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Can the Internet work for you? See page 488.

REGULARS

Two exclusive EW+WW reader offers – 20% discount on this LCR meter, page 487, and a £220 designer's controller board for £99, page 509 onwards.

Cover illustration – Jamel Akib

Rather than relying on large enclosure dimensions, this design extends bass via electronic compensation – page 469.

Next month:
Heart-rate monitor, dual current mirror for faster audio power amplifiers, designing interface cards for the pc.
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EDITOR
Martin Eccles
0181-652 3128

CONSULTANTS
Jonathan Campbell
Philip Darrington
Frank Ogden

DESIGN & PRODUCTION
Alan Kerr

EDITORIAL ADMINISTRATION
Jackie Lowe
0181-652 3614

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

E-MAIL ENQUIRIES
martin.eccles@rbp.co.uk

ADVERTISEMENT MANAGER
Richard Napier
0181-652 3620

DISPLAY SALES EXECUTIVE
Malcolm Wells
0181-652 3620

ADVERTISING PRODUCTION
Christina Budd
0181-652 8355

PUBLISHER
Mick Elliott

EDITORIAL FAX
0181-652 8956

CLASSIFIED FAX
0181-652 8956

SUBSCRIPTION HOTLINE
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Tired old cable

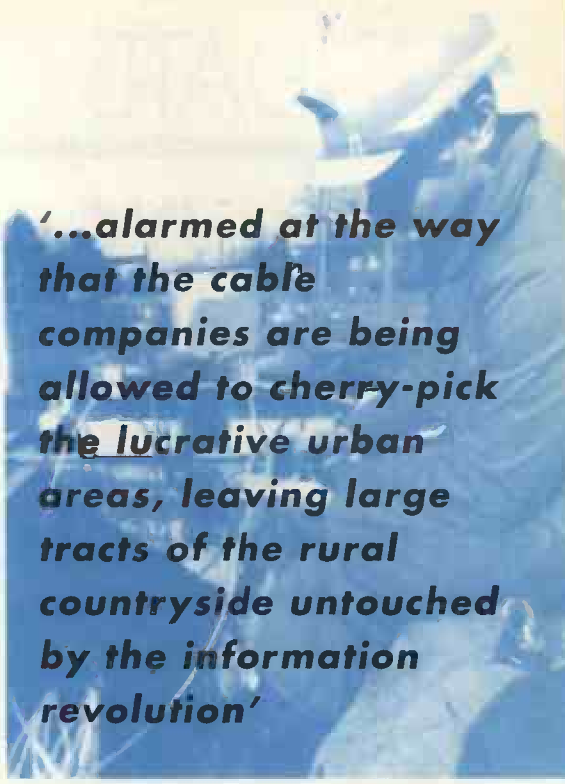
Why was cable described as 'Tired' by *Wired* magazine? Perhaps maps showing the coverage of the country by cable services give some clue. Penetration in the Cambridge area, for example, has the appearance of a monstrous spider with its body resting on the city and its legs trailing out into the surrounding countryside. Some key villages are missed out completely; others, with perhaps only one road connection, are served as they are on a straight line between Cambridge and another large town. The idea that today you telephone a person rather than a place seems to have lost something in the translation to cable.

While coverage is one issue, take up of services and churn rates (lapsed subscriptions) also cause concern. In the Cambridge area there are now 130,000 homes which could be connected to the cable network. So far 28,000 have taken television and 30,000 have taken telephone services. These figures fall short of those achieved by Bell Cablemedia who also operate in the East of England. They have take up of 22.3% for television and 25.7% for telephone – and still Bell Cablemedia made a loss last year of over $25 million.

Anne Campbell, who is Labour MP for Cambridge, said in a recent interview that she was "alarmed at the way that the cable companies are being allowed to 'cherry-pick' the lucrative urban areas, leaving large tracts of the rural countryside untouched by the information revolution". She feels that "We should be imitating the US, where some States have refused to allocate franchises unless the cable companies were prepared to cable up the loss-making areas as well as the profitable ones".

A few years ago the idea of doing to BT what the US administration did to AT&T seemed to make sense – unfortunately the world has moved on.

Globalisation has happened and Europe has happened. BT is about the right size



'...alarmed at the way that the cable companies are being allowed to cherry-pick the lucrative urban areas, leaving large tracts of the rural countryside untouched by the information revolution'

for a regional operator if the European Telecoms market is taken as a whole. However, it is unlikely the UK government sees it this way. The EU sees it this way but is too afraid of upsetting anyone to knock the whole thing into shape – too many equipment suppliers on too many committees. The American model will not work in the UK alone – in America a local call costs only the connection charge – whether a call is 3 minutes or 24 hours it costs the same. The UK's local tariffs are based on a cost per minute which stunts the growth of on-line services. BT could provide US style local tariffs if it was allowed to provide services to subsidise them.

Cable companies may, eventually, provide the US model but perhaps for only a maximum of 70% of the population. Whether they make a profit in doing so or even in some cases survive remains to be seen. For some, the battle to wire the country with cable may end up being a switch too far.

Peter Kruger

The full interview with Ann Campbell quoted here is on-line and can be accessed on the World Wide Web at http://www.gold.net/flames/=20 – ed.

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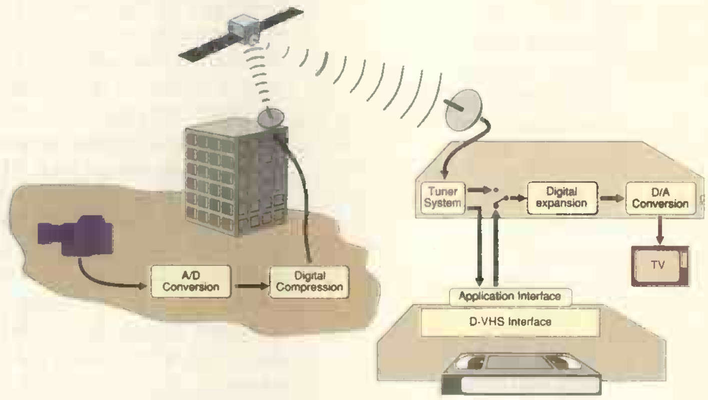
JVC announces D-VHS

JVC plans to launch a digital VHS system next year, for simple data stream recording and playback. It will not in itself produce pictures, but will have to be used in conjunction with other equipment to provide digital to analogue conversion. With encrypted broadcasts, such as DirecTV in the US, where it will first be marketed, the D-VHS VCR will record the compressed, encrypted data as received from the satellite, and play it back into the receiver at the same stage in the conversion process - before the signal is decrypted, re-expanded and converted to analogue for feeding to a TV set.

D-VHS uses virtually the same head mechanism as existing VHS and vcrs will still be able to play and record analogue material, though there will be no cross-over between the two modes. D-VHS capacity will add about £250 to the price of whatever type of vcr it is built into.

JVC believes D-VHS has computer and multimedia applications, but mainly as a back-up. "We are not trying to compete with the disc format in multimedia – disc has quick access, but tape has high capacity and low cost," said planning manager Kazuo Kohda.

Data capacity of a reusable E240-



D-VHS video recorders will record compressed, encrypted data as received from the satellite, and play it back into the receiver at the same stage in the conversion process – before the signal is decrypted, re-expanded and converted to analogue for feeding to a tv set.

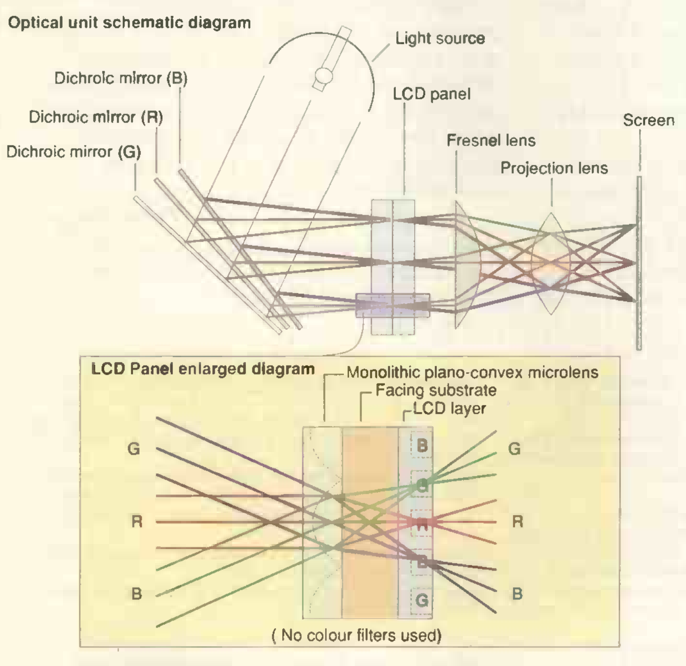
Single panel optical system

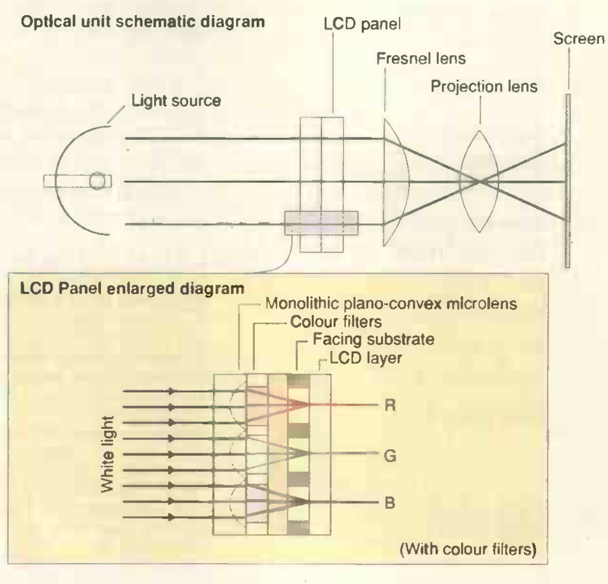
Video projectors have proved a boon to those needing a large screen display. Unfortunately, achieving a bright image has meant using a large and expensive system, while those without the space and cash have had to put up with viewing in a darkened room.

A new development from Sharp boosts light output of the compact and comparatively inexpensive single lcd panel projector. Incorporated in a recently launched projector unit, this technology involves a single-panel optical system featuring 'filter-less' technology. Instead of using a mosaic of red, green and blue filters over the pixels to provide colour from a white light source as is the case with conventional panels, the single panel system employs three dichroic mirrors to first separate the light into its primary colours and these pass through clear pixels.

The dichroic mirrors selectively reflect and transmit the light: the first

In conventional colour projectors, bottom right, rgb selection is done using colour filters inside the lcd panel. In the new Sharp projector, dichroic mirrors are used to separate the light into the three primary colours, eliminating the need for colour filters in the lcd panel. Will this bring down the cost of lcd projection systems?





length D-VHS cassette will be 44 gigabytes, compared with 10Gbyte on the proposed double-sided Toshiba DVD. The tape will be S-VHS quality ferric oxide.

Three recording modes are proposed: Standard, giving seven hours of recording at 14.1Mbit/s, suitable for MPEG-2; HD at 3.5 hours with doubled bit rate, 28.2 Mb/s, and LP at 2Mbit/s but with a recording time of 49 hours.

The LP mode, and the fact that D-VHS can store up to six simultaneous data streams, without loss of time, provided they are multiplexed before reaching the recorder, is already attracting the attention of the security industry, where a tape that can handle six cctv cameras, and only has to be changed once every two days has obvious advantages.

Support for the format has come from Thomson,Hitachi, Matsushita and Philips, all of whom have made technical contributions, and LG (Goldstar), Mitsubishi, Samsung, Sanyo, Sharp, Sony, Toshiba and major tape manufacturers.

Peter Willis

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reflects blue and transmits red and green to the second mirror which reflects the red and transmits the green to the third mirror which reflects it..

Angles of the individual mirrors relative to the 125W metal halide light source ensures the reflected primaries pass only through the appropriate pixels in the lcd panel – analogous to a shadow mask crt. A monolithic plano-convex microlens brings the light to a focus at the lcd layer and, having passed through the pixels, it goes to a fresnel lens and thence to the projection lens.

Gain in brightness achieved by this 'filter-less' technology is uncertain. Sharp claim it provides a four-fold increase over the preceding model. However, the nearest equivalent in their new range, the *XV-315P*, which employs the same 3.6in active matrix tft panel with 301,158 pixels. With filters, this unit gives a claimed luminance of 330lux at 30in screen size by comparison with 500lux at 40in for the *XV-370P*. The price premium is also uncertain: the *XV-315P* costs £1800, the XV-370P £2697, both including VAT – but the latter has more features. However, both share a 350-line horizontal resolution.

Telephone-line tv advances

Video compression and transmission technology must be available on a single chip costing less than $90 for video over telephone lines to become a commercial reality, according to Motorola.

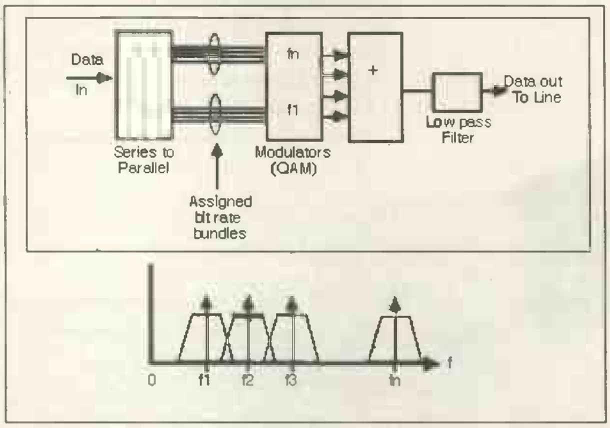
The company also believes it now has the technology to achieve this with the licensing of the discrete multi-tone modulation scheme, DMT, developed by specialist Californian designer Amarti Communications.

Amarti's DMT analogue line modulation scheme will be incorporated into a single chip transceiver for the asymmetrical digital subscriber line, ADSL, systems which operators like BT plan to use to deploy video-on-demand services operates over existing telephone lines.

Current ADSL systems, including those being evaluated by BT, support one-way video transmission to the subscriber using a 2Mbit/s digital channel. Amarti claims that its DMT-based technology will support a 6Mbit/s channel to the subscriber and a 640kbit/s return channel to the exchange. As well as supporting multiple tv channel transmission, an integrated DMT transceiver could reduce the cost of ADSL hardware. Motorola, which plans to market its first DMT chips in 1996, has set a target price for the silicon of under $90.

ADSL disproportionally limits the bandwidth of the telephone line between a narrow return channel and wideband subscriber channel. Its key advantage over alternative fixed band cable tv transmission schemes, such as quadrature amplitude modulation, QAM, is that the available line bandwidth is divided into a number of subchannels.

DMT generates 256 subchannels at 4kHz, each of which has be separately modulated with multiple carrier frequencies. A fast Fourier transform used to generate the 256 carrierless amplitude modulated and phase modulated subchannels. The compressed video data stream is divided between the subchannels



Discrete multi-tone, DMT, modulation optimises line bandwidth by dividing the channel into 256 sub-channels.

according to their available bandwidth, which can vary from 500kHz to 1MHz per subchannel due to the signal-noise performance of the copper pair cabling. As a result the digital capacity of the subchannel varies from one to 15 bits per symbol.

One possible drawback with DMT is the duplication of the dps functions across the 256 transmitters. But according to Amarti the FFT carrier generation is more efficient than the adaptive equalisation techniques used in fixed-band QAM transmission. The company suggests that its FFT-based ADSL design is five times less complex than a 200 tap equaliser needed to implement a 1.5Mbit/s QAM ADSL channel.

Richard Wilson, *Electronics Weekly*

Catseyes gain intelligence

An innovation from Doncaster-based r&d firm Astucia could save the European Union over £2bn, and more than 2500 lives per year, claims Martin Dicks, Astucia's managing director.

The invention is a light-emitting catseye, named Intelligent Road Stud, IRS, that can warn drivers of impending dangers on the road. The IRS circuit consists of couple of microcontrollers, a solar cell and an array of sensors powered by daylight, car headlights or a battery.

Depending the danger the IRS will emit red, blue, orange or white light respectively. Dicks said the DoT is interested in evaluating the device.

Silicon shortage scare

Another shortage scare looks set to hit the electronics industry, with reports that demand for polycrystalline silicon is about to outstrip supply.

Concern about the supply of polycrystalline silicon – the raw material for monocrystalline ingot production – is growing as the continuing boom in chip sales spurs semiconductor manufacturers to step up demand for wafers.

According to reports last week in Japanese trade paper *Japan Chemical Week*, the world demand for polycrystalline silicon this year is estimated to be 13,500 tonnes whereas total production is unlikely to exceed 12,000 tonnes.

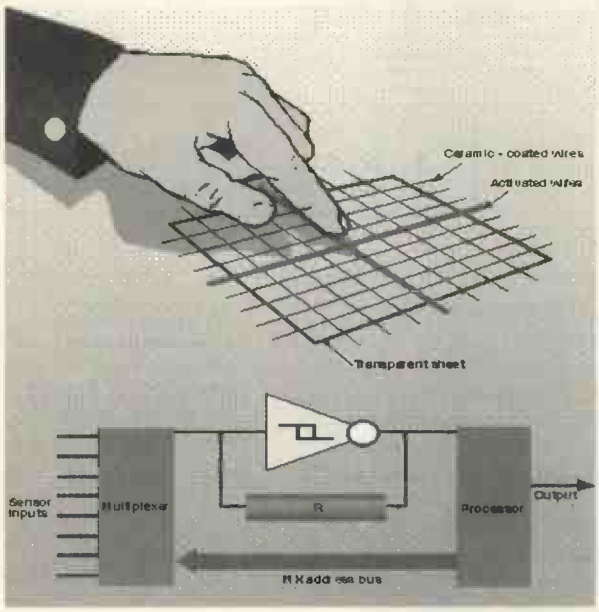
Ingo Reichel of wafer makers Freiberger Elektronikwerkstoffe, formerly the main wafer production operation for all of eastern Europe, said: We can see demand for polycrystal silicon rising and feel that there may not be enough on the market next year.

Polycrystalline silicon suppliers are reluctant to invest in new capacity as they were badly bitten in the silicon panic of ten years ago. Then they invested in new facilities to supply the expected surge in demand for 16 and 64kbit dynamic rams Unfortunately for them demand was entirely met by a three-fold increase in die yield on wafers.

The scare follows reports earlier this year that tantalum capacitors are in short supply, and last year's scare of a chip packaging shortage after a fire at the major resin supplier's factory. However, some observers suspect the shortage stories are a ploy designed to push up prices.

A source within the polycrystalline silicon supply industry said: "Polysilicon is not in such short supply that wafer manufacturers should worry." He went on to explain: "There is still room for process improvement and the major companies are expanding cautiously without discussing it."

Steve Bush, *Electronics Weekly*



A small UK r&d firm has developed a flexible keypad technology which it believes could be used to make robust, complex keypads for as little as £2 in volume.

Binstead Designs, based in Nottingham, is now looking for a major electronics firm to licence the technology for mass production. Binstead's keypads can be made as thin as 100μm or less and are flexible enough to be rolled up when not being used. They can be configured to trigger when a finger touches the keypad, or when a finger comes close to it. "It is sensitive enough to be used through double glazing," claims Ron Binstead, the company's founder.

The keypads are comprise a grid of 25μm thick ceramic-coated wires, embedded in a thin transparent plastic sheet. The wires are thin enough to make them virtually invisible.

The keypad detects when an area is activated by measuring the change in capacitance induced in a wire when a finger comes close to it. Each wire is scanned in turn by making it the capacitance in an RC oscillator. When capacitance of a wire changes, so does the the oscillation frequency. By finding which horizontal and vertical wires have been triggered the system can pinpoint the position of the user's finger.

The keypad is controlled from a small interface box, whose output goes to the RS232 port of a computer.

"The secret is in the processing," Binstead says. The change in signal can be as low as one per cent, but with temperature changes there are massive variations in the background levels. We have clever software in the sensing circuit that takes out these variables."

Karl Schneider

Bending superconductors around the corner?

US Government researchers claim to have created a breakthrough superconducting material that is flexible rather than brittle and can be used in a wide number of applications.

Scientists at the Los Alamos National Laboratory in New Mexico, described the new material at a meeting of the Materials Research Society in San Francisco recently.

They demonstrated a flexible metal and ceramic foil that they said can be made into wires with a huge current carrying capacity at liquid nitrogen temperatures.

The superconductor is a ceramic material based on yttrium barium copper oxide deposited on a nickel tape to give it flexibility. Previous superconducting materials have been too brittle to form wires.

Dean Peterson, head of the Los Alamos National Laboratories' Superconductivity Center, said that the superconducting material can carry more than a million amperes per square centimetre compared with No 12 copper wire that carries 800A/cm^2.

Blue lasers get the green light

The US Advanced Research Projects Agency is funding e development of blue-light laser diodes by Philips subsidiary, Philips Laboratories and Cree Research.

ARPA has given a $4m grant for a two year project to develop blue light laser diodes based on gallium-nitride materials on silicon carbide wafers. One focus of the project will be to develop higher capacity optical data storage devices.

"In a systems market that is constantly looking for ways to increase storage capacity, the blue laser is a significant missing link," said Neal Hunter, president of Cree Research.

Other firms are also trying to find ways to build blue laser diodes cheaply. Advanced Technology Materials is working with Hewlett-Packard to develop blue laser diodes, also using gallium nitride. Japanese firm Nichia Chemical Industries says it is already sampling blue laser diodes for about $30.

First distributor on the Net

SEI is the first European distributor to use Internet to make its product information available to customers across Europe. The company believes the Internet project will ultimately transform pan-European distribution. It will change how customers internationally do their business," said Wim Teunissen, a member of SEI's European managing board. A full system including on-line pricing and ordering will be available before the end of the year. ■

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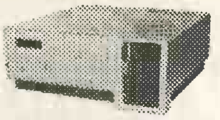

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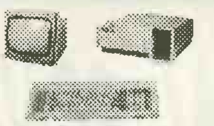

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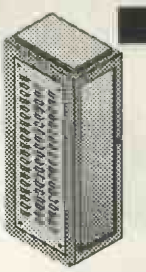

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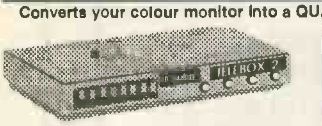

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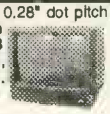

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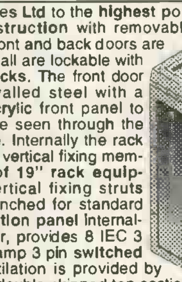

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CIRCLE NO. 107 ON REPLY CARD

Jonathan Campbell

Fibre comms comes to the front door

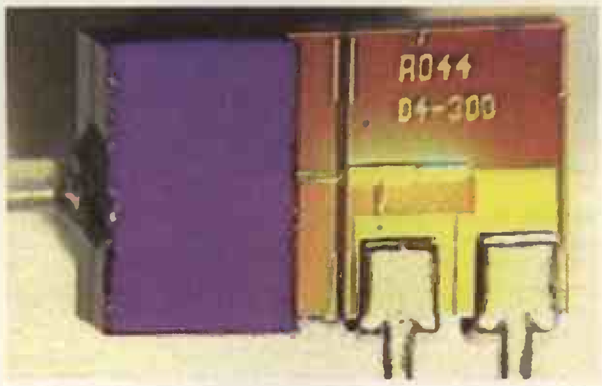
Simple laser/fibre linking technology, a tenth the cost of current approaches but putting into a fibre more than seven times the level of light than competing systems, has been developed by BT Laboratories, Martelsham (BTL). Advantage of the technique is that it overcomes the alignment problems which dog today's systems and make fibre links so expensive. With such a dramatic reduction in price – down from £100 to perhaps £10 – the prospect of high speed fibre optics finding their way into individual homes has come several steps closer. For consumers that means huge increases in the volumes of data they will be able to access down the telephone line.

The reason fibre optic links are so costly now is that optically-efficient attachment of the semiconductor laser to the fibre is difficult. To work effectively, optical fibres must be aligned to within less than 1μm. An offset of only 1.2μm halves the amount of light that can be coupled into the fibre. Normally this requires expensive active alignment, with each laser having to be turned on while the fibre is moved around in front of it to maximise the coupled light. Fixing the fibre in place then involves computer-controlled welding using a high power laser.

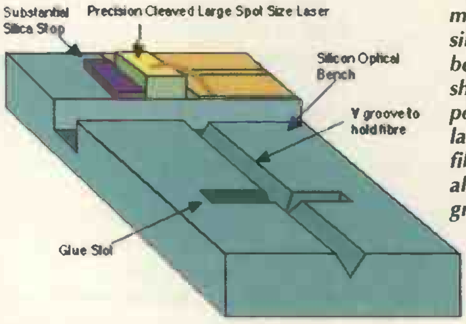
But BTL has redesigned the shape of the laser to incorporate of a taper allowing light to coupled directly to the cleaved fibre with efficiencies around 50%. At the same time, a special cleaving technique allows the position of the laser active region relative to the edge of the chip to be known to 0.25μm.

The final task has been developing a micro-machined silicon mount to sit the laser on. This incorporated a silica stop, to which the laser is aligned simply by pushing it in until contact is made, and a precision etched V-groove in which the fibre may be glued.

Ian Lealman, part of the BTL team, says so far research samples have been tested and the search is on for a 'down-stream' comms company to develop the technology.



An assembled silicon optical bench. The semiconductor laser is the small gold rectangle in the middle of the bench.



Micro machined silicon optical bench showing the position of the laser chip and fibre alignment groove.

Organic leds: colour limits pushed aside

Researchers in Japan and Sweden have pushed back the old colour limits on leds just a little further with announcement of an organic white light device that could glow as brightly as a fluorescent tube, and a polymer blend led that promises emission of any colour simply by adjusting the voltage.

Various multi-layer systems have been proposed before to obtain different colours. But white-light devices have always caused a problem because of the dearth of white fluorescent dyes.

Now Junji Kido, Masato Kimura and Katsutoshi Nagai at Yamagata University in Japan have used thin film technology to create a device that simultaneously emits blue, green and red wavelengths to produce bright white light (Multi-layer white light-emitting organic electroluminescent device, *Science*, 267, pp.1332-1334).

Conventional leds comprise an emitter layer and a carrier transport layer. But doping the emitter layer with a different coloured fluorescent dye can produce light that is mix of the two emissions.

Carrier recombination can also be controlled so that emission takes place in two different layers. A hole-blocking layer inserted between the electron transport layer and hole transport layer, can force carrier recombination – and so light emission – to occur in both layers.

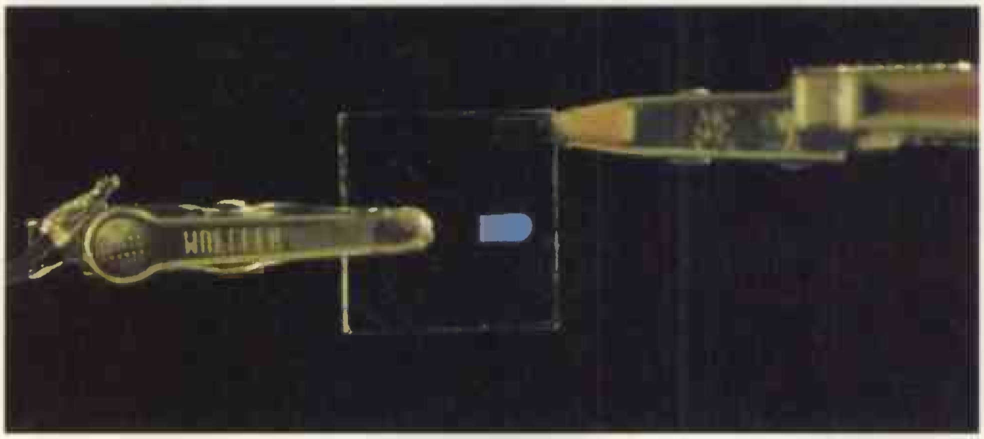
The Japanese white light led puts both methods to work.

Onto a hole-injecting indium-tin-oxide-coated glass substrate (ITO), are vacuum deposited a series of layers beginning with TPD (triphenyl-diamine derivative) showing emissions at 410-420nm, in the blue region. Next comes a 1,2,4-triazole derivative layer that transports electrons but blocks holes; and this is followed by three layers of an electron-transporting aluminium complex (Alq) that emits at 520nm, green. The middle Alq layer is also doped with nile red, emitting at 600nm. Finally, a magnesium-silver alloy is used as the hole injecting electrode.

