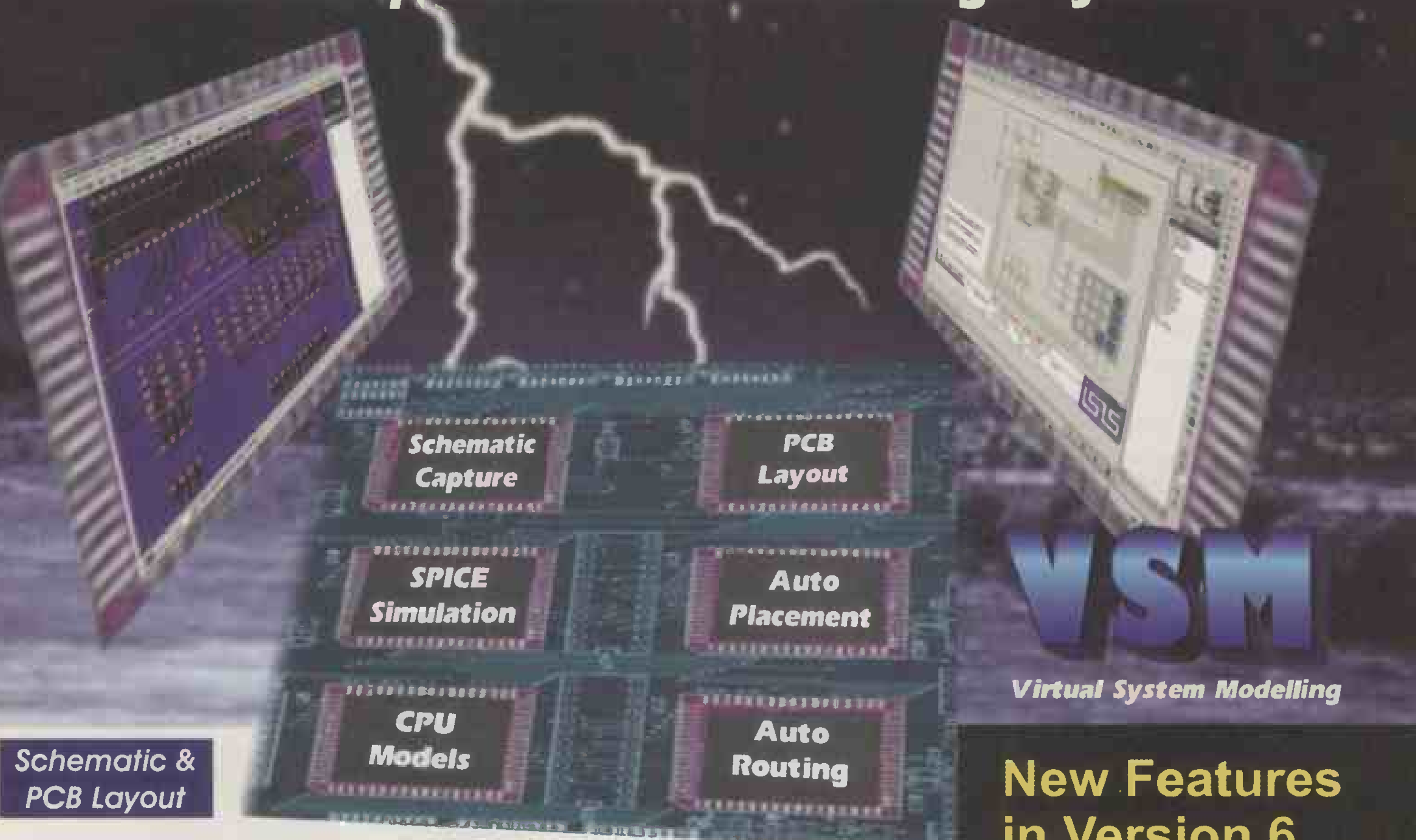

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info@labcenter.co.uk

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the need for frequency multiplication in 5GHz ISM band applications. Power consumption is 7.0mA. The LMX2433 PLL has an operating frequency of 3.6GHz RF with 1.7GHz IF and a power supply current of 5.2mA.
National Semiconductor
www.national.com
Tel: +44(0) 870 242171

Low ESR cylindrical SMT electrolytics

Embedded PC specialist, DSP Design, has introduced a PC/104

application development kit that supports Linux, the open source operating system. Known as LaunchPad LP400LIN, this application development kit is supplied with a version of the Linux operating system configured to meet the user's needs. It includes a PC/104 processor board, a Compact flash card pre-loaded with the Linux operating system, full documentation and sample software plus necessary hardware accessories. The complete system is housed in a smart laptop carrying case.
DSP Design
www.dspdesign.com

Analyser tests amplifiers in mobile handsets

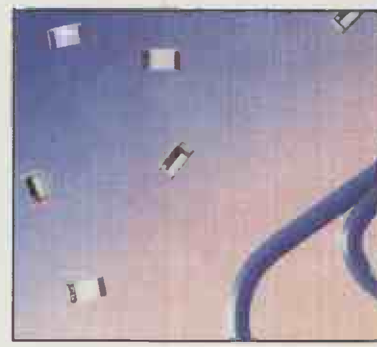
Anritsu has introduced the ME7840/4 handset power amplifier test system, that characterises the performance of next-generation handset power amplifiers. The device integrates the measurement capabilities of five instruments with a high power test set, said the supplier. The system handles 5W of power from 10MHz to 6GHz and includes measurement software including the ability to tune for swept power gain compression and intermodulation distortion (IMD). Other single connection tests include S-parameters, harmonics, and noise figure from 50MHz to 6GHz.
Anritsu
Tel: +44(0) 1582 433200

Kits are ready to roll

Trident microsystems has added a series of 'ready-to-roll' kits for battery-operated systems designed for implementing handheld PC-based instrumentation and equipment. Each of the kits includes a single-board computer and TFT display, together with the associated inverter, cables, software drivers and optional power supply and hard disk drive. Microsoft Windows can be pre-loaded. The kits are based on Toshiba's low power poly silicon 8.4" 800x600 TFT LCD. Designers can opt to have one of three 3.5" FDD-format Evalue SBCs in the kit: the ECM3412 uses a National Geode GX1 processor at 300MHz offering fanless low power operation; The ECM3610 has a 667MHz VIA Eden processor; and the ECM3615, with a Transmeta Crusoe TM5400, can run two displays simultaneously.
Trident Microsystems
www.tridentdisplays.co.uk
Tel: +44(0) 1737 780790

Chip inductor for 3GHz signals

The HPL range of thin film chip inductors, available from Rhopoint Components, operate



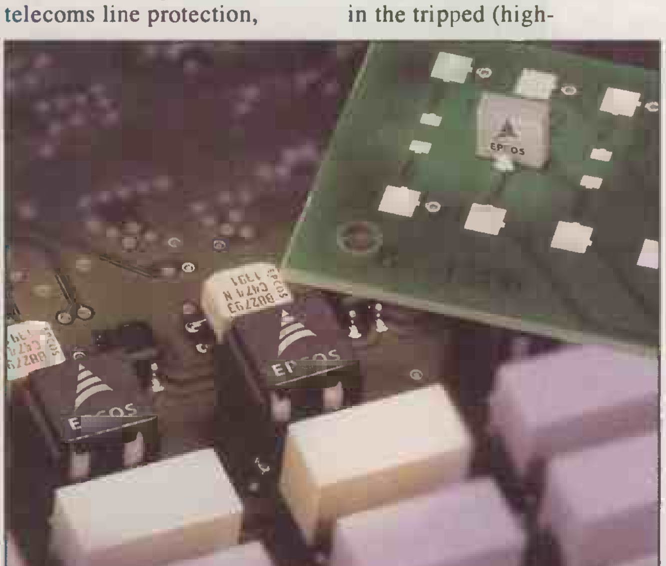
at frequencies as high as 3GHz and is designed for use in such applications as cellular phones, telecommunications networks, and filter circuits. These thin film inductors have a Q factor of up to 50 at 1.5GHz, which is higher than most multilayer devices. Available in the range 1.0nH to 15nH, this series of inductors has an inductance tolerance down to ±0.1nH. The temperature coefficient of inductance is from 0 to 125ppm/°C. The inductors are available in 0201 and 0402 sizes.
Rhopoint
Tel: +44(0) 1883 717988

DC-DC outputs with independent regulation

SynQor has a line of dual output quarter brick DC-DC converters with independent regulation. Called the DualQor Giga series, the quarter bricks are designed to deliver 15A of total output current on each rail simultaneously, without a heatsink. Profile on a standard footprint and pin out is 8.5mm. The series operates from an input voltage of 35V to 75V and offers output voltage combinations of 3.3/2.5V, 3.3/1.8V, 3.3/1.5V and 3.3/1.2V with other voltages to follow.

Line protection thermistor withstands lightning

Epcos has launched two PTC thermistor designs for telecoms line protection, which comply with ITU-T standards K20, K21 and K45 in the tripped (high-



resistance) state, they can even withstand continuous contact with power lines at 110/230VAC. The square PTC thermistor R212 is designed for the latest generation of main distribution frame modules. With overall dimensions of only 5.95 x 5.95 x 2.65mm, it is intended for vertical surface mounting which offers considerable space savings over conventional SMD solutions. Its electrodes are suitable for soldering as well as for clamping. Rated resistance is 12Ω at 120mA and 25°C. There is a telecoms pair protector (TPP) containing two PTC thermistors in a four-pin case.
Epcos
www.epcos.com

Please quote *Electronics World* when seeking further information

The unit measures 1.45 x 2.3 x 0.335in. All modules can deliver full load current of 15A on each rail at typical ambient environments of 55°C and 200 LFM airflow. The DualQor units essentially consist of two eight-brick modules placed in the space of a quarter-brick. According to the supplier, this allows it to develop any output voltage combination required by customers.

SynQor
www.synqor.com
Tel: +44(0) 49 9621 60700

Chipset for cordless telephones

Atmel has available a 2.4GHz DCT chip set, consisting of the LNA/PA IC T 7026 and the transceiver ICT 2803. It is designed for cordless phones, while also being used for wireless data applications such as internet access, wireless headsets, games and home entertainment. The chipset enables an extended distance between the handset of a cordless phone and the basestation, from 300m to 600m. the 2803's variable data transmit bit rate allows to adjust to any data transmit bit rate in a range of 9.6kbit/s to 1.152Mbit/s.

Atmel
www.atmel.com
Tel: 0033 442 53600

Module brings GPRS connectivity

Acte Components has available the WISMO Quik Q2406B module that is designed to bring voice, data, SMS and GPRS class 10 connectivity to applications including mobile phones, industrial machine-to-machine and automotive applications. The ETSI compliant Q2406B is loaded with complete 2.5G GSM/GPRS class 10 capability. It is designed for dual band operation and is available in either 900MHz/1800MHz 850MHz/1900MHz or CDMA/AMPS variants. With overall dimensions of 58 x 32 x 3.9mm thick, the module includes an optional embedded TCP/IP. It also has 32k of RAM and 4k EEPROM that enables customers to embed their own applications (via MUSE Open AT) and reduces the need for additional software and hardware that is external to the module. The Q2406B draws less than 3.5mA in GSM idle mode, and an average 300mA when communicating (GSM900, 0 per cent DTX, Pmax). The module requires a 3.6V DC power supply and can be connected directly to a Li-ion battery pack.

Acte Components
www.actecomponents.com
Tel: +44(0) 1256 84588

Analogue video signal analyser

Tektronix has an automated component analogue video signal analyser that is designed to measure high definition, progressive scan, and PC format signals for consumer video equipment manufacturers, video network operators, and others that require fast and repeatable testing. Designated the VM5000HD, the automated video measurement set is designed to provide accurate, and repeatable video measurements in 1080i, 720p, 480p, and SXGA formats making use of multiple industry-standard video parameters, without the need for complicated instrument set-ups, algorithm selection, time-consuming manual measurements or tedious results correlation. With a single button push, the VM5000HD can make 100 different parametric measurements in eight specific test categories within 10 seconds so that product performance can be objectively and reliably assessed. It offers a high-definition matrix test signal set for the creation of standardised test signals including colour bars, multi-burst, sweep, and five other signal types, for testing in Y/Pb/Pr and RGB colour space.

Tektronix
www.tektronix.com
Tel: +44(0) 1344 392241

Test software for fast serial data

Test equipment and software aimed at compliance and validation testing for emerging serial data standards has been unveiled by Tektronix. The RT-Eye serial data compliance and analysis software (TDSRT-Eye) package works with the firm's TDS6000 and TDS/CSA7000 oscilloscopes and the P7350SMA 5GHz differential probe. It is capable of testing serial links up to 3.2Gbit/s. Option RTE provides software clock recover, eye diagrams, and standard specific parametric measurements in the amplitude, timing, and jitter domain, including total jitter at 10^{12} BER. Beyond known standards, waveform masks, measurement

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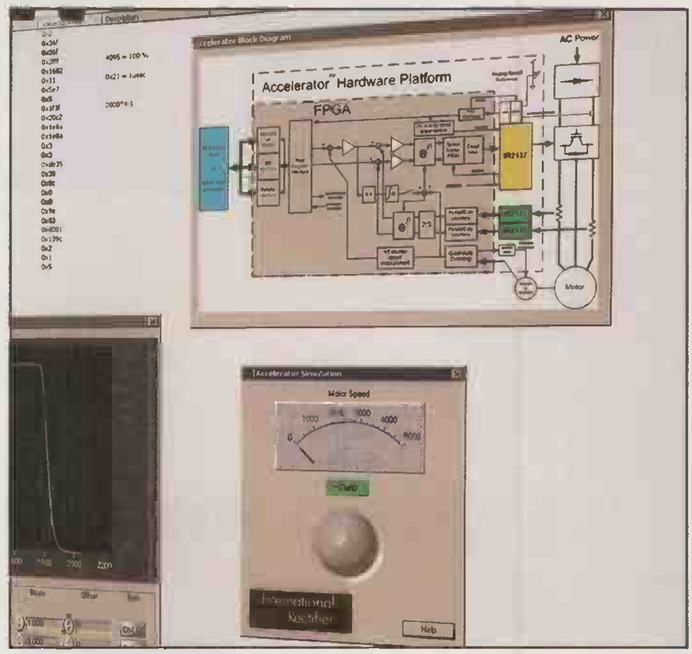
limits, and reporting functions can be customised by the user, enabling companies and industry groups working on proprietary and emerging standards to develop their own internal and external compliance programs and reports.
Tektronix
measurement.uk@tektronix.com
Tel: +44(0) 118 984 5515

Dual comparator and reference for 6.5µA

Linear Technology's LT6700 dual micropower comparator has an integrated 400mV reference and is available in the 6-lead ThinSOT. Aimed at monitoring supply voltages in portable devices, the chip has 40mA outputs for driving LEDs, opto-isolators and relays directly. Operating over 1.4V to 18V supply range, the device draws 6.5μA, has built-in comparator hysteresis, and guaranteed plus two per cent threshold accuracy to ensure stable operation over temperature and in noisy environments. The device provides two external comparator inputs with the other two inputs connected internally to the on-chip reference. These high impedance inputs need less than 10nA input bias current. The comparator outputs are open collector and the output load can be referred to any voltage up to 18V, also independent of supply voltage.
Linear Technology
www.linear.com
Tel: +44(0) 1276 677676



IP library for speedy motor control design

International Rectifier has announced the IRMCS203 development system and the IRMCO203 intellectual property (IP) library of motion control algorithms. This development system and object code IP library support the firm's iMOTION integrated design platform for applications such as high-speed spindle motor control. The development system comes with proven analogue and power stage hardware designs as well as the software needed to achieve a design flow that requires no coding to configure motor control algorithms. The digital control section uses the firm's Accelerator configurable control engine system architecture using parallel processing and customisable peripherals. The analogue section uses high voltage ICs for the gate drive and current sensing functions. The power stage is based on the latest non-punch-through (NPT) IGBT and PlugNDrive integrated power module technologies. The algorithm is a licensable object code IP library for sensor-less control of permanent magnet motors featuring high starting torque and smooth ramp-up. It is targeted for low cost FPGA applications, and is designed to reduce the need for complicated programming processes.
International Rectifier
www.lambda.com
Tel: +44(0) 20 8645 8003

DIN-rail AC-DC supplies for industry go lead-free

Lambda has developed the DLP series, a family of DIN-rail mounted AC-DC supplies aimed specifically at factory automated, industrial control and test and measurement applications. The 23V DLP series has a maximum efficiency of 87 per cent and power density of up to 0.21W/cm^3. Three models are being launched with output powers of 75, 120 and 240W. Output currents range from 3.1A on the DLP-75 to 10A on the DLP240-24. Input voltage covers the range 85 to 264AC. The supplies are fitted with a red alarm LED giving operators an immediate warning should a short circuit of the load occur. The units have a common height and depth of 97 x 110mm. They range in width from 50mm for the 75W unit to 120mm for the 240W supply. Operating temperature range is –10 to + 50C at full load derating to 60 per cent at +60°C. From October 2003 the DLP will be manufactured as lead-free to satisfy emerging legislation.
Lambda
www.lambda.com
Tel: +44(0) 1271 856666

Fact: most circuit ideas sent to *Electronics World* get published

The best circuit ideas are ones that save time or money, or stimulate the thought process.

This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

Don't forget to say why you think your idea is worthy.

Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly. Where software or files are available from us, please email Jackie Lowe with the circuit idea name as the subject.

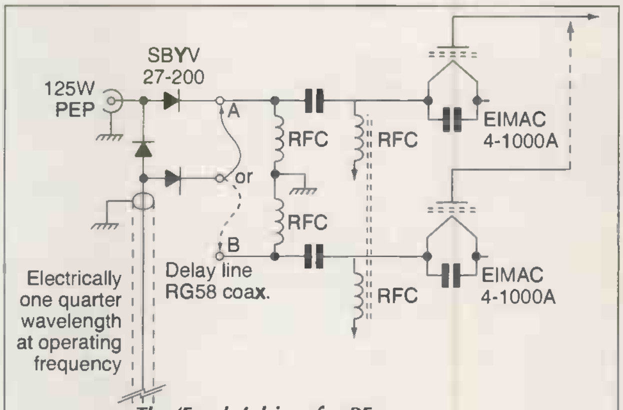
Send your ideas to: Phil Reed, Highbury Business Communications, Nexus House, Azalea Drive, Swanley, Kent, BR8 8HU *email ewcircuit@highburybiz.com*

RF power loader

This circuit offers a symmetrical load for the 125W exciter. It is equally suited to driving one or two loads but power is increased when one one load is driven. The configuration is broadband and inexpensive.

When driving two loads, outputs of the circuits are connected in parallel. The diodes are very high speed SBYV27-200 types capable of handling an amp. Their reverse recovery time is 15ns.For lower powered and VHF circuits a suitably rated 'hot carrier/Shottky' diodes would be suitable.

Frank Shaw, VK6ALF
Ballajura
Western Australia



The 'Franks' driver for RF power loading.

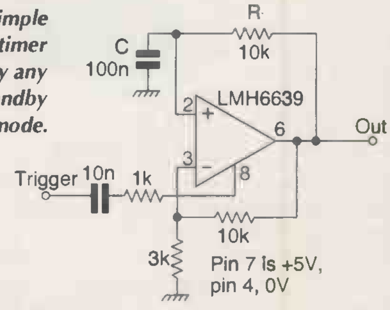
Simple triggered timer

A trigger pulse of 5V amplitude turns on the amplifier and pulls the non-inverting input high. Output from the amplifier is latched high until capacitor *C* charges to the voltage on the non-inverting input. It then switches low, turning off the amplifier and discharging the capacitor.

An advantage of this circuit is the high speed of the LMH6639. Also, the circuit draws almost no power supply current when not triggered.

Kamil Kraus,
Rokycany
Czech Republic

This simple triggered timer draws hardly any current in standby mode.



Electronic fluorescent lamp starter

I have found that conventional neon/bi-metal starters are not only a source of malfunction, but they can also cause tubes to suffer from heater failures. In this circuit, the heater pins are paralleled, effectively halving the current in the heaters during tube operation.

The circuit has been in operation for some years now and appears to extend tube life significantly. I adapted existing choke ballasts to permit the circuit shown to strike the tube repeatedly until sustained by normal choke action.

Resistor R_1 charges C_1 until, the transistor conducts and fires the SCR, discharging C_1 to provide a high voltage across the tube. Once the tube has ionised, the rectified AC voltage generated across the choke tap drives the transistor negative, and turns the starter off.

Components R_1, R_2 and P_3 are 350V rated. In addition, C_1 must be suitable for 400V DC.

In the case of one slim style choke, it was possible to remove the end cap from the connections and solder a tap to the existing winding. In another case, a choke wound on E and I laminations was adapted by winding 50 turns of 36 SWG (0.2mm) wire over the outer layer without dismantling. The number of turns required should develop 6V when run as a transformer across 240V mains.

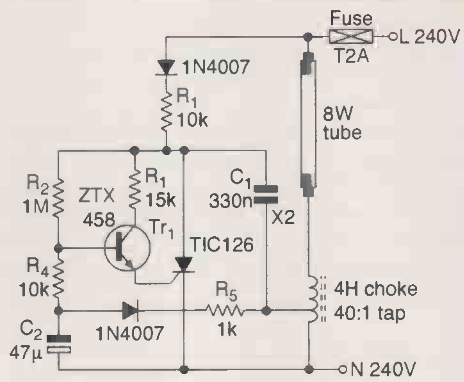
Observe safety precautions for mains operated equipment.

Henry Maidment
Salisbury Wiltshire UK

Warning

When powered up, this circuit must be considered live throughout and potentially lethal. Do not attempt to replicate it unless you are fully conversant with the regulations and safety guidelines associated with mains-rated circuitry.

Having found that conventional neon/bi-metal starters can cause tubes to suffer from heater failures I designed this electronic alternative, which effectively halves current in the heaters when the tube has fired.

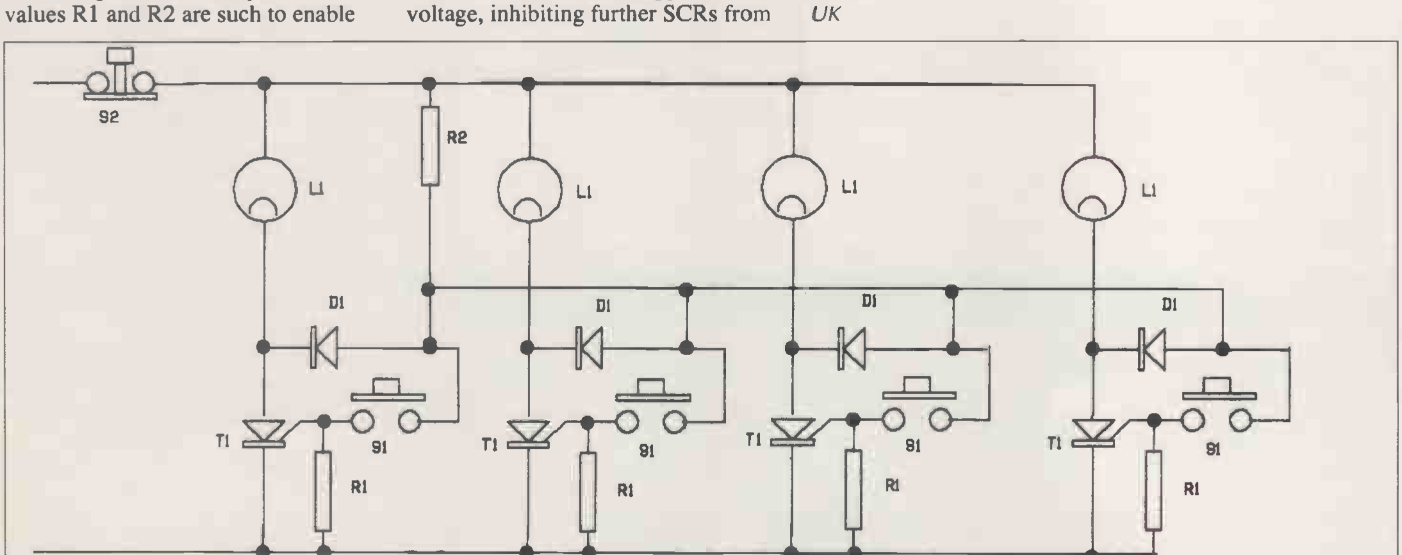


Simple fastest finger first circuit

This simple circuit indicates the first switch closed and will inhibit others until re-set.

A number of identical circuits can be built, four are shown, there are no critical components, the lamps L1 are low voltage filament lamps, resistor values R1 and R2 are such to enable the SCR trigger voltage to be available. The first switch of S1 to be closed will turn on the related SCR, illuminate the associated lamp, maintaining current flow. This will cause the diode gating arrangement to reduce the available trigger voltage, inhibiting further SCRs from being turned on. Momentary opening of S2 removing power and turning off the activated SCR resets the system.

M J Nicholas
Bournemouth
Dorset
UK



Simple oscilloscope DC offset adapter

This circuit was born out of nostalgia for some of the facilities available on my old Tektronix 535A oscilloscope, namely the 1A5 plug-in with its calibrated DC offset. These and other useful functions are not available on my current CRO, a Tektronix 2213. I have found this adapter useful for monitoring CRT video driver waveforms and improved amplitude accuracy by measurement of the offset voltage, while using the CRO as a null indicator.

The adapter is a 20dB pad for a 1MΩ system, with a DC injection point and frequency compensation. R1+R2 = 1M22; R3+R4 = 4M95 and R5+R6+R7 = 1M22

åWith these component ratings, the CRO adapter will have a maximum input rating of 250V (DC + AC peak). This being determined by the capacitor ratings, which can be increased by adding fixed 100V or 200V capacitors in series with C1 & C2. Select values about triple the total required value, this will give a total voltage rating approximately the sum of the two ratings. Alternatively, C1 & C2 could be constructed from copper or brass wire or strip, but the resistors then will limit the input to about 500V.

The offset generator can be either a 10kΩ 2W 10 turn pot connected across a ±40V regulated supply. Or a 40V floating supply with polarity reversing switch. A 1 volt input to the DC offset terminal appears as a 1 volt input to the adapter. When used with a 10:1 probe this gives an offset of 400V at the probe tip.

Setting up

A 1kHz square wave generator with low output impedance is required. First connect a 10:1 probe to the CRO and adjust the compensation. Then remove the probe and connect the adapter. Connect the generator to the input of the adapter via a suitable inline-terminating resistor and adjust C2 for optimum corners (the same as for probe compensation). Now connect the 10:1 probe to the adapter and adjust C1 for the correct waveform.

Operation

Use the adapter between the 10:1 probe and CRO to monitor signals with a high DC component. Set the CRO coupling switch to DC. And use the offset generator to bring the relevant part of the signal into view.

To measure a waveform

With the coupling switch set to ground centre the trace. Then set the V/div switch for the highest sensitivity without overload (distortion). Use the offset generator to bring the +ve peak to the centre and note the offset volts. Then do the same for the –ve peak, the pk – pk of the waveform equals the difference between the two readings.

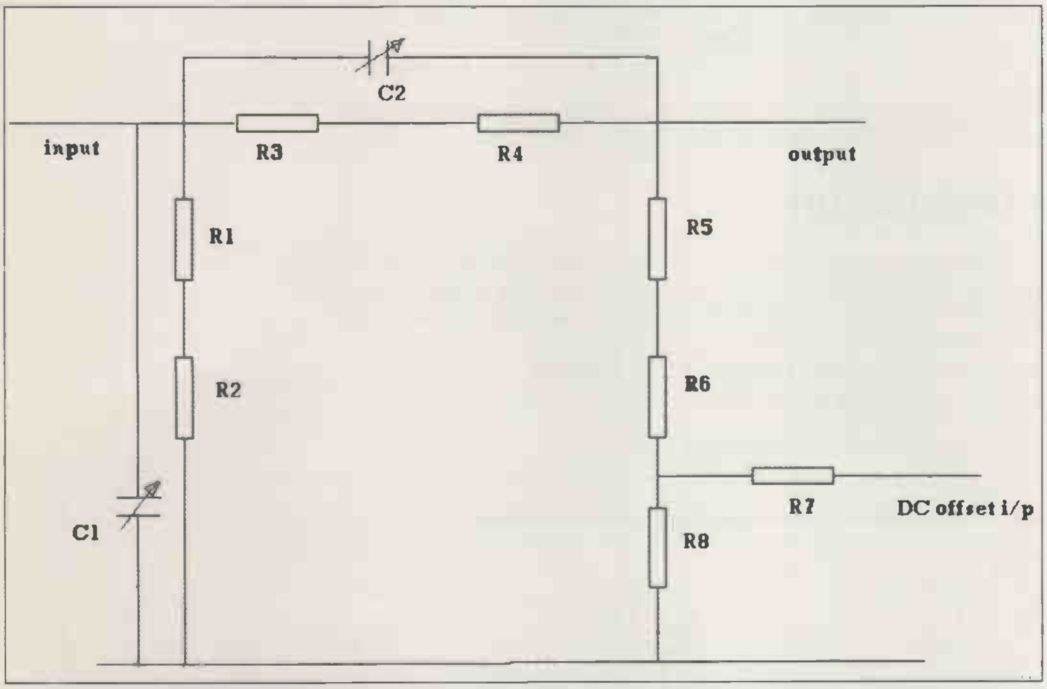
It can also be used to closely examine a signal. I have used it to check the output of a simple digital sine wave generator. With the V/div set to maximum (2mV/div) giving 20mV at the adapter input, the synthesised sine wave, (at 8V pk-pk) could be examined in detail over the whole cycle.

Construction

This adapter is mounted in a small box with a BNC male at one end and BNC female at the other. Farnell have a suitable item, 309-801. The components should be kept away from the sides of the box and the 1M22 resistors should preferably be at right angles to the 4M95 resistor. For improved accuracy, either use a suitable ohmmeter to select the value (the MRS25 resistors have good stability) or higher accuracy components, although getting the values above 1MΩ may be more difficult.

Caution: This unit will not increase the maximum input voltage rating of your oscilloscope or the probe

Rodger Bean
Watson
Australia



The resistors are all Philips MRS25 (0.6W, 250V max, 1%)

R1	698K	R5	681K
R2	523K	R6	536K
R3	2M2	R7	6K2
R4	2M7	R8	2K

C1 & 2 2 – 10pF trimming 25V, BC components 2222 808 11109, Farnell 303-811.

Missed call indicator

No answer phone or voice mail service? Need to be told you have been called? This handy missed call indicator will plug into a telephone socket and indicate if any calls have been received by a pulsing red LED and a unobtrusive chirp from the piezo sounder. All power is obtained from the telephone line.

Telephone systems can only supply a few milliamps without an unacceptable voltage drop so low power techniques are employed through out this design.

R5, D5,6,7,8,9 & C3 form a 5 Volt power supply. A bridge rectifier is employed to maintain supply when "ringing" is in progress and up to 100V AC is superimposed on the steady 50V DC line voltage.

R1, D1,2 R2, TR1, D3, & R3 detect the presence of the ringing voltage , D1 & 2 isolating C1 when off hook preventing dial tones from triggering the unit and any loading of the audio signal. When ringing is detected TR1 collector is pulled low on +ve going peaks, D3 providing AC continuity and protection for TR1.

R4 , C2 & IC1a are coupled into TR1 via D4 and in effect form a low pass filter.

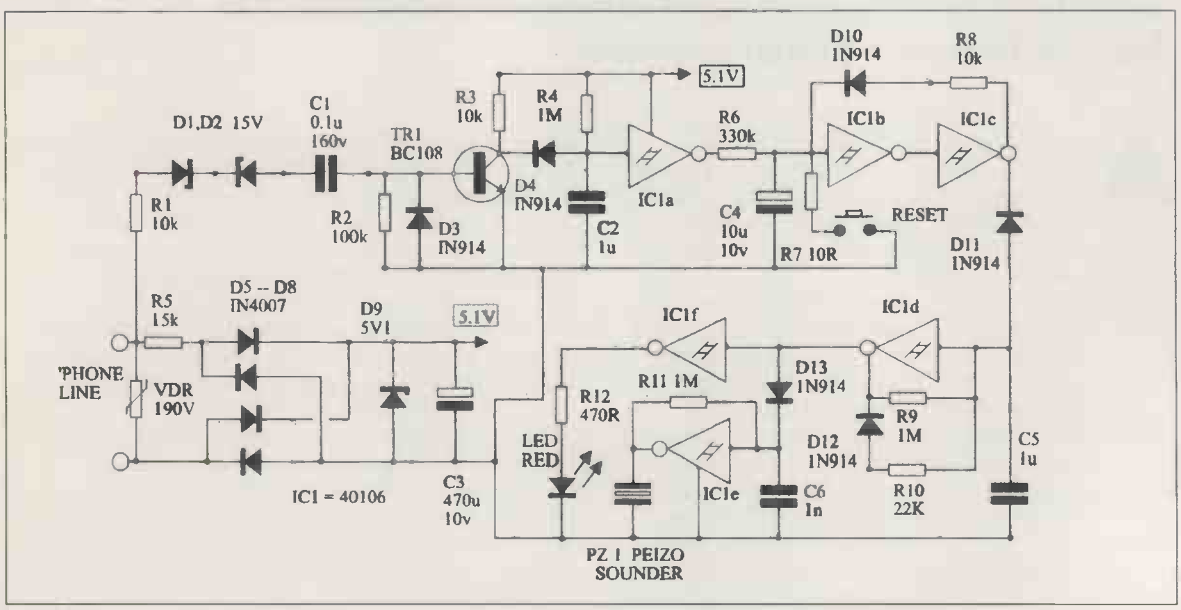
IC1a output goes high when "ringing" is first detected and does not go low until after ringing is finished, by when either the caller has rung off or voice mail has taken the call.

The output of IC1a is connected via R6 and C4, the time constant of which is approx. the delay before the unit triggers, to the latch formed by IC1b and IC1c and associated feedback components R8 and D10. The unit remains active until it is reset by pressing the reset switch. R7 reduces contact erosion by limiting the discharge current from C4.

When the Latch output is high Oscillator IC1d is enabled this gives a 20mS pulse every second, this is passed via buffer IC1f which in turn drives a high brightness LED. This pulse also enables the oscillator IC1e that chirps the piezo sounder.

NOTE: This unit is not type approved for connection to a PSTN and as such should only be used on private telephone circuits.

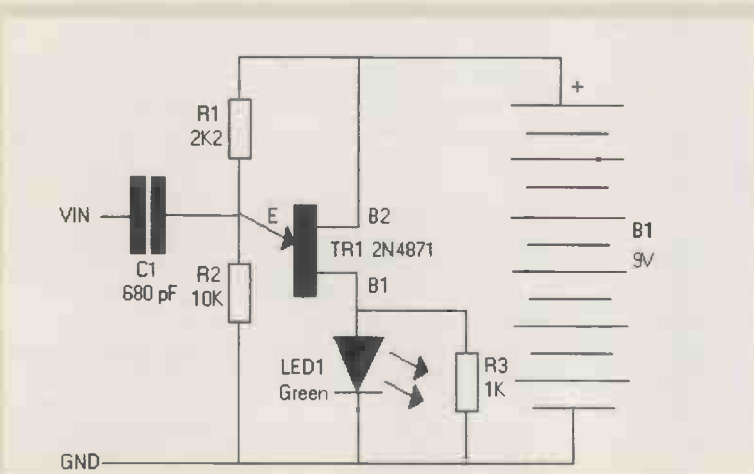
Nigel Goodman Eng tech MIIE
St. Leonards on sea
East Sussex



Unijunction memory

This circuit illustrates using a unijunction transistor as a bistable memory element. R1 and R2 set the DC bias at the emitter of TR1. If a positive pulse is applied to Vin, TR1 will switch to a low resistance state causing the LED to light. If a negative pulse is applied to Vin, TR1 will revert to a high resistance state. R3 prevents the LED glowing faintly when the transistor is in a high resistance state.

André de Guérin
Vale
Guernsey



Noise and moving-magnet cartridges

Marcel van de Gevel investigates noise optimisation in RIAA amplifiers for moving magnet cartridges and presents a high-performance design example.

As you can read in any textbook on low-noise design, the noise of an amplifier can be described with two equivalent input noise sources. These are an equivalent input noise voltage source in series with the input of the amplifier and an equivalent input noise current source shunted across its input[1].

Which of the two has the largest influence depends on the impedance of the signal source driving the input. For example, when the magnitude of the source impedance is 12kΩ, a noise current of 1pA/√Hz has the same influence as a noise voltage of 1pA/√Hz×12kΩ=12nV/√Hz.

Unfortunately, some of the measures that a designer can take to reduce the equivalent input noise-voltage density lead to an increase of the equivalent input noise current density and vice versa. This applies especially to the selection of the input device and its biasing conditions.