Applying dc voltage, with ITO positive, produces white light visible through the glass substrate.

The researchers report that luminance starts at around 6V, improving up to a maximum of 2200cd/m^2 at 16V. Optimisation of structure and materials could lead to devices exceeding the 8000cd/m^2 of fluorescent lamps.

Uses for such white light devices include lightweight applications such as aircraft or space shuttles. But they could be useful as backlights



Research in Sweden into polymer blends could enable any colour to be generated simply by changing the voltage.

for liquid crystal displays and, with suitable micropatterned colour filters, in full colour displays too. Another led advance aimed squarely at colour displays is a polymer-blend device being developed by Magnus Berggren and colleagues in Sweden (Light-emitting diodes with variable colours form polymer blends, *Nature*, 372, pp.444-446).

His team's design strategy has been to control the geometry of a thiophene polymer main chain, producing a family of blends combining materials with different band gaps. Colours ranging from blue to near infra-red, with green, orange and red as intermediate steps can be produced, with intensity ratio of the peaks being determined by the voltage applied and stochiometry of the polymer blend.

So far, the precise mechanism for the phenomenon has not been positively established, but the researchers say they can easily combine colours such as red and blue, green and red, orange and blue and expect soon to be able to combine red, green and blue.

When perfected, the simplicity of forming multi-colour screens with passive addressing of individual multi-colour pixels could make the technology irresistible to display engineers.

Nanowires must still have a little flab

Scientists at Georgia Tech in the US and Universidad Autonoma de Madrid in Spain are warning researchers that there are limits to how small the wires can be made in miniaturised components. Those sizes are still very small – electrical, mechanical and other properties of microscopic wires only change significantly as their width narrows to the nanoscale – less than ten atoms.

"Small is different", says Uzi Landman, director of Georgia Tech's Center for Computational Material Science.

The researchers found that under certain conditions, the ability of the nanowires to conduct electricity declines to the point that they resemble insulators. Conductance of such atomic-scale gold wires depends on their length, lateral dimensions, the state of atomic order and disorder and the elongation mechanism of the wires.

"If we are to reduce the size of microelectronics systems, connecting wires between elements of such devices must be reduced in size and therefore such quantisation patterns of conductance could start to appear," says Landman.

The combined experimental and theoretical investigations of electronic transport and mechanical elongation in ultra-thin metallic wires carried out by the scientists are the first to measure, in three-dimensional wires at room temperature, a localisation phenomena previously seen only in one-dimensional "whiskers" at cryogenic temperatures.

The phenomena are expected to occur when the physical dimensions of the systems approach that of the electronic wavelength.

Adding a spark to video games

Aura Systems has developed a vest that vibrates in response to sounds from video games in an attempt to make the play more lifelike.

Picture Sega Saturn.

Virtual reality may have brought added realism to computer combat games. But there is still one aspect of life in silicon city that just doesn't ring true: where's the pain? Exchanging karate kicks with on-screen adversaries is only another empty experience without the physical jolt of heads cracking and ribs breaking.

Fortunately, El Segundo-based Aura Systems may have shown us the solution.

Aura has developed a special combat vest to be worn during game playing. It responds to sound, so that when, for example, a fist thuds into vital organs, the vest vibrates to give the player a stronger taste of the action. Unfortunately, some of the first kids to try out the new hardware were a little less than grateful, plainly expecting more from a former defence company. One teenage tester commented disappointedly that it was hard to tell a punch from a cheering crowd.

So the race is still on to develop a computer peripheral that can convey to the compulsive electronic combat kid some of the real fun and excitement of going to war, whether it's with alien invaders or local bandanna-wearing street fighters.

Maybe *EW + WW* readers could connect up something these desperate children really need. Though surely those reaching for their March issue and the article on Tesla coils have got completely the wrong idea.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New leds promising for flat panel displays

Organic leds demonstrating a 30-fold improvement in stability and significantly lower operating voltages – and so power consumption – have been developed by researchers at AT&T Bell Laboratories in the US.

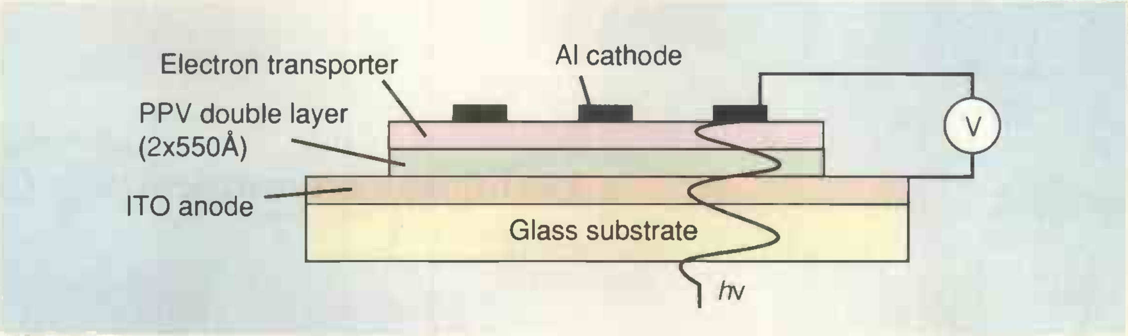
The improvements were achieved in a class of devices where an electron transporting layer (ET) is incorporated into the design to improves quantum efficiency by confining holes to the emissive layer and ensuring that both holes and electrons are generated. The ET layer also boosts power efficiency by aiding electron injection from the cathode.

What the Bell team has done (*Science*, Vol 267, pp.1969-1971) is to develop a new ET material that boosts power efficiency by almost a factor of 10, producing devices that have a low turn-on voltage of 6 to 10V, compared to 30V normally. The figure is similar to that for devices without an ET layer, but of course with all the efficiency advantages of the layer retained. The researchers also established conclusively that the most important factor in determining diode stability is the electron transporter used.

Several new ET materials were investigated as part of the study and compared with conventional ET layers. Best performance was achieved with a poly(aryl ether) layer and this compound was also able to pass as much as 3A/cm^2 before failure.

All the leds were composed of the two thin layers of organic material – PPV (poly(*p*-phenylenevinylene) and the polymeric ET layer – sandwiched between indium tin oxide (ITO) and aluminium electrodes.

The scientists say the improvements demonstrate the strong promise of this type of device in indicators and flat panel displays etc.



PPV and the new organic ET layer sandwiched between ITO and aluminium electrodes. The device could be ideal for flat panel displays.

Video stills lose their fuzz

Many video enhancement techniques take account of motion that occurs between frames. But a new technique developed by workers at the University of Rochester, New York, and Eastman Kodak attempts to compensate for movements *within* a single frame.

Results are said to dramatically improve picture quality and allow single images from fast moving scenes to be output to a printer without blurring. Such a facility is likely to become more important with the growing integration between tvs, videos and computer systems. The method should also reduce some of the drawbacks experienced in transferring film to tv, problems that are not discernible on a normal tv set but which could become apparent on hdtv.

Rochester postgrad Andrew Patti, who with a Kodak colleague has filed four patents related to the technique, summarises how the process operates: " The one image you want is related to all the images before and after it. We extract that information and use it to clarify our image", he says.

The technique could also find application in forensics and satellite imaging; or anywhere there is need to generate a clear frame from a video. ■

Defuzzing video still images by compensating for movement within a frame produces dramatic results.

Trimodal Audio power

I present here my own contribution to global warming in the form of an improved Class-A amplifier that I believe is unique. It not only copes with load impedance dips by means of an unusually linear form of Class-AB, but will also operate as a 'blameless' Class-B engine. The power output in pure Class-A is 20 to 30W into 8Ω, depending on the exact supply rails chosen.

Initially, I simply intended to provide an updated version of the Class-A circuit published in reference 1, in response to requests for a pcb for the Class-A amplifier designed with my methodology. I decided to use a complementary-feedback-pair, or cfp, output stage for best possible linearity, and some incremental improvements have been made to noise, slew rate and maximum dc offset.

***Douglas Self's** latest audio power amplifier can be switched between Class-A/AB or Class-B to provide remarkable performance over a wide range of operating conditions. In Class-A, it operates with ultra-low distortion, but presented with a low-impedance load, it has recourse to an unusually linear AB configuration.*

Naturally, the Class-A circuit bears a very close resemblance to a 'blameless' Class-B amplifier. As a result, I decided to retain the Class-B V_{be} multiplier, and use it as a safety-circuit to prevent catastrophe if the relatively complex Class-A current-regulator failed. From this the idea arose of making the amplifier instantly switchable between Class-A/AB and Class-B modes. This gives two kinds of amplifier for the price of one, and permits of some interesting listening tests. Now you really can do an A/B comparison...

In the Class-B mode the amplifier has the usual negligible quiescent dissipation, but in Class-A the thermal efflux is naturally considerable. This is because true Class-A operation is extended down to 6Ω resistive loads for the full output voltage swing, by suitable choice of the quiescent current.

With heavier loading the amplifier gracefully enters Class-AB, in which it will give full output down to 3Ω before the safe-operating-area, SOAR, limiting begins to act. Output into

2Ω is severely curtailed, as it must be with only one output pair, and this kind of load is not advisable.

In short, the amplifier allows a choice between being firstly very linear all the time – blameless Class-B – and secondly ultra-linear most of the time – Class-A – with occasional excursions into Class-AB.

The amplifier's AB mode is still extremely linear by current standards, though inherently it can never be as good as properly-handled Class-B, and nothing like as good as A. Since there are three possible classes of operation I have decided to call the design a Trimodal power amplifier. It is impossible to be sure that you have read all the literature on an area of technology; however, to the best of my knowledge this is the first ever Trimodal amplifier.

As I said earlier, designing a low-distortion Class-A amplifier is in general a good deal simpler than the same exercise for Class-B. All the difficulties of arranging the best possible crossover between the output devices disappear. Because of this it is hard to define exactly what 'blameless' means for a Class-A amplifier.

In Class-B the situation is quite different, and 'blameless' has a very specific meaning; when each of the eight or more distortion mechanisms has been minimised in effect, there always remains the crossover distortion inherent in Class-B. There appears to be no way to reduce it without departing radically from what might be called the generic Lin amplifier configuration. Therefore the 'blameless' state appears to represent some sort of theoretical limit for Class-B, but not for Class-A.

However, Class-B considerations cannot be ignored, even in a design intended to be Class-A only, because if the amplifier does find itself driving a lower load impedance than expected, it will move into Class-AB. In this case, all the additional Class-B requirements are just as significant as for a Class-B design proper. Class-AB can never give distortion as low as optimally-biased Class-B, but it can be made comparable if the extra distortion mechanisms are correctly handled.

My correspondence has made it abundantly clear that *EW* readers are not going to be satisfied with anything less than state-of-the-art linearity, and so the amplifier described here uses the complementary-feedback-pair type of output stage, which has the lowest distortion due to the local feedback loops enclosing the output devices. It also has the advantage of better output efficiency than the emitter-follower version, and inherently superior quiescent current stability. It will shortly be seen that these are both important for this design.

Half-serious thought was given to labelling the Class-A mode 'distortionless' as the thd is completely unmeasurable across most of the audio band. However, detectable distortion products do exist above 10kHz, so sadly, I abandoned this provocative idea.

Before putting cursor to CAD, it seemed appropriate to take another look at the Class-A design, to see if it could be inched a few steps nearer perfection. The result is a slight improvement in efficiency, and a 2dB improvement in noise performance. In addition the expected range of output dc offset has been reduced from ±50mV to ±15mV, still without any adjustment.

The power and the glory

The amplifier is 4Ω capable in both A/AB and B operating modes, though it is the nature of things that the distortion performance is not quite so good. All solid-state amplifiers – without qualification, as far as I am aware – are much happier with an 8Ω load, both in terms of linearity and efficiency; loudspeaker designers please note.

With a 4Ω load, Class–B operation gives better thd than Class-A/AB, because the latter will always be in AB mode, and therefore generating extra output stage distortion through gm-doubling. This should really be called gain-deficit-halving, but somehow I don't see this term catching on. These not entirely obvious relationships are summarised on the right.

Figure 1 attempts to show diagrammatically just how power, load resistance, and operating mode are related. The rails have been set to ±20V, which just allows 20W into 8Ω in Class-A. The curves are lines of constant power, ie $V{\times}I$ in the load, the upper horizontal line represents maximum voltage output, allowing for $V_{ce(sat)}$s, and the sloping line on the right is the SOAR protection locus; the output can never move outside this area in either mode. The intersection between the load resistance lines sloping up from the origin and the ultimate limits of voltage-clip and SOAR protection define which of the curved constant-power lines is reached.

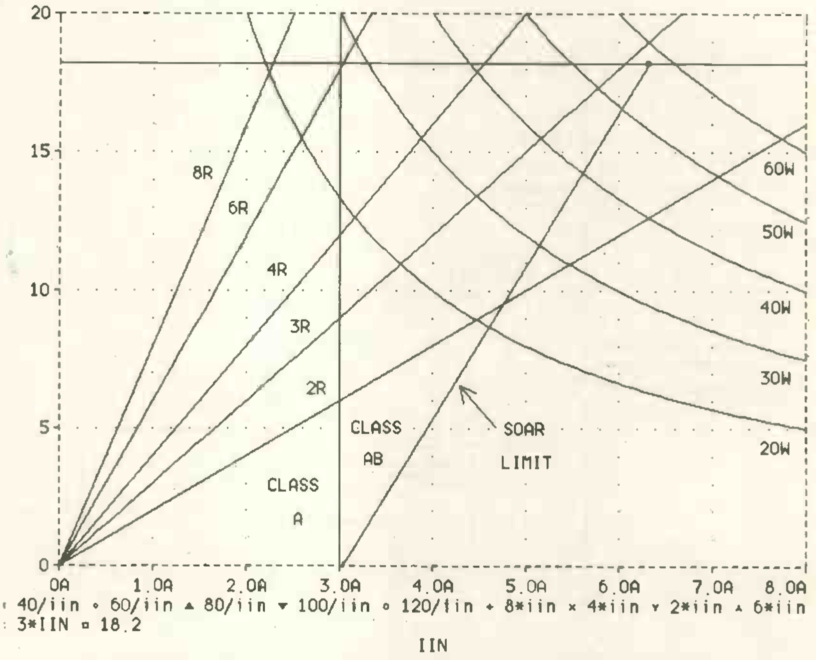


Fig. 1. Relationships between load, mode, and power output. The intersection between the sloping load resistance lines and the ultimate limits of voltage-clipping and SOAR protection define which of the curved constant-power lines is reached. In A/AB mode, the operating point must be to the left of the vertical push-pull current-limit line for true Class-A.

In A/AB mode, the operating point must be left of the vertical push-pull current-limit line (at 3A, ie twice the quiescent current) for Class-A. If we move along one of the impedance lines, when we pass to the right of the push-pull limit the output devices will begin turning off for part of the cycle; this is the AB operation zone. In Class-B mode, the 3A line has no significance and the amplifier remains in optimal Class-B until clipping or SOAR limiting occurs. Note that the diagram axes represent instantaneous power in the load, but the curves show sine-wave rms power, and that is the reason for the apparent factor-of-two discrepancy between them.

Health and efficiency

Concern for efficiency in Class-A may seem paradoxical, but one way of looking at it is that Class-A watts are precious things, wrought in great heat and dissipation, and so for a given quiescent power it makes sense to ensure that the amplifier approaches its limited theoretical efficiency as closely as possible. I was confirmed in this course by reading

Load	Mode	Distortion	Dissipation
8Ω	A/AB	very low	high
4Ω	A/AB	high	high
8Ω	B	low	low
4Ω	B	medium	medium

Note that in the context of this sort of amplifier, 'high' means about 0.002% thd at 1kHz and 0.01% at 10kHz.

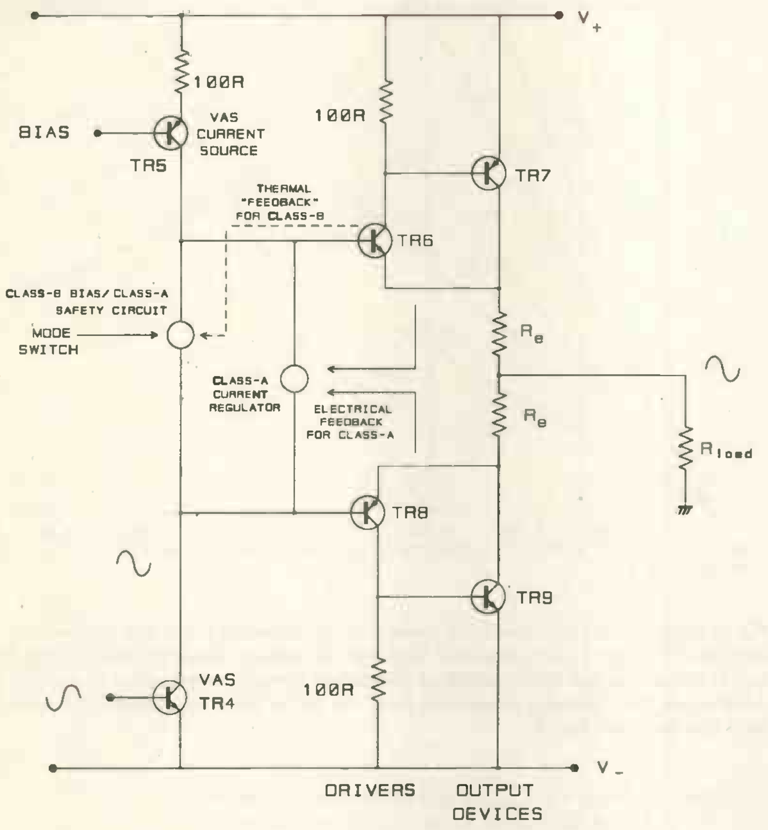


Fig. 2. Basic current feedback output stage, equally suited to operating Class B, AB and A, depending the magnitude of V_{bias}. The emitter resistors R_e may be from 0.1 to 0.47Ω.

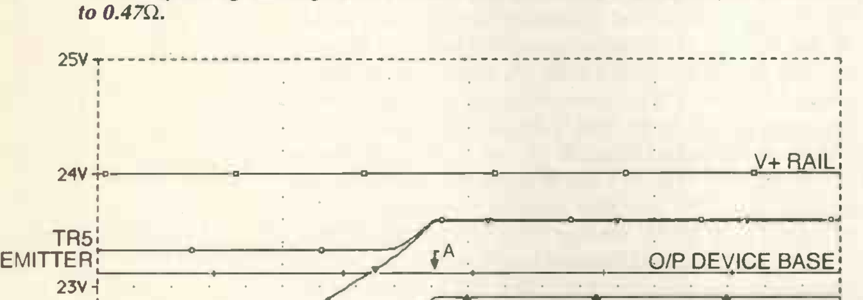


Fig. 3. PSpice simulation showing how positive clipping occurs in the current feedback output. A higher sub-rail for the voltage amplifier cannot increase the output swing, as the limit is set by the minimum driver V_{ce}, and not the voltage amplifier output swing.

of another recent design[2] which seems to throw efficiency to the winds by using a hybrid bjt/fet cascode output stage. The voltage losses inherent in this arrangement demand ±50V rails and sixfold output devices for a 100W Class-A capability; such rail voltages would give 156W from a 100% efficient amplifier.

Voltage efficiency of a power amplifier is the fraction of the supply-rail voltage which can actually be delivered as peak-to-peak voltage swing into a specified load; efficiency is invariably less into 4Ω due to the greater resistive voltage drops with increased current.

The Class-B amplifier I described in reference 3 has a voltage efficiency of 91.7% for positive swings, and 92.5% for negative, into 8Ω. Amplifiers are not in general completely symmetrical, and so two figures need to be quoted; alternatively the lower of the two can be given as this defines the maximum undistorted sine-wave. These figures above are for an emitter-follower output stage, and a complementary-feedback pair output does better, the positive and negative efficiencies being 94.0% and 94.7% respectively.

The emitter follower version gives a lower output swing because it has two more V_{be} drops in series to be accommodated between the supply rails; the complementary-feedback pair is always more voltage-efficient, and so selecting it over the emitter follower for the current Class-A design is the first step in maximising efficiency.

Figure 2 shows the basic complementary-feedback pair output stage, together with its two biasing elements. In Class-A the quiescent current is rigidly controlled by negative-feedback; this is possible because in Class-A the total voltage across both emitter resistors R_e is constant throughout the cycle. In Class-B this is not the case, and we must rely on 'thermal feedback' from the output stage, though to be strictly accurate this is not 'feedback' at all, but a kind of feedforward.

It is a big advantage of the complementary-feedback pair configuration that quiescent current, I_q depends only on driver temperature, and this is important in the Class-B mode, where true feedback control of quiescent current is not possible. This has special force if low-value emitter resistors such as 0.1Ω, are chosen, rather than the more usual 0.22Ω; the motivation for doing this will soon become clear.

Voltage efficiency for the quasi-complementary Class-A circuit of reference 1 into 8Ω is 89.8% positive and 92.2% negative. Converting this to the complementary-feedback pair output stage increases this to 92.9% positive and 93.6% negative. Note that a Class-A I_q of 1.5A is assumed throughout; this allows 31W into 8Ω in push-pull, if the supply rails are adequately high. However the assumption that loudspeaker impedance never drops below 8Ω is distinctly doubtful, to put it mildly, and so as before this design allows for full Class-A output voltage swing into loads down to 6Ω.

So how else can we improve efficiency? The addition of extra and higher supply rails for the small-signal section of the amplifier surprisingly does not give a significant increase in output; examination of Fig. 3 shows why. In this region of operation, the output device Tr_7 base is at a virtually constant 880mV below the positive rail, and as Tr_6 driver base rises it passes this level, and keeps going up; clipping has not yet occurred.

The driver emitter follows the driver base up, until the voltage difference between this emitter and the output base, ie the driver V_{ce}, becomes too small to allow further conduction; this choke point is indicated by the arrows A-A. At this point the driver base is forced to level off, although it is still about 500mV below the level of the positive rail. Note also how the voltage between the positive rail and Tr_5 emitter collapses. Thus a higher rail will give no extra voltage swing, which I must admit came as something of a surprise. Higher sub-rails for small-signal sections only come into their own in fet amplifiers, where the high V_{gs} for fet conduction (5V or more) makes their use almost mandatory.

Efficiency figures given so far are all greater for negative rather than positive voltage swings. The approach to the rail for negative clipping is slightly closer because there is no equivalent to the 0.6V bias established across R_{13}; however this advantage is absorbed by the need to lose a little voltage in the *RC* filtering of the negative supply to the current-mirror and voltage amplifier stage. This filtering is essential if really good ripple/hum performance is to be obtained.[3]

In the quest for efficiency, an obvious variable is the value of the output emitter resistors R_e. The performance of the current-regulator described, especially when combined with a complementary-feedback pair output stage, is more than good enough to allow these resistors to be reduced while retaining first-class I_q stability. I took 0.1Ω as the lowest practicable value, and even this is comparable with pcb track resistance, so some care in the exact details of physical layout is essential; in particular the emitter resistors must be treated as four-terminal components to exclude unwanted voltage drops in the tracks leading to the resistor pads.

If R_e is reduced from 0.22Ω to 0.1Ω then voltage efficiency improves from 92.9%/93.6%, to 94.2%/95.0%. Is this improvement worth having? Well, the voltage-limited power output into 8Ω is increased from 31.2 to 32.2W with ±24V rails, at absolutely zero cost, but it would be idle to pretend that the resulting increase in sound-pressure level is highly significant. It does however provide the philosophical satisfaction that as much Class-A power as possible is being produced for a given dissipation; a delicate pleasure.

The linearity of the complementary-feedback pair output stage in Class-A is very slightly worse with 0.1Ω emitter resistors, though the difference is small and only detectable open-loop; the simulated thd of an output stage alone (for 20V pk-pk in 8Ω) is only increased from 0.0027% to 0.0029% This is probably due simply to the slightly lower total resistance seen by the output stage.

However, at the same time, reducing the emitter resistors to 0.1Ω provides much lower distortion when the amplifier runs out of Class-A; it halves the size of the step gain changes inherent in Class-AB, and so effectively reduces distortion into 4Ω loads.

Figures 4 & 5 are output linearity simulations; the measured results from a real and 'blameless' Trimodal amplifier are shown in **Fig. 6**, where it can be clearly seen that thd has been halved by this simple change. To the best of my knowledge this is a new result; my conclusion is that if you must work in Class-AB, keep the emitter resistors as low as possible, to minimise the gain changes.

Having considered the linearity of Class-A and AB, we must not neglect what effect this radical Re change has on Class-B linearity. The answer is, not very much, but there is a slight reduction in thd, **Fig. 7**, where crossover distortion seems to be slightly higher with R_e at 0.2Ω than for either 0.1 or 0.4Ω. Whether this is a consistent effect – for complementary-feedback pair stages anyway – remains to be seen.

The detailed mechanisms of bias control and mode-switching are described in the second part of this article.

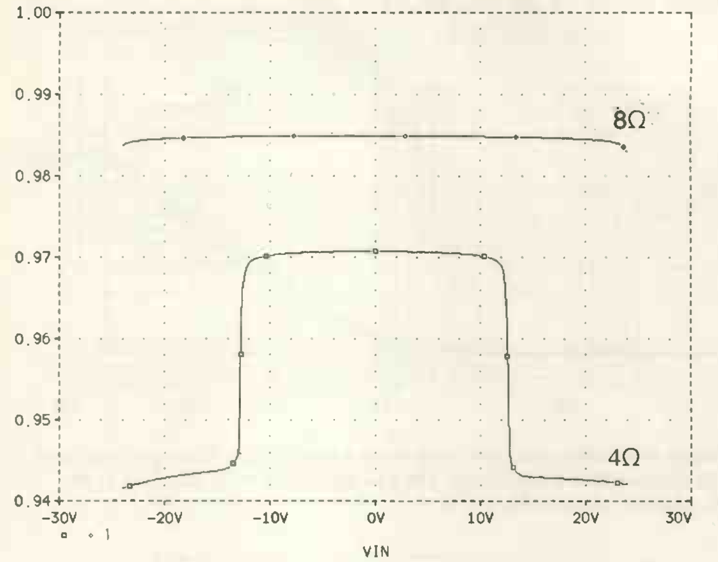


Fig. 4. Complementary feedback pair output stage linearity with R_e set at 0.22Ω. Upper trace is Class-A into 8Ω, lower is Class-AB operation into 4Ω, showing step changes in gain of 0.024 units.

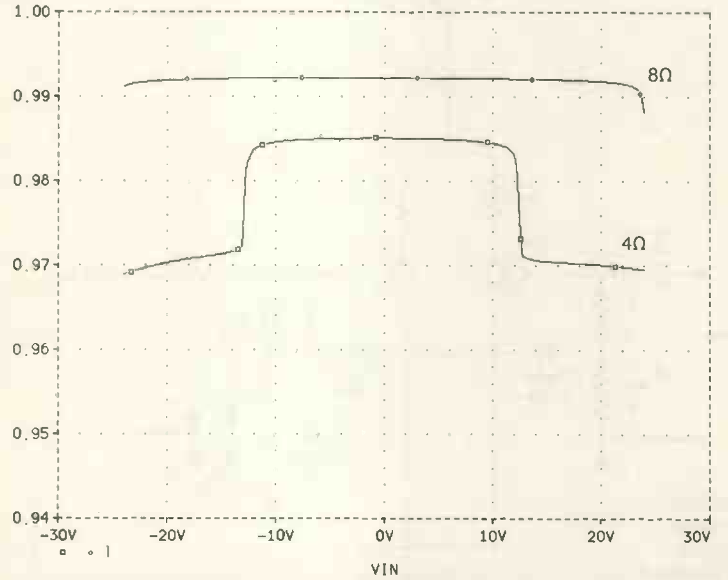


Fig. 5. Current feedback output linearity with R_e set at 0.1Ω, re-biased to keep Iq at 1.5A. There is slightly poorer linearity in the flat-topped Class-A region than for an R_e of 0.22Ω, but the 4Ω AB steps are halved in size at .012 units. Note that both gains are now closer to unity; same scale as Fig. 4.

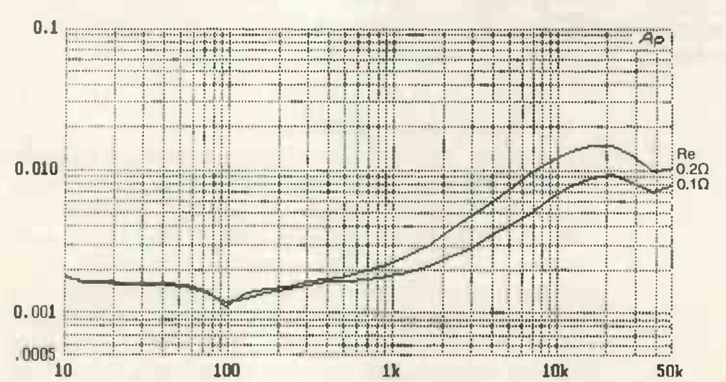


Fig. 6. Proving that emitter resistor value really matters in Class-AB. Output was 20W in 4Ω, so amplifier was leaving Class-A for about 50% of the time. Changing emitter resistors from 0.2 to 0.1Ω halves the distortion. Current I_q is 1.5A for both cases.

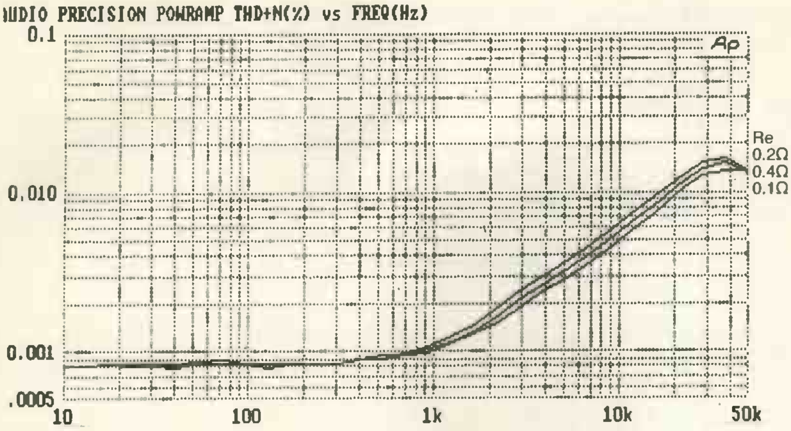


Fig. 7. Proving that emitter resistors matter much less in Class-B. Output was 20W in 8Ω, with optimal bias. Interestingly, the bias does NOT need adjusting as the value of R_e changes. Bandwidth 80kHz.

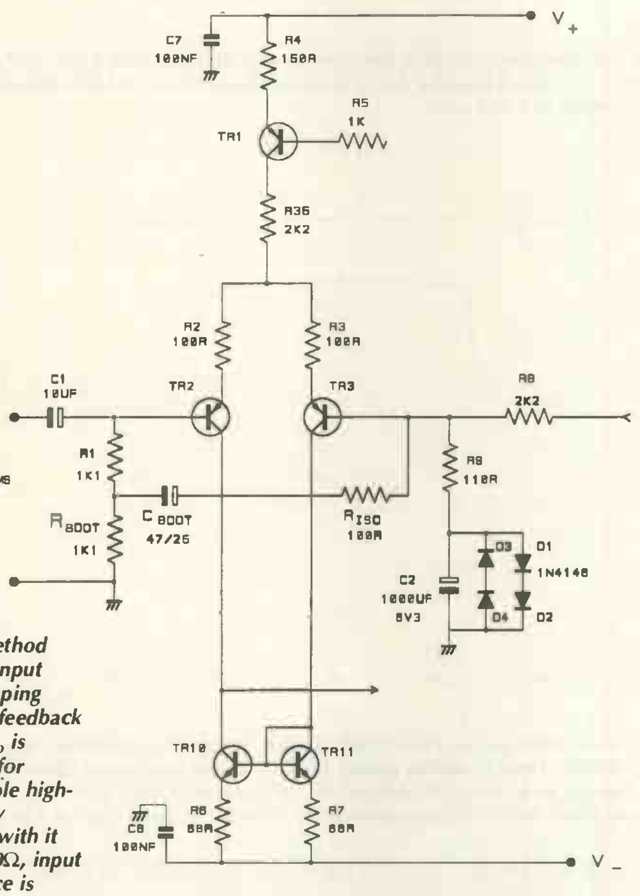


Fig. 8. Method used for input bootstrapping from the feedback point. R_{iso} is essential for dependable high-frequency stability; with it set to 100Ω, input impedance is 13kΩ.

Improving noise performance

In a power amplifier, noise performance is not an irrelevance.[4] It is well worth examining just how good it can be. As in most amplifiers, noise is set here by a combination of the active devices at the input and the surrounding resistances.

Operating conditions of the input transistors themselves are set by the demands of linearity and slew-rate, and there is little freedom of design here; however the collector currents are already high enough to give near-optimal noise figures with the low source impedances – a few hundred ohms – that we have here, so this is not too great a problem. Also remember that noise figure is a weak function of I_c, so minor tweaking makes no detectable difference. We certainly have the choice of input device type; there are many more possibilities now that we have relatively low rail voltages. Noise performance is, however, closely bound up with source impedance, and we need to define this before device selection.

Looking therefore to the passives, there are several resistances generating Johnson noise in the input, and the only way to reduce this noise is to reduce them in value. The obvious candidates are input stage degeneration resistors $R_{2,3}$ and R_9, which determines the output impedance of the negative-feedback network. There is also another unseen component; the source resistance of the preamplifier or whatever upstream.

Even if this equipment were miraculously noise-free, its output resistance would still generate Johnson noise. If the preamplifier had, say, a 20kΩ volume pot at its output – not a good idea, as this gives a poor gain structure and cable dependent hf losses, but that is another story[5] – then the source resistance could be a maximum of 5kΩ, which would almost certainly generate enough Johnson Noise to dominate the power-amplifier's noise behaviour. However, there is nothing that power-amp designers can do about this, so we must content ourselves with minimising the noise-generating resistances we do have control over.

The presence of input degeneration resistors $R_{2,3}$ is the price we pay for linearising the input stage by running it at a high current, and then bringing its transconductance down to a useable value by adding linearising local negative feedback. These resistors cannot be reduced, for if the hf negative-feedback factor is then to remain constant, C_{dom} would have to be proportionally increased, with a consequent reduction in slew rate. Used with the original negative feedback network, these resistors degrade the noise performance by 1.7dB. Like all the other noise measurements given here, this figure assumes a 50Ω external source resistance.

If we cannot alter the input degeneration resistors, then the only course left is the reduction of the feedback network impedance, and this sets off a whole train of consequences. If R_8 is reduced to 2.2kΩ, then R_9 becomes 110Ω, and this reduces noise output from –93.5dBu to –95.4dBu. Note that if $R_{2,3}$ were not present, the respective figures would be –95.2 and –98.2dBu. However, R_1 must also be reduced to 2.2kΩ to maintain dc balance, and this is too low an input impedance for direct connection to the outside world.

If we accept that the basic amplifier will have a low input impedance, there are two ways to deal with it. The simplest is to decide that a balanced line input is essential; this puts an opamp stage before the amplifier proper, buffers the low input impedance, and can provide a fixed source impedance to allow the high and low-frequency bandwidths to be properly defined by an *RC* network using non-electrolytic capacitors. The common practice of slapping an *RC* network on an unbuffered amplifier input must be roundly condemned as the source impedance is unknown, and so therefore is the roll-off point. A major stumbling block for subjectivist reviewing, one would have thought.

The other approach is to have a low resistance dc path at the input but maintain a high ac impedance; in other words to use the fine old practice of input bootstrapping. Now this requires a low-impedance unity-gain-with-respect-to-input point to drive the bootstrap capacitor, and the only one available is at the amplifier inverting input, ie the base of Tr_3. While this node has historically been used for the purpose of input bootstrapping[6] it has only been done with simple circuitry employing very low feedback factors.

There is good reason to fear that any monkey business with the feedback point, at Tr_3's base, will add shunt capacitance, creating a feedback pole that will degrade hf stability. There is also the awkward question of what will happen if the input is left open-circuit...

Figure 8 shows how the input can be safely bootstrapped.

The total dc resistance of R_1 and R_{boot} equals R_8, and their centre point is driven by C_{boot}. Connecting C_{boot} directly to the feedback point did not produce gross instability, but it did seem to increase susceptibility to sporadic parasitic oscillation. Resistor R_{iso} was added to isolate the feedback point from stray capacitance: this seemed to effect a complete cure.

The input could be left open-circuit without any apparent ill-effects, though this is not exactly good practice if loudspeakers are connected. A value for R_{iso} of 220Ω increases the input impedance to 7.5kΩ, and 100Ω raises it to 13.3kΩ, safely above the 10kΩ standard value for a bridging impedance. Despite successful tests, I must admit to a few lingering doubts about the high-frequency stability of this approach, and it might be as well to consider it as experimental until more experience is gained.

Another consequence of a low-impedance negative feedback network is the need for feedback capacitor C_2 to be proportionally increased to maintain the low-frequency response, and prevent capacitor distortion from causing a rise in thd at low frequencies; it is the latter constraint that determines the value. This is a separate distortion mechanism from the seven previously considered, and I think deserves the title Distortion 8. This criterion gives a value of 1000μF, which necessitates a low rated voltage such as 6.3V if the component is to be of reasonable size. As a result, C_2 now needs protective shunt diodes in both directions, because if the amplifier fails it may saturate in either direction.

Close examination of the distortion residual shows that the onset of conduction of back-to-back diodes will cause a minor increase in thd at 10Hz, from less than 0.001% to 0.002%, even at the low power of 20W/8Ω. It is not my practice to tolerate such gross non-linearity, and therefore four diodes are used in the final circuit, and this eliminates the distortion effect, Fig. 8. It could be argued that a possible reverse-bias of 1.2V does not protect C_2 very well, but at least there will be no explosion.

We can now consider alternative input devices to the *MPSA56*, which was never intended as a low-noise device. Several high-beta low-noise types such as *2SA970* give an improvement of about 1.8dB with the low-impedance negative feedback network. Specialised low-R_b devices like *2SB737* give little further advantage – possibly 0.1dB – and it is probably better to go for one of the high-beta types; the reason why will soon emerge.

It could be argued that the complications of a low-impedance negative feedback network are a high price to pay for a noise reduction of some 2dB; however, there is a countervailing advantage, for the above negative feedback network modification significantly improves the output dc offset performance. The second and final part of this article shows how, and also gives full details of the mode-switching and bias control systems, and the performance of the complete amplifier. ■

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Bigger bass smaller box

Considering the speaker as part of the electronics leads to a tailored amplifier frequency response, extended bass and relatively small size — as Jeff Macaulay demonstrates with his 'Microreflex' full-range loudspeaker design.

Seventy years after its invention, the loudspeaker is still the weakest link in the audio chain. In particular, the bass response is usually severely compromised; the bottom two octaves are a special problem. It is hard to see why the audio fraternity places so much emphasis on reducing distortion levels in power amplifiers and yet ignores the gross errors inherent in speakers.

Typically, 99% of the carefully cultivated signal delivered to a speaker heats the voice coil, the remaining 0.1% being mangled by the phase shifts and amplitude variations imposed by loudspeaker and associated crossover. What emerges from the speaker is a distorted version of the driving signal, no matter how perfect the input may be.

At the risk of being lynched by irate audiophiles and engineers, I must point out that the laws of physics dictate that a flat-response audio system cannot be produced simply by driving speaker systems from flat-response amplifiers! With current speaker systems, the only way to produce a system with a flat frequency response is to use amplifiers with a non-linear response.

An ideal speaker would behave as a pure piston, regardless of the signal frequency. No such animal exists. Practical speaker units have a cone with mass which resonates with the compliance of the surround to produce a fundamental resonance. Below this resonance, the response falls away rapidly, while above it pure piston operation is maintained over a restricted band of frequencies before the response starts to roll off again.

Designing an enclosure

At frequencies where the speaker's diameter is less than a wavelength of the sound emitted, antiphase waves from the rear of the cone diffract around it to cancel out the wanted radiation from the front.

This is the reason why some kind of enclosure has to be used, the simplest method being to mount the speaker in a sealed enclosure. Unfortunately, the enclosed air possesses stiffness which adds to that of the speaker surround and raises the resonant frequency, which is obviously undesirable as well as unavoidable. To get an extended low-frequency response one has to enlarge the enclosure. An alternative would be to use a

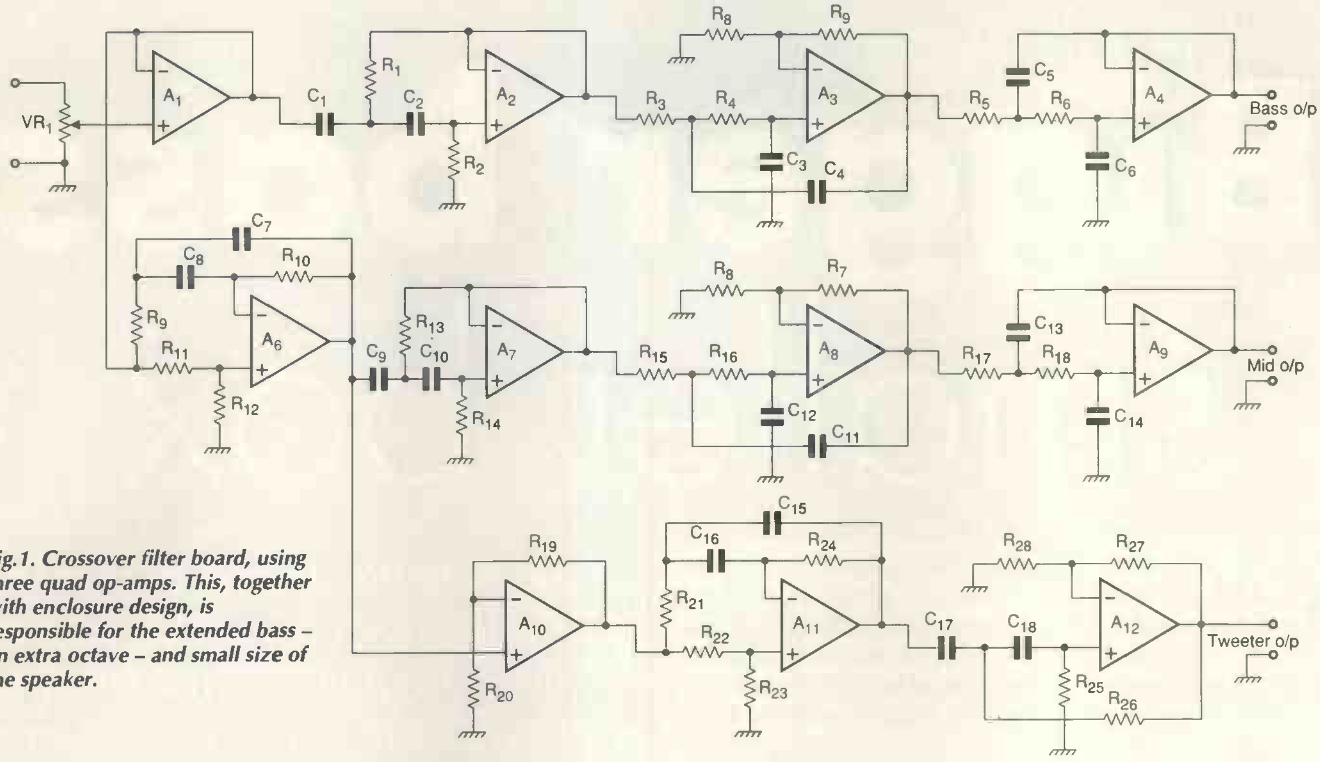

Fig.1. Crossover filter board, using three quad op-amps. This, together with enclosure design, is responsible for the extended bass – an extra octave – and small size of the speaker.

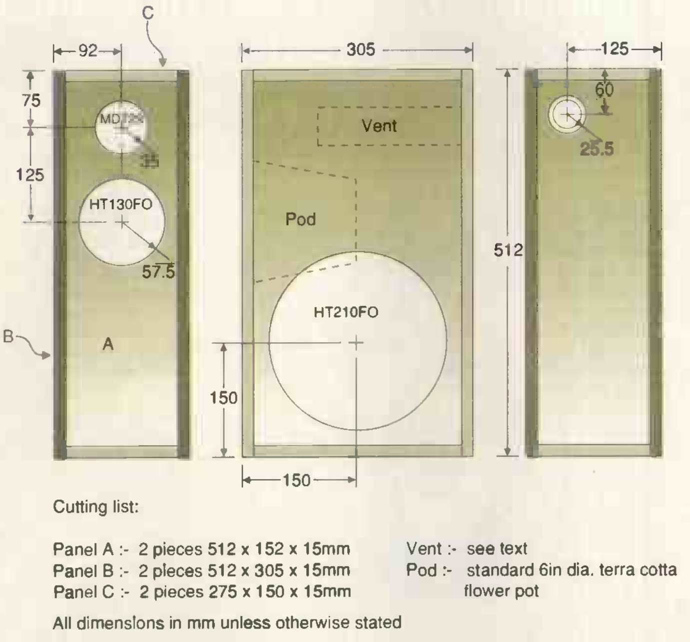

Fig.2. Enclosure construction, using medium-density chipboard and lots of glue, not to mention a flowerpot.

speaker with a lower free air resonant frequency, but the efficiency of a speaker is proportional to the cube of that frequency.

Alternatively, there is the reflex enclosure, in which a duct is cut into the enclosure. The mass of air in the duct and the compliance of the air in the enclosure form a mechanical tuned circuit which is excited by the cone's rear radiation. Duct output is out of phase with the rear radiation from the cone and in phase with the wanted output from the front. Hence, over a restricted range of frequencies, the duct or vent augments the bass output from the speaker.

Below the enclosure resonance, the radiation from the vent moves out of phase with the speaker and in consequence the extreme bass output falls off more rapidly than that from a sealed box. The transient response of the system is therefore poorer. On the plus side, though, the resonant frequency of the speaker is hardly raised from its free air value, leading to lower distortion.

A further advantage of reflex operation is that the cone excursion for a given output is greatly reduced at and near the enclosure resonance. This is because the speaker 'sees' the high mechanical impedance of the enclosure's resonant circuit.

Owing to the pioneering work of Theile[1], extended by Small[2], it is a simple matter to design a good reflex speaker, although design is constrained by the characteristics of the bass units available. Again, a greatly extended low-frequency response is usually only obtainable at the expense of a large enclosure.

This state of affairs dictates that nearly all available speaker systems of a reasonable size exhibit a tendency to have little or no useable output below 60Hz, thereby losing almost two octaves of the audio band. What can be done?

There is a widespread belief that real bass cannot be generated in small boxes. However, as my neighbours will testify, this is not the case. It is simply that such performance is impossible using the techniques already described. However, several solutions have been devised, the best known probably being motional feedback, in which a small transducer is attached to the speaker cone and the resulting signal fed back into the driving amplifier's feedback loop. This is used both to correct the low frequency roll-off and reduce harmonic distortion. A good example of this technique is shown in reference 3.

Another method, used in this design, is the 6th-order reflex speaker system in which the low-frequency response is extended by the use of an underdamped high-pass filter, commonly in the form of an op-amp circuit between preamplifier and power amps. According to Keele, the response can be extended by half an octave in exchange for 3dB less maximum drive signal[4]. Other variations on the theme can be found, for example, sub-resonant speaker systems after Linkwitz[5] and Harcourt[6].

What really limits the bass extension of a driver is the volume of air that can be shifted – a direct function of cone area and peak-to-peak excursion limits. So long as the unit can be equalised to a flat response, the cabinet dimensions can be kept small without affecting the overall response. As the response curves of both sealed and reflex cabinets can now be accurately calculated, we are in a position to extend the low-frequency response of small speaker systems. All that is required is to sacrifice the sacred cow of flat-response electronics.

In reality, using filter techniques to flatten system response is both simple and inexpensive. Intellectually, it is no stranger than equalising the response of a magnetic pickup cartridge. It does, however, require a leap in thinking from the current piecemeal approach to designing a system to a more holistic view in which the acoustic performance is incorporated into the electronic design process. From such an approach comes the realisation of overall system responses that are simply impossible to achieve by purely mechanical means.

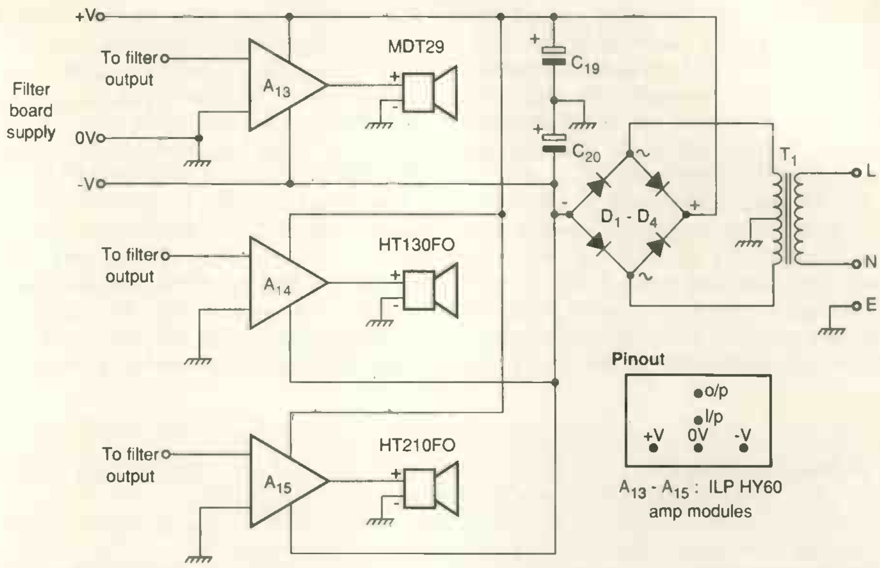

Fig.3. Power supply board. All the electronic part of the speaker is in a plastic box mounted on the cabinet, input being at line level from the radio or cd player, with no further amplification.

Choosing drivers

Designing a speaker system is the simultaneous solution of several, often mutually entangled problems. At best, the individual responses of the drivers used resemble asymmetrical band-pass filters, with unwanted resonances thrown in for good measure. These responses need to be modified and harnessed so that the system response resembles a band-pass filter with a flat response across the audio frequency range.

From the wide choice of possible drivers for the system, I chose Audax units for their consistent Theile/Small parameters. The *HT210FO* bass and *HT130FO* were designed to be used together, as is evidenced by their closely matched reference efficiencies. They are also supplied in pairs, computer matched to within <0.3dB, so that gain matching between them is unnecessary.

One of the design aims was a smooth treble response and accurate integration of driver responses. After trying several tweeters, including titanium and hard-dome types, I chose the Morel *MDT29*, a soft-dome unit with an excellent, smooth and resonance-free response combined with an equally smooth low-frequency roll-off. It is also a robust ferro fluid-cooled unit capable of high power handling on transients.

It is now well established that a wide stereo

Due to the electronic bass-driver compensation, this full-range loudspeaker has a –3dB point of 25Hz. Response is ±3dB over the whole 25Hz to 20kHz range while power handling of the bass unit is 70W continuous.

image requires good horizontal sound dispersion. This requires a narrow enclosure, and to obtain the minimum front baffle dimensions, the bass driver is mounted on the side of the enclosure. This does not cause problems because, below 100Hz, the bass driver's response is omnidirectional.

The mid-range driver needs its own enclosure. In three-way designs, it is common practice to mount this sub-enclosure within the main enclosure and, after a lot of head-scratching, I chose a common-or-garden 6in diameter terracotta flower pot. Although the choice may seem strange, the non-parallel shape ensures that, within the mid driver's range, standing waves cannot occur within the sub-enclosure. This is the major cause of coloration in most enclosures and its removal tightens the sound considerably.

Mounted in this way, the *HT13OFO*'s bass resonance is raised from 48Hz to 144Hz, with a consequent increase in Q from 0.25 to approximately 0.73, the natural choice for the crossover point between the mid-range and bass drivers. Rolling off the bass unit at this frequency also ensures that the cabinet is acoustically small. That is to say, the wavelengths of the sound radiated by the driver are much longer than the largest enclosure dimension. In consequence, standing waves cannot be generated within it.

Crossover considerations

Designed in this way, the system is without standing-wave problems and needs no esoteric construction technique. A further advantage of this crossover point is that the peak output power in musical and speech signals occur around this point. Since both drivers are radiating, peak levels are some 6dB greater than

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Active crossover

In a conventional passive crossover, the designer alters the Q of the network by varying the ratios of the reactive components, bearing in mind the – hopefully – resistive load presented by the speaker. In practice, this is hard to achieve because of reactive effects in the drivers.

In contrast, active crossovers are easily fabricated without recourse to inductors and are independent of driver loading. Although several possible filter configurations exist, the most suitable for active crossovers are the Sallen and Key types, in particular the 'equal-component' and 'unity-gain' variations, shown in **Fig. A1**.

Standard op-amps are used for the active elements and a large variety of types are available; the *TLO* series, used in this design, are well tried and tested and recommended for development work. In the unity-gain circuit, the op-amp is wired as a buffer. Component values for the high-pass version can be determined by the following equations.

1. Choose a convenient value for C, then: $R_2=a/(1.257\times10^{-6} f_o C)$.

2. $R_1=1/(3.142\times10^{-6} a f_o C)$, where a is $1/Q$ and C is expressed in µF.

For the low-pass version:

3. Choose a convenient value for R, then: $C_1=a/(12.57 f_o R)$.

4. $C_2=1/(3.142 a f_o R)$.

In the equal component filter:

5. $f=159155/RC$, where R is expressed in ohms and C in µF.

This equation holds true for both high and low-pass filters. The Q of the filter is set by the voltage gain of the circuit, set by the ratio of R_3 and R_4. R_3 should be $(3-(1/Q+1))R_4$.

Higher order filters are obtained by cascading 2nd-order filters; the Q of a cascaded pair is equal to the product of the Qs of each section. Figure A1 shows the effect of Q on the response shape of a 2nd-order filter, in this case a high-pass type; the response of low-pass filters is a mirror image. Underdamped filters of Q>0.7 show a peak in the passband, while overdamped filters with Q<0.7 do not.

Filters with a Q of 0.7 are Butterworth types, which possess the flattest passband response combined with no peak. However, it can be shown that best transient response is obtained with a Q of 0.5, regardless of filter order; hence the popularity of crossovers with this Q. The standard Linkwitz-Riley crossover uses a 4th-order filter for both high and low-pass sections. A further advantage of the 4th-order filter is that the phase difference between sections is zero.

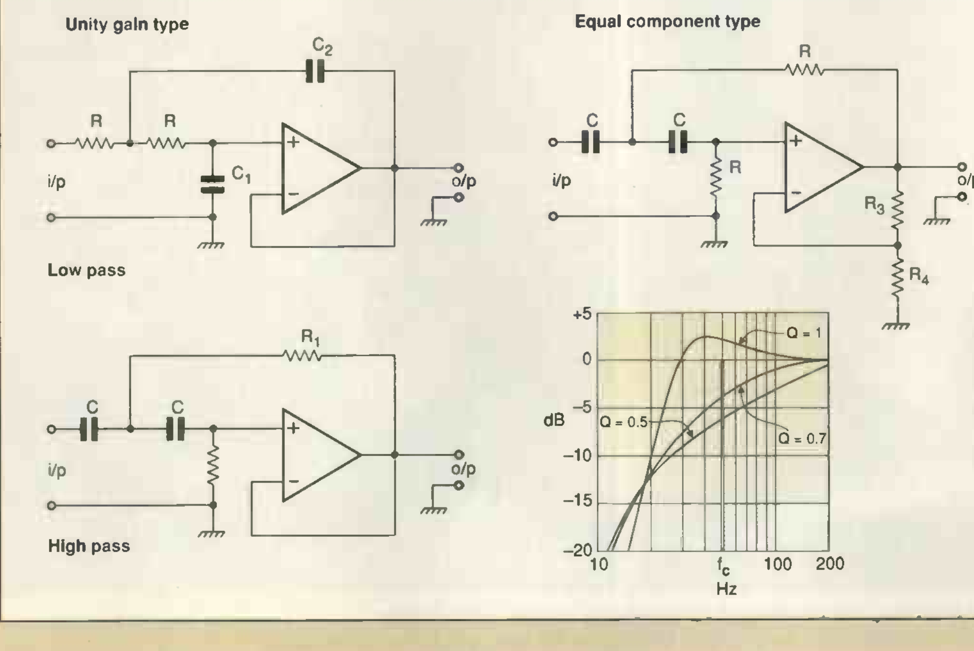

***Fig. A.** Several possible filter configurations exist, but the most suitable for active loudspeaker crossovers are Sallen and Key types, in particular the 'equal-component' and 'unity-gain' variations.*

Best rf article '95

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

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

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

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

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

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Components

$R_{1/7/8/20/27\text{-}30}$	10k
R_2	330k
$R_{3\text{-}6}$	11k
R_9	22k
R_{10}	91k
$R_{11/12}$	180k
R_{13}	15k
R_{14}	8.2k
$R_{15\text{-}18/25/26}$	82k
R_{19}	12k
R_{21}	33k
$R_{22/23}$	270k
R_{24}	130k
$R_{31\text{-}36}$	100
VR_1	47k, log pot
$C_{1\text{-}6/9/10}$	100nF
$C_{7/8}$	4.7nF
$C_{11\text{-}14/17/18}$	2.2nF
$C_{15/16}$	1nF
$C_{19/20}$	3300μF, 35V
$C_{26/27}$	100nF, ceram disc
$C_{28\text{-}33}$	100μF, 25V
$IC_{1\text{-}3}$	*TLO74*
IC_4	*7812*
IC_5	*7912*
$A_{13\text{-}15}$	*HY60* modules – see text
$D_{1\text{-}4}$	*1N4002*
T_1	22-0-22, 80VA toroidal

could be supplied by a single driver.