It is important for a designer of low-noise electronics to know the source impedance, so that he or she can choose the device and biasing that gives the smallest total noise for that specific source impedance. Because noise optima are usually rather broad, there is no need to know the source impedance with a high accuracy; a reasonable estimate will do.

In the case of RIAA amplifiers for moving magnet (MM) cartridges, the source impedance varies enormously over the band of interest. A typical cartridge with 1kΩ of DC resistance and 0.5H of inductance has an impedance of 1kΩ at low frequencies and almost 63kΩ at 20kHz. The question is: for what impedance should a RIAA amplifier be optimised if we want to get the smallest amount of audible noise? It will be shown in this article that the cartridge impedance at 3852Hz is a good estimate. Also, an easy way to account for 1/*f*-noise will be given.

When the input stage and feedback network are properly designed, the thermal noise of the 47kΩ resistor shunted across the input is usually the largest remaining noise contribution in the RIAA amplifier itself.

Just leaving this resistor out is not an option, as this would reduce the damping of the resonant circuit consisting of the cartridge inductance and the load capacitance, seriously affecting the response between 10kHz and 20kHz. However, a lower noise 47kΩ input resistance can be realised by using combinations of series and parallel feedback.

This technique is also known as the active input impedance or 'electronic cooling' technique. Two examples of such RIAA amplifiers, one discrete circuit and one with op-amps, will be given.

Finally, the improvement which can be obtained by replacing the 47kΩ resistor with an active input resistance will be estimated for a typical cartridge.

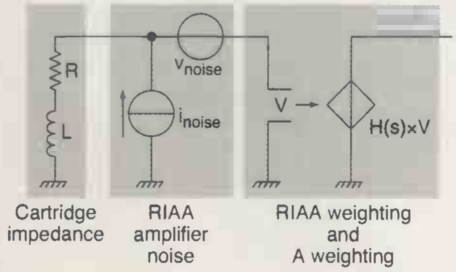


Fig. 1. Simple model of a cartridge connected to an RIAA amplifier.

The 3852Hz rule

Figure 1 is a simple model of a cartridge connected to an RIAA amplifier, followed by an A-weighting filter. The L-R series network models the impedance of a moving magnet cartridge with inductance L and DC resistance R.

The noise of the RIAA amplifier is represented by the noise voltage source v_{noise} and the noise current source i_{noise}. These noise sources are assumed to be white and uncorrelated. The noise of the cartridge itself is not included in this model, because the amplifier designer cannot optimise this anyway. The voltage controlled voltage source with transfer H(s) models the RIAA- and the A-weighting.

Note that a real cartridge has a frequency-dependent effective series resistance, unlike a simple L-R series network. However, this mainly affects the phase of the cartridge impedance, while only the magnitude will be used in the following calculation.

The contribution of v_{noise} to the total integrated noise at the output is:

$$A = \int_0^\infty |H(j2\pi f)|^2 S(v_{noise}) df$$

$$= \int_0^\infty |H(j2\pi f)|^2 df \bullet S(v_{noise}) \quad (1)$$

The second equality sign applies when $S(v_{noise})$ is independent of frequency, that is, when the noise is white. $S(v_{noise})$ is the noise voltage spectral density, that is, the number of squared volts per hertz. A is the part of the mean-square noise voltage (square of the RMS noise voltage) at the output which is caused by v_{noise}. $|H(j2\pi f)|$ is the magnitude of the RIAA and A filter transfer at frequency f.

The contribution of i_{noise} equals:

$$B = \int_0^\infty |H(j2\pi f)|^2 (4\pi^2 f^2 L^2 + R^2) S(i_{noise}) df$$

$$= \int_0^\infty |H(j2\pi f)|^2 (4\pi^2 f^2 L^2 + R^2) df \bullet S(i_{noise}) \quad (2)$$

assuming the current noise to be white. $S(i_{noise})$ is the noise-current spectral density, that is, the number of squared amperes per hertz. B is the part of the mean-square noise voltage (square of the RMS noise voltage) at the output which is caused by i_{noise}.

Hence,

$$\frac{B}{A} = \frac{\int_0^\infty |H(j2\pi f)|^2 (4\pi^2 f^2 L^2 + R^2) df}{\int_0^\infty |H(j2\pi f)|^2 df} \bullet \frac{S(i_{noise})}{S(v_{noise})}$$

$$= \left(\frac{\int_0^\infty |H(j2\pi f)|^2 f^2 df}{\int_0^\infty |H(j2\pi f)|^2 df} \bullet 4\pi^2 L^2 + R^2 \right) \frac{S(i_{noise})}{S(v_{noise})} \quad (3)$$

At a single frequency $f = f_1$, the ratio of the contribution of i_{noise} to the contribution of v_{noise} equals:

$$\frac{D}{C} = (4\pi^2 f_1^2 L^2 + R^2) \bullet \frac{S(i_{noise})}{S(v_{noise})} \quad (4)$$

Here, D is the contribution of i_{noise} and C is the contribution of v_{noise} to the total noise density at f_1.

Equations (3) and (4) are equal when:

$$f_1 = \pm \sqrt{\frac{\int_0^\infty |H(j2\pi f)|^2 f^2 df}{\int_0^\infty |H(j2\pi f)|^2 df}} \quad (5)$$

Equation (5) can be evaluated numerically. This can easily be done with PSpice and Probe, using an analogue behavioural model of an idealised RIAA amplifier with an A-weighting filter behind it.

You can first simulate the AC transfer and use Probe to evaluate the denominator integral. Then put a differentiator in the circuit with unity transfer at 1Hz, and do the same for the numerator. The result is $f_1 \approx 3852$Hz. I used the IEC modified RIAA response, which has an extra roll-off below 20Hz, but this should make very little difference.

This proves that the total integrated weighted noise is a constant factor times the noise in a 1Hz bandwidth around 3852Hz. 'A constant factor' means that this factor is independent of $S(v_{noise})$ and $S(i_{noise})$. Hence, the combination of $S(v_{noise})$ and $S(i_{noise})$ which is optimal at 3852Hz is also optimal for the integrated weighted noise.

In other words, optimising the integrated A- and IEC modified RIAA-weighted noise is the same as optimising the noise at 3852Hz. When R=1kΩ and L=494mH, which are typical values for a MM cartridge, the impedance is about 12kΩ at this frequency. We will therefore assume a 12kΩ cartridge impedance in the remainder of this article, unless otherwise noted.

1/*f*-noise and noise densities relationships

The calculations above apply only to white noise, while many active devices produce substantial 1/*f*-noise in the audio frequency band. To what white noise level does a given amount of 1/*f*-noise correspond? Further, if you measure the total A-weighted noise level, how can you compare the results to the theoretical noise densities?

Calculations which are similar to those in the previous section show that 1/*f* voltage noise has the same influence on the total RIAA/IEC- and A-weighted noise level as white voltage noise with the same density at 1169Hz. In other words, if you determine the noise voltage density at 1169Hz, it doesn't matter what part is white and what part is 1/*f* noise.

The situation is a bit more complicated for 1/*f* current noise. If the cartridge impedance were constant, the same rule that applies to 1/*f* voltage noise would also apply to 1/*f* current noise. If the magnitude of the cartridge impedance were exactly proportional to frequency, a similar rule would apply, but at a frequency of 5398Hz instead of 1169Hz.

The frequency dependence of the magnitude of the impedance of a real cartridge is something in between these two cases, closer to being proportional to frequency than to being constant. Hence, the current noise density should be determined at a frequency of about 5kHz.

In order to be able to compare measured A- and RIAA/IEC-weighted RMS noise levels to theoretical noise voltage and current densities, one needs to know the noise bandwidth of the cascade of a RIAA/IEC and an A filter. Simulations show that this bandwidth is 3219Hz referred to the transfer at 1kHz. In other words, if one divides the A-weighted RMS output noise voltage by the voltage gain at 1kHz and by the square root of 3219Hz, one finds the average equivalent input number of volts per root Hz.

In order to find the equivalent input noise voltage and the equivalent input noise current, one can measure the output noise voltage twice, once with short-circuited and once with open input.

The equivalent input noise voltage per root hertz can then be calculated directly from the measurement with shorted input, as explained above. As long as the input capacitance is so small that the input impedance can be assumed to be 47kΩ over the entire audio band, the equivalent input noise current is simply the input noise voltage measured with open input, divided by 47kΩ.

'Electronic cooling'

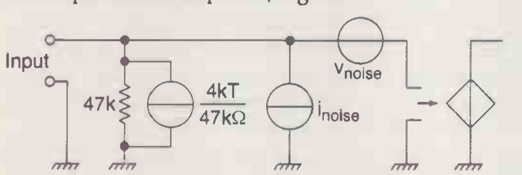
The thermal noise of a resistor can be modelled with a noise voltage source with a spectral density of 4kTR (RMS value $\sqrt{4kTR\Delta f}$ over a bandwidth Δf) in series with a noiseless resistor (the Thévenin equivalent). In this equation, T is the absolute temperature and k is

Boltzmann's constant (1.38065×10^{-23}J/K).

Alternatively, the noise can be modelled with a noise current source with a spectral density $4kT/R$ (RMS value $\sqrt{(4kT/R)\Delta f}$ over a bandwidth Δf) in parallel with a noiseless resistor (the Norton equivalent). Both models behave in exactly the same way at their terminals. A designer can choose whichever model is the most convenient for his or her calculations.

Modelling the 47kΩ resistor with its Norton equivalent, it becomes clear that this resistor contributes $4kT/47k\Omega\approx3.445\times10^{-25}A^2/\sqrt{Hz}$, or 0.5869pA/√Hz, to the equivalent input noise current, **Fig. 2**. With a typical effective cartridge impedance of 12kΩ, 0.5869pA/√Hz has the same influence as 7.043nV/√Hz, which is substantial compared to a carefully optimised input stage.

Fortunately, there's a simple solution. Build an accurate and low-noise inverting amplifier with a high input impedance and a voltage gain of –K times. Then connect a resistor with a value of (K+1)×47kΩ between the input and output of this amplifier, **Fig. 3**.



Note: i_{noise} is noise from the rest of the RIAA amplifier.

Fig. 2. RIAA amplifier with a noisy 47kΩ resistor.

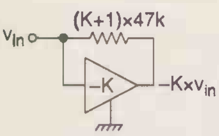


Fig. 3. Low-noise active input resistance.

The voltage across the resistor is K+1 times the input voltage and its value is K+1 times 47kΩ. Hence, the current flowing through the resistor equals the current that would flow through a 47kΩ resistor connected between the input and ground.

If the noise of the inverting amplifier is negligible though, the noise current spectral density is K+1 times as small. This technique is known as the active input impedance technique or as 'electronic cooling', as it has the same influence on the thermal noise as cooling the resistor down to a K+1 times lower absolute temperature.

Both terms can cause confusion, the first because it has a completely different meaning in network theory and the second because the resistor is not really cooled down.

Discrete RIAA amplifier

Figure 4 depicts a circuit in which the inverting amplification and the RIAA equalised amplification are combined[2]. Block N_1 is a two-port with very high voltage, current, transimpedance and transadmittance gain.

Because of the negative feedback around this high-gain block, the voltage v_{in} at the amplifier input also occurs at R_1. As a result, a current $i=v_{in}/R_1$ flows through R_1, through the RIAA equalising network R_2, C_2, R_3, C_3 and through the output port of the amplifying block. Resistors R_1 and R_4 are normally made much smaller than R_5. Hence, the voltage on the upper side of R_4 becomes approximately,

$$-R_4 \times i = -\frac{R_4}{R_1} v_{in}$$

realising a negative voltage gain $-R_4/R_1$.

A more accurate analysis shows that if the amplifying block has infinite gain and if the current flowing into one output pin exactly equals the current coming out of the other, the input resistance equals:

$$R_{in} = \frac{R_4 + R_5}{\frac{R_4}{R_1} + 1} \tag{6}$$

Unfortunately, the input impedance changes when a load

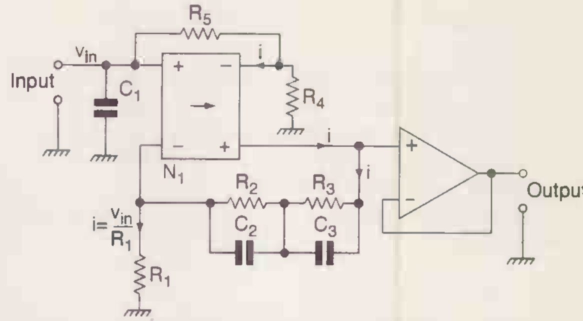


Fig. 4. RIAA equalised amplifier with active input resistance according to reference 2

is connected to the output of the two-port. That is why there is a voltage follower connected behind the RIAA equalising stage.

A discrete implementation is shown in **Fig. 5**. I use it as a part of a wholly discrete preamplifier with ±14V regulated power supply voltages, but it should also work at the more usual ±15V. It has an IEC modified RIAA response, that is, it includes a first-order roll-off below 20Hz.

If desired, this corner frequency can be lowered by increasing the values of the bipolar electrolytic capacitors, although lowering it too much can make the start-up time rather long. The input resistance increases from 47kΩ to 10MΩ for subsonic frequencies. This is no problem, as the purpose of the 47kΩ input resistance is to damp a resonance in the 10kHz to 20kHz frequency range.

The voltage follower at the output is a simple emitter follower. The high-gain block is made of two differential pairs. For reasons of noise optimisation, the input pair is asymmetrical, consisting of a JFET and a bipolar transistor.

It can be shown[1] that the optimal bias current for minimum noise is,

$$I_{C,opt} \approx \frac{\frac{kT}{q}\cdot\sqrt{h_{FE}}}{|Z_S + r_b|} \tag{7}$$

when a bipolar transistor is driven from a source impedance Z_S, assuming that $1/f$ noise is negligible and that the frequency lies well below $f_T/\sqrt{h_{FE}}$. In this equation, q is the electron charge (1.6022×10^{-19}C), h_{FE} the DC current gain factor and r_b the parasitic base resistance of the transistor. The corresponding transconductance equals,

$$g_{m,opt} \approx \frac{\sqrt{h_{FE}}}{|Z_S + r_b|} \tag{8}$$

and the total contribution to the noise is,

$$S(v_{n,bip,inc.currentnoise}) \approx 4kT\left(\frac{|Z_s + r_b|}{\sqrt{h_{FE}}} + r_b\right) \tag{9}$$

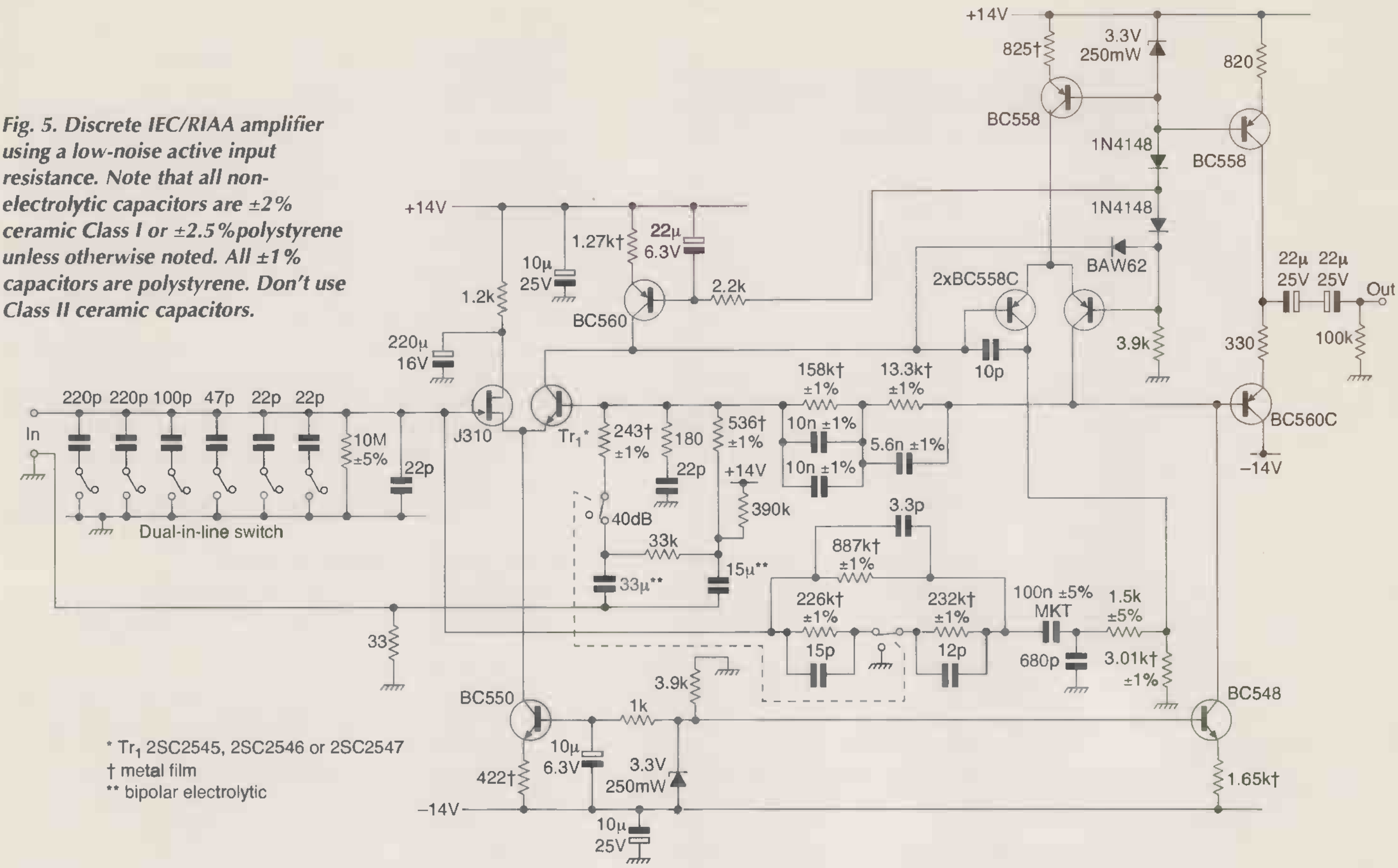
This includes the current noise term, which has been transformed into an equivalent noise voltage across the given source impedance Z_S. The 2SC2545, 2SC2546 and 2SC2547, which are good low-noise low-frequency transistors, have $r_b\approx14\Omega$ and $h_{FE}\approx600$.