At the high end, the *MDT29*s have their fundamental 'bass' resonance at 900Hz. These units possess the frequency response of a high-pass filter which, since Q is 0.5, is critically damped. To obtain the best transient response from the system as a whole, pay careful attention to the crossover. Of the available alternatives, the best is the 4th-order Linkwitz-Riley filter, which ensures the minimum overlap between driver outputs, maintains in-phase operation and is critically damped for best transient response. Cascading two 2nd-order sections forms the 4th-order filter; in the case of the mid-range and tweeter, one of these sections is the mechanical roll-offs imposed by the mid-range enclosure and tweeter roll-offs.

In most multiway speaker systems, the crossover frequency between the mid-range and tweeter is set too high, caused by fears over the power handling of modern-day 25mm tweeters. The result is that the mid-range driver has to respond at frequencies above its piston range, where 'cone break-up' can occur; sections of the cone resonate, producing a rough response. Obviously, this is to be avoided and one cure is to lower the crossover frequency.

With the cone tweeters used in the 'fifties and early 'sixties, crossover was usually done at as low a frequency as possible. Modern dome tweeters are not so robust, although their response is considerably better, but it is still perfectly possible to run a dome tweeter down to its fundamental resonant frequency, provided that the rate of roll-off is sufficiently fast to avoid overload by low frequencies. A sealed dome tweeter acts as a 2nd-order high-pass filter. The *MDT29* resonates at 900Hz, where the average power contained in a musical signal is about 10dB down on the peak, which occurs about 120Hz. By feeding the tweeter from a 2nd-order high-pass filter, the power required is reduced by a further 6dB. This is well within the rating of the unit and no overloading occurs.

A further advantage of using a lower crossover, around the 1kHz point, is that the critical upper mid-range band, 1 to 5kHz, is handled by a unit with low moving mass, ensuring superior transient response. Phase anomalies due to the usual 2-3kHz crossover are also reduced.

Electronics

I intended the speakers to be used directly from the line outputs of a pre-amp or cd player, the entire audio system consisting of just the signal source and the speakers, and decided that the easiest way to achieve this is to mount all the electronics within the enclosure. To avoid reinventing the wheel, I used *ILP HY60* amplifier modules, which provide a wide frequency range, low distortion and noise levels, and integral heat sinking. Of course, if you have three stereo amplifiers already these can be pressed into service, fed by the active filter, but anyone who does this must take responsibility for equalising signal levels etc. themselves.

With only five connections apiece, wiring is reduced to manageable levels. I built the equalisation and crossover electronics on a

Extending small-speaker bass

A myth has grown up about the supposed inability of small speaker systems to produce bass in large quantities. Probably the best way of dispelling this myth is to examine the behaviour of a small speaker radiating bass in a domestic environment.

At low frequencies we want our speaker to operate as a piston. Its maximum output is determined by the volume of air that can be displaced which, in turn, is determined by the maximum cone excursion. For the typical small driver, the maximum undistorted excursion is 6mm pk-pk.

Sound pressure level, measured at 1m in free-space conditions, is determined from the formula

$$p=-86+40\log_{10}f+40\log_{10}d+20\log_{10}a_{pp},$$

where p is the spl measured at 1m, a_{pp} is the peak to peak excursion and d is the diameter of the driver, *app* and d being expressed in millimetres. To take the case of a 30Hz input and 6mm a_{pp}, a 200mm diameter driver will generate an spl of 80dB – hardly impressive. However, in practice, at very low frequencies, both woofer channels radiate in phase, so we can add 6B to the figure to give an spl of 86dB.

This calculated value assumes free space radiation in a sealed box. Making use of floor and wall reflections by putting the drivers within a wavelength of the floor and wall, 5.74m at 30Hz, will add another 6dB to the spl, giving a very respectable 92dB spl. Again, in practice there will also be reinforcement from ceiling side and rear walls adding to the spl generated.

As is indicated in the equation, the lower in frequency you go the less spl you get for a given excursion; the converse is also true. Since very few recordings, with the possible exception of organ works, generate full power at these low frequencies, this is the worst case. Maximum sound-pressure level increases at 12dB/octave up to the limit set by the available electrical power.

The second misconception regarding speaker behaviour at low frequencies is that an enormous amount of power is required. In fact, above resonance, for every halving of frequency with a given drive voltage, cone excursion will increase fourfold. Below resonance the cone excursion levels off and to compensate the drive voltage must rise proportionally. This means increased gain at low frequencies with a consequent increase in power supplied. In practice, it is very seldom that the amplifier runs out of steam. What tends to happen is that the cone excursion exceeds prudent limits on overload unless care is taken .

With a reflex enclosure, the situation is similar, although complicated by both the sound contribution of the port and the reactive loading of the cone. Generally, the same arguments apply, but the available spl can be some 6dB higher in the deep bass.

By whatever means bass extension is produced, the clinching argument for it is that, without it, unacceptable errors of 20dB or more have to tolerated.

piece of stripboard; using the well known *TL074* quad op-amp allows the whole circuit to be constructed with three chips per channel.

Circuit details

Figure 1 shows that line-level inputs are applied to the volume control VR_1 and thence to the buffer amplifier built around A_1. From here the signal is fed three ways.

Amplifier A_2 is configured as a high-pass filter with a turnover frequency of 28Hz and a Q of 2.82, primarily to provide bass equalisation for the woofer. Bass loading is 6th-order and the bass enclosure is tuned to a low frequency, 35Hz, which produces an overdamped 4th-order filter response. Low-frequency boost applied by A_2 levels the response, ensuring a –3dB point for the system at 32Hz. Reflex cabinets are not loaded acoustically at subsonic frequencies and, with conventional systems, driver excursion is wasted by subsonic disturbances. A sharp filter roll-off ensures that subsonic frequencies are sufficiently attenuated to avoid overload. Attempts to push the response down below this frequency result in excursion limit problems.

Crossover between the bass and mid-range drivers is handled by A_3 and A_4, which form a 4th-order, low-pass filter from two 2nd-order sections. Q of the final filter is 0.5, the condition of critical damping combining the best transient response with rapid stop-band attenuation.

Characteristics of the mid-range enclosure dictate the choice of crossover point. The mid-range driver, mounted in its ceramic pod, rolls off at the bass end at 144Hz, producing much the same response as a 2nd-order filter with a turnover frequency of 144Hz and a Q of 0.73. On its own, the high Q of this resonance would give rise to an undesirable peak just above the resonant frequency. To tame the response the drive signal goes through a high-pass filter, A_7, which has the same turnover frequency but a Q of 0.68. Resulting acoustic output is the required 4th-order high-pass crossover response.

To avoid cone breakup effects in the *HT130F0*, the crossover between this and the *MDT29* is set at as low a frequency possible. Since the *MDT29* has a high-pass acoustic response centred at 900Hz and with a Q of 0.5, this is the natural frequency to choose. To provide the required 4th-order response, the tweeter is driven from the output of A_{12}, which is wired as a 2nd-order, high-pass filter with a Q of 1. Mid-range output is rolled off at 900Hz by the 4th-order filter comprising A_8 and A_9.

Because the driver's acoustic centres are in different planes, time delay using all-pass filters is needed to compensate. (This apparent contradiction in terms is applied to circuits which provide a flat frequency response but a fixed time delay.)

Two of these filters are used in the circuit. The first, built around A_6 compensates for the time delay between the woofer and mid-range units. This is equivalent to, a –24° phase shift at 144Hz. The second compensates for the time delay between mid range and tweeter. Here a –50mm offset is compensated for by A11 and the associated circuitry.

Finally, the tweeter has a 1dB higher sensitivity than the *HT130F0* and this is compensated for by the gain of A_{10}.

Implementing the Microreflex

Because my woodworking skills end at butt joints, I kept the enclosure, shown in **Fig. 2**, very simple. I used standard 15mm thick medium-density chipboard, which is available from local hardware stores in a variety of finishes. Unless you have an extensive range of woodworking tools and enjoy using them it is best to go to a store where you can get the panels cut accurately to size. I realise there are constructors who would rather use a different type or thickness of timber for this project; this is not a problem provided that the internal volume of the cabinet is kept at 19 litres. Whatever timber you choose, accuracy is very important. Flat-pack cabinets for this project are available from Wilmslow Audio.

To make it easy to access the electronics and to keep the enclosure airtight, amplifiers, filter and power supply were mounted in an ABS box, 220 by 150 by 60mm, on the outside of the rear panel. ABS is an extremely easy material to work with and the necessary holes can easily be made. Connection to the drivers is by six M5 by 40mm-long screws fitted into the rear of the speaker enclosure; electrical connections are taken via solder tags. These screws also hold the electronic package in place on the rear panel, 'piggy-back' fashion.

Construction proper begins with the front baffle. This has to take both the *HT130F0* and the *MDT29* as well as the mid-range pod. Mark out the apertures for both the drivers on the inside surface. However do not drill the mounting holes yet. This is important, particularly in the case of *HT130F0*. The T bolts provided are not used because they would foul the ceramic pod; instead, both the drivers are secured by 12mm long No 6 self-tapping screws. Before cutting the wood, position the pod over the *HT130F0* aperture and draw around it. Roughen the panel's surface where the pod's rim is to be attached to provide a key. At this stage cut out the apertures for the *HT130F0* and the tweeter; position both drivers and mark out the mounting holes using the drivers as templates.

Drill pilot holes to a depth of 10mm with a 3mm drill but, before mounting the units, mount the pod using Araldite Rapid spread around the roughened surface; mix enough to ensure a gap-free bond. Position the pod centrally and place a weight on the top while it dries. An hour or so will allow handling.

It seems to be unnecessary to use absorbent materials in the pod. Absorbents are good at soaking up high frequency signals but are useless at low to medium frequencies; they are definitely not required in the bass enclosure, where adding any quantity will prevent proper operation. After much experimentation, I decided not to use any, since I couldn't hear any difference with or without. It is, however, vital to ensure that there are no air gaps around the pod.

Solder the speaker leads to the drivers, remembering to observe polarity and fit the drivers into position using the self-tapping screws. Feed the leads to the *HT130F0* through the small hole in the bottom of the pod, leaving a loop of wire in the pod to facilitate easy removal of the driver. Lastly, apply filler to the hole and allow it to dry.

Cut the vent aperture on the rear baffle with a 2in hole cutter, which will produce a slightly oversized hole. Wrap a couple of layers of masking tape round the 225mm length of pipe to take up the excess. The aim is a tight-fitting vent.

The rest of the cabinet construction is straightforward. Use butt joints and plenty of glue to obtain an airtight case, except for the vent.

Aftermath

After all the work involved, is it worth it? Definitely. The stereo imaging of this system is excellent and the absence of standing waves improves the detail rendition of the system. Using the drivers substantially within their piston regions and the sharp roll-offs of the crossover contribute to seamless driver integration. The extra bass octave delivered allows music to reproduced with the correct weight and authority.

In short, I have listened to the speakers for nearly a year now and have had no urge to change them for any others, regardless of price. ■

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Data rater:

power ICs on trial

How good is the current crop of power ICs? Ben Duncan finds some claims for high quality music reproduction simply do not measure up.

How do the "new" devices stack up against older ones previously reviewed[1], with each other and with a "benchmark" *LM12*?

In every case, the advertised slew rates are a poor relation in comparison to the 300V/µs from modern mosfet topologies; or even the 50V/µs from the common Lin topology with Self's improvement[2]. This may not matter if all portal rf filtration is good and you avoid certain classes of music[3]

All of the circuits (**Figs 1, 2 & 4 to 7**) except *PA42* (Fig. 7) show some benefit from using a regulated (or just strongly ripple-filtered) psu. In every case, the effects are strongest below 1kHz – but only because the psu noise spectrum is itself strongest from 50Hz to 500Hz.

Recently published psr improvement techniques[4] are of scant use here, but might be useful later, if such ideas are incorporated into future integrated circuits.

Several measurements show distinct 30Hz features that cannot be ac power harmonics and appear to be the artefacts of thermal distortion.

The high quality audio claim for the Boomer ics (*4860/61*) does not appear justifiable as %thd is always over 0.1%, worst of all in the mid-band. However, the spectra show dominant 2nd and 3rd, and all harmonics at levels no worse than several of the previously tested devices[1].

The remaining ics seem to be on a par with the best of the previous contest. But when compared to the purest discrete and part-discrete designs, *PA42* (at least with lateral mosfets) is the only ic here with performance that comes close, if the absolute level and structure of the harmonics is anything to go by.

Protection against death by adverse loading and output abuse is important whatever the ic cost, and even though music drive is usually benign. The *LM3875* and *3886* have the most comprehensive and believable all round soa/adverse load protection I have seen in any ic [5]. They and the *Boomers* and *LM12* are also thermally protected. The *LM12* also shuts down if the supply exceeds 60V but is under 80V. But all seven ics can be killed by marginally excess supply volts. The *PA42* is well protected against output abuse by proven, minimum mosfet techniques. *PA45*'s one-slope protection solely averts over-current into a resistive load. Without added protection, a prolonged bad phase angle (such as driving bass-heavy music into a difficult speaker) might destroy it.

Noise performance

The *LM4861* with input shorted, (**Fig. 9**) shows quite high, ragged noise despite the quiet regulated supply and the output being loaded with 16Ω. Connecting a 100n+5R zobel to the output greatly reduces noise, especially above 200Hz. An unregulated supply – lightly loaded so it has only 7mV ripple – does not improve matters and a +5V pc supply could be far noisier. A large reservoir and correct supply noding will help. The rf oscillation in the absence of an output zobel naturally increases the unregulated noise plot greatly.

These noise plots were initially disputed by NSC, as they could not be corroborated. But retesting demonstrated that the noise is correct for the conditions. The excess at lf, particularly at 50 and 100Hz is caused solely by the 5m of unshielded test load cabling.

Devices on test: 0.5W LM4860 Boomer, 1W LM4861 Boomer and the 40W LM12 (reference) from NSC; 40W LM3875 Overture and 50W LM3886 Overture , NSC; 2kW PA42 and 200W PA45, Apex.

Fig. 1: Parts under test and key specifications.

		Nom Pout into 8Ω W	%thd at nom P_o %	Slew limit	Abs max* Vrail ±V
LM4860 Boomer	NSC	0.5	1	1.8	6
LM4861 Boomer	NSC	1	1	1.8	6
LM12 -	NSC	40	0.01	9	30
LM3875 Overture	NSC	40	0.1	5-11	42
LM3886 Overture	NSC	50	0.03	8-19	42
PA42 -	Apex	(to 1k)	-	40	150
PA45 -	Apex	(to 100)	-	36	75

*: Assumes worst likely worst case ie. signal present & over temperature.
- = not stated by maker
( ) = estimated in lieu.

That just two makers are represented wasn't for lack of trying. One range of power ICs were excluded, because after having identified high levels of crossover distortion, the maker held that music use was not suggested. A European maker had an interesting data sheet, but no silicon. One Japanese maker didn't respond; one didn't sell its ICs in Europe; one no longer made audio ICs.

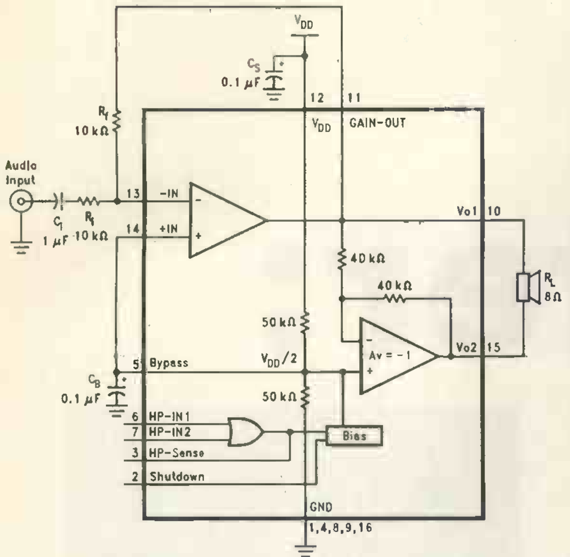


Fig. 2. The 4860 and 61 Boomer ics are supplied ready connected. Note the inverting gain configuration of x1 (Ri/Rf), with the working gain of +6dB being achieved by the bridged output. To make the ics rf stable from 24in of lab supply cabling, a 100μF capacitor was needed across the supply pins. A standard zobel network (100n+5R) was also deployed to quell rf oscillation appearing when the test load was opened, leaving just the AP's analyser loading the output. Parviz Ghaffaripour, NSC's audio design manager suggested a 1k resistive termination would suffice.

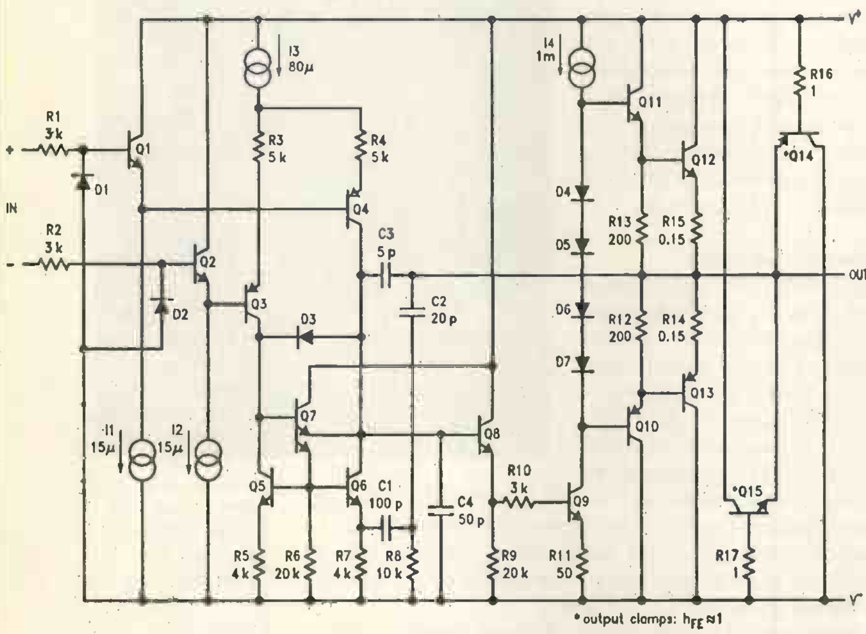


Fig. 3. National's LM12 *– a 40W device with a quoted 0.01% thd at full power.*

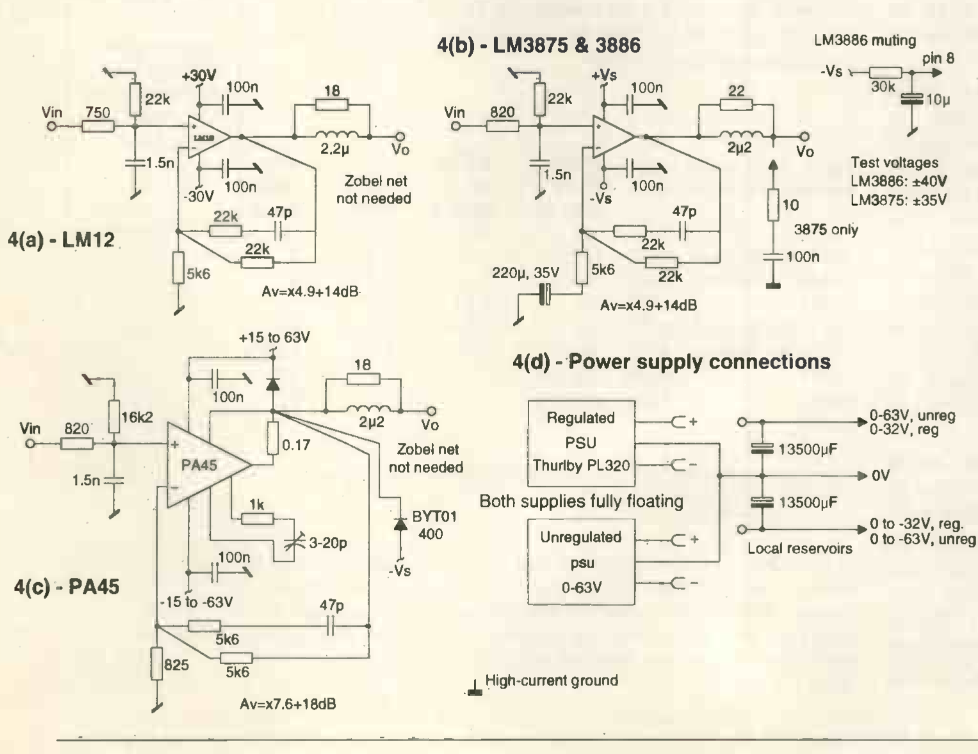


Fig. 4. Test circuits Section (a) LM12. Gain is set low at 14dB to preserve loop gain, as gbwp is only 700kHz. Other IC gains are set within 6dB of this. Granted that the non-inverting configuration is the norm, then some CM distortion is inevitable in this situation towards clipping, with >1V at input. Section (b) is LM3886 and 3875. The 3886 requires current from –Vs to mute pin as shown, to un-mute. With LM3875, the zobel was required for rf stability. The others did not need a zobel though in other conditions, they might. In (c) the PA45 requires added RC compensation and a series current limiting R. The fast clamp diodes are advisable considering IC cost. Input protection diodes are worth considering. In (d), all ICs were powered via local reservoirs within 150mm, with either regulated 150W (0-32V) or unregulated 500VA (0-65V) supplies being plugged in.

In effect, the balanced output, while rejecting rail noise, is unusually emi sensitive. The voltage dip on load demonstrates that Z_o is quite high, about 0.5Ω, giving a damping factor of 20. This may frustrate bass performance into the better miniature speaker designs.

LM12, tested with supply ripple at 6mV p-p and 20mV (+10dB) shows a noise increase of at least 5dB below 500Hz **(Fig. 10)**. The *LM3886*'s noise is lower than the *LM12* at hf and slightly higher where it matters most, in the mid-band **(Fig. 11)**. The high 50Hz spike may be ameliorated by refined layout and optimised shielding. PSR is generally better than *LM12*.

When the *PA42* is used alone, as an high voltage op-amp, its performance is exemplary **(Fig. 12)**. Noise is uniformly low across the spectrum, particularly the low sensitivity to 50/75/150Hz magnetic field frequencies and no special precautions needed to be taken. Similarly, the *PA45* shows a strong 50Hz sensitivity but commendably low noise (–140dBr) at the mid/high frequencies where it matters most **(Fig. 13)**. Degradation that occurs by using an unregulated supply indicates limited psr and a high sensitivity to magnetic fields – though a large hum spike at –125dBr is probably not so audible as it appears.

Total harmonic distortion and power

Power (right hand axis) and %thd (left hand axis) were plotted for each device.

With a regulated supply, thd of *LM4860* shows only small differences between 8 and 16Ω loads **(Fig. 14)**, at 0.5dB below clip. but more than halves with the same loads at –5.5dB below clip. The 16kHz spike suggests some emi susceptibility to the pc's vdu, 0.7m away.

Power is confirmed at about 800mW and 300mW into 8Ω, at the two test levels.

Using an unregulated supply, **(Fig. 15)** %thd figures are about 3-6dB greater, particularly at lf. A dynamic %thd plot shows that the %thd baseline is higher when the supply is unregulated (cf Fig. 9). The poor psr is unusual for a bridged output and is not caused by load cable pickup.

For *LM12*, just below clip, %thd is passable **(Fig. 16)**. But a rise below 50Hz could be thermal distortion or lf-triggered rf instability.

Mid-band output delivery during the %thd plot is confirmed at about +27.5dBu, alias 42W into 8Ω. Dynamically, %THD at 1kHz changes threefold in the top 18dB below clip, whereas at 10kHz, the change over 18dB is barely x1.5.

The *LM3886*'s %thd is much better than the *LM12* at hf **(Fig. 17)**, at just 0.02%. But residue at 10kHz has club spikes,. There are also traces of narrow cross-over spikes in the noise at 1kHz, while lf distortion now rises

Data sheet warnings

- All the power ic makers are guilty of overstating the quality of their slew limits. Today, when even vfb op amps and even audio amps are slewing at 350V/μs and more, designating a slew limit of under 20V/μs "fast" raises my eyebrows.
- Other than the Apex ics (which are not power specified), there isn't the rail voltage leeway that discrete component power amplifier designers are used to. Due to the over-focus on power output figures, the ics have power output specifications that are only available close to the part's limits. Attaining these powers proves dicey without supply regulation or over-voltage protection. Also, the max supply specification for *LM12, 3875* and *3886* is literally dangerous, being quoted higher without signal – hardly a valid test condition. Even when *LM3875* is operated below its maximum realistic rail rating of ±42V, and is mounted on the generous heat sink used, it can't deliver a continuous test signal cleanly into 8Ω until the supply is set at or below ±35V, a consequence of the Spike soa protection working properly.
- The *Boomer* data clearly states that no (output) zobel is needed, apparently as the design team expect amplifiers will always be used with an adjacent, permanently connected speaker. How can they be so sure? **BD**

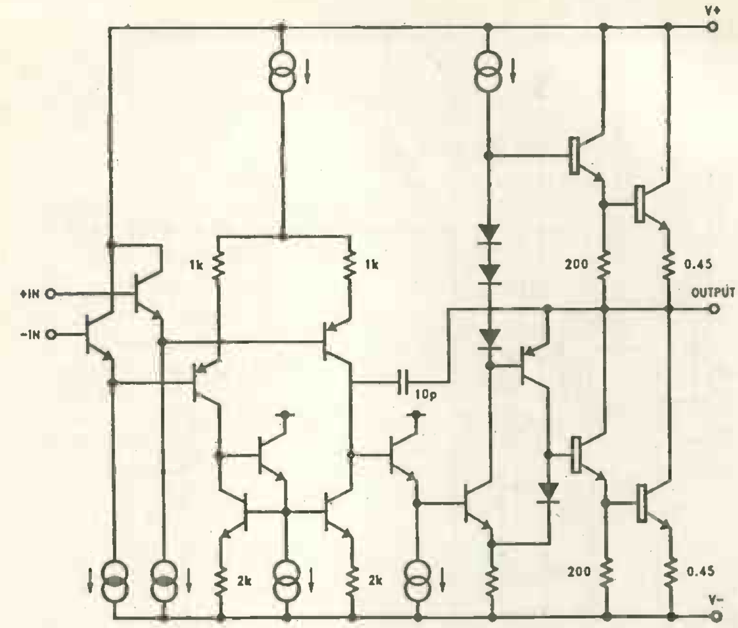


Fig. 5. LM3875's lineage is apparent after reviewing the LM12. Most changes are details around vas. The related LM3886 is almost identical in the lower half, but above the input pair are nested long tailed pairs and sources, and output monitoring, all added for protective muting purposes.

gently from 200Hz, the residue at this point being highly angular.

As one of few major differences between the *LM3886* and *LM12* circuits is the output devices and their heat transfer values, thermally induced cross-over distortion is a reasonable hypothesis – though the repeated 30Hz spike seen also on the *LM12*'s response begs explanation.

Dynamic %THD behaviour was similar to *LM12*.

%thd for the *PA42*'s solo is good below 1kHz, and quite passable at 20kHz (**Fig. 18**) – provided the current limit is set generously enough to suit the load; Near vertical take-off at 14kHz indicates how a 680Ω *CL* resistor can make the ic proof against continuing shorting with dc but also starves the output at <8mA, the load being just 100kΩ and 1nF of analyser plus cable capacitance.

Setting *CL* to 30Ω, a value chosen for driving the external mosfets, produces no vertical take off, but an intermediate value would be needed to provide some abuse protection, at

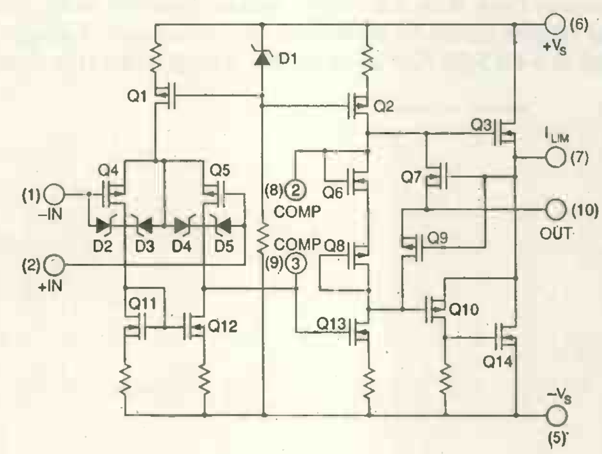


Fig. 6. Inside the PA42 driver ic, all is mos. Input protection zeners prevent excess Vgs. The source biasing is shown as an unspecified (and possibly noisy?) zener. Note the output is really at pin 7, which is linked to pin 10 via a protective, current sensing resistor.

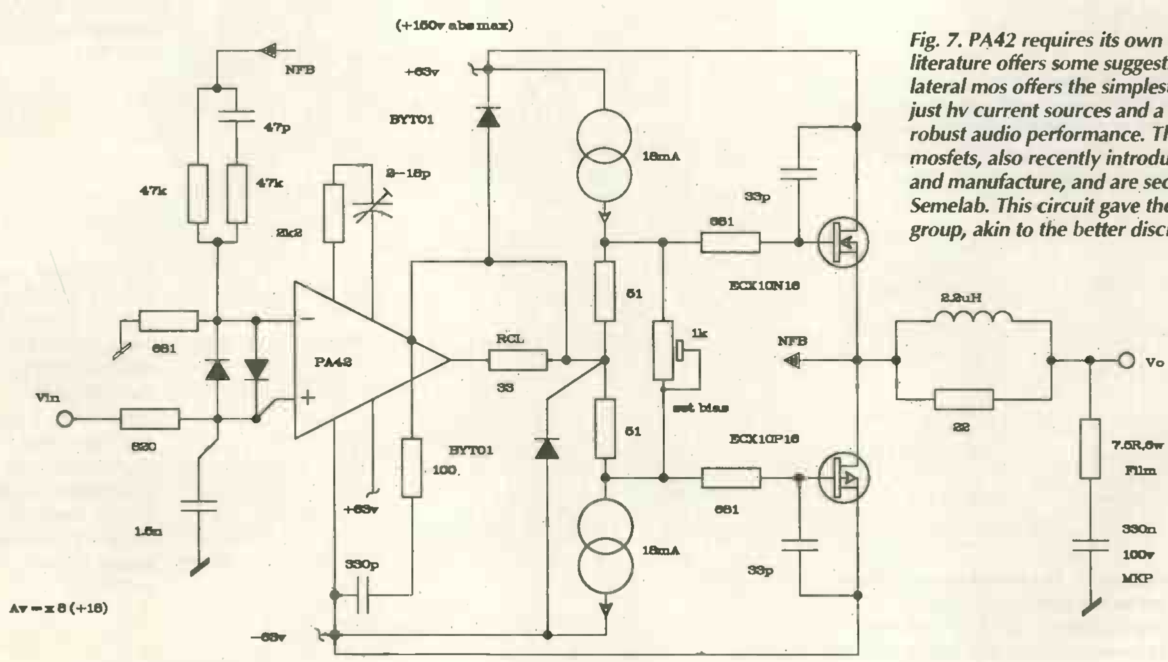


Fig. 7. PA42 requires its own output stage. Apex literature offers some suggestions with v-fets, but lateral mos offers the simplest solution, requiring just hv current sources and a gate spreader for robust audio performance. The Exicon ECF output mosfets, also recently introduced, are of UK design and manufacture, and are second sourced by Semelab. This circuit gave the cleanest results in the group, akin to the better discrete designs.

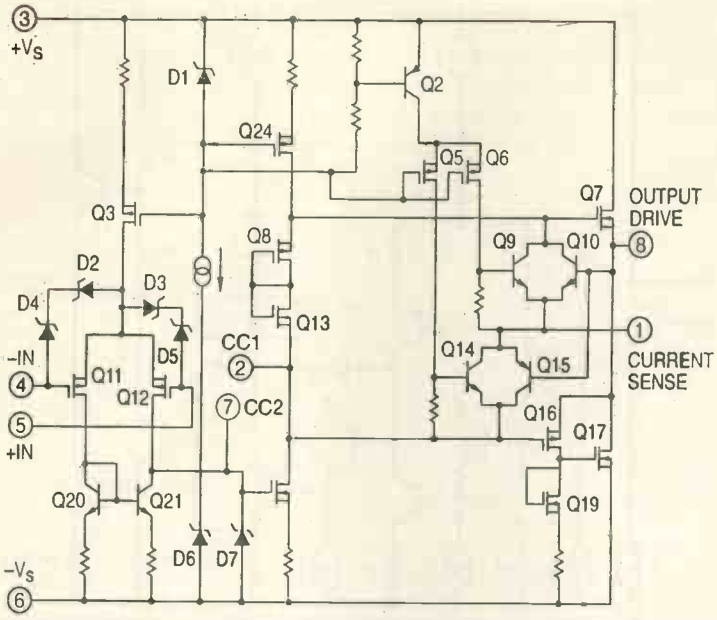


Fig. 8. PA45. A mixture of mosfets and bjts. It differs from the PA42 by naturally having larger die area output mosfets (Tr7,17), more elaborate protection (Tr2,5,9,14, etc), and more biasing and clamping zeners.

least above 20Hz. The dynamic plot showed that both 1kHz and 10kHz %thd hardly varied in the 18dB below clipping.

With lateral mosfets added, and *CL* set at 30Ω (a safe value when driving one pair of lateral mosfets) the *PA42*'s %thd is hardly changed (**Fig. 19**). %thd reduces only minutely when the biasing (for one output pair) is increased from 40mA to 75mA, and is only slightly lower into 16Ω.

Noise (not shown) is very similar to Fig 16, and a 7dB increase in ripple has no effect except at 100 and 200Hz, where noise increases by some 7dB.

The PA45's %thd is impressive when unloaded (**Fig. 20**), at ±63V and 1dB < clip, with a notch in the residue decreasing markedly as the supply increases past ±60V. Alas, the highest current limit *R* value (*0R17*) that still assures short protection (5A) is not able to handle continuous drive into 8Ω at these voltages (56Vpk/8 = 7A) – at least without a fatter heat transfer bracket. Loaded percentage thd is more acceptable when drive is –11dB below clip, 16Ω, rather than –1dB < clip. An lf rise is assumed to be a thermal distortion artefact.

The power plots show about 120W wrt 8Ω when loaded with 16Ω, hence really 60W.

Continued on pages 482 & 531

Noise performance: Figs 9 to 13 show noise spectra with shorted inputs, with differing power supplies and amounts of ripple used to demonstrate spot psr. Except for Fig. 9, each graph has a 40dB window. The AP's residue is –135dB flat across the band with the test set-up environment.

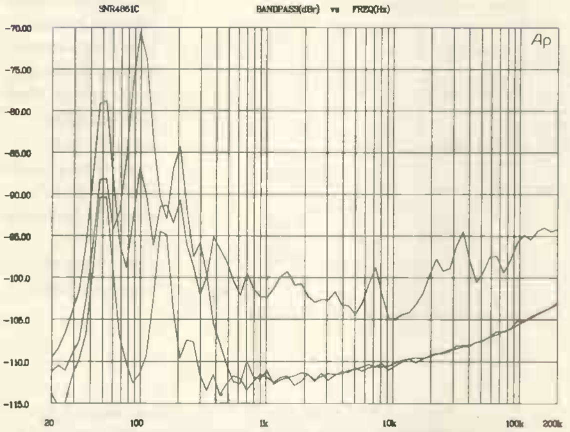


Fig. 9. Output noise spectra, showing effects of rf instability and poor psu rejection for LM4861: upper with input shorted and regulated psu, middle, with 100n+5R zobel added, lowermost, using an unregulated supply. The excess at lf particularly is caused solely by the 5m of unshielded test load cabling.

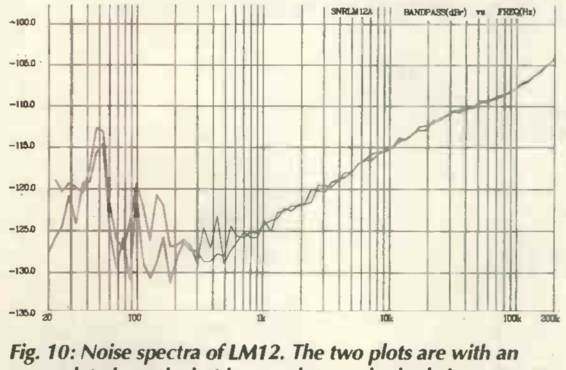


Fig. 10: Noise spectra of LM12. The two plots are with an unregulated supply, but in one plot supply ripple is increased from 6mV p-p to 20mV (+10dB) by resistive loading. There is an increase of at least 5dB below 500Hz.

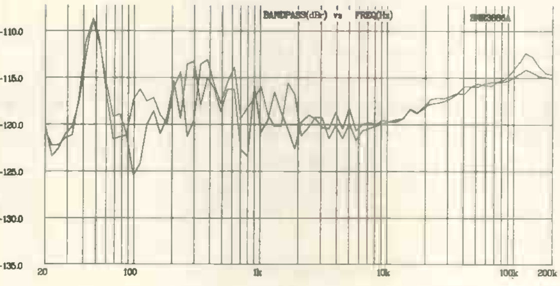


Fig. 11. The LM3886's noise is lower than the LM12 at hf and slightly higher in the mid-band. Powering is from an unregulated supply,: A is 5mV ripple voltage and B is 15mV pk-pk.

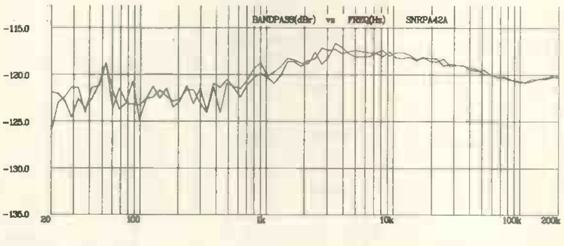


Fig. 12. Exemplary noise performance of the PA42.

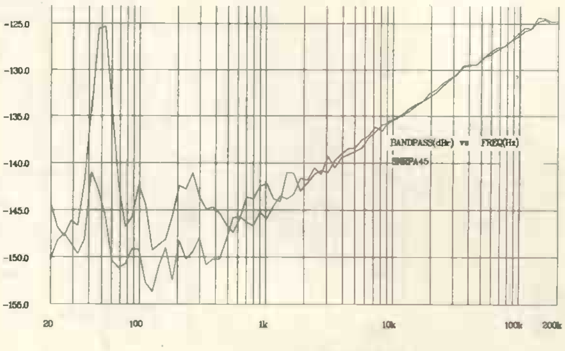


Fig. 13: PA45 noise spectra. A strong 50Hz sensitivity but commendably low noise (–140dBr) at the mid/high frequencies. Upper plot shows the degradation with the unregulated supply having 30mV pk-pk ripple (at least 30dB worse) and set at ±63V.

Percentage thd and power bandwidth: Figs 14 to 20 show %thd vs frequency and also the power into 8Ω (right hand side) at just below clip, and at some lower output. Analyser bandwidth is 80kHz, so the 20kHz sum truncates above the 4th harmonic. Several plots demonstrate hikes at lf that are most easily explained as thermal distortion. Others demonstrate clear differences as supply ripple changes on one or both rails, demonstrating that psrr is a real issue too.

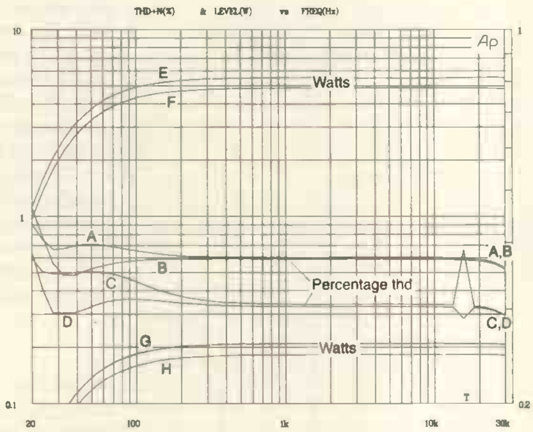


Fig. 14. LM4860 with regulated supply. A (8Ω) and B (16Ω) show small differences at 0.5dB below clip. C and D show how %thd more than halves with the same respective loads at -5.5dB below clip. Power curves E and F, G and H respectively confirm about 800mW and 300mW into 8Ω, at the two test levels. Retesting demonstrated that thd was not influenced by noise pickup susceptibility in the load cabling.

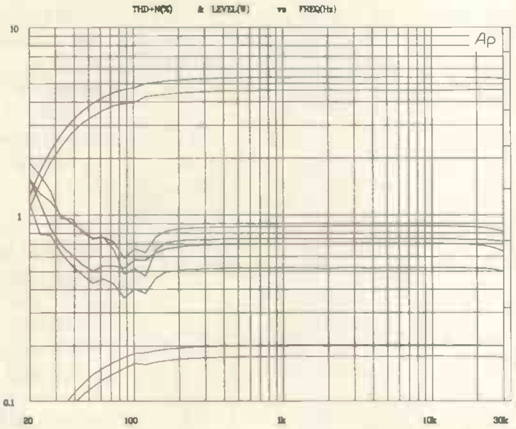


Fig. 15. Unregulated supply and LM4860. All-round %THD figures are about 3 to 6dB greater.

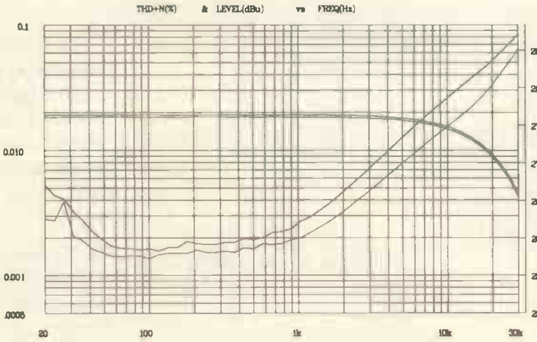


Fig. 16: LM12's %thd is passable at below 0.06%, 20kHz into 8Ω just below clip but shows a rise below 50Hz. Lower curve shows 16Ω response. The upper pair of curves confirm a mid-band output delivery during the %thd plot of about +27.5dBu.

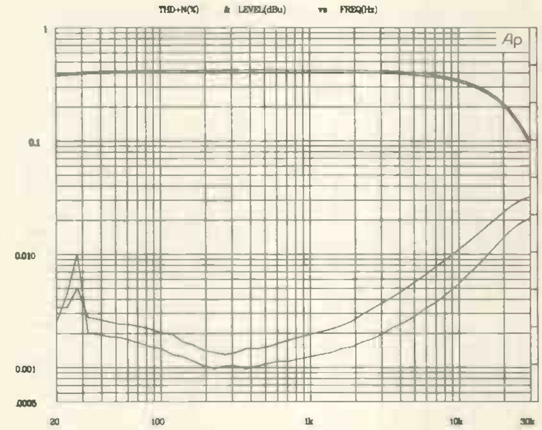


Fig. 17: Compared to LM12, the LM3886's %thd is better at hf, at 20kHz into 8 Ω. Thermally induced cross-over distortion could be causing problems.

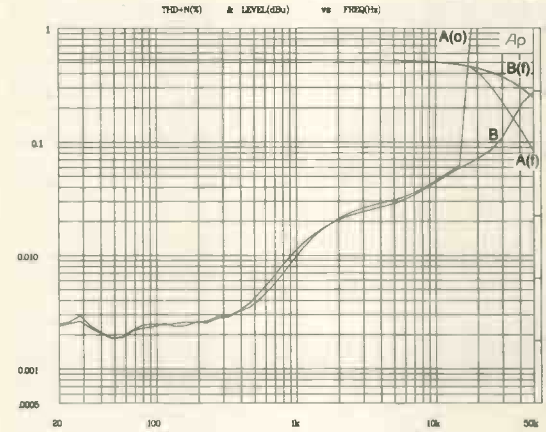


Fig. 18. PA42 %thd is satisfactory though the current limit must suit the load. Curves A show result of a 680Ω CL resistor; and curve B shows CL set to 30Ω. A ±61V unregulated supply was used. Residue was mostly quite angular, and increasingly complex above 2kHz. A(F) and B(F) show frequency response.

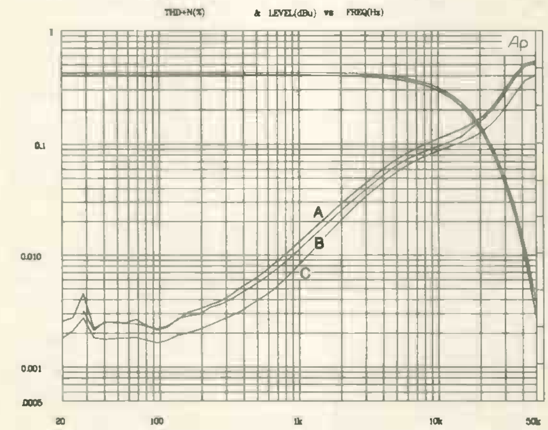


Fig. 19: PA42 with lateral mosfets added, and CL set to 30Ω: %THD is hardly changed. It reduces only minutely when the biasing (for one output pair) is increased from 40mA (A) to 75mA (B). (C) shows only slightly lower %THD into 16Ω.

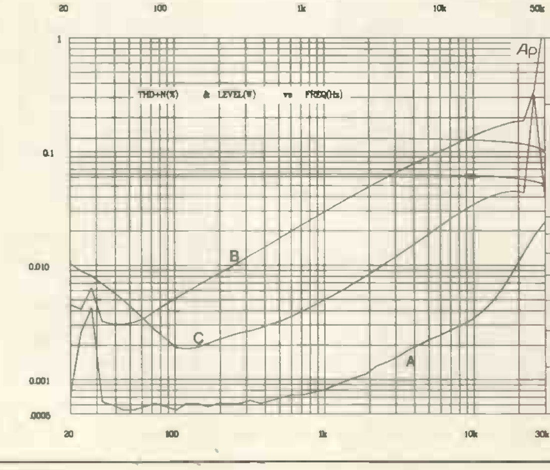


Fig. 20. PA45's impressive %THD when unloaded (lower plot, A) at ±63v and 1dB < clip. Upper %thd plot (B) shows drive into 16Ω -1dB < clip. Middle curve (C) shows a more acceptable result when drive is -11dB below clip. Power bandwidth is plotted on the right.

Spectral behaviour: These plots show harmonic spectral behaviour –1dB below onset of clip (judged from spikes in the thd residue), then at the lower level more typical of most listening levels, together with some examples of supply regulation effects. Each graph has a 100dB window. All dBs are referred to the fundamental.

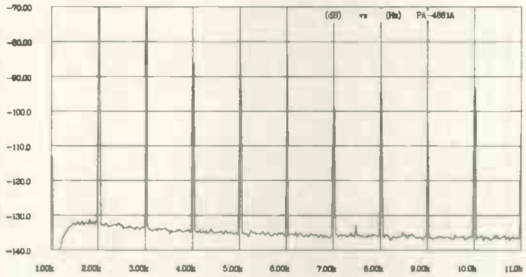


Fig. 21: LM4861 spectra 1dB below clip into 8Ω.

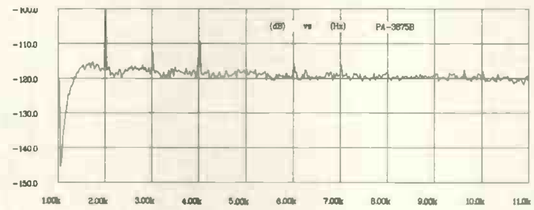


Fig. 27: At –25dB below clip, LM3875 spectra are very like LM12 under similar conditions. LM3886 performance was also similar, but more like the LM12 at high levels, while at –25dB down, the 3rd, 5th and 7th are stronger and replace the 4th and 6th. This small but possibly crucial difference may result from the VCA type elements used for muting.

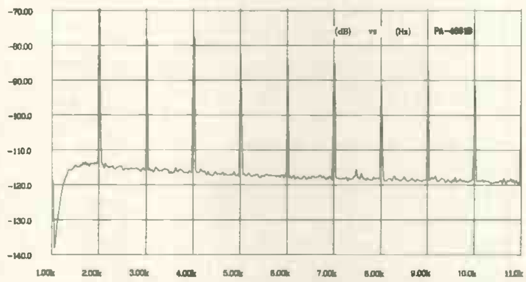


Fig. 22. Fatiguing sonics suggested by 4861's spectra at 18dB below clip.

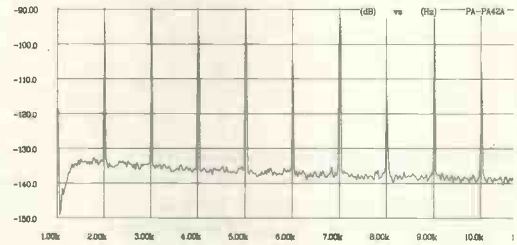


Fig. 28: Just below clip, spectra of PA42 and added mos output stage is ragged but nearly all below –90dBr.

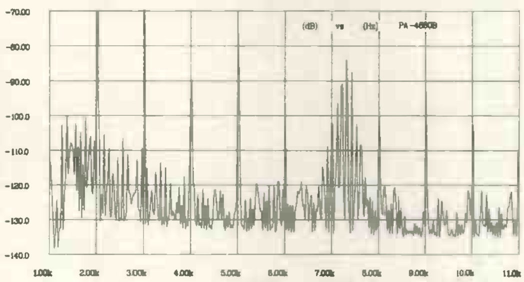


Fig. 23: LM4860 spectra possibly as a result of a faulty IC.

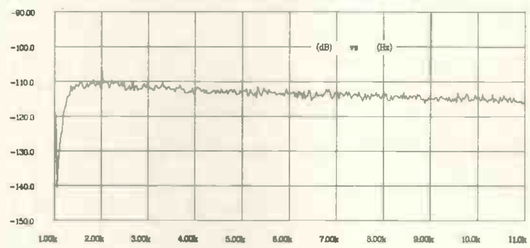


Fig. 29. At –26dB below clip the only PA42 harmonic readable is a tiny amount of second.

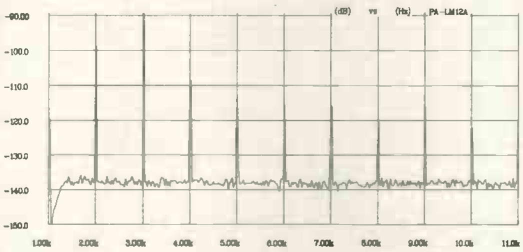


Fig. 24. At onset of clip, most of the LM12 products are just below 100dB.

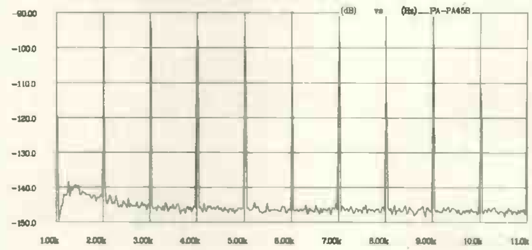


Fig. 30. 0.5dB below clip with +51V rails, and a 16Ω, the PA45 mainly makes odd harmonics which will be very prominent by ear.

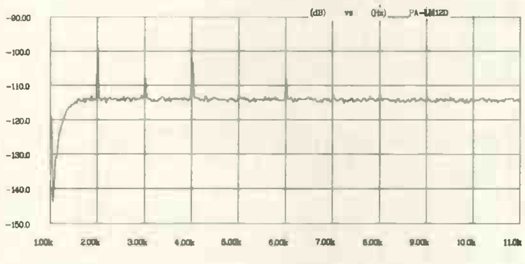


Fig. 25. 25dB below clip, the noise floor has increased, and harmonics above it have changed.

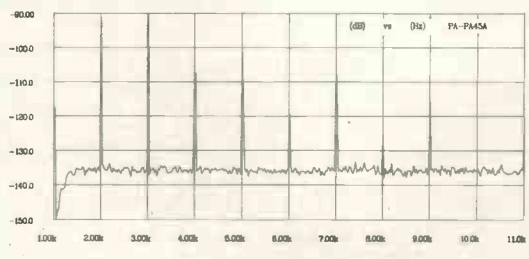


Fig. 31: PA45 with 16Ω load, but at –25dB below clip with ±31V rails. Odd harmonics still dominate the evens.

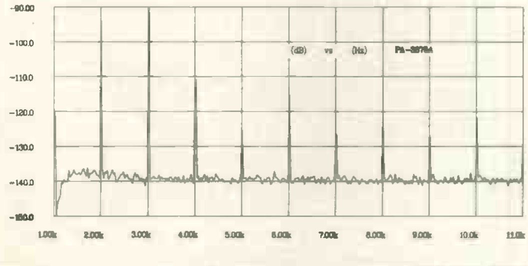


Fig. 26. The LM3875's spectra just below clip are similar to LM12.

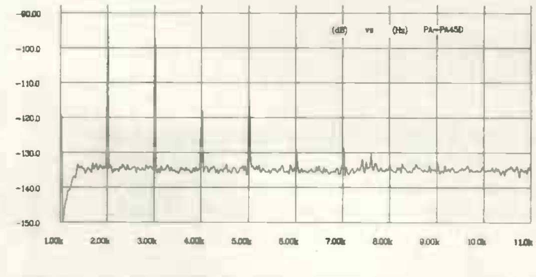


Fig. 32. PA45 with ±62V rails, other conditions the same as Fig. 31, all the harmonics are reduced – excellent sonics should be the result.

Two chip VIDEO digitiser

Given a pc of reasonable performance, it is possible to grab 25 frames a second of standard composite video frame using little more than a flash converter and LPT port D connector. Steve Webb explains how most of the image reconstruction work is done via software.

Commercially available video digitisers are an expensive luxury, with most models costing upward of two hundred pounds. For experimenters, and applications such as shape recognition, this may be off-putting. This article describes how anyone with a 286 or better can capture video on a shoe string.

Performance of the digitiser is not brilliant. But on the other hand, useful results are possible, **Fig. 1**, and the cost makes the circuit ideal for applications such as counting cars or intruder detection. The design was conceived to provide a simple means for experimenting with computer vision.

A key design criterion was minimum hardware, in order to keep costs down and to allow the interface to be easily removed. This lead to a circuit based on the pc LPT printer port. The port is pushed to its limit, but the results are rewarding.

Fig. 1. A typical screen shot, fed into the pc via the low-cost video grabber.

Design overview

The interface acts as a free-running a-to-d converter providing asynchronous samples to the printer port. The whole video signal – including sync pulses – is digitised to five-bit reso-

Video grabber specifications

Host computer:	286 pc, or better
Connection:	LPT1 printer port
Eff. frame size:	150 hor. (depends on bus speed) 100 vert.
Grey levels:	20
Frame rate:	25/s (486DX33, 3×1 filtering, all assembler)
Power supply:	Obtained from port (15mA)

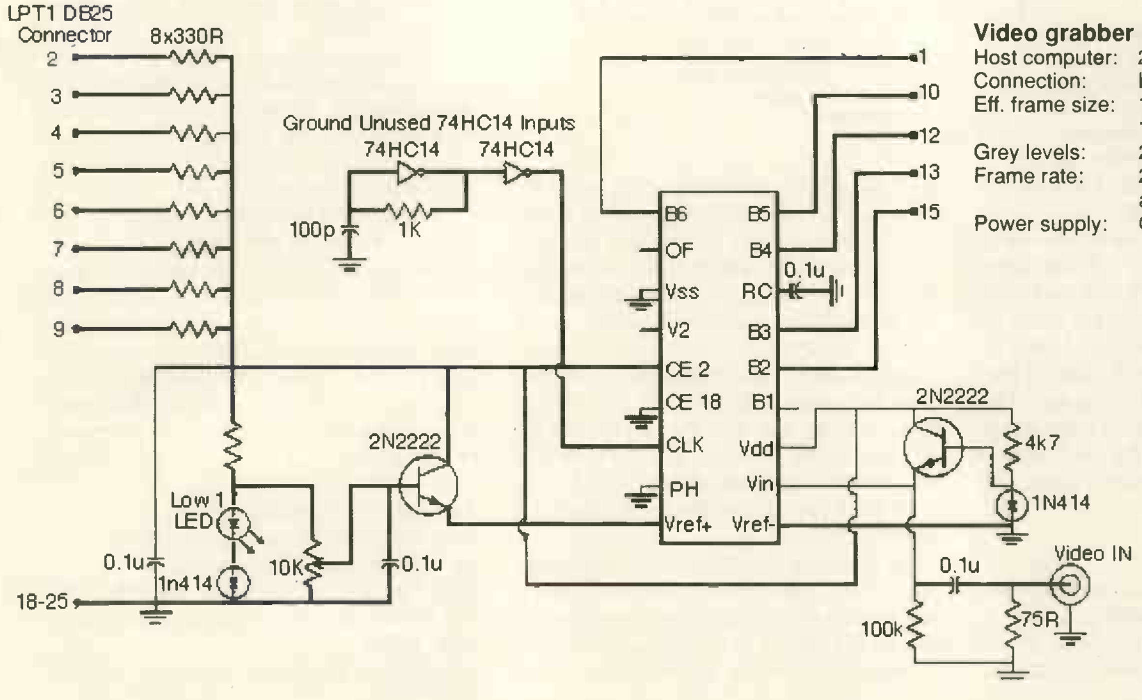


Fig. 2. With little more than a CA3306 a-to-d converter, it is possible to capture video signals for displaying or processing on a pc. Although low in resolution, images captured can be used for recognition.