The input device is driven from a source impedance of about 12kΩ, which is largely reactive. If a 2SC2545 were used here, the resulting values would be:

$I_{C,opt}\approx51.56\mu A$
$g_{m,opt}\approx2.041mS$
$S(v_{n,bip,inc.currentnoise})\approx8.158\times10^{-18}V^2/Hz$ or 2.856nV/√Hz

According to the graphs in reference 3, a high

Fig. 5. Discrete IEC/RIAA amplifier using a low-noise active input resistance. Note that all non-electrolytic capacitors are ±2% ceramic Class I or ±2.5%polystyrene unless otherwise noted. All ±1% capacitors are polystyrene. Don't use Class II ceramic capacitors.



transconductance JFET such as the J310 biased at several milliamperes should be able to outperform this bipolar transistor, having a noise of about 2nV/√Hz at 1169Hz (for a large part $1/f$).

At the same time, the JFET provides more transconductance and a better high-frequency behaviour than a 2SC2545 biased at about 52μA. The J310 in Fig. 5 is biased at about 3.3mA. A higher bias current (10mA to 15mA) would be somewhat better.

The situation is different for the transistor driven from the feedback network. In the 40dB gain position, its base is driven from a source impedance of about 165Ω, while its emitter 'sees' the output resistance of the source follower, about 145Ω. This results in a 1.913mA optimal bias current and a 0.6637nV/√Hz total noise contribution. This cannot be outperformed by a single J310 – although it should still be possible with a group of paralleled J310s.

The 887kΩ, 226kΩ and 232kΩ feedback resistors have been shunted with capacitors. This has been done to swamp the few tenths of a picofarad of capacitance associated with a discrete resistor.

The corner frequency lies at about 52kHz. The 680pF capacitor attenuates the voltage on the feedback resistors above 52kHz, eliminating the Miller effect which would otherwise multiply the influence of the capacitances across the feedback resistors.

Gain at 1kHz can be switched between about 30 and 40dB to account for different cartridge sensitivities. The input resistance is designed to be about 3% too low in the 30dB position. It should damp the resonance of the cartridge inductance and load capacitance a bit more than normal, resulting in a small high-frequency loss. This should more or less compensate for the error of +0.37dB at 20kHz caused by the fact that the gain drops to unity rather than zero at high frequencies.

At low frequencies, the maximum output signal level without clipping is about 16.7V peak to peak, or 5.9V RMS in the case of a sine wave. The tail current of the output stage and the feedback network impedance determine the headroom at high signal frequencies.

The output stage can deliver about 1.33mA peak to the feedback network. In the 40dB position, the resistance between the base of the 2SC2545 and ground is about 167Ω, corresponding to 222mV maximum peak input voltage at high audio frequencies. This changes into 711mV peak in the 30dB position.

To put these figures in perspective: my record player produces peak signal levels of 30mV to 35mV when it plays popular music records with a large high-frequency content, but according to reference 4, peak signal levels up to 100mV can occur at the output of sensitive cartridges playing loud records.

Measured results

Several measurements have been made using an HM1505 oscilloscope with cursors and a ±3% vertical accuracy specification. A 1:1 probe has been used whenever this could be expected to give the most accurate results.

When possible, relative measurements have been made using the same oscilloscope channel to minimise inaccuracies. In most cases, the test signals have been generated with a CD player playing a test CD and with 0.01% accurate resistors to attenuate the signal or to convert it into a current (for the impedance measurements). For the noise measurements, a separate 100 times amplifier and A-weighting filter have been used.

Response when driven from a low source impedance has been measured at 15 different frequencies and compared to the theoretical IEC modified RIAA response. The voltage gain at 1kHz has been measured to be 30.17dB in the 30dB-mode and 40.27dB in the 40dB-mode.

Compared to 1kHz, the response was accurate to within

Table 1. Equivalent parallel resistances at various frequencies calculated from measurement of input impedance magnitude and phase.

Frequency	R_{in}, 30dB	R_{in}, 40dB
1kHz	46.64kΩ	48.85kΩ
10kHz	45.54kΩ	46.99kΩ
15kHz	45.5kΩ	46.76kΩ
20kHz	45.79kΩ	46.3kΩ

±0.1dB for frequencies between 30Hz and 20kHz in the 40dB-mode. With 30dB nominal gain, the response was accurate to within ±0.14dB between 30Hz and 10kHz, and at 20kHz, the measured error was +0.24dB. Theoretically, this should be +0.37dB. At 20Hz, the error was +0.18dB in the 30dB and +0.25dB in the 40dB position, probably due to the relatively large tolerances of the electrolytic capacitors.

The input impedance magnitude and phase have been measured at various frequencies and the equivalent parallel resistance has been calculated from the measurement results, see the **Table 1**.

Input capacitance is in the order of 30pF at the minimum input capacitance setting – i.e. just the RIAA amplifier, not including the probe capacitance.

A-weighted noise has been measured with a shorted input and with an open input with 30dB and 40dB mid-band gain, in all cases with the minimum input capacitance setting.

The measurements can be about 20% off, because I had to estimate the RMS noise from the quasi peak to peak values measured with the oscilloscope. Besides, the noise current (open input) measurement has been affected by interference from the mains.

In the 30dB mode, the A-weighted noise at the RIAA amplifier output was – fortunately – 27.67μV RMS with open input and 9.85μV RMS with shorted input. This corresponds to 5.381nV/√Hz and 0.3216pA/√Hz at the input.

In the 40dB mode, the results were 26.83μV A-weighted RMS at the output with shorted input and 43.67μV RMS at the output with open input, corresponding to 4.729nV/√Hz and 0.1638pA/√Hz at the input.

All measured results are reasonably close to the expected values, except for the equivalent input noise voltage, which is higher than expected. With the source of the J310 AC shorted to the input ground, it drops to somewhere between 2.2 and 2.9nV/√Hz in the 40dB gain mode.

Most probably, the unexpectedly high noise voltage is due to the $1/f$ noise of the J310 being higher than expected.

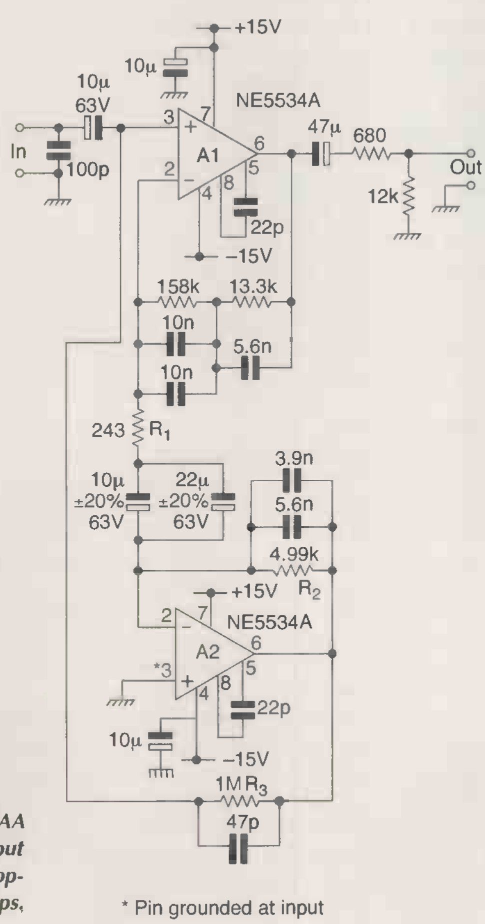


Fig. 6. IEC modified RIAA amplifier with an active input resistance made from op-amps.

RIAA amplifier using op-amps

The network in Fig. 4 is not very suitable for implementation with op-amps. After all, an op-amp does not have a pin carrying a current which is equal but opposite to the output (signal) current.

Although it is possible to use floating power supplies to solve this problem[5], it is much simpler to use a different topology, **Fig. 6**. A $-R_2/R_1$ times amplified version of the input signal occurs at the output of A2, resulting in an input resistance of $R_3/(1+R_2/R_1)$. I have only subjected this circuit to a functional test, not to measurements. It seems to work fine.

Noise can be slightly reduced by replacing A2 with an ultra-low noise op-amp such as the LT1028, and adjusting its frequency compensation. The LT1028 may need a compensation capacitor between pin 5 and its output when the output signal is fully fed back to the inverting input at high frequencies. The data sheet is not very clear about the required value[6].

On the other hand, replacing A1 with an ultra low noise op-amp would make the noise much worse. Ultra-low noise op-amps usually have a bipolar input stage biased at a high bias current. This results in a low input-noise voltage density, but also in a high input noise current density.

To make matters worse, they often feature a base current compensation circuit which injects large fully correlated noise currents in the inverting and non-inverting input terminals.

In the case of the LT1028, the input noise voltage is specified as 0.9nV/√Hz, with a noise current of 1pA/√Hz typical at 1kHz[6]. Carefully reading the small print and the applications information shows that the noise current value only applies when both inputs are driven from exactly the same impedance.

In the case of an RIAA amplifier, this would mean that a dummy cartridge would have to be included in the feedback network. Fortunately, there is also a graph of the total noise versus unmatched source impedance, which shows that the actual input noise current is about 3.25pA/√Hz.

With a source impedance of 12kΩ, 3.25pA/√Hz has the same influence as 39nV/√Hz, showing that even a 741 would be better suited for a moving-magnet RIAA amplifier.

How much improvement can be obtained?

As a best-case approximation for the improvement due to using an active input resistance, I'll compare the sum of

Table 2. Impedance for a Shure V15 III cartridge and for an LR network.

f	$\lvert Z_{measured} \rvert$	$\lvert Z_{theory} \rvert$	$\varphi_{measured}$	φ_{theory}	R*	G*	F*	F_{theory}
2kHz	6kΩ	5934Ω	72°	76.96°	1.85kΩ	51.5μS	1.5dB	1.93dB
5kHz	14kΩ	14513Ω	72°	84.71°	4.33kΩ	22.1μS	2.93dB	6.38dB
7kHz	20kΩ	20276Ω	70°	86.21°	6.84kΩ	17.1μS	3.51dB	8.77dB
10kHz	28kΩ	28934Ω	66°	87.35°	11.4kΩ	14.5μS	3.92dB	11.55dB
14kHz	37kΩ	40486Ω	61°	88.1°	17.9kΩ	13.1μS	4.19dB	14.32dB
20kHz	50kΩ	57821Ω	53°	88.67°	30.1kΩ	12μS	4.42dB	17.33dB

*calculated from measured results

the cartridge noise and the amplifier noise for two cases.

The first case is with a RIAA amplifier with the 47kΩ resistor as its only noise source, the second with a RIAA amplifier which does not generate any noise at all.

In both cases, assume that the noise coming out of the cartridge is only thermal noise. This implies that the record player is not playing a record, otherwise there would also be record noise and rumble.

If the cartridge had a frequency-independent effective series resistance, calculating the difference in A-weighted integrated noise level would be easy. For the example of a cartridge with a 1kΩ effective series resistance and 12kΩ impedance at 3852Hz:

Cartridge noise: 4.024nV/√Hz
Noise due to 47kΩ: 7.043nV/√Hz
Total of the cartridge and resistor: 8.111nV/√Hz
Difference between total and cartridge noise: 6.089dB.

However, the effective series resistance of a real cartridge rises with frequency. Richard Visée[7] measured the impedance of a Shure V15 III cartridge using an HP4194A impedance gain and phase analyser. The DC resistance and the inductance were 1.3388kΩ and 460mH, respectively. See **Fig. 7** and **Table 2**.

In Table 2, the value $|Z_{measured}|$ is the magnitude of the measured impedance, $\varphi_{measured}$ is its phase. R is the effective series resistance at a frequency f, while G is the effective parallel conductance. F is the ratio of the thermal noise of the cartridge and a 47kΩ load to the noise of the cartridge alone at a frequency f, expressed in dB.

For comparison, $|Z|$, φ and F have also been calculated

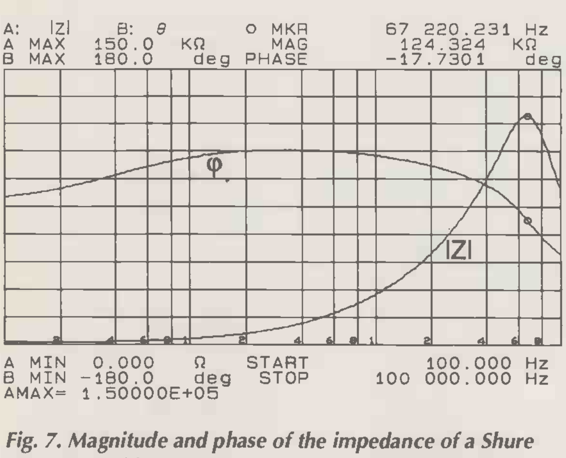


Fig. 7. Magnitude and phase of the impedance of a Shure V15 III cartridge. Frequency scale is 100Hz to 100kHz while the magnitude scale is 0Ω to 150kΩ and the phase scale is –180° to +180°. The phased hardly exceeds 72° at any frequency.

for a theoretical cartridge with 460mH of inductance and a constant 1.3388kΩ series resistance. Obviously, only the magnitude of the impedance can be reasonably modelled with a simple *LR* series network.

It is clear that the improvement averaged over the audio band lies somewhere around 3dB, rather than the 6.089dB calculated before.

Similar results have been found for a Marantz cartridge by measuring its A and IEC/RIAA weighted noise using the amplifier in Fig, 5. After correcting for the amplifier noise, this relatively low-impedance cartridge with 349mH inductance and 350Ω DC resistance turned out to have a noise equivalent to roughly 2215Ω weighted average effective series resistance. The noise of the cartridge and a passive 47kΩ resistor together would have been 2.27dB higher than the thermal noise of the cartridge itself.

Whether about 3dB best-case improvement is worth the effort depends on your point of view. It will not be immediately noticed by the average user. But then again, it is common practice to try to keep the distortion, frequency response errors and noise levels introduced by audio amplifiers well below the errors introduced in other parts of the audio signal chain or well below the threshold of audibility.

In this way, the final quality is determined by the parts of the audio chain which are the hardest to get right: microphones, loudspeakers, cartridges, the acoustics of the listening room and the bad habit of some recording and broadcasting people to use unnecessarily high amounts of dynamic compression. ■

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Simple formulae for skin effect

As Leslie Green explains, although the exact formula for the AC resistance of an isolated piece of wire was developed over a century ago, a modern graduate electrical engineer is not equipped to calculate the resistance of a piece of copper wire to within a factor of three! This gap is filled by some new formulae.

Text books have mentioned the idea that alternating current preferentially flows on the outer surface of a conductor since at least as early as James Clerk Maxwell's 1864 essay 'A Dynamical Theory of the Electromagnetic Field'.

Joseph Henry provided early qualitative confirmation of this preferential current flow in 1876. Lord Kelvin eventually beat the mathematics into submission, solving the second order partial differential equation in terms of a zero order Bessel function with a complex argument.

It turns out that a Bessel function of complex argument is itself too complicated to immediately evaluate, so Kelvin had to invent a new pair of functions to deal with the problem. The Kelvin-Bessel functions *ber* and *bei*, meaning the real and imaginary parts of a Bessel function, were the result.

A complete expression for the AC resistance of an isolated round wire could then be given in terms of *ber* and *bei*, and

Table 1. The formulae, together with their error limits and range of validity.

Formula	Region	Error
$\delta = \sqrt{\dfrac{\rho}{\pi f \mu}}$	all	— none — definition of skin depth
$R_{AC} = R_{DC} = \dfrac{4\rho \cdot L}{\pi \cdot d^2}$	$\left(\frac{d}{\delta}\right) \le 0.93$	<0.1% always low
$R_{AC} = R_{DC} = \dfrac{4\rho \cdot L}{\pi \cdot d^2}$	$\left(\frac{d}{\delta}\right) \le 1.66$	<1% always low
$R_{AC} = R_{DC} \cdot \left[1 + \frac{1}{772} \cdot \left(\frac{d}{\delta}\right)^4 \left(1 - \frac{1}{1079} \cdot \left(\frac{d}{\delta}\right)^4 \left[1 - \frac{1}{1687} \cdot \left(\frac{d}{\delta}\right)^4\right]\right)\right]$	$\left(\frac{d}{\delta}\right) \le 4.7$	<±0.04%
$R_{AC} = \frac{R_{DC}}{4} \cdot \left[1 + \left(\frac{d}{\delta}\right) + \ln\left[1 + \left(\frac{\delta}{d}\right)\right] \cdot \left(0.77 - 15 \cdot \exp\left[-0.645\left(\frac{d}{\delta}\right)\right]\right)\right]$	$\left(\frac{d}{\delta}\right) > 4.7$	< ± 0.25%
Butterworth's approximation $R_{AC} = \frac{R_{DC}}{4}\left[1 + \left(\frac{d}{\delta}\right)\right]$	$\left(\frac{d}{\delta}\right) \ge 3.8$	< 2% always low
$R_{AC} = \dfrac{\rho \cdot L}{\pi \cdot (d - 0.955 \times \delta) \cdot \delta}$	$\left(\frac{d}{\delta}\right) \ge 6.8$	< ± 0.2%
$R_{AC} = \dfrac{\rho \cdot L}{\pi \cdot (d - \delta) \cdot \delta}$	$\left(\frac{d}{\delta}\right) \ge 6.8$	< 1% always high
Butterworth's approximation $R_{AC} = \frac{R_{DC}}{4}\left[1 + \left(\frac{d}{\delta}\right)\right]$	$\left(\frac{d}{\delta}\right) \ge 8.2$	< 1% always low
$R_{AC} = \frac{R_{DC}}{4} \cdot \left[1 + \left(\frac{d}{\delta}\right) + 0.77 \cdot \ln\left[1 + \left(\frac{\delta}{d}\right)\right]\right]$	$\left(\frac{d}{\delta}\right) > 8.6$	< ± 0.04%

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their first derivatives, *ber´* and *bei´*.

$$R_{AC} = R_{DC} \times \frac{x}{2} \cdot \frac{ber(x)\,bei'(x) - bei(x)\,ber'(x)}{\left[ber'(x)\right]^2 + \left[bei'(x)\right]^2}$$

The x in the equation contains information about the operating frequency, the wire diameter, the conductivity of the material and the permeability of the material.

'Approximations'

Now Kelvin's formula was useful because the *ber* and *bei* functions, and their derivatives, were tabulated. Furthermore, the overall function was tabulated and graphed so engineers could look up the values and get immediate accurate answers.