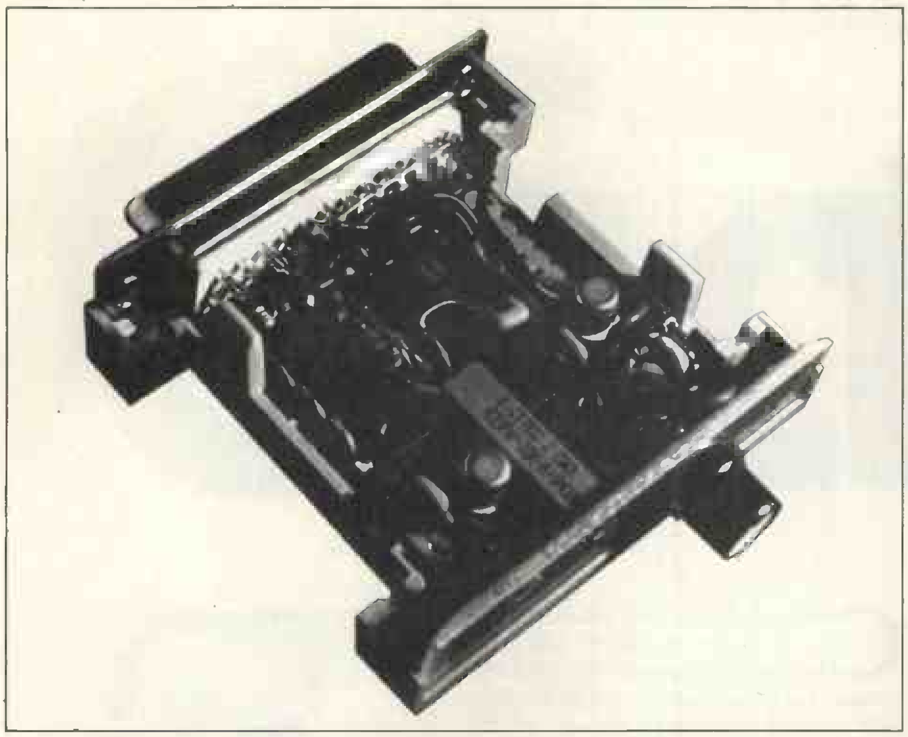


Fig. 3. Video digitiser prototype plugs directly into the printer port.

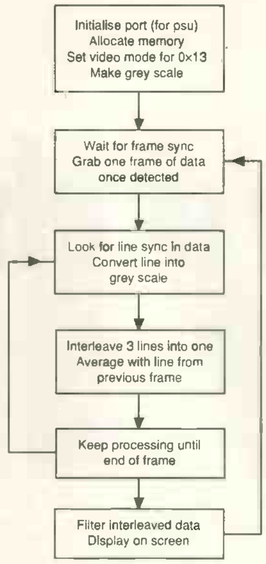


Fig. 6 Flowchart for digitiser software

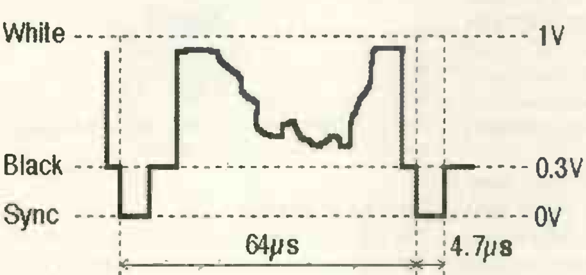


Fig. 4. Representation of a monochrome composite video signal. The digitiser captures the whole signal and it is up to the software to separate the syncs.

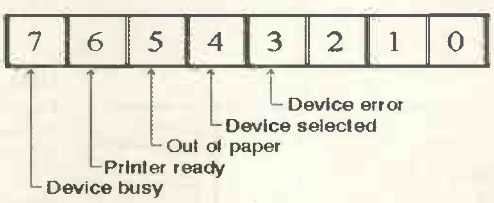


Fig. 5. Digitised video is made available in the status register at port base address plus 1. This is the bit structure.

lution. Power is derived from the eight port data output lines leading to a very compact dongle-sized design.

Software is arranged to synchronise to the video signal and then grab one complete frame of data. Due to the relatively low port bandwidth, only 40 or 50 samples are taken for each video line. Post processing, during the interlace period, converts three sampled lines into one display line. Additionally digital filtering is used to make the results more acceptable. The whole process is repeated giving a real-time display rate of up to 25 frames per second.

Electrical design

Figure 2 is the full interface schematic showing how the design revolving around a *CA3306* six-bit flash a-to-d converter. Power supply is derived from the eight port data output lines, suitably ballasted together via 330Ω resistors. Total current consumption is around 15mA. This method of deriving power is unconventional, but I have found it more than adequate. Assuming the port uses a standard *74LS* latch driver, then 2.5mA per line is typically available.

An adjustable voltage reference is derived from the series-connected led and *1N4148* diode, feeding the a-to-d converter via the *2N2222* emitter follower. The led also indicates interface operation when activated by the software. A low current device was selected to conserve current.

The reference is adjusted for a typical 1V pk-pk video signal. A standard video signal is terminated with 75Ω and capacitively coupled into the *CA3306*. A low impedance dc clamp is formed by another emitter follower and diode combination. Thus the tips of the video synchronising pulses are referenced to ground, and in a suitable form for the data converter. Decoupling is provided by 0.1μF capacitors as shown.

In practice, the port is the limiting factor, so I decided to make the a-to-d converter free run and provide samples asynchronously to the pc processor. A simple clock is generated by the *74HC14* schmitt inverter, and set to an arbitrary 7MHz.

The clock was chosen to be greater than the bus sampling rate to help reduce patterning. On my pc, port read times of around 1.2μs per sample were typical, but this is likely be machine specific.

There are only five input data lines available for reading via the status register. These would normally indicate PAPER OUT, DEVICE ERROR, etc, but in this application they are connected to the five most significant bits of the data converter. In this way the whole video signal – sync pulses and all, is digitised to a five-bit resolution.

The decision to digitise rather than extract the sync pulse will make sense once you have studied section. It should be noted that the most-significant bit connected to DEVICE BUSY is inverted by the port circuitry so the software takes this into account. Using one of the spare inverters to correct this situation, was unsuccessful due to gate delay.

There is little more to the hardware, other than what has been described here. Prototypes have been built in the small gender-changer sized boxes as part of a DB25 connector. It is possible to do this with Vero-board although some unconventional construction techniques and patience are required as it can take some time, **Fig.3**.

Analysis of digitised signals

A typical monochrome video signal is 1V pk-pk with the top 70% representing the grey level where white is 1V and black is 0.3V. The blacker than black sync pulses are below 0.3V, **Fig. 4**.

For the 625 line system, a complete image is formed by two interlaced frames each of 312.5 lines. A longer frame sync pulse is at the start of each frame. The whole video signal is digitised by the circuit to five bits resolution, so allowing 21 effective grey levels to be represented. It is up to the software to detect and synchronise to the sync pulses in order to display a stable picture.

The digitised signal is made available via the status register. This is port base address plus 1, or $378_{16}+1$ on most computers, an address obtainable via the bios, **Fig. 5**. As stated, bit seven is inverted by the port hardware, so this needs to be accounted for.

Software

The software reads a packet of data from the port and then post processes it during the interlace frame. In effect, this gives 312 lines each of around 53 samples including sync pulses. Obviously such an aspect ratio is undesirable, but it is the post processing that makes the scheme viable.

In order to read the port and process data at an appropriate rate, it is necessary to use assembler. 'C' has been used for higher level set-up where speed is not critical. The simplest way of explaining the digitiser software is to start with a flow-chart, **Fig. 6**.

The port is initialised by outputting FF_{16} to the Data register at address 378_{16}, so setting all data lines high for the interface psu. Correct operation is indicated by the led.

Suitable sized arrays are allocated for sampled data and a workspace. The screen mode is set to vga mode 0×13 (320×200 256 colours) by a dos call 0×10. The 21 level grey scale is also generated as required.

Once initialised, the port is read repeatedly until the start of a sync pulse is detected. Reading continues until the software is sure that a frame sync has been detected. A complete frame's worth of data is sampled immediately after the sync. pulse end.

Disabling system interrupts ensures the processor gives undivided attention to the port using the efficient **REP INSB** instruction, Fig. 7.

Due to the low port bandwidth, no information should be discarded, despite the elongated aspect ratio. The chosen algorithm involves taking the first pixel from three successive lines and depositing them, in order, to the workspace. This is repeated for the second pixels and so-on, **Fig. 8**.

Additionally pixels from the current frame are averaged with those of the previous to reduce 'twinkle' noise. Samples from successive lines are staggered relative to each other either to the left or the right, the trick being to decide which way. Suggestions on how to improve the scheme would be welcome.

Interleaved data contains striations that are a natural product of the process. To make a picture more acceptable it is recommended that some form of low-pass filtering is implemented. In practice the picture will have comparably good vertical resolution compared to the horizontal, so you will find that a simple 3×1 average is more than adequate. A 3×3 average tends to defocus the image. A 3×3

Fig. 7. Simplified source code for driving the low-cost video grabber via LPT1.

```
//-----------------------------------------------------------
//      LPT1 Video Digitiser
//      (c) Copyright by SM Webb February 1995
//      Program to operate a real-time video digitiser interface
//      connected to the parallel port.
//      Compiled with Borland Turbo C using Large Model//
//-----------------------------------------------------------
#include <stdlib.h>
#include <dos.h>
#include <malloc.h>
#include <conio.h>

#define DataLPT1            0x378          //Base address of LPT1
#define StatusLPT1          DataLPT1+1
#define GRABSIZE            17000          //Samples for one frame
#define TIMEOUT             20000          //No sync detected timeout
#define LINELENGTH          155            //Length of displayed line
#define MAXLINES            625/2/3        //312.5 lines interleaved 3 into 1
#define SYNCLVL             7              //Digitized sync threshold
#define BLKLVL              10             //Black level

void DoScreen(void);
char Grab(void);
void Interleave(void);
void Display(void);

unsigned char far *video=(char *)0xa0000000L;        //Base address of screen
unsigned char *grab;                                 //Sample array
unsigned char *interleave;                           //Interleave workspace array
//-----------------------------------------------------------
//Main routine to repeatedly grab, interleave, and display smoothed frame
void main(void)
{
        grab=(unsigned char *)malloc(GRABSIZE*sizeof(char));
        interleave=(unsigned char *)malloc(GRABSIZE*sizeof(char));
        if((grab==NULL)||(interleave==NULL))
        {
                printf("Unable to allocate memory.\n\n");
                free(grab);
                free(interleave);
                exit(1);
        }
        DoScreen();                             //Setup display
        outportb(DataLPT1,0xff);                //Set data lines high (for PSU)
        while(!kbhit())                         //Keep going until keypress
        {
                if(Grab())                      //Grab one frame of data
                {
                        Interleave();           //Interleave 3 lines to make one
                        Display();              //Filter and display interleaved data
                }
                else
                {
                        gotoxy(5,5);
                        printf("** No Sync *");
                }
        }

        getch();                                //Clear keyboard buffer
        outportb(DataLPT1,0x00);                //Switch digitiser off
        textmode(3);                            //Revert to text mode
        free(grab);                             //Free memory
        free(interleave);
        exit(0);
}
//-----------------------------------------------------------
//Setup display to VGA320x200x256 and make palette
void DoScreen()
{
        static union REGS In_Regs;
        static union REGS Out_Regs;
        int i;
        float j=0;
        In_Regs.h.ah=0;
        In_Regs.h.al=0x13;                      //VGA mode 19
        int86(0x10,&In_Regs,&Out_Regs);
        for(i=0;i<32-BLKLVL;i++)
        {
                outportb(0x03C8,i+BLKLVL);
                outportb(0x03C9,(int)(j+0.5));  //Red
                outportb(0x03C9,(int)(j+0.5));  //Green
                outportb(0x03C9,(int)(j+0.5));  //Blue
                j+=(float)(64/(31-BLKLVL));
        }
}
//-----------------------------------------------------------
//Wait for V-Sync and Grab frame to buffer array
char Grab()
{
        asm {
                cli             //Disable interrupts to reduce jitter
                mov bx,TIMEOUT                  //Max number of samples with no sync
                mov dx,StatusLPT1               //Port address
        }
        vsloop1:
        asm {
                mov ah,10                       //Sync must be here for 10 loops
        }
        vsloop2:
```

median filter has been performed with good results, but is computationally intensive for real-time applications. Results of simple filtering are shown in Fig. 9.

Code execution will obviously be processor related. In the interests of brevity and understanding, the source given in this article is not the full assembly language program I have developed. Nevertheless, frame rates of 10Hz will be realised with a typical set-up. It is possible with optimisation and assembler to reach 25Hz on a 486DX33 with filtering. In the program, the choice of the constants GRABSIZE and LINELENGTH may have to be adjusted for faster/slower ports.

The design is a basis for a very cheap computer imaging system. Results may be limited, but will certainly allow experimentation with different enhancement and filtering techniques. An interesting application for the digitiser could be as an intelligent trigger for a security system video recorder. Colour capture is another possibility, by taking three successive red, green and blue frames.

More detailed assembler source code, including a simple movement tracking program, is available from the author. Send cheque for £12.50 UK or £15.00 overseas to, S.M. Webb, Selborne, Station Road, Clive, Shrewsbury, Shropshire SY4 3LD. A 3.5in disk will be dispatched, unless otherwise requested. Allow 28 days for delivery. Suggestions for improvements or optimisations are welcome. ■

Please direct all enquiries regarding this software, accompanied by an sae please, to Steve Webb at the above address – Ed.

```
        asm {
                dec bx
                jz no_sync                      //No sync found
                in al,dx                        //Get 5 Bits
                xor al,128                      //Hardware inverts MSB so change it back
                shr al,3                        //Move bits down giving a number from 0-31
                cmp al,SYNCLVL
                jnl vsloop1                     //Keep sampling until sync detected
                dec ah          //Sync found - check it's here for 10 loops
                jnz vsloop2
        }
        vsloop3:
        asm {
                dec bx
                jz no_sync                      //No sync found yet
                in al,dx
                xor al,128
                shr al,3
                cmp al,SYNCLVL
                jle vsloop3                     //End of sync not found yet
        } vdone:
        asm {
                les di,grab                     //Store data to grab array
                mov cx,GRABSIZE                 //Number of samples required
                cld                             //Make sure we increment di
                rep insb                        //Go get them!
                sti                             //Enable interrupts
        }
        return 1;                               //Sync was found
        no_sync:
        return 0;                               //Sync not found
}
//----------------------------------------------------------
//Routine to interleave grabbed frame
//
void Interleave()
{
        int x,y,l,i=0,b,s_line=0,pos;
        for(y=0;y<MAXLINES;y++)
        {
                for(l=0;l<=2;l++)               //3 source lines per one display
                {
                        pos=s_line+l;
                        while(((grab[i++]^128)>>3)>SYNCLVL);     //Search for start of sync
                        while(((grab[i++]^128)>>3)<=SYNCLVL);    //Search for end of sync
                        for(x=0;x<LINELENGTH;x+=3)
                        {
                                b=(((grab[i++]^128)>>3)+interleave[pos])>>1;
                                                        //Average current with
                                                        //previous sample
                                interleave[pos]=b;
                                pos+=3;
                        }
                }
                s_line+=LINELENGTH;             //Increment by one line
        }
}
//----------------------------------------------------------
//Display routine
void Display()
{
        int b,x,y,i=0,j=0;
        for(y=0;y<MAXLINES;y++)
        {
                for(x=0;x<LINELENGTH;x++)
                {
                        b=interleave[i];                        //No filtering

                        //b=(interleave[i-LINELENGTH]
                        //+interleave[i-LINELENGTH-1]
                        //+interleave[i-LINELENGTH+1]
                        //+interleave[i]
                        //+interleave[i-1]
                        //+interleave[i+1]
                        //+interleave[i+LINELENGTH]
                        //+interleave[i+LINELENGTH-1]
                        //+interleave[i+LINELENGTH+1])/9;
                        //      uncomment above for 3x3 filtering

                        //b=(interleave[i]
                        //+interleave[i+1]
                        //+interleave[i-1])/3;
                        //      uncomment above for 3x1 filtering

                        i++;
                        video[j++]=b;                           //Put on screen
                }
                j+=320-LINELENGTH;                              //Increment by one line
        }
}
```

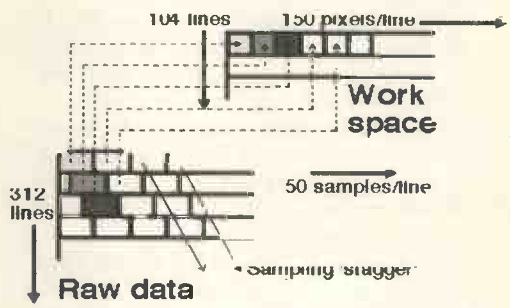


Fig. 8. Technique of interleaving sampled data. Samples from successive lines are staggered relative to each other, to the left or to the right. The trick is to finding out which direction the stagger is.

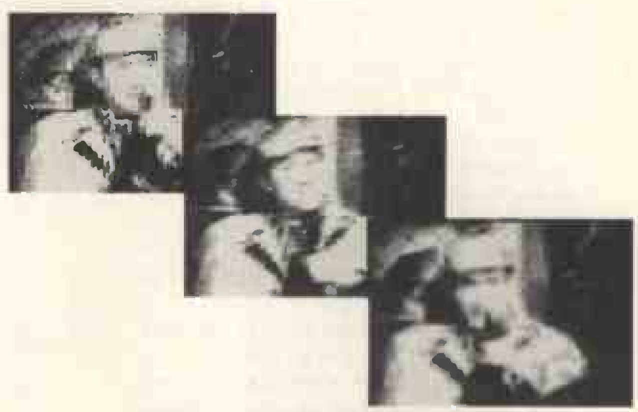
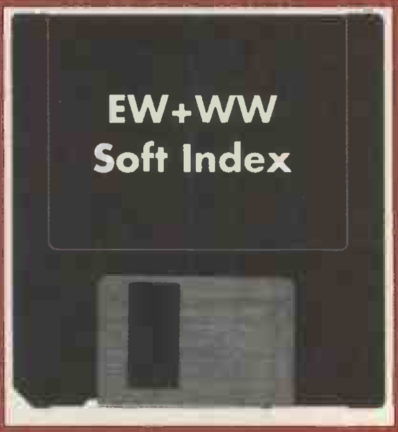


Fig. 9. "Its perfectly simple Watson" – no filter compared with averaging at 3x1 and 3x3 .

SOFT INDEX ON DISK



A computerised index of *Electronics World+Wireless World* magazine is now available. It covers the five years 1990 to 1994 – volumes 96 to 100 – and contains over 1400 references to feature articles, circuit ideas and applications, with a synopsis for each. The software is easy to use and very quick. It runs on any IBM or compatible PC with 512K ram and a hard disk. Each disk is scanned before shipping with the current version of Dr Solomon's Anti-Virus Toolkit.

For the UK, the five year index is priced at £20. Please specify $5^1/_4$ or $3^1/_2$in format. This price includes UK postage and VAT. Add an extra £1 for overseas EC orders or £5 for non-EC overseas orders.

Photo copies from back issues of *EW+WW* are available at 50p per page plus VAT (in EC) and a flat postage charge of 50p (UK), £1 (rest of EC), and £2 (rest of world). For enquiries about photo copies send an sae to Video Interface Products.

Please allow up to 28 days for delivery. Cheques should be made payable to **Video Interface Products,** not *EW&WW* or Reed Business Publishing.

Please post your request to **Video Interface Products Ltd, 1 Vineries Close, Cheltenham GL53 0NU, UK.**

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200pF	0.1pF	1%±1 dig.
2nF	1pF	1%±1 dig.
20nF	10pF	1%±1 dig.
200nF	100pF	1%±1 dig.
2µF	1nF	1%±1 dig.
20µF	10nF	1%±1 dig.
200µF	100nF	1%±1 dig.
2mF	1µF	2%±10 dig.
20mF	10µF	2%±10 dig.

Inductance

Range	Resolution	Accuracy
200µH	0.1µH	2%±2 dig.
2mH	1µH	1%±2 dig.
20mH	10µH	1%±2 dig.
200mH	100µH	1%±2 dig.
2H	1mH	2%±2 dig.
20H	10mH	2%±2 dig.
200H	100mH	3%±2 dig.

Resistance

Range	Resolution	Accuracy
2Ω	1mΩ	1%±5 dig.
20Ω	10mΩ	1%±2 dig.
200Ω	100mΩ	1%±2 dig.
2kΩ	1Ω	1%±2 dig.
20kΩ	10Ω	1%±2 dig.
200kΩ	100Ω	1%±2 dig.
2MΩ	1kΩ	2%±2 dig.
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Name ____________________

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____

____

Credit Card name, no ____________________

Expiry date ____________ **Tel:** ____________

Signed ____________ **Date** ____________

SURFING with intent

Accesing Internet for useful research is easier than you might think, as consultant Cyril Bateman has been finding out.

With a modern computer and a telephone line, the world's largest reference library is available 24 hours a day. This library – the Internet – resides on a large number of computer systems, in nearly every country in the world, and is available almost for free.

If it is the world's largest, just how big is it? In England, for public reference, we have the Science Reference Library and the Patent Office, at Holborn in London. Both house large collections of books and papers going back over many years. Imagine this information, indexed in computer searchable format, accessible by telephone, then multiply the amount of information a thousand fold.

For me as a consultant, losing the Norwich Library and sub-Patent Office by fire last autumn was a disaster. My nearest public reference alternatives are a three hour journey away. So I decided to find out whether the Internet would provide an alternative source.

Reading the various specialist computer magazines, and the two most recommended books, left me with the impression this was not so. It seemed the Internet had three main uses – sending E mail, obtaining shareware programs and participating in newsgroups.[1,2]

Encouraged by last autumn's price reductions for modems however, I decided to find out for myself. I bought a *Zoom 14.4X* modem and installed the easy to use and excellent communications software included with *OS/2 Warp* in the Bonus Pack – free.

I found that the Internet was not only the largest library, but was also easily accessible. What then is Internet? It is the name for a group of world-wide information resources, located in universities, technical colleges, schools, public libraries, businesses, government offices, patent offices. These resources, or their indexes, are stored on computer systems, linked by networks.

Amazingly, no one owns or controls Internet

Fig. 1. IBM New York World Wide Web home page. The starting point of our journey. Click on third menu item to access reference files sub-menu. Central bar shows present location, bottom shows moving to.

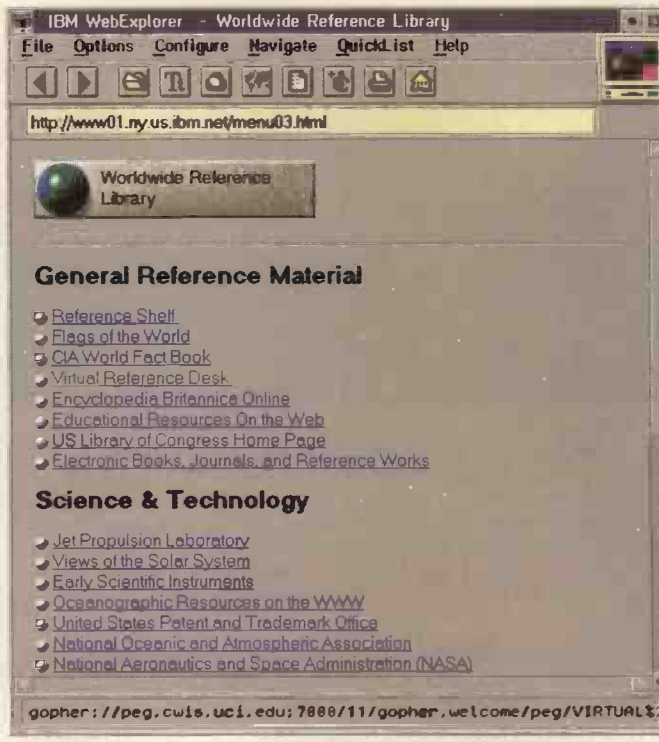


Fig. 2. Still at IBM New York, second click on fourth menu item, to access Reference Desk at 'peg.cwis.uci.edu' University California, Irvine College.

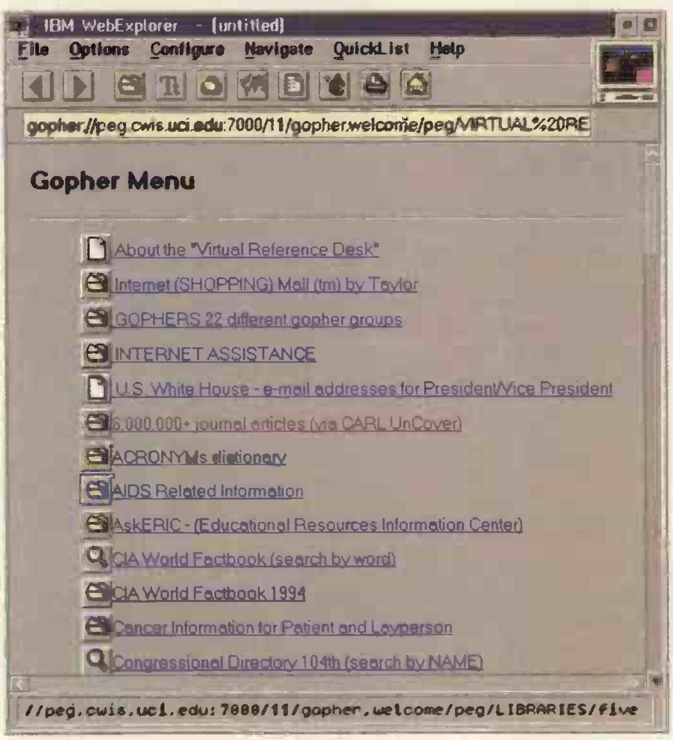


Fig. 3. Now at 'peg.cwis.uci.edu'. Third click, on sixth menu item to change host computer. University of California, Irvine College.

Fig. 4. Now at Carl UnCover Libraries'. Gopher computer. Fourth click on second menu, seamlessly starts 'Telnet' session.

and by and large the services are available for the cost of the telephone call to the service provider used. Each resource e.g. at a university, is created, controlled, and managed for use on the campus by their own staff. Much of the information is supplied voluntarily.

Obviously, in order for each computer to communicate, an operating-system independent 'protocol' has been devised. This, supervised by the National Science Foundation, is perhaps the only controlled aspect of the Internet.[3]

When visualising a network, most people assume that there is one central controlling server computer, implying a 'big-brother' supervising and controlling access. So why does no one control Internet?