Unfortunately all of these texts have been out of print for more than 50 years and are no longer readily available. Additionally, these old books necessarily use an old system of units; conversion from which allows plenty of opportunity for errors.

Kelvin's formula contains mathematics which is too advanced for undergraduate courses, which means that it is not taught. Instead, the graduating electrical engineer is equipped with two approximations, one for 'low frequency' and one for 'high frequency'.

At DC the resistance is given exactly by:

$$R_{DC} = \frac{4\rho \cdot L}{\pi \cdot d^2}$$

where ρ is the electrical resistivity, L is the length and d is the diameter.

The AC resistance is evaluated using a quantity called the *skin depth*. This skin depth is an *equivalent* depth in the material, where, for the purposes of calculation, one can assume that the current is evenly distributed throughout this thin surface layer, being zero elsewhere. The skin depth is given by:

$$\delta = \sqrt{\frac{\rho}{\pi f \mu}}$$

where ρ is the electrical resistivity, f is the frequency and μ is the (absolute) permeability.

The 'high frequency' resistance of a piece of wire is then given as:

$$R_{HF} = \frac{\rho \cdot L}{\pi \cdot d \cdot \delta}$$

where it is stated that this approximation is only valid when the diameter is 'much greater; than the skin depth.

Suppose you wish to calculate the resistance of a 10mm diameter copper bus bar for use in a power system at 50Hz. Which formula should you use and what is the error in the approximation? The first step must be to calculate the skin depth. For copper we have 17nΩ/m resistivity and a skin depth at 50Hz of 9.3mm.

The DC resistance works out as 0.233Ω/km. The 'HF' resistance works out as 0.0582Ω/km, which is less than the DC value and therefore evidently completely wrong. Of course you could argue with my interpretation of the HF formula. I evaluated the effective cross-sectional area as the circumference times the skin depth. Perhaps it would have been more accurate to use the 'true' cross-sectional current path as:

$$\frac{\pi}{4} \cdot d^2 - \frac{\pi}{4} \cdot (d - 2\delta)^2 = \pi \cdot \delta(d - \delta)$$

giving,

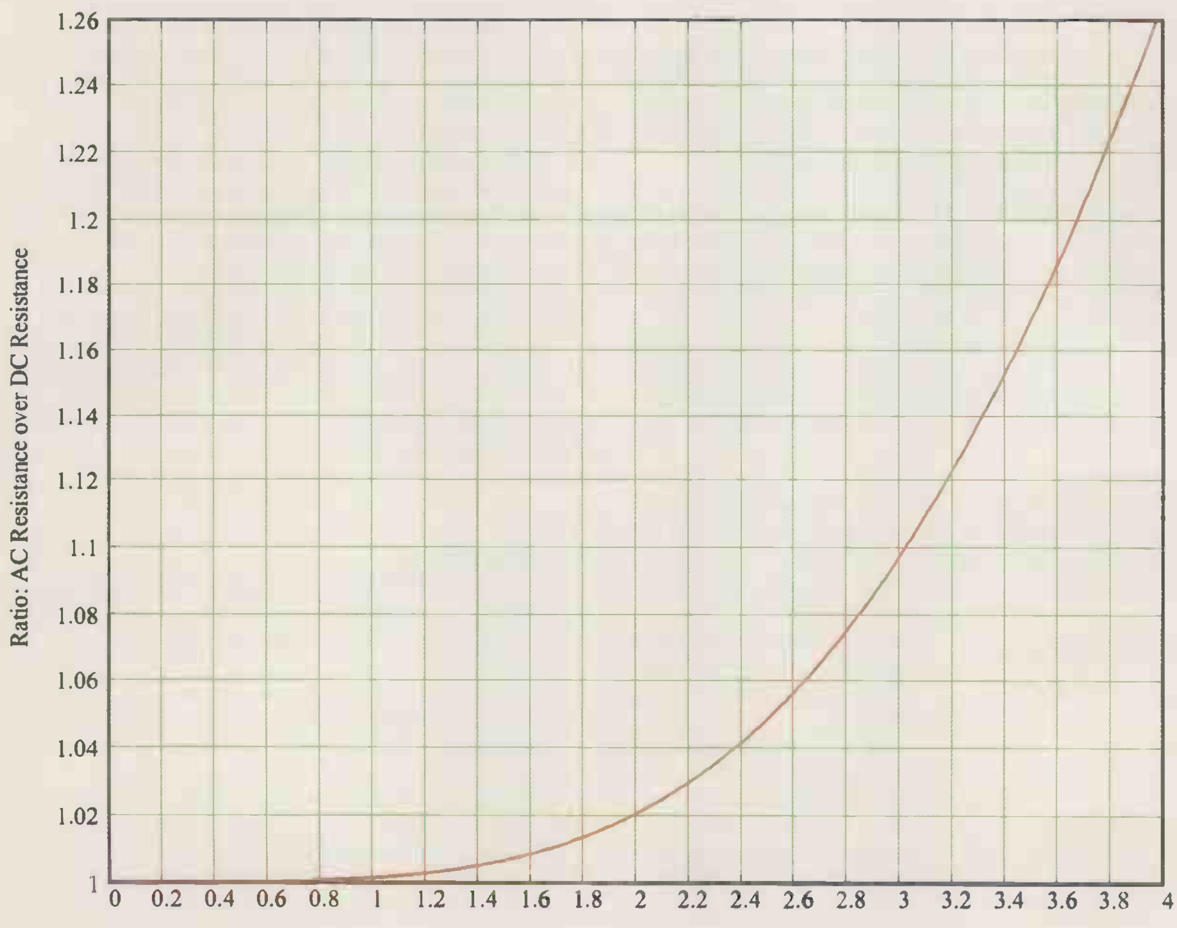


Fig. 1. Graph of the transition region between the low frequency and high frequency formulae.

$$R_{HF} = \frac{\rho \cdot L}{\pi \cdot (d - \delta) \cdot \delta}$$

This revised formula gives the HF resistance as 0.831Ω/km. Among these three 'approximations' we have a greater than 14:1 spread, and we have no way of knowing which one is the best approximation!

Accuracy

No engineering formula or approximation is useful unless its range of validity and error limits are given. Since we have an exact equation for this problem, it is only necessary to evaluate the function numerically and compare the approximations with the exact answer.

The formulae, error limits, and range of validity are best given in a table for ease of comparison and use. See **Table 1**. Butterworth's approximation, as given in the table, was published in 1921. The two complicated approximations, which between them cover the entire range of possible values, are new.

While the title of this article relates to 'simple' formulae, these two new formulae still qualify on the grounds that they use standard calculator functions, which are also found in mathematical libraries for compilers. This ease of calculation is in stark contrast to the Kelvin-Bessel function formula, which uses functions that can only be evaluated as power series solutions, and which have a nasty habit of causing overflows due to the large factorial numbers used.

If you happen across a table of values of the AC/DC resistance ratio of a cylindrical conductor it will probably be given as function of the variable x, as given in Kelvin's formula. The relationship to the present usage is:

$$x\sqrt{2} = \left(\frac{d}{\delta}\right)$$

It is very easy to make a typing mistake when applying a formula, so a few exact 'check values' of the function are useful. These are provided in **Table 2**. A graph of the transition region between the low frequency and high frequency formulae is also useful, see **Fig. 1**.

I have also included a short table of resistivity and skin depth, **Table 3**. The skin depth is given for 1Hz, so for any other frequency just divide the figure in the table by the square root of the actual frequency in Hertz. Note that all resistivity values are very approximate, since the purity of the specimen affects the resistivity.

The skin depth for iron is considerably more uncertain than for the rest however. The reason is that the relative permeability of iron changes by orders of magnitude according to its purity and its crystal structure.

Table 2. It's easy to make a typing mistake when applying a formula, so a few exact 'check values' of the function are useful.

0	1.0000
1	1.0013
2	1.0205
5	1.5050
10	2.7681
20	5.2593

Table 3. resistivity versus skin depth.

Material	Resistivity	Skin depth @ 1Hz
Iron	97	10
Copper	17	65.6
Aluminium	26	81.2
Brass	39	99.4
Tin	115	171
Lead	220	236

Errors

It is important to note that when the approximation formulae state errors, these errors are mathematical values derived using the exact formula. Consider the case for the new formula giving errors of less than ±0.04% for $d/\delta \leq 4.7$. The error in the approximation is so low that the approximation can be considered as exact for all practical purposes.

In a real situation, the tolerance on the resistivity of the conductor is likely to be of the order of several percent at least. Furthermore, the loss resistance will increase when other conductors are brought up close to the ideal isolated conductor. This increased loss is attributed to further current crowding within the conductor, such additional current crowding being known as the *proximity effect*.

Text books love to simplify the theory of the skin effect in order to make it 'understandable' to mathematically challenged engineers. The problem with these simple explanations is that they can be grossly misleading.

One popular 'theory' for the skin effect says that current does not flow down the middle of a conductor at high frequency because of the increased inductance at the centre of the conductor. This theory is amplified by drawings showing concentric circles of flux around the conductor.

It is then evident that a current filament in the centre of the conductor links to more of these flux lines than a current filament at the perimeter.

It does not take much thought to shoot this 'increased inductance' theory down in flames. The internal inductance of a piece of wire, that is the inductance due to the flux within the wire, is a very small proportion of the total inductance.

The formula is given in all basic electromagnetics text books. And yet this miserably small difference in inductance is supposed to account for the *exponential* drop in current density which occurs in practice. Furthermore, an increased inductance would have the effect of merely reducing the current density.

In reality, any good electromagnetics text book shows that the current density actually reverses a small distance from the surface; that's right, there is a subsurface current flowing backwards! Some simple theories allow teachers to silence their student's questions, but fill the students with incorrect 'facts' that subsequently need to be unlearned.

It can only be hoped that future teachers present the facts, state that the maths is rather hard, and present a graph rather than a formula. Anyone can understand a graph. Showing that there is something more to learn is, to my way of thinking, better than stating that everything is exact and complete.

Earlier in this article the problem was posed as to the exact resistance of a copper bus bar 10mm in diameter at 50Hz. The ratio d/δ is 1.075, so it is clear from Fig. 1 that the increase from the DC resistance is certainly considerably less than one percent.

Using the appropriate new formula, the AC/DC resistance ratio is found to be 1.0017, so the resistance effectively remains unchanged at 0.233Ω/km. Of course 10mm copper bus bars are probably of little interest to you, but I trust that you have realised by now that the 3:1 uncertainty I mentioned when using the previous LF/HF approximations also applies for 1mm wires at 5kHz. ■

Further reading

Circular No.74 of the Bureau of Standards: "Radio Instruments and Measurements", 1918.

Principles of Radio Communication by J.H. Morecroft (2nd edition), 1927.

Wireless: A treatise on the theory and practice of high-frequency electric signalling by L.B. Turner, 1931.

Radio Frequency Measurements by E.B. Moullin, (2nd edition), 1931.

Radio Engineer's Handbook by F.E. Terman, 1943.

A useful list of references and formulas for other situations is provided in:

H.A Wheeler, 'Formulas for the Skin Effect', in *Proceedings of the Institute of Radio Engineers*, 30 (Sept 1942), pp. 412-424.

Measuring Amplifier Distortions

Cyril Bateman continues to describe his improved real-time hardware distortion measuring system

Pre-amplifier and power amplifier distortions, to a maximum of five volts RMS, can be measured directly using my Real Time second and third harmonic distortion analyser described in the July issue, together with my 1ppm low distortion 1kHz generator, buffer amplifier and notch filter preamplifier using a soundcard and FFT software or my ADC-100 AD converter[1]. The addition of a simple resistive input attenuator in front of the notch filter, quickly extends this measurement capability to encompass even the highest output power amplifiers.

I seldom need to measure at powers exceeding 100 watts into an eight Ohm load so have assembled a simple, easily replicated dummy load complete with a built in attenuator matching the switchable input voltage ranges of my LC960[2] cased, Standalone Distortion Meter, which was described in my last article[3].

Apart from being able to handle the amplifier output power without overheating, a dummy load should be totally resistive, exhibiting no significant self-inductance at least for audio frequencies. Special non-inductive resistors are available but these are far too expensive for use by many enthusiasts, so I decided to examine whether a less expensive approach, based on low cost commercial 'aluminium clad' wirewound resistors would suffice. I already had a number of 25 watt 8.2Ω resistors and a 2°C/W heatsink. Mounting two series pairs of 8.2Ω resistors in parallel would reduce their self inductance and mounting on this heatsink would permit measurement to 50 watts which could then be extended up to my 100 watts maximum when needed by using a small 'snail' cooling fan to dissipate excess heat.

The attenuator included in this dummy load would easily connect to the 'V – in' and 'V – low' BNC sockets on my LC960 cased distortion meter, the 'I – out' and 'I – low' sockets providing the test signal needed by the amplifier under test. Using an 8Ω dummy test load, 100 watts develops some 28.3 volts, 50 watts requires 20 volts and 25 watts just 14.14 volts. Using a simple fixed 7.075:1 attenuator and extending the 50 watts to just 56 watts (21.25 volts) reduces these load voltages to 4, 3 and 2 volts respectively, exactly matching my switchable test levels. Inclusion of one additional low value attenuator resistor would also allow measurements to be made at 50 watts when needed.

Exploratory measurements using one of my elderly laboratory workhorse Maplin Mosfet 100W amplifier modules confirmed this approach would be more than satisfactory, so I proceeded to build a permanent attenuator/dummy test load for routine use with my LC960 cased standalone distortion meter, described in my last article[3]. **Fig. 1.**

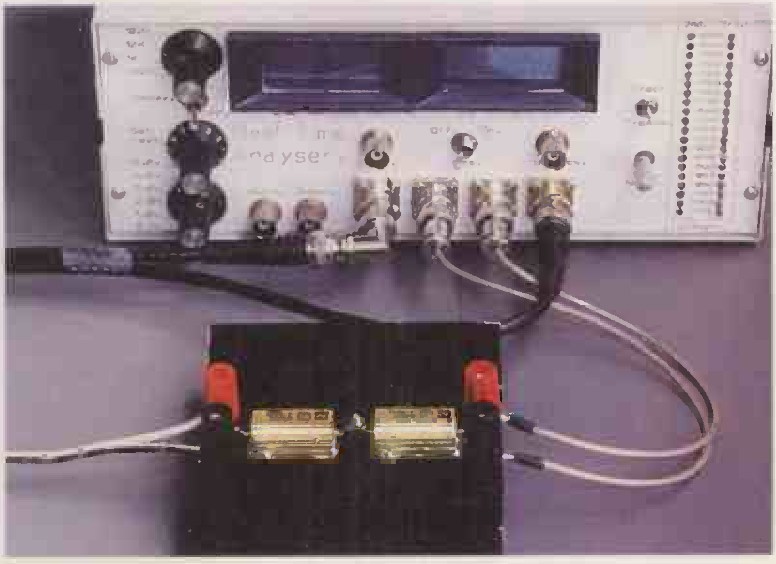


Fig. 1. The Real Time hardware system assembled into the low cost LC960, a 250 x 180 x 100 mm case as described in the September issue, is shown using my dummy test load/attenuator to measure the distortions produced by my Maplin Mosfet 100W amplifier.

Experimental circuits

Measuring experimental low level circuits and pre-amplifiers requires a different approach. In the past I have used my uncased test generator and notch/filter assembly with the FFT software to measure experimental circuits assembled on plug in proto lab boards, but this arrangement poses considerable difficulties with mains pickup. Also the proto board circuit assemblies are easily and unwittingly disturbed. I decided to use a small die cast box to contain my circuits, which would solve the mains pickup problem. Mounting my designs on an easily re-configurable PCB would make for an easily adapted but stable test system.

Mounting four short BNC cables at 22mm centres, to match the front panel of my standalone distortion meter, along one side of a die-cast box, would facilitate the required connections and ensure a repeatable, consistent test method. This boxed PCB would include its own power supply three leg stabilisers and a power switch, disabling the circuit to allow change of components or integrated circuit. All ICs would be fitted in turned pin sockets and resistor and capacitor components which may need changing, would be soldered onto Vero pins, accessible without dismantling the PCB from the test box, facilitating experimental circuit changes. **Fig. 2.**

I decided to provide for two amplifier stages, similar to the two stage output buffer used in my test equipment, to allow experimental verification of performance especially when driving into low impedance loads. Replacing the second amplifier with a DIL header which has its pins three and six shorted together, would permit comparison measurements of single IC performance against the same amplifier with separate output buffer, all without changing this PCB layout. **Fig. 3.**

Because many readers prefer to use dual IC amplifiers to save space and cost, I provided a second similar two IC PCB which used a dual amplifier for the first stage. While this alternative board could be quickly exchanged in the box by unsoldering a few wires and removing six nuts, I decided it would be simpler and better to assemble this alternate PCB into a second die cast box, the small extra expense being quickly recouped by saving much time when comparing and designing just a couple of circuits. **Fig. 4.**

The input amplifier for both PCB variants can be arranged either in non-inverting or inverting mode simply by changing a couple of link wires and resistors. Additionally a capacitor-coupled input with variable level potentiometer can also be quickly linked into circuit when desired. This flexibility allows many different circuit configurations also performance comparisons between different ICs. For this reason space

Fig. 2. The four BNC connectors mounted at 22mm centres, provide true four terminal measurements, isolating the test current from the voltage being measured. These enable a quick change from measuring capacitor distortions to evaluating low power amplifier circuits, simply by change of test jig as shown, with no change of the distortion meter internal circuitry.

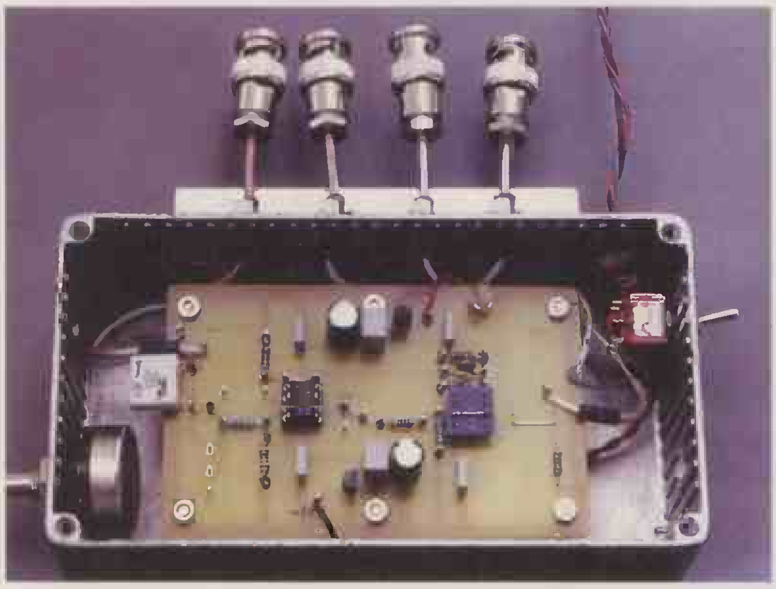


Fig. 3. Using a simple adaptable printed circuit board with components mounted on Vero pins permits change of circuit configuration as well as component values, without access to the track side of the PCB. The DIL header connecting pins three and six in the right hand op-amp position as shown, permits evaluation of single stage circuits, easily extended to two IC stages, by replacing the DIL header with an IC as in figure 4.