Very simple. The 'Internet' has no central server. It is organised like a 'peer-to-peer' network, computers both giving and receiving outside access. Each has a unique address. Data is sorted and passed on to its destination. Public access to the Internet requires the services of a 'service' provider, acting either as a

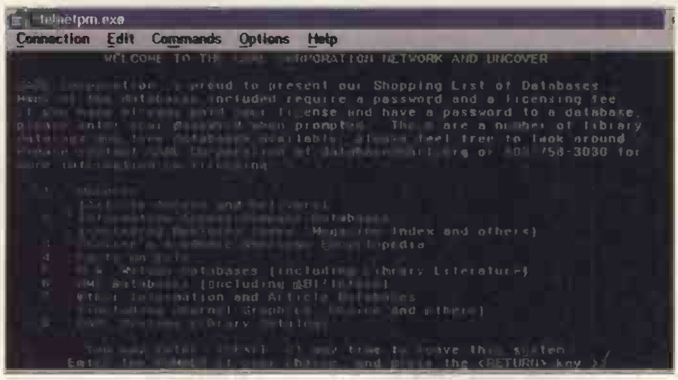


Fig. 5. Telnet' program started transparently on local computer, for user. Telnet permits access to the remote computer as though seated at Carl UnCover. This is the Welcome screen viewed on the local computer. Menu item 1 is for public access to browse, free of charge, the 6 million articles held. Search tools are provided, on-line, to search by titles, authors and text body of articles. Having identified the required article, it might be available locally. Alternatively requested articles can be sent by fax for nominal charges.

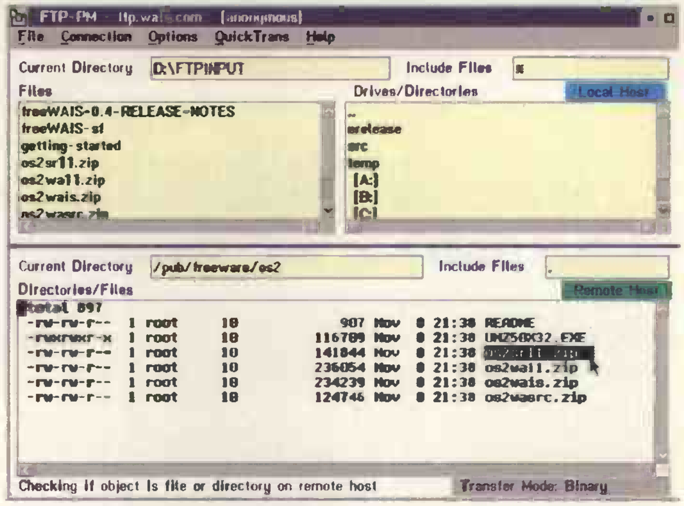


Fig. 7. Screen showing file-transfer program, FTP, downloading files. Upper screen shows directories of local computer, lower shows shows files to be transferred to provide full WAIS searches by local computer. Host is wais.com (WAIS Inc) – a Unix system.

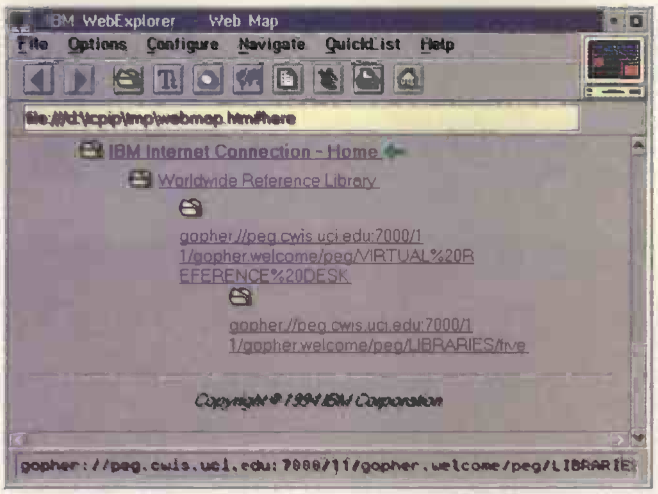


Fig. 6. The 'road map log' of World-Wide Web computers visited. As a result of simply clicking on four menu items. No other actions were required of the user. The Web automatically inputs the necessary addresses. Note that the Unix directory is accessed automatically.

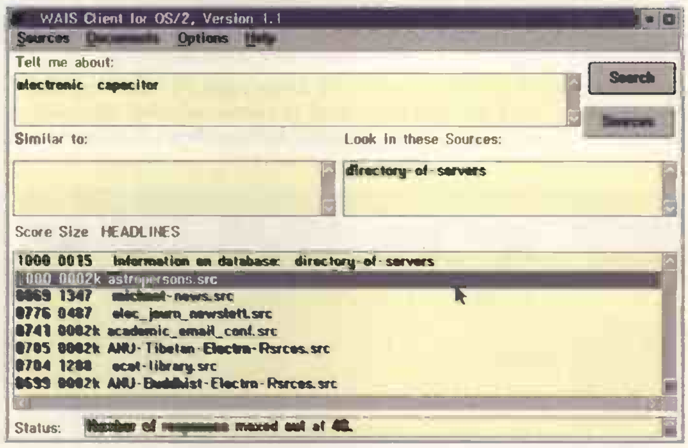


Fig. 8. Down-load WAIS searche software in use. Result of first-stage search of 620 computers, lists of servers holding files related to the 'tell me about query "electronic capacitor"'. Additional searches of a selection of servers are needed to get to file level.

'poste-restante' mailbox or telephone switch-board.

One's costs are the telephone calls to this service provider and the providers charges. Two levels of connectivity are commonly available:

● BBS 'dial-up' is generally restricted to E-mail, a number of news groups, file transfers, and Telnet. Access is indirect, the BBS computer intercepting all data, for subsequent retransmission both to and from the Net.

● SLIP/PPP 'dial-in' provides direct access to all Internet services. The 'provider' acts as a switch-board connecting your computer directly to the Internet. All data is handled in both directions in real time. Local use of the graphical browser called the World-Wide Web is used to access sources. This level of service is essential for any serious scientific reference searching.[4]

Origins of the Internet

In 1969 the US military needed to connect four computers in such a way as to maintain a level of service assuming one or more computers or links, were damaged. It was to be known as the Darpanet. With time, the name changed to Arpanet as more computers were added.

With the number of University sites connected by the early '80s, this network had grown to such extent it was decided to segregate military and research systems. In 1984, the National Scientific Foundation NSFNet was established to link together five supercomputer centres, each located at a university, making the information contained accessible to any desiring US educational facility that needed it.

This access was later broadened to include other countries, allies of the USA. By 1990, the Internet had begun, and access was opened up to anyone having the means to connect. From a beginning with some 5000 users it has grown to the present estimate of over 30 million.

While Unix was dominant initially, with this level of expansion all operating systems are now included across some 62 countries.

Hypertext and World-Wide Web

The idea of hypertext is not particularly new. Hypertext is data that contains links to other relevant data. The World-Wide Web concept was developed at the CERN research centre, Switzerland, to disseminate information.

In 1980, Tim Berners-Lee devised Enquire Within – a program for his own use, designed to facilitate insertion and cross-referencing of data links in technical reports. These basic ideas expanded in 1991 into a text based access system on Unix for the Internet. This became the Mosaic graphical user interface for Unix in February 1993, and was subsequently translated for all other platforms.

In the present day World-Wide Web, clicking on hypertext highlighted keywords takes one directly to the next link. This happens regardless of changes in host computer, country or even communication methods. All are transparent to the user, **Fig. 1**.

As an example, you want to transfer a file to your computer. The Web link will dial the required host computer, log in to the file-transfer service FTP, commence FTP on your computer and download the file for you. All of this is autonomous – even if the host is running Unix and the receiving computer is running Windows, OS/2, System7, or whatever.[5] Photo sequence **Figs 1-4** demonstrates this.

Internet service providers

While Internet can be accessed in many ways, for Scientific reference, two methods only are considered. The preferred 'dial-in' route requires a minimum of computer skills. Access through a BBS is generally more restricting, and can require memorising a number of Unix commands to perform searches. Note the Unix directory structure, involving a forward slash as opposed to the backslash used in dos, Figs 6,7.

Some popular national providers, taken from the Paola Kathuria list, Feb. 95 are,

Dial-in providers (SLIP/PPP)
Atlas, £12/month and £25 start up fee
BBC, £12/month and £25/35 start-up
Cityscape, £180/year and £50 start-up fee
Demon, £10/month, unlimited time, no hourly charge, software less refined, £12.50 start-up fee.
Direct, £10/month, unlimited time, provide your own software.
IBMnet, £10/month to 3h/month, then £3 subsequent, super software free with OS/2Warp, 1st month/3h free, no start-up fee. BBS providers
CompuServe, £6.65/month and £3.20/h CIX, £6.25/m to 1h 45 min (2h 36min offpeak) then £0.06 (£0.04 off peak) per minute, £25 start-up fee.
DELPHI, £10/month to 4h, then £4/h, 1st 5h free, no start-up fee.

How do I join Internet?

Given a suitable computer, for example a pc running *Windows*, *OS/2 Warp*, Macintosh *System 7*, Archimedes *Risc-OS*, etc. and a modem are the only essentials.

Examine your intended use. Most facsimile machines work at 9600baud. Many BBS now work at 28,800baud. Most UK Internet providers work at 14,400baud but depending on 'traffic' the 'Internet' connection can be slower.

If running *Windows*, unless your computer uses a 16550A serial chip, your system can limit the actual run rate. *OS/2 Warp* accesses the serial port more efficiently than does Windows. It provides a 5Kbyte receive buffer, better interrupt handling, resulting in faster serial data transfer (see the modem panel).

Choose a service provider, tell the provider your credit card details for billing service charges, install the software and log on (see the providers panel)

Bit rates in practice

Accessing a UK bulletin board at 14,400baud using Zmodem protocol, data transfers at around 100Kbyte per minute. With the same hardware, Internet achieves around 60 to 100Kbyte per minute.

One aspect that confused me initially was the relationship between the Modem's claimed data rate and the computer serial port transfer rate. This is a setting required to install most communications software. As a rule of thumb, with a 9,600baud modem, set 9,600 bps rate. With a 14,400baud modem set to three or four times this rate, 57,600 bit/s for example, or as fast as the system will accept in practise.

Data transfer

All Internet transfers make use of the Transmission Control Protocol/Internet Protocol, TCP/IP, which started life as the UNIX networking standard. As a protocol for networking it has spread into conventional networks. During 1992, an open standard for TCP/IP under Windows was defined, now known as *Winsock*. Today versions exist for most operating systems.

Using the system is like sending a letter, one page at a time in individually addressed envelopes. One page is a maximum of 1500 bytes of data, addressed to the receiving computer. Pages can arrive out of sequence and must be sorted by the receiving computer. Any garbled pages are automatically resent, transparently to the operator, who sees a record of the reception in bytes.

What can be found on the Internet?

Basically data of any form can be found on the Internet. This may be correspondence, program files, databases, graphics, audio, video – in fact anything which can be stored or processed in a computer. In addition, there are over 7000 news groups.

Internet has many libraries. One of these, the Carl Corporation, has a system called Uncover, which can be freely searched. Any required document can be faxed to you for a nominal charge. This data-base presently holds some 20,000 Journal Titles – mainly scientific – spanning 1988 to date. In total this represents some 6 million articles.

Searching can take many forms:

- by journal, title and contents page
- by authors of articles
- by keywords within the article
- by article title or summary.

See **Fig. 5**.

Modems

Typical serial port performance of pc compatible computers running Windows is shown in the following table,

Modem data transfers may be limited by the uart pc combination.

Processor	Uart	Speed in bit/s
386 SX	8250/16450	19,200
386 SX	16550A	38,400
386 DX	8250/16450	19,200
386 DX	16550A	57,600
486 (all)	8250/16450	38,400
486 (all)	16550A	115,200

Source – *Windows International*, Dec. 1993 p. 208.

Modem choice to access the Internet using World-Wide Web browsers is basically a choice between 14400baud as a minimum, or faster. Depending on the uart in your computer, you might have to buy an internal modem with a *16550A* chip built in, or a high-speed serial interface for external modems. Modern modems have built in hardware compression, hence to minimise initial and on-line costs, the computer uart must output data fast enough to use this.

Typical modem prices

Modem baud	Max. serial-port bit rate	Typ. price (BABT app.)
9,600	19,200bit/s	£90
14,400	57,600bit/s	£135
28,800	115,200bit/s	£200

Modems are labelled with many 'V' claims, however simply base your choice on a fax-modem with the desired bit rate. By and large all other functions follow suit.

Searching for resources

Certain specialist computers are called servers since by acting as a librarian, searching records and pointing you in the right direction, they 'serve' the 'Internet' to you. Each maintains very large databases routinely updated by accessing each data computer's files. These servers are dedicated to perform specific searches.

Archie. Many computers allow the public to log on anonymously, read their directories, and download files. Some estimates suggest over 1000 such computers now exist, housing over 2.5 million files, some 50 gigabytes in total. Selecting one computer at random cannot be expected to locate a specific filename. A group at McGill University has created a program for tracking each computer's content. On request, it lists computer addresses where the file can be accessed. This search tool, restricted to single word searches, is Archie. If the filename is unknown, the files' descriptive-field indexes can be searched.

Gopher. In 1991 the University of Minnesota developed a menu based search system to 'Go For' information. The college team was the Golden Gophers and the name Gopher stuck. The resulting information base on some 5000 servers, is known as Gopherspace.

Gopher is multipurpose: it provides access to information changes, Gopher servers or performs transactions in response to your menu choices and search requests.

Veronica. To search Gopherspace, a variation of Archie, developed at the University of Nevada and called 'Veronica', allows multiple word-search strings with Boolean controls. It assumes implicit 'and' unless otherwise instructed. To search a specific Gopher only, another variation called Jughead was developed. Fortunately this name doesn't appear in menus; it is implied by, for example[6], 'Search Gopher Menus at the University of Minnesota'.

Wais. The final search tool is of a much different form. Its name, Wais, is pronounced Ways, for Wide-Area information service. This tool grew out of a project from three companies – Apple, Thinking Machines and Dow Jones. It performs a full document search. In response to keywords, Wais searches all text, recording occurrences of specified keywords in each article, and normalising this 'score' such that the document with most hits rates 1000. It then presents you with a list of higher scoring documents. This can be highly beneficial when searching for topical references, but since 'Wais' is not context sensitive it can also provide some high scoring irrelevancies. At present Wais is the ultimate 'tell me about xxx' tool, freely available, **Figs 7,8**.

Having identified and located the file, the actual transfer is performed by a program called file-transfer protocol, FTP. In December 1994, FTP accounted for 31% of total Internet activity, Fig. 7.[3]

World-Wide Web

While not in itself a search tool, the World-Wide Web, is the glue that holds together all these resources. Using menus and hypertext links, the front end Unix graphical browser Mosaic, now ported to Windows, Macintosh machines, etc., resulted in the recent radical explosion in Internet accesses.

In December 1994, the World-Wide Web was the second largest activity, accounting for 16% of all data transmitted. In all this represents some 3,314Gbytes out of a total activity of 20,743Gbytes.[3] See **Figs 1-6**. ■

References.

1. Hahn & Stout, The Internet Complete Reference, Osborne McGraw-Hill
2. Ed Krol, The Whole Internet, O'Reilly & Associates Inc.
3. NSF Statistics, gvu.center.nsf.statistics
4. WWW – Frequently asked questions, http://sunsite.unc.edu/boutell/faq/
5. FTP – Frequently asked questions, comp.sources.wanted
6. Veronica – How to compose queries, gopher://veronica.scs.unr.edu

Useful documents

CICNets Electronic Journal Archives
gopher.cicnet/00/e-serials/readme
Hytelnet database of publicly accessible sites, ftp.usask.ca/pub/hytelnet/pc/latest
JANET - OPACS in the UK, nic.funet.fi/pub/doc/library
Library resources on Internet, dia.ucop.edu/pub/internet/libcat_guide
Paola Kathuria UK access providers, ftp.demon.co.uk/pub/archives/

TRANSISTORS

Part	Price
AAY32	9p
AC107	40p
AC125	30p
AC126	30p
AC127	30p
AC128K	40p
AC141K	45p
AC176	22p
ACY18	48p
ACY19	48p
AD149	60p
AF125	50p
AF139	30p
AF239	30p
BB105B	18p
BB205B	24p
BC107	8p
BC108	8p
BC109	8p
BC109C	10p
BC140	20p
BC142	20p
BC143	20p
BC147	8p
BC149	8p
BC159	8p
BC160	30p
BC171	10p
BC172	10p
BC177	14p
BC178	14p
BC179	14p
BC182	7p
BC182L	7p
BC183	7p
BC183L	7p
BC184	7p
BC184L	7p
BC212	7p
BC212L	7p
BC213	7p
BC213L	7p
BC214	7p
BC214L	7p
BC237	7p
BC238	7p
BC239	7p
BC300	20p
BC301	20p
BC302	20p
BC303	20p
BC304	25p
BC327	7p
BC328	7p
BC337	7p
BC338	7p
BC441	28p
BC446	8p
BC477	18p
BC516	22p
BC537	25p
BC546	8p
BC547	8p
BC548	8p
BC549	8p
BC550	8p
BC556	8p
BC557	7p
BC558	8p
BC559	8p
BC560	8p
BC637	20p
BC639	20p
BC640	20p
BCY33	200p
BCY34	200p
BCY70	16p
BCY71	16p
BCY72	16p
BD115	30p
BD124P	50p
BD131	25p
BD132	25p
BD133	50p
BD135	20p
BD136	20p
BD137	20p
BD138	20p
BD139	20p
BD140	20p
BD144	90p
BD157	38p
BD166	30p
BD175	30p
BD177	30p
BD179	32p
BD181	45p
BD182	60p
BD184	60p
BD187	30p
BD201	33p
BD202	38p
BD203	42p
BD204	42p
BD222	31p
BD225	31p
BD232	31p
BD233	30p
BD234	32p
BD235	28p
BD236	30p
BD237	21p
BD238	24p
BD239	30p
BD240	40p
BD241A	40p
BD243A	50p
BD244	50p
BD245	50p
BD246A	50p

Part	Price
BD265	45p
BD267	45p
BD269	45p
BD278	50p
BD311	100p
BD314	100p
BD315	150p
BD317	150p
BD331	40p
BD332	40p
BD361	60p
BD362	60p
BD370	30p
BD371	30p
BD410	50p
BD433	28p
BD434	30p
BD435	31p
BD436	30p
BD437	28p
BD438	36p
BD439	40p
BD440	40p
BD441	40p
BD533	50p
BD534	38p
BD535	38p
BD536	38p
BD537	40p
BD538	40p
BD643	50p
BD645	50p
BD647	50p
BD649	50p
BD675	40p
BD676	40p
BD677	38p
BD678	40p
BD679	40p
BD680	40p
BD681	45p
BD682	45p
BD705	50p
BD707	50p
BD709	50p
BD711	50p
BD736	50p
BD826	50p
BD828	50p
BD839	55p
BD897	50p
BD899	50p
BD977	50p
BDX33	60p
BDX65	80p
BDW24	55p
BDW93	50p
BDW94	50p
BDY92	100p
BF137	35p
BF167	30p
BF181	18p
BF183	20p
BF195	7p
BF199	8p
BF200	16p
BF225	30p
BF240	16p
BF245	25p
BF254	15p
BF255	12p
BF256	18p
BF257	18p
BF259	18p
BF262	25p
BF270	15p
BF273	15p
BF311	21p
BF336	20p
BF337	20p
BF338	20p
BF362	30p
BF367	30p
BF371	17p
BF421	18p
BF422	21p
BF423	25p
BF455	18p
BF458	19p
BF462	50p
BF471	28p
BF472	28p
BF479	30p
BF494	16p
BF495	16p
BF595	16p
BF596	16p
BF615	30p
BF617	30p
BF760	40p
BF763	20p
BF870	22p
BF871	22p
BF960	38p
BF961	35p
BF964	38p
BFR90	85p
BFR91	99p
BFT43	30p
BFX29	20p
BFX84	20p
BFX85	20p
BFX87	15p
BFX88	15p
BFX89	60p
BFY50	14p
BFY51	14p
BFY52	14p
BFY56	25p
BFY64	25p

Part	Price
BFY90	45p
BLY48	85p
BR100	14p
BR103	37p
BR303	85p
BSS74	33p
BSX20	15p
BT100A	70p
BT106	180p
BT109	90p
BT119	100p
BT146	99p
BTY79	140p
BU105	80p
BU108	100p
BU109	80p
BU110	90p
BU111	100p
BU124	60p
BU126	65p
BU180	100p
BU184	100p
BU204	65p
BU205	70p
BU206	100p
BU208	70p
BU208A	75p
BU208AT	200p
BU208D	130p
BU209	90p
BU225	120p
BU226	120p
BU312	90p
BU325	55p
BU326A	75p
BU406	60p
BU406D	85p
BU407	55p
BU407D	75p
BU408	60p
BU408D	75p
BU409	85p
BU426A	70p
BU500	100p
BU505	90p
BU505D	90p
BU505DF	90p
BU506	100p
BU506D	90p
BU506DF	120p
BU508A	75p
BU508AF	95p
BU508D	75p
BU508DF	115p
BU508V	110p
BU508VF	100p
BU526	70p
BU536	100p
BU546	125p
BU608	120p
BU626	120p
BU705	130p
BU706DF	175p
BU706F	150p
BU801	70p
BU806	70p
BU807	60p
BU902	110p
BU903	110p
BU920	100p
BU922	110p
BU930	130p
BU2508A	130p
BU2508AF	130p
BU2508D	130p
BU2508DF	150p
BU2520AF	225p
BU2520DF	225p
BU2525AF	325p
BUH515	200p
BUT11AF	55p
BUT12	80p
BUT56A	75p
BU18	80p
BU18AF	80p
BUX10	150p
BUX11	200p
BUX12	150p
BUX20	350p
BUX21	450p
BUX22	450p
BUX37	220p
BUX40	210p
BUX41	200p
BUX42	200p
BUX47A	220p
BUX48A	150p
BUX80	180p
BUX84	50p
BUX85	50p
BUX86	30p
BUX87	50p
BUX98A	350p
BUY69A	200p
BUY71	250p
BUZ11	200p
BUZ71	75p
BUZ80	200p
BY448	20p
BYT11	25p
C106D	20p
IRF630	150p
J174	38p
J300	50p
MJ900	200p
MJ1000	200p
MJ1001	200p
MJ10012	300p
MJ15003	250p
MJ15004	300p

Part	Price
MJ2501	100p
MJ2955	55p
MJ3000	100p
MJ3001	100p
MJE29A	30p
MJE30A	30p
MJE340	25p
MJE350	80p
MJE520	30p
MP8112	45p
MPSA05	15p
MPSA06	15p
MPSA13	15p
MPSA20	15p
MPSA42	15p
MPSA43	15p
MPSA70	15p
MPSA92	20p
MPSA93	20p
MR510	35p
MR856	36p
OC28	350p
OC29	250p
OC35	350p
OC36	250p
OC45	50p
OC200	180p
R2008B	100p
R2010B	100p
S2000A3	175p
S2000AF	175p
S2055A	175p
S2055AF	200p
S2530A	100p
S2800M	72p
TIP29	15p
TIP29A	22p
TIP29C	25p
TIP29E	40p
TIP30	25p
TIP30C	25p
TIP31A	25p
TIP31C	27p
TIP32	24p
TIP32A	21p
TIP32C	28p
TIP33	50p
TIP33C	60p
TIP34	50p
TIP34C	60p
TIP35C	65p
TIP36C	65p
TIP41A	20p
TIP41C	22p
TIP42A	20p
TIP42C	22p
TIP47	40p
TIP48	40p
TIP50	60p
TIP51	80p
TIP52	80p
TIP54	85p
TIP105	65p
TIP106	65p
TIP107	65p
TIP110	40p
TIP111	40p
TIP112	35p
TIP112H	50p
TIP115	30p
TIP116	30p
TIP117	30p
TIP120	37p
TIP121	35p
TIP122	30p
TIP125	30p
TIP126	40p
TIP127	35p
TIP130	30p
TIP131	30p
TIP132	30p
TIP141	65p
TIP142	75p
TIP145	70p
TIP146	70p
TIP147	80p
TIP150	90p
TIP151	60p
TIP2955	42p
TIP3055	42p
TIPL760	100p
TIPL763A	200p
TIPL791A	80p
TIS61	15p
TIS90	15p
TIS93	20p
ZTX107	11p
ZTX108	11p
ZTX109	12p
ZTX212	20p
ZTX300	10p
ZTX301	16p
ZTX302	10p
ZTX303	20p
ZTX304	10p
ZTX320	20p
ZTX501	13p
ZTX502	10p
ZTX503	18p
ZTX504	25p
2N696	26p
2N697	22p
2N698	40p
2N78	22p
2N914	28p
2N930	18p
2N1131	28p
2N1132	28p
2N1613	24p
2N1711	24p
2N1893	30p

Part	Price
2N2102	50p
2N2218A	24p
2N2219	24p
2N2221	23p
2N2222	23p
2N2369	15p
2N2484	15p
2N2646	40p
2N2904	20p
2N2905	20p
2N2906	18p
2N2907	18p
2N3019	28p
2N3053	18p
2N3054	40p
2N3055	38p
2N3055H	50p
2N3442	85p
2N3702	9p
2N3703	9p
2N3704	9p
2N3705	9p
2N3706	9p
2N3707	9p
2N3710	12p
2N3711	12p
2N3771	85p
2N3772	90p
2N3773	100p
2N3799	18p
2N3819	29p
2N3903	11p
2N3906	11p
2N4031	25p
2N4401	12p
2N4403	12p
2N5061	20p
2N5088	20p
2N5192	50p
2N5241	500p
2N5245	45p
2N5294	30p
2N5296	30p
2N5448	12p
2N6107	40p
2N6292	40p
2N6385	120p
2N6403	160p

RECTIFIER DIODES

Part	Price
BY127	8p
BY133	8p
BY164	40p
BY179	35p
BY184	32p
BY206	11p
BY207	9p
BY227	19p
BY228	28p
BY298	15p
BY299	18p
BY448	20p
BYX10	15p
BYX55/600	25p
BYX70/500	
OA47	10p
OA91	10p
OA202	10p
IN4001	3p
IN4002	3p
IN4003	3p
IN4004	3p
IN4005	3p
IN4006	3p
IN4007	4p
IN4148	2p
IN5400	9p
IN5401	8p
IN5402	8p
IN5403	8p
IN5404	8p
IN5405	11p
IN5406	12p
IN5407	12p
IN5408	12p
RGP15	25p
RGP30	16p
SKE4F2/06	60p
SKE4F2/08	80p
SKE4F2/10	100p
SR2M	60p

I.C. SOCKETS

Part	Price
8 PIN	5p
14PIN	6p
16PIN	7p
18PIN	10p
20PIN	12p
22PIN	13p
24PIN	14p
28PIN	16p
40PIN	18p

ZENERS

Part	Price
400 mWatts	
2V7 to 39V	5p
1.3 Watts	
2V7 to 39V	9p

VOLTAGE REGULATORS

Part	Price
7805	25p
7806	25p
7808	25p
7812	25p
7815	25p
7818	25p
7824	25p
7905	25p
7906	30p
7908	30p
7912	30p
7915	30p
7918	30p
7924	30p
78L05	24p
78L08	24p
78L12	24p
78L15	24p
78L18	24p
78L24	24p
79L05	35p
79L08	35p
79L12	35p
79L15	35p
LM309K	100p
LM317T	100p
LM323K	350p
78H08KC	800p
79H12KC	700p
79HGKC	800p

LEDs

Part	Price
3mm	
RED	5p
YELLOW	8p
GREEN	8p
5mm	
RED	5p
YELLOW	8p
GREEN	8p

RECTANGULAR LEDs

Part	Price
5mm × 2.5mm	
RED	5p
YELLOW	8p
GREEN	8p

OPTO COUPLERS

Part	Price
4N37	58p
4N38	68p

BRIDGE RECTIFIERS

Part	Price
W005	16p
1A/50V	
W01	18p
1A/100V	
W02	19p
1A/200V	
W04	21p
1A/400V	
W06	23p
1A/600V	
W08	28p
1A/800V	
BR81D	33p
2A/100V	
BR82D	33p
2A/200V	
BR84D	37p
2A/400V	
BR86D	43p
2A/600V	
BR88D	43p
2A/800V	
BR32	43p
2A/200V	
BR34	43p
2A/400V	
BR36	44p
2A/600V	
BR62	80p
6A/200V	
BR64	72p
6A/400V	
BR251	150p
25A/100V	
BR252	165p
25A/200V	
BR254	185p
25A/400V	
BR256	200p
25A/600V	
BR258	240p
25A/800V	
BR351	185p
35V/100V	
BR352	200p
35V/200V	
BR354	220p
35V/400V	
BR356	230p
35V/600V	
BR358	260p
35V/800V	
BY164	40p
1.5A/100V	
BY176	40p
1.5A/800V	