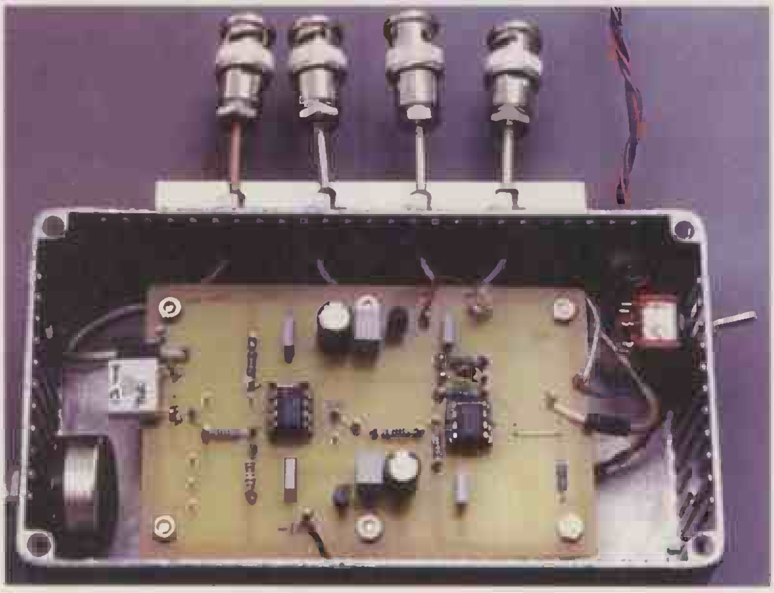


Fig. 4. The figure three PCB but now arranged as an amplifier with an output buffer IC. The potentiometer, bottom left allows measuring the differing distortions produced by both inverting and non-inverting variable level voltage followers.

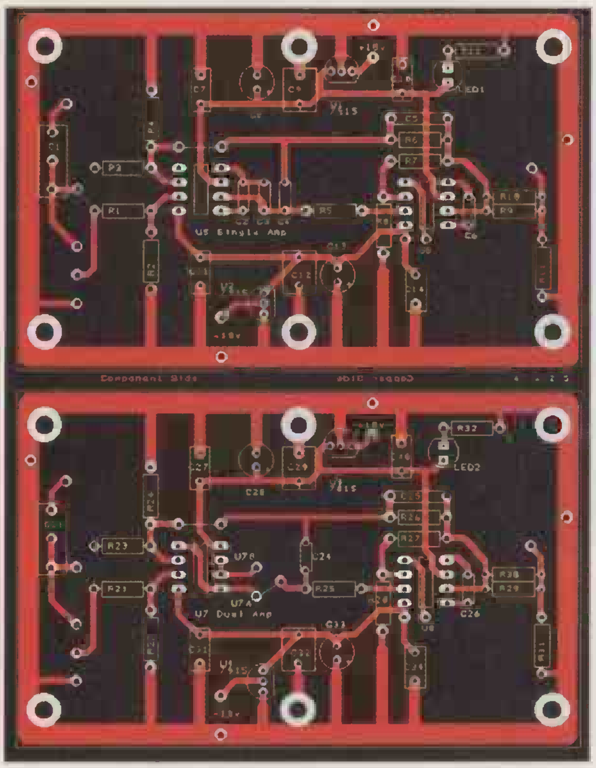


Fig. 5. I designed two PCB versions, the top version accepts a variety of single op-amps, and provides mounting positions for the 50pF capacitor needed for the AD797 also the 22pF needed to allow using an NE5534 at unity gain. The lower circuit board is similar but allows using a dual op-amp either as a single or two stage amplifier with or without a separate output buffer stage.

has been provided for the 22pF capacitor from pin 5 to pin 8, needed when using an NE5534 at unity gain, also the 50pF capacitor from pin 6 to pin 8 when using an AD797 IC. Other options allow using a BUF634P as the output stage as well as the AD811 used in my test oscillator output buffer. No doubt readers will think of many other variations they would like to incorporate for their own experiments. **Fig. 5.**

Use with Uncased test modules

Perhaps you are happy to use my test equipment as a number of uncased modules, assembled on your workbench, exactly as I did for all published measurements, prior to my September 2003 issue article. These test box methods and power amplifier dummy test load approaches are equally valid when using my uncased test modules.

The general procedure was described in earlier articles, the user manual supplied with my PCBs as well as in my CD ROM. When testing capacitors the test signal output from the buffer amplifier is connected to the notch filter input via the capacitor test jig. When measuring amplifiers this common connection must be broken, the test signal being sent to the amplifier input and the notch filter connected to the amplifier output.

This is easily arranged by turning the 'switch41' fully anticlockwise to the AOT position, having first removed any resistor from the R77 position. The test signal is sent to the test amplifier input using a short length of coax cable connected to the R77 resistor Vero pin nearest to 'switch41'. The test amplifier output, is connected to the capacitor test jig using a second coax cable. This general arrangement is clearly shown in **Fig. 6.**

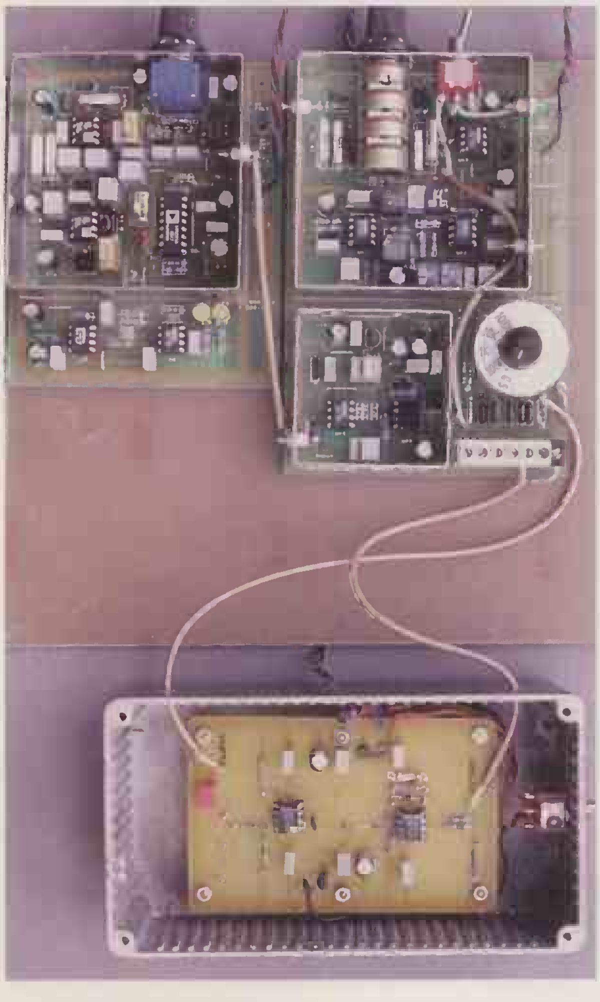


Fig. 6. The test arrangements shown for my LC960 cased distortion tester can equally be applied to the original bench top assembly of discrete modules, as shown.

Amplifier Dummy Load

Four 25 watt 8.2Ω aluminium clad power wirewound resistors were connected as two pairs in series and parallel to provide a 100 watt capable low inductance 8Ω dummy load. One pair of resistors was mounted above the heatsink using thermal paste as shown in the photo, the second pair beneath the heatsink, so are not visible. Also underneath the heatsink are fitted the attenuator resistors needed to reduce the test amplifier output to match the 0.5V, 1V through 4V test levels provided in my LC960 cased, standalone Real Time test system. For most purposes using a simple single stage attenuator to measure amplifier powers of 1.5W, 6.5W, 25W, 56W and 100W would suffice, however adding one extra attenuating resistor and output contact as in my photo, visible between the two 8.2Ω resistors, provides also for measurements at 50W. **Fig. 7.**

I said earlier that a dummy test load ideally should be non-inductive. To ascertain the behaviour using these two pairs of series/parallel 8.2Ω aluminium clad commercial 25W wirewound resistors, I made a series of impedance measurements by frequency using the method described in my SMPS capacitor article[4]. From 1kHz to 100kHz the impedance increased by less than one decimal place, from 8.02Ω to 8.06Ω, less than 0.5%. At 1MHz the impedance increased to 9.94Ω indicating my dummy test load's self inductance was slightly less than 1μH at audio frequencies.

A simple attenuator was assembled using three 1kΩ 1% 1W resistors in parallel to drop the excess voltage, then through the 2.7Ω compensation resistor needed when measuring amplifiers at 50W into a 56Ω tail resistor. The measurement voltage being taken across this 56Ω to earth for all measurement except for 50W, when the output is taken from the junction of the 2.7Ω and 1kΩ resistors. In this way we only need to adjust the test levels to match our 0.5V, 1, 2, 3 and 4V switchable test levels to measure at the desired amplifier power. Fig. 8.

As can be seen, I have provided two ground returns into the standalone distortion tester, one for the test current input to the test amplifier, the second for the attenuated amplifier output voltage, both grounds are of course commoned inside the distortion meter. Depending on the earthing details used for the amplifier under test, lower distortion meter readings may be obtained if both grounds on the heatsink are commoned, and the I-low ground to the tester is not used. This has not been found necessary for my tests, both earths having been connected exactly as shown in figure 1.

How does this work in practise?

Using a conventional notch filter THD distortion meter requires a certain delicacy of touch to first set the correct gains or test voltage levels, then more importantly to ensure optimum nulling of the test fundamental, which using my HP331A distortion meter can take several minutes. During this time the amplifier must continue producing the desired power level. Even when testing at only 25W both amplifier and dummy load can become hot.

Because of the excellent stability of my test oscillator and notch filter after a short warm up, having once carefully matched the notch filter tuning to the test oscillator, further tuning adjustments are not needed. This notch filter tuning adjustment is easily performed when using the capacitor test jig. Simply monitor the AC voltage at the 'Harmonics out' front panel BNC socket and slowly adjust the notch filter tuning potentiometer so as to attain the lowest possible voltage.

With my Real Time second and third harmonic distortion method, measurements at a number of amplifier power levels can be performed very quickly, simply select the desired test voltage range and adjust the test level control. A matter of a few seconds only of amplifier power is needed for each measurement, reducing the possibility of amplifier damage though overheating, even when testing inadequately heatsinked amplifiers. Much quicker and far easier than when using my conventional Hewlett Packard distortion meter.

Bi-polar electrolytics

In my article about aluminium electrolytic capacitor distortions[5]. I recommended using bi-polar and even two bi-polar types in series, which for convenience I labelled as a 'double bi-polar series pair' (Patent application GB0227606.1), in place of a conventional polar capacitor to reduce distortion.

Using the workhorse Maplin Mosfet 100W amplifier mentioned above, I performed a quick experiment to confirm the improvements gained. Originally this amplifier was built using bi-polar aluminium electrolytic capacitors for both its 10µF input coupling and the 47µF in the negative feedback loop. I replaced these and both power supply rail decoupling capacitors with new polarised aluminium electrolytic capacitors, as originally specified in the parts list. With its bias current carefully reset to the recommended 100mA, I measured its distortion. Built using the Maplin capacitor types, its measured second harmonic was -81.5dB and -91.4dB third harmonic when tested at 25 watts, confirming the design did meet its claimed less than 0.01% distortion.

I then replaced these four polar aluminium electrolytics, with the same value and voltage bi-polar types and re-measured distortion at 25 watts. For both tests the power supply used was built using two 10,000µF 63V conventional polar aluminium electrolytic capacitors.

With bi-polar capacitors in the module only, second harmonic improved significantly to -92.1dB and third to -94.3dB, a three fold improvement in distortion gained by using four bi-polar capacitors, a total cost of £2.10, to now measure just 0.0031%. No doubt replacing the remaining capacitors, especially the metallised PET items specified for this design with better types and replacing the input bi-polar electrolytic with a good film capacitor would further reduce this amplifier's measured distortion. These additional changes require modifications to the printed board layout, so have yet to be tried. No doubt redesigning the power supply to also use bi-polar electrolytic capacitors would further reduce measured distortions.

Distortion measurements

The conventional notch filter style distortion meter provides a single measure of distortion, including circuit noise which can be significant when using a wide measurement bandwidth. To clarify whether second or third harmonic distortion is dominant requires using an oscilloscope to identify the distortion meter residuals. At best this can only

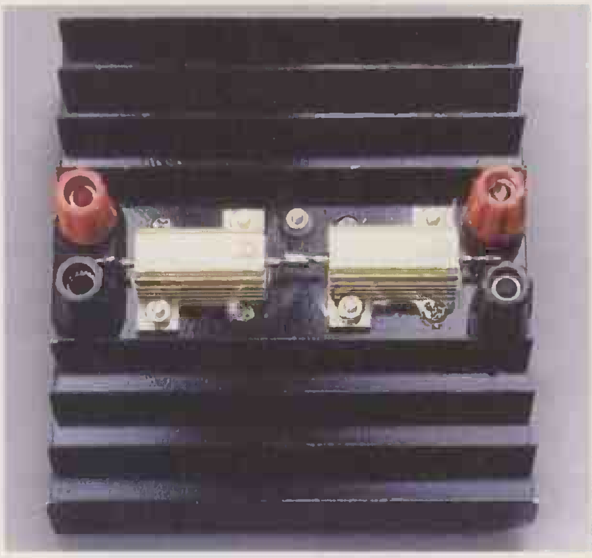


Fig. 7. The dummy load/attenuator shown was built to match my LC960 cased test meter's switched test levels, enabling distortion testing at 1.5, 6.5, 25, 50 and 100 watt amplifier powers. This attenuator and two of the four 8.2Ω load resistors used are mounted on the underside of the heatsink, so not visible in this photo.

Technical support

Full details of this new hardware test method and my original Capacitor Sounds series 1ppm low distortion oscillator, buffer amplifier, notch filter/preamplifier and DC bias assemblies, together with parts lists, assembly manuals and full size printed circuit board drawings, all as .PDF files arranged for easy viewing on screen or hardcopy, are provided in my new 'Capacitor Sounds' CD.

This CD ROM includes updated and much expanded re-writes with very many more figures, of my recent series of six 'Capacitor Sounds' articles, supported now by some ninety capacitor distortion measurement plots. Also on the CD are PDF re-writes of my earlier 'Understand Capacitors' series together with articles on how to diagnose failed capacitors while still mounted on printed circuit boards and essential low cost capacitor measurement methods, more than twenty popular articles.

The CD is now available, cost £15 Sterling including post packing. Send cheques or postal/money orders in Pounds Sterling only to:-

C. Bateman.
'Nimrod' New Road. ACLE. Norfolk. NR13 3BD. England.

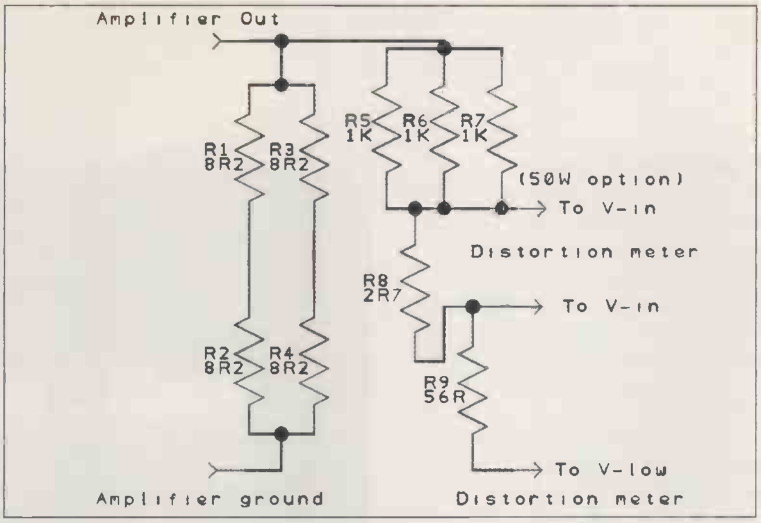


Fig. 8. Schematic of my non-inductive dummy test load and the resistive attenuator used with my LC960 cased distortion meter. The four 8.2Ω resistors are standard commercial 25W aluminium clad wirewound, the three 1kΩ were 1% 1W metal film types and the remaining resistors were 0.5W 1% metal film.

become an educated guess.

Using my Real Time method, the exact levels of second and third harmonic are directly indicated on the panel meters. By also using the ADC100 or computer soundcard and FFT software to monitor the front panel 'Harmonics out' BNC connector, other harmonics and intermodulation products can be measured together with total distortion with and without noise levels as desired.

The main disadvantage of my equipment may be seen as its fixed frequency when compared to some commercial instruments having comparable low distortion performance, but even second hand, these commercial distortion testers will be at least 25 times more expensive than building my published designs and vastly more complex in use.

I know of no commercial distortion measuring instruments that provides similar DC bias facilities, usable when testing either amplifiers or capacitors. For this reason a number of amplifier designers of repute have recently replicated my test instruments.

In my next article I explore using this equipment to look more closely into capacitor implications in amplifiers and to measure distortions in I.C. amplifier circuits. ■

References.

1. Capacitor Sounds. C. Bateman, *Electronics World*, July, September, October 02.
2. DIL/C-I Electronics, Holland. email to dil@euronet.nl
3. Capacitor SoundsII. C. Bateman, *Electronics World*, September 2003.
4. Evaluate capacitors for SMPS designs. C. Bateman, *Electronics World*, September 2000.
5. Capacitor Sounds. C. Bateman, *Electronics World*, January 2003.

LETTERS to the editor

Letters to "Electronics World" Highbury Business Communications, Nexus House, Azalea Drive, Swanley, Kent, BR8 8Hu
e-mail EWletters@highburybiz.com using subject heading 'Letters'.

Easy job?

I read with interest Ian Hickman's preface to the July edition of *Electronics World*, and being a contract electronic design engineer for the last 20 years, whose work environment was dealt a double death blow (GSM and 3G hit the buffers in 2001, followed by the aircraft industry after September 11th), I now work as a factory maintenance electrician for a local company.

My main criticism with Ian's piece is his assumption that his future son-in-law "given a few days study of the latest edition of the IEE Wiring Regulations, could probably make a comfortable living at the trade". This is, I feel, misleading at best and could have people who are out of work in the electronics and software industries thinking it will be an easy way to get a job. After spending the last two years collecting an armful of certificates and work experience, I am only now approaching the point where I can apply for a Joint Industries Board (JIB) card. Without this card, the majority of electrical contractors will not even interview you, let alone set you on. As an alternative, he could set up his own business as an electrician. Great, this works fine until he comes to the question of company and public liability insurance. The first question the insurance company asks is 'How much experience have you in this industry. (Are you getting my drift yet?) It is because of the (apparent) risk, the insurance will (of course) cost a large four figure sum, probably more than the work will generate - not a good position to be in. (exactly the same reason that the electrical contractors aren't interested!!). If he decides to work without insurance, then remember that the first time he makes a mistake, the 'no win, no fee' lawyers are always looking for clients, particularly with what they deem as a 'soft target' - hence the insurance.

Sorry if I seem to be discouraging what is probably a quite able and competent person from filling a gap in the skills shortage, but I think that it must be quite a few years since Ian last looked at the electrical trade. Like all other trades (plumbers, gas fitters, etc.) the electrical industry is trying to rid itself of 'cowboys' and, as such, does not easily accept anyone who has not come the traditional route through a trade apprenticeship. In addition, I have found that you need to be able to do a great deal more than read and interpret the regulations, particularly when it comes to practical installation and trade tests. However, on the bright side, if you would like information on how you can get through the maze, let me know (Davew@davroengineering.co.uk) and I will forward you a list of requirements put out by the National Electro-technical Training Scheme (NETS). Also, if you can get any job in the electrical trade, this will help enormously when you need to start amassing an 'evidence' file to back up the qualifications.

***Dave Wroe** BA (Hons), C.Eng., MIEE, MIDiagE*
By email

Technicians

I read with interest the critique by Bob Nelson (*Electronics World*, Letters - June 2003). He stated that not enough high-level technical articles are being published in *Electronics World*. Perhaps you should try to attract people at different levels, from hobbyist to professional. Also periodic tutorials on different topics could be included, such as op amp, digital, and industrial electronic controls. Many readers may want more articles on electronics in industrial automation, robotics, and interfacing with programmable logic controllers (PLCs). Automated inspection with lasers and electronic eyes, motor controllers, interface electronics for flow control (hydraulics and pneumatics), and computer numeric controlled (CNC) machining are especially important. I don't know what other publications in the UK or Europe cover this, but industrial machine automation is especially important to a nation's technical and economic standing. You should also cover the state of the electronics industry both in the UK and in other countries. Especially important are employment levels, different careers in electronics, and training programs. Such articles would be valuable for those with children at an age to decide on a career path. In the past, engineers were usually the people emphasised.

Now technicians and technologists are assuming more of the high-level design and implementation traditionally done by engineers. In America, the two-year associate's degree from a community college or state technical school is becoming ever more important. Associate level programs are offered in a variety of technical options, including electronic engineering technology, robotics, advanced manufacturing, and quality control. Topics covered in great depth include analogue, digital and

Hard times

With reference the article Hard times by Ian Hickman, he refers to the fact that software engineers are going through a black patch with regards to employment. He also suggests that his son-in-law who is a software engineer and finds himself in this position takes up employment as an electrician suggesting that his son-in-law has an appreciation of just dangerous 240V not to mention 415V three phase can be.

He also suggests after a few days study of the IEE regulations he could make a comfortable living at the trade. As a retired electrician he would require a little while longer than that.

He refers to the shortage of skilled men in various trades – electricians, plumbers and welders. He also refers to our unbalanced education system with politicians of all shades wanting everyone to finish with a degree resulting in many people finishing with 'Mickey-Mouse' degrees.

Another view is that not many young people today are willing to get their hands dirty doing something whichever way you look at it can be a dirty job!

Brian Corbett
Oldham
Lancashire UK

Self powered amp

In Doug Self's article 'Power amplifier input currents and their troubles' he mentions using FETs instead of bi-polars to reduce input current non-linearities in the power amp. One way to use them is to cascode the difference amp with the base bias of the cascode transistors taken from a voltage divider between the output and a DC bias source, so as to track input and feedback. This might get rid of the main source of capacitive non-linearity in the difference amp stage.

William C. Cross
Denver
Colorado
USA

microprocessor systems, programmable logic controllers, fluid power control, and semiconductor manufacturing technology. When Congress passed the $125 million TECH PREP training bill in the early 1990's, it stated that America's technical leadership depends more on technicians than engineers. As *Fortune* magazine (August 22, 1994) stated in the feature article "Technicians - The New Worker Elite", as companies rely on technology to improve product quality, technicians are the front-line workers they depend on. Also, some four year bachelor's degree programs in industrial technology cover industrial electronic controls, as well as robotics and computer aided design and manufacturing (CAD/CAM). Technicians and technologists serve as the important link between high-level design and world-class quality factory-floor manufacturing. Perhaps *Electronics World* could cover both the Associate's degree options in America and the HNC programs in the UK. Different specialties (computer, automation systems, communication) could be included. Some magazines are too oriented toward computers or internet technology. A broader coverage would attract more readers.

Glen W. Spielbauer
Dallas,
Texas
USA

Another pile of good ideas to put into my melting pot. Ed.

Electronic Balun

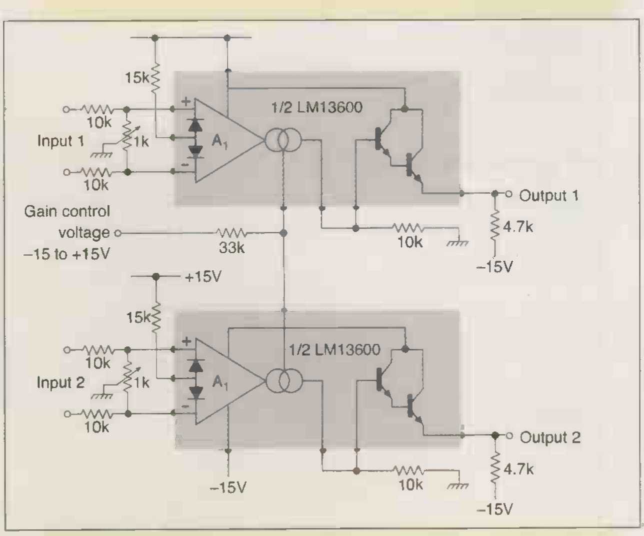
Referring to Ian Hickman's "An Electronic Balun" article in the June 03 *EW*, putting in C3 (and C4) in Fig. 2 introduces an asymmetry that degrades the CMRR at high frequencies. Presumably it is introduced to limit the bandwidth and improve the stability margin. You can restore symmetry by bypassing R11 and R14 with 100pF caps.

David Hadaway
By email

With regards to Ian Hickman's electronic balun article, I have an alternative solution: It is well known that Ota's differential input is converted in single ended at the output with good common mode rejection without any need to carefully match resistors.

For example, adding two resistors to a circuit idea of September 2002 (2 channels one gain pot), you obtain a stereo balun with some 60 dB CMRR and a very low distortion. The gain of each LM13600 goes from 0 to 900 when the command voltage is varied from -15 to + 15 volts.

Jean-Marc Brassart
SAINT-LAURENT-DU-VAR
France



In circuit RF measurement

When I saw the third page of the article by Tuck Choy in *EW* July 2003, I was surprised at its low standard. I decided to write a proper article on capacitor-diode RF voltage measurement based on a correct mathematical solution to the relevant equations and a C-programming language program to calculate the correct DC output voltage, none of which I have seen published before.

I then noticed that Tuck Choy had used an incorrect equation taken from Wes Hayward's *Introduction to radio frequency design*. Wrong in Hayward and just copied in your issue; is this plagiarism any way one spells it?

Next I noticed that you ridicule 'hand written letters' on page 56 of the same issue. What now am I to do with hand written original work not wanted by you but technically useless printed work copied from Wes Hayward's mistake is acceptable to you?

William Coughlan
Dublin
Ireland

The point I was trying to make about hand-written material is that mistakes can and will occur. It also takes a lot longer to get a handwritten contribution on the printed page as it has to be typed in (usually by me). As regards accepting hand-written feature contributions – that's fine, but the remuneration will be less as it costs me more to get it in the magazine and it will take longer to get it published as rather than relying on fast email – the printed page and 'snail mail' will have to be used instead. But surely one could toddle off to the local library, type it into a computer and email it? In any case, send it in. Ed.

Tuck Choy replies:

Regarding the 'error' – I'm not sure what William means, if it's the factor (-1) then the exponential prevails. The diode equation at high frequencies contain much more of course. For details of an equivalent circuit, see S.M. Sze "Physics of semiconductor Devices".

The allegation of plagiarism is a very serious one to be put in print. I happen to be acquainted with Hayward and he has not complained to me so far. In any case the formula and its related derivations are found in many standard texts. No one owns the diode equation nor any one equation for that matter. If Einstein owns $E=mc^2$, then many many more people would have been in trouble for plagiarism!

Measurements using Slotted Lines

Congratulations on the very fine July 2003 *EW* article! As both a professional engineer and a radio amateur since 1952, I passed the majority of my professional career, i.e. 37 years, at the laboratories of the Belgian state broadcasting institute, formerly NIR/INR later BRT and now VRT for the Dutch speaking part.

Although my lab occupations at the beginning were strictly low frequency (recording lab from which I became Laboratory Chief in 1963) as my interests as a radio amateur shifted in the early sixties away from the HF bands to VHF and UHF, I spent a lot of my spare time in the microwave labs of my friends at the institute. Later on as I became Laboratory Group Leader, I also had the responsibility over the radio studios and our mobile recording vans, in which VHF/UHF equipment was used for communication. So I could use my experience on VHF/UHF from my hobby for my professional job.

In the early seventies I had the opportunity to buy in a local Ghent surplus store a General Microwave Corporation (Farmingdale USA) coaxial slide screw tuner type N253. This serves normally to correct a high SWR by moving on a carriage along a slab type coaxial line a capacitive acting mass from which the distance to the centre conductor of the line can be varied with a micrometer-gauge.

The instrument is similar in construction to Hewlett-Packard slotted lines 805 C and General Microwave N200. This type of construction, side plates of approximately 115 mm wide, spaced at ± 20mm, has many advantages over classic 'slotted lines' as the one described in the *EW* article .To quote an old HP catalogue "This configuration results in negligible slot radiation, minimum sensitivity to variation in probe depth or centering and greater structural stability" (as the backsides or outersides of the 'slabs' are reinforced by ribs.)

I soon modified my slide screw tuner into a real slotted line. The carriage travel is ± 350mm, the vernier reads to 0.1mm, so the usable frequency range is about 400 to 4000MHz. I modified the set as follows: The micrometer drive serves for the capacitive probe penetration (a thin silver brazing rod passes through a slide contact near the grounded edge in a half-wave resonator tuned at the centre by a capacitive plunger from about 1280 to 2500MHz. The probe is isolated for UHF from the micrometer gauge above it by some ferrite beads.) The other grounded side of the Lambda/2 line has a BNC socket on which a homemade Schottky-barrier diode detector is seated.

I do have much experience in mechanical engineering, as one of my uncles had a well equipped workshop nearby with six lathes, three milling machines, four planing machines, etc. etc. in which I spent great part of my summer holidays.

Later we had a small workshop at the recording lab including a Myford Super 7 Lathe and grinding machine, etc.

So the authors of the *EW* article must have had an very well equipped machine shop, far more sophisticated than my laboratory possibilities at the time. They chose for the construction of a very difficult project: drilling a very long precise hole in the axis of a 300mm long brass square and cutting a precise slot in it, etc. Sincere congratulations to their work! The slab type slotted line is much easier for home construction. I started constructing one for VHF (slabline with two large aluminium square U profiles) but never finished it because of its bulkyness and other measuring apparatus with more possibilities (antenna analysers from Bird and/or MFJ.)

Originally I made my 1296MHz measurements using a second hand General Radio 1021-AU HF generator (250 MHz -960 MHz, piston type attenuator < 0,5 micro volts to 2V) succeeded by a diode frequency tripler and a homemade coaxial cavity.

For the readout instrument I constructed along the lines of *Ham Radio Magazine* October 1970 "The SWR Meter" (p6 to 14 incl.) using European transistors instead of the original American, a square law reading peaked 1kHz microvoltmeter with a sensitivity of 0.15 microvolts, with attenuators seven steps of 10dB. This with the help of 'DATA' publications and a Philips transistor curve tracer. I had also much help from a later *HR* article - May 1977 "How to use the slotted line for transmission line measurements".

Later I constructed a simpler SWR indicator in a small die-cast box ("Wide range power meter" *HR* April 1986) using some op-amps as a tuned amplifier with HSCH-3486 Schottky barrier diode.

I also constructed two small VCO units following another *HR* magazine article from July 1985. This uses a pair of BFR 93s in push-pull for 880 to 1350MHz and a pair of BFG 65s for the VCO for 1.6 to 2.5GHz The VCOs are tuned by a variable emitter voltage from -1 to -11V. For the 2.5GHz unit I had to use thin Rogers RT duroid, the lower frequency unit uses ordinary G4-1.5mm epoxy. Each VCO is followed by a MMIC amplifier from Minicircuits, which can be fed by +12V or a square-wave at 1kHz from a 555 timer-oscillator and a pass transistor. Both include small attenuators to isolate the VCOs proper, to stabilise the MMIC and also attenuators before the 'device under test'. Output from the small devices is 0.5 to 0.8mW at the lower tuning frequency and around 1.5 (1.3GHz VCO) to 4mW from the 2.5GHz one flat $\pm$1dB between 1GHz and 1.3GHz respective 2 to 2.5GHz. The phase noise is high, about -40dB as seen on a spectrum analyser, but they are satisfactory for wide-band measurements such as antenna-alignment, insertion loss of coaxial cables, isolation of coaxial switches, wideband preamps etc. The power and readout unit contains in a die-cast box fixed supplies for +12V, variable ones for -1 V to - 12V for the VCOs (10 turn helipots plus LM337T), +2 to +30V for the tuning voltage of transistor GDO units tuned by vari-caps (10 turn helipot + LM317 T), a small DVM for readout of tuning voltages plus a small meter for GDOs with DC-amplifier, the 555 oscillator, etc. This construction from 1985 I baptised 'The UHF- tester'.

I used extensively the information of

Metal detection

In the article 'Metal detection' in the July issue, the author firstly experimented with a square loop measuring its inductance in presence of metal objects, and found that the iron object did not increase the inductance, the inductance being reduced both with iron objects and with metallic non ferromagnetic objects. The author attributes (rightly, I think) this behaviour to the eddy currents, "predominant in both materials". His point is confirmed by the experiments made with ferrite rods, where eddy currents are negligible and the increase in the inductance is effectively seen.

His tests with the assembled metal detector confirm all of that.

However, having done some consulting work for induction heaters, I know that the increase in the inductance, in a coil near iron objects, is a real word effect, as is introducing into the induction coil some iron to make the resonant frequency decrease.

So I started to think that the frequency makes the difference. At high frequencies, the eddy currents are strong enough to prevent the magnetic field effectively entering the metal object (be it iron or non ferrous), but at some lower frequency the inductance should increase in the presence of iron. The experiments of the article where made at 10kHz (with measuring bridge) and at 460kHz (with the assembled metal detector).

Having at hand a LCR bridge with selectable measuring frequency of 10, 1 and 0.1kHz and a big air-cored coil I normally use as dummy load for testing magnetotherapy devices, I made some measurements with them. Here are the results, for anybody that may be interested.

Coil: about 27 turns, 2.5mm^2, on a square 240x240mm cardboard box; resistance 0.29 ohm, inductance 618μH. Metal objects are placed about 80mm over the centre of the winding.

Measuring instrument: Wayne Kerr automatic LCR meter 4210, in 'series' mode measurement. All values are μH.

Measurements on 100Hz show some variations, about 0.5μH around the mean value, presumably because of disturbances from 50Hz fields. The above table confirms that the iron can be distinguished from non-ferromagnetic metals, if the frequency is low enough. Besides, there is no 'magic' frequency; one needs to resort to lower and lower frequencies, as the iron sample becomes bigger.

Examining the table, the first line (free air) tells us that the inductance is about the same at all the three measuring frequencies. This should mean that the measuring instrument and the measuring mode (series inductance) are both adequate. The second line (aluminium) confirms, as we expected, that non-iron metals make the inductance decrease; but, too, it shows that at 100Hz the reduction is less, as presumably the eddy currents are less effective in keeping the magnetic field entirely out of the metal body.

The third line (iron can) shows what I suspected; at 10kHz the inductance decreases, as in the experiments of the article, but at 1kHz it increases (and at 100Hz even more).

The fourth line (iron nails) shows that, for objects as small as a nail, the inductance increases even at 10kHz (though less than at a lower frequency). This could well explain the claims of the capability to discriminate iron, made by some authors and not confirmed by other experiments. The difference could be in the dimension of the metal object. I had to test with a number of nails, because a single nail would be not detected by the resolution of the LCR bridge.

Incidentally, this test tells us another thing: it is the dimension of the single object (nail) that makes it detectable as iron at a given frequency, not the total weight; the total weight of the nails is well over the weight of the metal can of the preceding line.

The fifth line (large iron box) confirms that the discriminating frequency is lower for a bigger object; at 1kHz the inductance still decreases, only at 100Hz it increases.

Ezio Rizzo *Italy*

	10KHz	1KHz	100Hz
Free air	618.00	618.20	619.10
115x200mm aluminium heat sink	606.20	607.70	613.30
Iron tin-plated cylindrical can 65mm diam., 25mm length	615.70	620.60	622.30
Plastic box with iron nails	618.80	619.40	620.70
Large empty iron box 365x150x170mm	589.70	606.60	631.20

several other articles from the late *HR* magazine e.g. to align a homebuilt 1.2 meter diameter parabolic antenna with log periodic primary feed with the slotted line and the two UHF indicators (one for the slotted line, one for a remote field strength meter) to find a compromise position relative to the focus of the dish and the phase centre for 1296 and 2320MHz for the LPD. It is a pity that this fine magazine ceased publication in the late 80's.

I understand that *Ham Radio Magazine* is now available on several CD ROMs from the ARRL. Their own QEX is interesting but fails in being a valid substitute.

Ing Walter Empsten, ON4ZN
Chief of Service (retired)
External Radio Studios,
Recording and Outside Broadcasting
VRT
Brussels

Illogicalities

With Ian Johnson, I too feel most gratified that Ivor Catt is alive and well - as with his name, he should have at least nine goes at staying alive. On the other hand, though concurring with Ian's support of independent points of view and argument, I found his letter, (July issue, *EW*) strangely pessimistic - amounting at the end to a feeling of depression!

Yes of course, the World, the Universe and Everything is imperfect (a human value judgement, and therefore subjective, by the way). We all know how boring a so-called perfect world would be. Everything would be done, it would all be designed, no free will - a perfect perpetual motion machine with not a single statistical variation or wear-out phenomenon. Those poor old capacitors in Ian's world, what have they done to be sooo evil and faulty? I have always found them remarkable, from my first Leyden Jar (which I thought was perfect - it gave me and pals a mighty shock when we held hands in series to 'test the possible charge', two days after charging it) to the extraordinary technology of the microwave surface mount caps we routinely solder onto our boards. Every time I open, say, a distant Radio Repeater with a little stub antenna radiating 2.5 watts (Oh, sorry, the little imperfect PA cannot be more than about 50% efficient - therefore about a watt may go out.), I marvel at those 'Laws of Physics' that Ian would like to be 'flexible', (what a mess we would be in if they were.) Just what does radiate all those miles - at that speed - to open the repeater? That mystery is what got me into musing for a lifetime about just what is out there, and how do 'waves' travel through nothing? (Don't mention the Ether - but...?) I find it no good to listen to the reply, "Well, when the electric field lines rapidly change their strength, (i.e. $\frac{\partial E}{\partial t}$) you get magnetic field lines at right angles - and they in turn give you more electric lines. You should know that!" I reply that I do, but just what are those 'lines'? Why, they're just made up models or pictures (and therefore completely false) that we have invented for our poor minds to see something. Then we believe it? But what does penetrate empty space - reach out over a distance to 'know' a lump of iron is there, for example, and pull on it, without losing any energy? (Even a child's little magnet has this deep mystery still taken for granted.) Oh, yes, we have Maxwell Equations (a tiny bit is that differential above) and we get surprisingly accurate design stuff from them - notwithstanding the depressingly imperfect world offered by Ian.

But though I mightily appreciate them, I cannot see an explanation from the equations of the real physical mystery of how and why all that design out there fits together so well, in spite of, er, imperfections. Of course, we could very easily slide into the Creationists' 'argument from design' stuff at this point, with all the illogicalities, wild assumptions, untestable assertions with which that bunch troubles certain areas of the United States. Nevertheless, it is the remarkable success of the discoveries of the givens in the natural world that is so exhilarating, not allowing the imperfections to depress us (I suspect Ian is one of the closeted perfectionists at heart - and they are always most unhappy.). It would be a tonic for Ian and any other readers to peruse a lovely little book called It Must Be Beautiful Great Equations of Modem Science, ed. Graham Farmilo, Granta Publications. No bemoaning the imperfections there.

'Joules Watt'
By email

Wireless networking

The Ian Poole article on 802.11 wireless networks explained how these networks offer many benefits over conventional wired networks but failed to explain how these benefits come with added risks for the unwary user.

As with many modern PC installations the default install does little to protect the user from prying eyes, even worse he can be open to hacking even stealing his data.

The August issues of both *Personal Computer World* and *PCPro* magazines include comparative performance reviews of many wireless kits, which might assist any reader wanting to install a wireless network.

A single page article "Turn on, tune in, get hacked" in *PCPro* p.11, by James Morris explains how he was able to obtain free broadband internet access using a nearby 'open' network without the owners permission. I paraphrase: "I turned on my notebook and found I'd magically acquired internet access. Someone in the vicinity had a wireless network connected to their broadband and it was completely insecure. One mouse click and I was able to connect for free."

This should present a sobering warning to all wireless network users. This default install lack of security problem is not exclusive to wireless networks. All internet users are commended to log on to the Gibson Research site: http://grc.com (note no www needed) This site provides a wealth of data about how easily most internet enabled computers can be accessed. In addition one can request a practical test to prove whether ones own computer is open or secure.

Gibson Research also offers a free for individual users local firewall software. Having once tested my computer using the Gibson site, I now personally never access the internet, using Windows, without first enabling his free firewall software.

Cyril Bateman
Acle
Norfolk
UK

Self made project

Can I reply to Sébastien Veyrin-Forrer's letter in the July issue of *EW* as it criticises me? (The floor is yours – Ed.)

Seb's letter gives out the impression of arrogance; condescendence, snobbery and elitism, I can live with the first two, but I am totally against the last two. The equation for snobbery is: the more the snob, the lesser the intellect.

Subjective! "relating to the subject – existing in one's own consciousness" *Chamber's Dictionary*. A designer designs a power amp – a manufacturer makes the amp, the amp is sold to hi-fi shop and who do they sell it too? That's right, Mr. & Mrs. Joe Public. And how do they judge the amp? Right again – subjectively. It is played to them to judge it, specifications come second. We are not robots, cut us and we bleed, something funny we laugh – sad we cry. Modulated airwaves go down our ear and are converted to electrical pulses and sent to our brain, Seb – we are 100% analogue, not digital. All out actions on this planet are subjective. We marry "I love this woman", "I eat fish and chips because I love them", "Mr. X punched me – so I punched him back". Try telling a psychologist subjectivism isn't built into human beings and shouldn't be used as a means for us to judge other beings or things.

Don't get me wrong – in my last letter I acknowledged that I used two of D. Self's means to lower distortion in my amp, the man is a genius, nil distortion amps using cheap components – I hold my hand up to him as he well deserves world-wide praise and will go down in history books. But Seb, although he is your 'God' it does not necessarily follow that his amps will sound better than anyone else's.

Taking "a leaf out of your book" I will be 60 in a year and a half, and suspect that I have been building amps long before you were born and have repaired more than "you've had hot dinners".

You think the hi-fi industry is perfect? Eight months ago a friend gave me his amp to repair, it had already been to a Scottish hi-fi shop to repair. This amp had no output when the CD input was selected. The fault was that a square of plastic carrying four phono sockets had sheared the steel connectors to the PCB and also internally due to the plastic nor being rigid enough. I replaced the phono sockets and hard wired then back to the mode switch. Now this a bottom of the range amp costing about £200, I have no problem with cheap sockets or cheap ALPS volume controls due to the price, but coming to the power section, there sat a well known audio chip - not JLH's way of doing things, or mine, but again the amp was cheap, but there staring at me at the output were two power chips, not by any stretch of the imagination could a designer spend more than five minutes planning the amp. Page 8 of my "How to build a power amp for beginners" shows the same circuit. My friend said it was smooth, and so it was smooth in the bass, smooth in the mid-range and smooth in the top range, so smooth that the amp was flat as a pancake – no dynamics, no imaging, no emotion. Yes, it had a very low distortion and a nice square wave response, but the amp was a rip-off. If anybody recognises it, they should hang their heads in shame. What really made me angry was when I gave it back to my friend, he again said it was smooth and I had to bite my tongue rather than tell him the truth. I am a straight forward person, what you see is what you get and in then past I have been brutally honest – but have learned how to keep my mouth shut.

On another tack, I read Douglas Self's article on input impedance. Again, I hold my hand up and say that I measured the bandwidth and the state of the square wave at just over three quarters revolution of a 50k Ohm variable resistor, it gave me a flat bandwidth to 100kHz, putting it back to quarter revolution produced a square wave very badly rounded at the leading edges and solid verticals with the speed of a dead snail, so out they came after a check for output impedance on my CD player and preamp and the 2 volts (RMS) outputs. I fitted two (one each channel) 20k Ohm pots. I turned them to very near the end stops and adjusted the amp's input gain to correspond to full output. There I re-tested after adjustment of the 1950s 'beehive' air spaced trimmer also used as compensation capacitor, as per D. Self, I found the frequency response was flat to an amazing 1.116MHz. No wonder I had so much trouble with HF oscillations, although now cured, I thought I was dealing with a 100kHz band width – not over 1MHz. The square wave was now excellent, the verticals can only be seen if you press your nose up to the scope screen.

The mains transformer I put in for Class A operation was upright and I noticed some AC current due to leakage. I decided to take a leaf out of the top of the range model and build an outboard power supply. In the transformer space I grouped all eight 10,000μF capacitors around the star earth and commoned the earth returns via a 30A cable, requiring run of only 3" to reach the star earth. I built the power unit into a small heavy gauge case with a 25A rectifier and two 4,000μF capacitors at its output (the regulation is still the amp case). As there is a very large RF pulse of approximately 2MHz in my area, I added very heavy current ferrite cored coils in series with the pos/neg output. Also 0V/earth, which is fully floating 12" of 30A cable connected by plug and socket to the stabilised section and to the bottom of the star earth. D. Self is quite correct about where you apply wires to. No mains earth is applied to the power supply, amp, or preamp earth fully floating, the AC current has now vanished. I took off the fan and fitted a very large heatsink, but was disappointed as noise has now doubled at the output. A previous amp design in *EW* had an output noise of 40μV, I now had 2mV. I then realised I wasn't judging like with like. I switched the amp off for two hours and on switching on I adjusted the quiescent current to 180mA, shorting the input, I put an 8Ω wirewound resistor and connecting up my millivolt meter and oscilloscope, did I feel good? You betcha – staring back at me was 90μV of noise, a world first for me. I then tested for distortion and another world first, 1kHz was 0.002% and 20kHz was 0.007%. Now I realise what was making the Class A distortion look bad.

Several points: I realise new regulations on earthing, i.e. mains isolated PCB/double insulation, but the amp has no longer any AC current running around it. Each channel now has six earth returns. No problem now with low frequency 'grunt' due the large storage capacity of 80,000μF. I admit that the circuit is now nothing JLH's original design, using the best points of D. Self and others in the amp, but the earth returns are totally down to me. Over the top? I refuse to accept that on two counts: firstly it works and secondly I was surprised at the large amount of audio designers who own Krell amplification equipment. An excellent amp, but it can cost £8000+. Just look inside one and tell me I've been OTT.

Personally, I think it's time *EW* got all the anti/pro subjectives together and hired a hall in London to thrash this out to a conclusion. D. Self and his disciples, JLH and his.

I now have my best ever imagery. It sounds like I am in large concert hall, it's pitch black and all of a sudden vocals and instruments come towards me in mid air, fast, clear and strong.

I want to thank *EW* for giving me my three minutes of fame. I will now die in peace.

D. Lucas
Pittenweem
Anstruther
Scotland
KY10 2lh

Help Wanted

Old radios

Having just read the article on the old German VE301 receivers in August edition, I write looking for help in trying to establish the origin of an old British broadcast receiver which I have recently revived.

Recently I was asked by a friend, (G0DKW) to attempt the restoration of a very old broadcast receiver which had apparently been languishing in a barn for unknown years. The set was a three valve TRF design and despite much dirt, nearly all of the components were in working order. After a week's work in cleaning and re-wiring it worked amazingly well bringing in 23 stations at loudspeaker level in the mid afternoon on medium wave. The aerial was a 20 metre wire.

Unfortunately, the receiver bears no identification marks apart from the component manufacturer's names and I would really like to know who made it and when. Perhaps someone with a grey beard that is somewhat longer than mine might be able to recognise the design.

The receiver is built on a half inch thick wooden base board of side 12 x 14". It has an ebonite panel, 14 x 7", with four controls, i.e. LT switch, reaction and twin edge-wise tuning controls with scales marked 0 to 180. The circuit is a 1V1 style using triode valves of the types210LF, ???LF and 215P of Cossor manufacture. As a triode is used in the RF amplifier it has an internally set neutralising capacitor adjustable with a short ebonite rod. The plug-in tuning coils are 2.125" dia. and 2.4" in length, Litz wound and are labelled 'LEW - The London Electric Wire Company and Smiths Ltd.' The RF coil is type SP A5 and the detector SP T5. The three fixed capacitors are made by Dubilier, two are mica breadboard fixing types and the third a 2μF paper decoupler. The only resistor, the 2MΩ grid leak, is held in a clamp made by DUMETOHM. It has a single audio interstage coupling transformer made by Varley which is also inscribed Oliver Pell Manufacture. It has no audio output transformer and I guess it was intended to operate with a high impedance balanced reed loudspeaker. All external connections are via row of ten screw terminals along the rear edge of the board. The neutralising capacitor is by Igranic but the main tuning capacitors bear no inscription. Valve holders by LOTUS, and RF choke by RI Ltd.

Was it a kit? Certainly the front panel mounted tuning capacitors look factory prepared. As no cabinet or cover was recovered, I wonder if it was originally fitted as one shelf in a cabinet which may have also housed a phonograph. The battery supplies look as if they were fed from below. This receiver works very well but needs a fair bit of expertise in order to adjust it to best results, so I guess it must have come early in the broadcast reception era. Can anyone please cast any light on this old mystery? It certainly took me back to my first encounter with TRF b/c receivers in 1944 with I guess much the same pleasure when it worked.

Chas F Fletcher. G3DXZ
g3dxz@thersgb.net

HP Manual needed

I need a user manual for an HP Digitising Oscilloscope type 54501A. Photocopy would be very acceptable. Will cover costs.

Charles Coultas
0118-9785713 or email: charles@dataskil.com

Data sheet needed

I need a datasheet for an obsolete Philips IC TDA 2581 or source.

Stephen Tames
0118 9875123 ext7998 or email: s.r.tames@rdg.ac.uk

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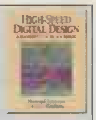

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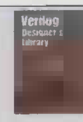

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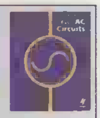

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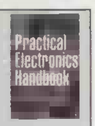

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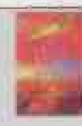

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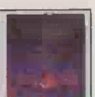

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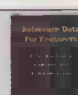

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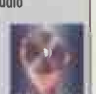

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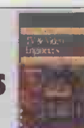

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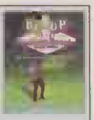

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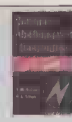

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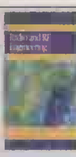

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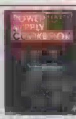

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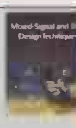

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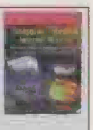

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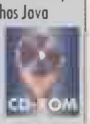

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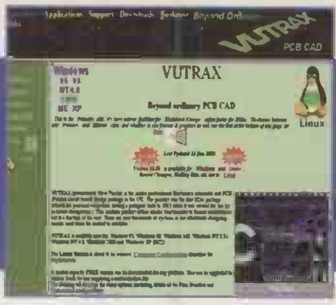

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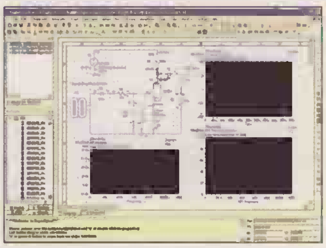

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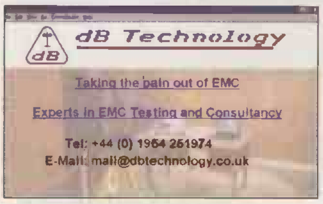

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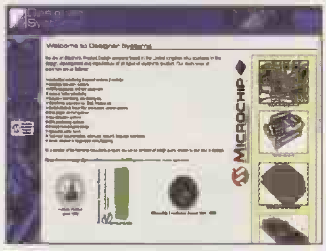

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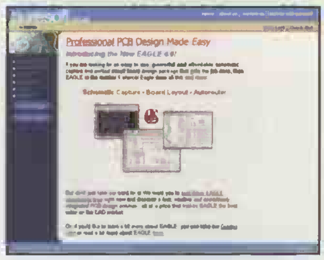

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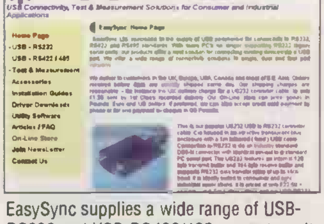

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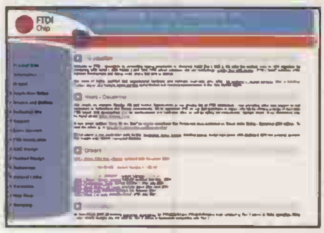

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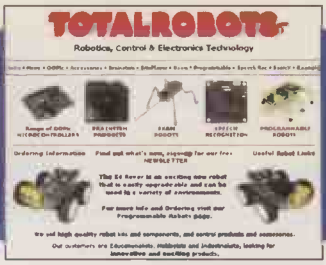

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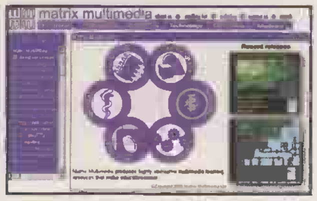

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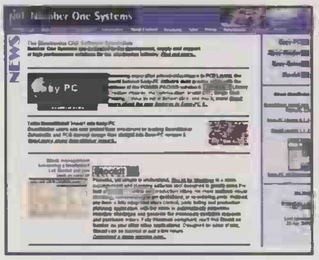

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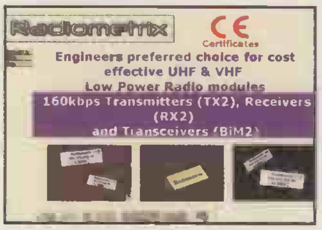

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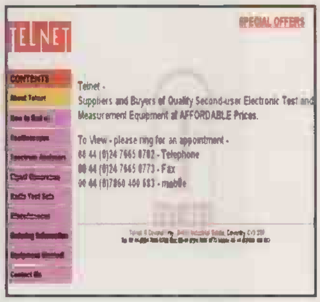

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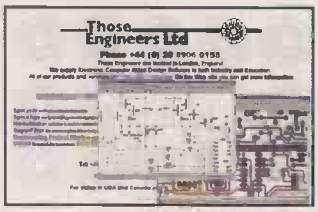

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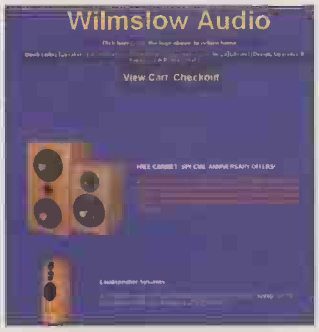

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