TRIACS

Part	Price
TIC206D	60p
4A/400V	
TIC225D	69p
6A/400V	
TIC226D	68p
8A/400V	
TIC236D	85p
12A/400V	
TIC246D	105p
16A/400V	
TIC253D	190p
20A/400V	
TIC263D	205p
25A/400V	

THYRISTORS

Part	Price
2N5061	20p
0.8A/60V	
TIC116C	59p
8A/300V	
TIC116D	70p
8A/400V	
TIC126D	75p
12A/400V	
TIC126M	90p
12A/600V	
C106D	28p
4A/400V	
BR103	37p
BR303	85p
BT106	180p
BT119	100p
17088	200p
17089	200p
17127	200p
15/80H	230p
15/85R	230p
SG264	800p
SG613	1500p

COMPUTER ICs

Part	Price
Z80ACPU	100p
Z80ADMA	200p
Z80ACTC	140p
Z80ASIO-1	210p
Z80ASIO-2	210p
75107	65p
75110	75p
75113	100p
75122	110p
75154	100p
75162	700p
75182	95p
75183	95p
75195	185p
2114	150p
2532	200p
2716	100p
2732	200p
2732A	220p
2764	150p
27C64	200p
27128	150p
27256-25	150p
27512	300p
4116	40p
4164-15	80p
4164-12	90p
41256-15	80p
41256-12	100p
41256-10	110p
41464-12	150p
6116	80p
6264-10	210p
62256-12	300p
6502A	360p
65C02	930p
6522	280p
6800	210p
6802	220p
680	500p
36808	500p
6809	500p
6810	150p
6818	380p
6821	130p
6840	290p
6845	200p
6850	90p
8085A	300p
8086	500p
8088	480p
8156	300p
8224	240p
8226	240p
8250	750p
8251	200p
8253	160p
8257	220p
8271	3400p
8279	270p
8283	400p
8284	440p
8287	260p
8288	650p
8748	700p
8755	800p
8T26	95p
8T28	110p

LINEAR ICs

Part	Price
AN203	210p
AN210	165p
AN214Q	170p
AN228	280p
AN252	150p
AN259	250p
AN262	140p
AN271	230p
AN274	250p
AN301	330p
AN303	250p
AN304	360p

Part	Price
AN315	210p
AN316	350p
AN360	100p
AN362	140p
AN366	150p
AN610	160p
AN3312	350p
AN3821K	600p
AN3822K	600p
AN3990K	300p
AN3991K	400p
AN5025	250p
AN5033	400p
AN5132	250p
AN5150	400p
AN5151	600p
AN5215	100p
AN5256	150p
AN5262	175p
AN5265	80p
AN5352	600p
AN5411	450p
AN5421	150p
AN5429	420p
AN5512	100p
AN5515	160p
AN5520	550p
AN5521	100p
AN5612	200p
AN5613	200p
AN5615	300p
AN5620	250p
AN5622	275p
AN5625	400p
AN5712	180p
AN5722	140p
AN5730	160p
AN5732	120p
AN5753	130p
AN5763	450p
AN5790	240p
AN5791	225p
AN5836	450p
AN5900	130p
AN6135	120p
AN6247	200p
AN6270	400p
AN6300	600p
AN6306	380p
AN6320	180p
AN6332	320p
AN6341	200p
AN6344	440p
AN6350	610p
AN6359	500p
AN6360	320p
AN6362	400p
AN6371	350p
AN6387	480p
AN6884	200p
AN7105	170p
AN7110	75p
AN7114	120p
AN7115	110p
AN7116	90p
AN7120	100p
AN7130	75p
AN7140	170p
AN7145	195p
AN7146	210p
AN7154	180p
AN7156	240p
AN7168	200p
AN7178	180p
AN7222	75p
AN7254	150p
AN7256	250p
AN7310	60p
AN7311	90p
AN7410	150p
AY3-1015	290p
AY3-1270	800p
AY3-1350	450p
AY3-8910	360p
AY3-8912	400p
BA301	55p
BA311	80p
BA313	60p
BA333	80p
BA401	60p
BA402	50p
BA511	145p
BA514	160p
BA516	150p
BA521	100p
BA524	240p
BA526	180p
BA527	95p
BA532	100p
BA534	220p
BA536	150p
BA546	160p
BA612	120p
BA656	110p
BA658	350p
BA684	400p
BA685	400p
BA1310	160p
BA1320	75p
BA1330	120p
BA1360	160p
BA4403	220p
BA5101	350p
BA5102	140p
BA5204	200p
BA5402	180p
BA5406	180p
BA5408	180p
BA6104	250p
BA6208	175p

Part	Price
BA6209	85p
BA6304	120p
BA6305	140p
BA6410	220p
BA6411	250p
BA6993	150p
BA7001	150p
BA7004	200p
BA7007	200p
BA7021	180p
BA7022	350p
BA7751LS	150p
BA7752	250p
BA7755	150p
BA7767AS	155p
CA3011	110p
CA3048	190p
CA3052	190p
CA3054	95p
CA3085	135p
CA3088E	200p
CA3089E	150p
CA3090Q	250p
CA3130S	100p
CA3134E	280p
CA3140E	38p
CA3160	85p
CA3189E	230p
CA3193E	230p
CA3260E	170p
CA3290E	150p
CX108	950p
CX136	600p
CX139A	750p
CX141	750p
CX145	725p
CX150B	325p
CX175	325p
CX187	825p
CX804A	775p
CX867	575p
CX868	525p
CX877	300p
HA1125	120p
HA1197	130p
HA1199	130p
HA1319	200p
HA1338	300p
HA1339A	350p
HA1377	120p
HA1388	320p
HA1389	210p
HA1392	120p
HA1394	170p
HA1397	200p
HA1398	240p
HA11219	280p
HA11221	180p
HA11225	130p
HA11235	120p
HA11251	190p
HA11423	140p
HA11724	650p
HA12002	220p
HA12003	250p
HA12005	180p
HA12017	100p
HA13001	110p
HA13002	200p
HA13006	400p
HA13007	400p
HA13108	350p
HA13412	600p
HA13432	400p
HA17524	250p
ICL7106	650p
ICL7660	240p
KA2102	150p
KA2130	150p
KA2206	150p
KA2209	125p
KA2210	230p
KA2212	80p
KA2213	130p
KA2214	150p
KA2261	100p
KA2263	100p
KA2264	100p
KA2284	100p
KA2401	150p
KA2412	350p
KA2912	125p
KA2914A	300p
LA1130	240p
LA1150	150p
LA1185	150p
LA1201	75p
LA1210	140p
LA1222	80p
LA1230	130p
LA1364	200p
LA1365	120p
LA1368	220p
LA1385	170p
LA2000	150p
LA2101	270p
LA2200	190p
LA3160	120p
LA3210	65p
LA3300	140p
LA3301	110p
LA3361	100p
LA3375	300p
LA4030	180p
LA4031	140p
LA4032	140p
LA4051	160p
LA4100	85p
LA4101	80p
LA4102	100p

Part	Price
LA4110	120p
LA4120	270p
LA4140	60p
LA4160	100p
LA4182	180p
LA4190	300p
LA4192	140p
LA4200	130p
LA4201	120p
LA4260	230p
LA4261	300p
LA4270	300p
LA4420	140p
LA4422	130p
LA4430	130p
LA4440	150p
LA4445	150p
LA4460	120p
LA4461	120p
LA4500	200p
LA4505	220p
LA4508	200p
LA4510	100p
LA4520	170p
LA4550	200p
LA4555	120p
LA4570	130p
LA5112	200p
LA5523	150p
LA5527	150p
LA5700	300p
LA7011	220p
LA7033	400p
LA7042	280p
LA7046	300p
LA7224	150p
LA7505	250p
LA7507	250p
LA7520	200p
LA7620	500p
LA7800	90p
LA7801	100p
LA7802	300p
LA7806	260p
LA7808	250p
LA7820	100p
LA7823	200p
LA7910	150p
LA7940	200p
LC7131	260p
LC7132	400p
LC7137	450p
LF347	110p
LF353	48p
LF355	60p
LF357	70p
LF398	300p
LM301	26p
LM311	35p
LM319	165p
LM324	30p
LM335Z	120p
LM339	35p
LM348	50p
LM358	45p
LM380	80p
LM381	150p
LM382	130p
LM386	60p
LM387	100p
LM393	45p
LM431	50p
LM710	45p
LM723	40p
LM741DIL	18p
LM741MET	45p
LM747	55p
LM1889	300p
LM1894N	200p
LM3900	40p
LM3909	100p
LM3914	160p
LM3915	160p
LM3916	270p
L200	200p
M491BB1	500p
M494B1	700p
M50115P	320p
M50117P	500p
M50119P	525p
M50784	300p
M50786	500p
M50790	600p
M51161	300p
M51381P	200p
M51387P	800p
M51544	150p
M51848	150p
M54523P	200p
M54563P	200p
M58484	500p
M51516	260p
M51518	200p
MB3712	140p
MB3713	130p
MB3714	270p
MB3715	250p
MB3722	280p
MB3730	160p
MB3731	220p
MB3756	160p
MB3759	200p
MB8719	360p
MC1455	45p
MC1496	65p
MC3401	45p
NE555	20p
NE556	40p
NE558	80p
NE565	110p

CIRCLE NO. 115 ON REPLY CARD

Whose heterodyne?

Watt's steam engine? Baird's television? Bardeen, Brattain and Shockley's transistor? The truth is that all advances in technology really depend on multiple contributions and previous research. With this in mind Tom O'Dell surveys the archives to find where the real credit belongs for the birth of the heterodyne.

Heterodyning, mixing a weak incoming signal with a local oscillator to produce an intermediate frequency, is taken for granted today. The front ends of nearly all radio, television and radar receivers exploit it and it as also been applied to optical comms receivers.

According to the Oxford English Dictionary, John Erskine-Murray coined the word heterodyne to describe "one of the most interesting of Professor Fessenden's many inventions". He was referring to an electrodynamic telephone receiver that R A Fessenden (1866-1932) patented in 1913.

Certainly, the original patent specification[1] does describe the idea of producing a beat-frequency. But despite what Fessenden claimed (see **Fessenden's heterodyne**), it is unlikely that his device ever worked at radio frequencies.

Instead, the first successful heterodyne system to be used in wireless telegraphy appears to have been developed in Germany by Rudolf Goldschmidt.

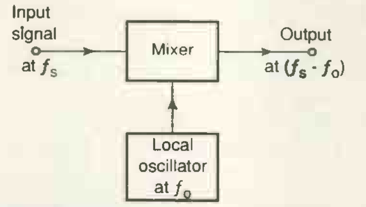
Sound-wheel rectification

Rudolf Goldschmidt graduated as an engineer in 1900 and joined AEG's engineering laboratories, first in Berlin and later in Prague. In 1907 he left industry to become a lecturer at the Technical University in Darmstadt, and it was from there, in 1909, that his most important invention was patented as the Goldschmidt high-frequency alternator.

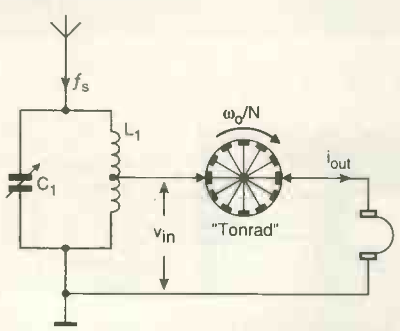
Goldschmidt alternators were used as transmitters for the German wireless telegraphy service between Germany and the USA, set up in late 1913. The machine at Tuckerton, New Jersey, had an output of 120kW at 50kHz, while the Eilvese transmitter, near Hannover, had a power of 150kW. During the first months of the Great War, after the British had cut German transatlantic submarine cables, this link became vital: the US did not join the War until April 1917.

In Goldschmidt's system, the cw signal from an aerial is connected to an *LC* input circuit tuned to signal frequency, f_s. A tap on the tuning inductor takes the signal on to the tonrad (sound-wheel), which is a commutator of *N* segments driven at an angular velocity $2f_0/N$. f_0 corresponds to the local oscillator frequency. Goldschmidt's tonrad had 800 segments[2] and was driven at 3750rpm.

The tonrad connects headphones to the radio frequency source for only half the time, at frequency f_0. If f_0 is made identical to f_s, then the incoming signal undergoes synchronous rectification. Under these conditions, no audible



The heterodyne technique familiar today.



Goldschmidt's tonrad heterodyne receiver of 1913 connects the headphones to the rf source during only half the time. It had 800 segments but only 12 are shown for simplicity.

sound would be produced in the high-impedance headphones because the mean current flowing would be a constant.

If the tonrad is now slowed down slightly, the current flowing in the headphones will have a mean value that varies at the beat-note frequency, $f_s - f_0$. In the original system this frequency was 1kHz, at which the human ear is most sensitive. The high-impedance resonant-reed type headphones could then be tuned for maximum sensitivity at the beat-note frequency.

Synchronous detection and heterodyning is very familiar to us today, with bilateral switches operating at up to several MHz and available as cheap integrated circuits in cmos technology. But in 1913 all this kind of signal switching had to be done mechanically.

However, mechanical rectifiers were well established in power engineering at that time, providing low voltage dc supplies for battery charging and electroplating. No other rectifier had such a low voltage drop across it during conduction.

Goldschmidt was a power engineer and, doubtless, would have considered his tonrad to be an extension, to radio frequencies, of the well-established technique of mechanical rectification. Although it was a first-class piece of mechanical design, the tonrad did not introduce any really new ideas, and no application was made for a patent.

His high-frequency alternator, on the other hands, was a radically new step forward and was patented in several countries.

Voyage of discovery

About one year after the Goldschmidt system began working traffic between the USA and Germany, a completely different kind of 'heterodyne' circuit began to appear in the literature. In it, the input circuit is tuned to signal frequency, f_s, and a local oscillator, at frequency f_0, is coupled into the diode circuit and is of sufficient strength to turn the diode on hard during alternate half cycles. Effectively, this is Rudolf Goldschmidt's tonrad realised electronically. A potentially fast diode switch has replaced the slow mechanical one,

Fessenden's heterodyne

In his patents[1] Fessenden gave no design detail on the electrodynamic headphones needed for his heterodyne system. But in the text he describes them – and represents them in his circuits – as two simple coils.

One coil was to be mounted on the headphone diaphragm cone, and the other to be fixed close to the first. Fessenden's reasoning was that the force between the two coils would be proportional to the product of the two currents, i_1 and i_2, flowing in the two coils. If i_1 were made the rf signal current from the aerial, and i_2 was obtained from the local oscillator, a multiplicative mixer would result as far as the force on the cone was concerned, and a beat note at the difference frequency would be produced as audio output.

By making the cone couple into a Helmholtz resonator and tuning the resonator to the beat note, Fessenden imagined excellent results would be obtained with a powerful local oscillator. He, and Hogan[3], wrote of an "amplification" of the small signal i_1 by the local oscillator current i_2 because audio output would depend on the product $i_1 i_2$.

But the argument collapses when practical design ideas are introduced into the problem.

The best geometry would be two really thin, flat, pancake-style coils, separated by as small a distance as possible. The force between two such coils is:

$$F_1 = \pi\mu_0 i_1 i_2 N_1 N_2 (r/a)$$

where N_1 and N_2 are the number of turns on each coil, a is the width of the coils and r is their mean radius.

In a conventional moving coil headphone that uses a permanent magnet, the moving coil again has N_1 turns and carries current i_1. The magnetic field around the moving coil, B_0, is now constant instead of varying at the local oscillator frequency and is provided by the permanent magnet. Signal current in this second case must be the rectified rf signal current and carry only the low frequency modulation on this rf. Force on the cone is now

$$F_2 = 2\pi r i_1 N_1 B_0$$

Comparing F_1 and F_2 shows that the 'amplification" claimed by Fessenden and Hogan could just as easily be claimed for the permanent magnet design. The constant magnetic field, H_o, is producing the same multiplying effect as i_2.

The designs can be compared by making the above equations for F_1 and F_2 equal to one another and solving for i_2. $i_1 N_1$ may be cancelled out because it is going to be about the same in both designs, even though impedances, current levels and frequencies will be quite different.

In the two-coil design, i_1 is the microamp level aerial current, and N_1 must be small because of the high frequency. In the permanent magnet set-up, i_1 is the nanoamp level crystal detector current, but N_1 may be several thousand turns.

The result is:

$$i_2 = 2B_0 a/\mu_0 N_2$$

and this shows how large local oscillator current i_2 must be to make Fessenden's design as useful as the conventional moving coil headphone.

B_0 could be made as high as 1Tesla, and a 10mm, accommodating 100 turns of 100μm wire. Such a winding would not have too high an inductance, but the above equation shows that i_2 would have to be over 150A if the same order of audio output were to be obtained from both Fessenden and conventional designs.

Hardly surprising then that Fessenden's heterodyne never saw wide application.

To have 150A of rf current so close to one's ear – even if the associated cooling problems could be solved – is not a good idea.

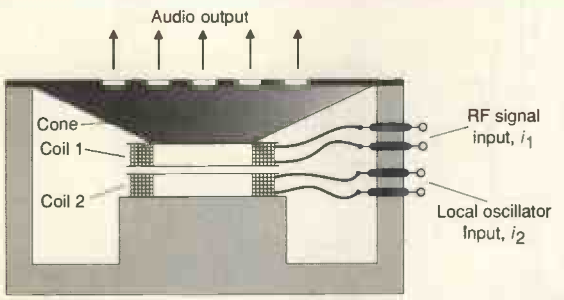


Fig. A. Sectional view of Fessenden's heterodyne headphone.

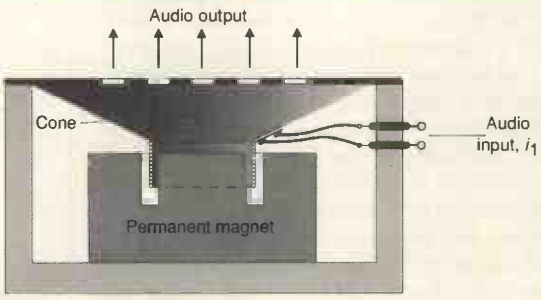


Fig. B. Conventional moving coil headphone.

opening up the high-frequency and microwave applications that are now so familiar.

In 1913, JL Hogan published a paper[3] mainly concerned with Fessenden's electrodynamic telephone idea but briefly mentioning the diode mixer in connection with some tests that had been made on behalf of his employers, The National Electric Signalling Co of Pittsburgh, Pennsylvania, during a voyage across the Atlantic.

That was the voyage of the *USS Salem*[4], which sailed from Philadelphia on February 15, 1913, for Gibraltar. On board were National Electric Signalling personnel and Navy radio specialists. Their purpose was to test new receiving equipment, working with two transmitters – a Federal Telegraph 35kW arc and an NES 100kW rotary spark – installed on the military reservation at Arlington, Virginia. The three receivers on board the *Salem* were a Fessenden heterodyne from NES, a Wireless Speciality Apparatus Co crystal receiver and a 'tikker' receiver from Federal.

Tests were to be made as the *Salem* crossed the Atlantic and, when she arrived at the British base in Gibraltar, arrangements had been made with the Royal Navy for experiments to be continued using the very large aerials that were available there on shore.

As part of the deal with the British, the US Navy had agreed to allow a Royal Naval Officer to board the *Salem* once she arrived at Gibraltar and work alongside the Americans during March 8-11, 1913. The officer concerned was Captain Willis, RN from HMS Vernon, Portsmouth, which at that time was a major Royal Navy r&d establishment.

Captain Willis' report[4] concentrated on what he called "the heterodyne". But from his account of the apparatus, this clearly was not the Fessenden heterodyne. Instead, he describes a local oscillator loosely coupled to "the usual crystallite receiving circuit".

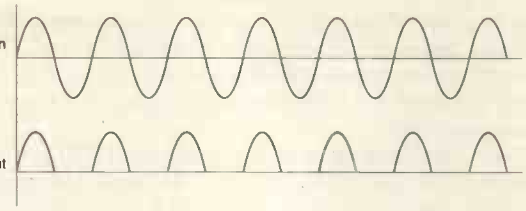
The local oscillator – a pretty fearsome arrangement 'placed as far as possible' from the receiving circuits – was almost certainly the one belonging to Fessenden's heterodyne system. Captain Willis describes it as an "apparatus... to produce undamped continuous oscillations by means of the electric arc in a hydrocarbon atmosphere". DC power input to this arc was over 100W.

But what appears to have happened on the *Salem* as she crossed the Atlantic was that the operators found experimentally that the Wireless Speciality Apparatus crystal receiver worked far better when Fessenden's heterodyne system was working at the same time. What Willis was actually seeing was the birth of the local oscillator/crystal receiver combination that later began to find its way into the literature.

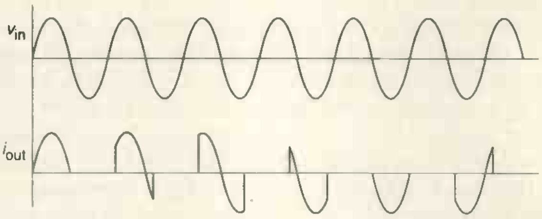
The suggestion that the experts on the *Salem* were engaging in much more general research than merely testing for the best commercial receiver is supported by the fact that Captain Willis tells us that the local oscillator and crystal receiver combination was being tested with signals coming from the Federal arc transmitter at Arlington.

This is very strange. It meant a combination of all three private radio companies' equipment was being used in the trials, when the goal was to select only one system for use by the US Navy. The Federal arc transmitter was supposed to be heard with the Federal tikker receiver, while the NES spark transmitter surely belonged with the NES version of the Fessenden heterodyne.

Another clue that things were still very much in a state of development on board the *Salem* is Captain Willis' surprise at the way the local oscillator /crystal receiver combination had to be operated. There was "no doubt", he wrote, that the local oscillator "should have been completely enclosed in an earthed metal case". But when he witnessed the experimental work, the local oscillator was placed as far as possible from the receiving circuits and the operator who wore the headphones was unable to make any adjustments to the local oscillator, relying on a "signal to an additional man to do this for him". The arrangement is surely not the kind of set-up expected during a demonstration to a potential customer.



Synchronous rectification with the tonrad running at angular velocity w = $2f_s/N$



Headphone current produced when the tonrad is slowed down slightly to run at angular velocity w < $2f_s/N$. Current in the headphones now has a mean value that varies at the beat-note frequency at which the ear is most sensitive.

Later, British copies of the American local oscillator design gave very good results used with crystal detectors in a heterodyne system. But the excessive local oscillator power was a problem. The solution, according to the 1914 report from HMS Vernon, was to mount the local oscillator outside the radio cabin "which is lead lined in modern ships".

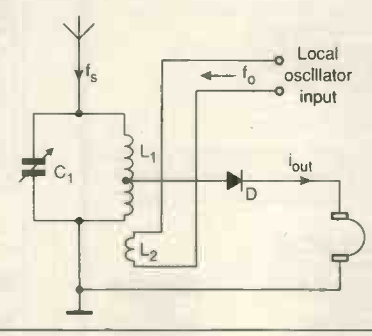
By 1915 the availability of valves resolved such difficulties and low power local oscillators could be made more easily. In any case, by then the working of the heterodyne system was beginning to be better understood.

How does it work?

Fessenden's two patents[1] on his heterodyne, assigned to the NES Co, were made final and published on January 14, 1913, just a few weeks before the *USS Salem* sailed. The patents made no mention of crystal diodes being used as heterodyne mixer devices. The conclusion must be that NES did not intend to use the crystal diode as a mixer because this would have exposed this new and unpatented idea to potential customers, to rivals and to the Royal Navy.

There is evidence that Hogan made experiments with crystal diode mixers just before the *Salem* sailed because he filed a patent application with his colleague, J W Lee, on November 16, 1912. In this he described the idea of two signals together – incoming signal and local oscillator signal – and their sum with a crystal or electrolytic detector. Lee and Hogan did not hurry to make this patent final, however, and it was not published[6] until June, 1915.

Even as late as that, Lee and Hogan did not understand how the heterodyne worked. They wrote: "It is well known in acoustics that the amplitude of the beats [resulting when



Simple heterodyne receiver using a diode mixer.

two notes are added together] may be considerably greater than the amplitude of one of the component waves... We have frequently obtained amplifications of as much as ten times in current and correspondingly as much as a hundred times in energy".

Their conclusion is, of course, nonsense. Only a multiplicative mixer can produce effects of this kind. Adding and rectifying can, at best, produce a beat of the same amplitude as the weakest input signal.

But the results of these early experiments should not be discounted today as "nothing but a bit of rf bias". The use of bias on crystal detectors to take the detector onto the most sensitive point of its characteristic was well established at the time, as Stanley's well-known *Text book on wireless telephony*[7] published in 1914 shows. *USS Salem's* crystal receiver would have had an adjustable dc bias for this very purpose.

The correct explanation for the remarkable improvement in sensitivity discovered by these early workers in radio when using the local oscillator/crystal detector combination has already been mentioned: electronic switching.

By using the detector diode at high forward current, instead of at the nanoamp levels normally found in a crystal receiver, Hogan and his US Navy colleagues removed the enormous power loss that was normally associated with simple crystal detectors. A much greater fraction of the received power could now be passed on to the headphones, translated by the heterodyne technique from rf to audio.

It took several years for this to be understood. Visualising the simple heterodyne as an example of electronic switching seems first to have occurred to LB Turner in 1921[8].

Other authors of the time attempted to explain the action of the diode mixer by using a square law model for its forward characteristic[9, 10, 11], an approach that continued for some time and can still be found in some student texts[12]. ■

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LETTERS

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

Dynamic cut...

As a recording engineer and designer I follow the subjectivist/objectivist audio battle with interest, with some sympathy for both camps.

However, in his article on slew rates (April *EW+WW*) Ben Duncan shoots himself in the foot by overstating his case; where on Earth does he get the idea that 20 years ago classical recordings were made with dynamic mics? As far back as 1959 I was using Neumann *M49* and *U47* condenser mics as well as AKG *C12*s and Telefunken *M251*s all of which are still used by many classical engineers today. The Ampex *300* valve tape machine had a record /replay response extending beyond 20 kHz (the test tape stopped at 20k) while 20 years ago Ortofon offered a disc cutting head and amplifier set (*G0741; DSS 731*) with a range of 10Hz-25kHz

In the next breath he states that on direct cut vinyl discs information can extend to 200kHz. I have to say that no cutterhead has been designed with such a range; the Ortofon above has the widest range of a production head. In any case, the polishing facets of the cutting stylus would erase any signal with the curvature of a 200kHz wavelength at any reproducible level.

S W Davis
Wembley

...heading for change...

Ironically, Ben Duncan's article ('Simulated attack on slew rates', *EW+WW* April 1995, p. 303) shows exactly why the slew rates of audio amplifiers are of limited relevance. According to Mr Duncan, a very unpleasant distortion can begin when an amplifier's slew limit is approached by a factor of two, or even ten. An amplifier with a 12V/µs slew rate which does not generate substantial sub-slewing or soft TIM up to half its slew rate must be superior to an amplifier with 50V/µs slew rate which already generates substantial TIM at a tenth of its slew rate.

Hence, applying a test signal whose rate of change is equal to the worst rate of change expected and measuring the distortion generated under these conditions, gives far more useful information than a slew rate measurement. The simplest usable test is a thd test with a full-amplitude sine wave of sufficiently high frequency†. Douglas Self's amplifiers perform quite well under these tests, as the graphs in his 'Distortion in power amplifiers' series show.

Marcel van de Gevel
Haarlem,
The Netherlands

†. Robert R.Cordell, "A fully in-band multitone test for transient intermodulation distortion", Journal of the Audio Engineering Society, vol 29, no 9, September 1981, pp578-586.

... slewing from reality

Doug Self comments on Ben Duncan's article, Simulated attack on slew rates (April 1995).

I read Duncan's essay at self-rebuttal with mounting alarm as it veered further and further from reality. I suppose I should be flattered that my work has received so much attention, but I'm not sure that disseminating Duncan's material, which is neither accurate nor constructive, has done audio much of a service.

Duncan's first contention seems to be that hf levels in music are higher than conventional wisdom suggests. Facts are facts. It is wrong to suggest that hf levels in music, live or otherwise, are anywhere near those at the bass end; if they were, it would simply be intolerable to listen to. As for using an Iron Maiden gig as an audio reference – words fail me.

It is fallacious to say that the rise of digital keyboards has brought about a significant increase in the hf content of musical material. Digital keyboards, being digital, have reconstruction filters after their d-to-a converters, and usually internal sampling frequencies lower than 44kHz cd standard.

As a believer in reason and experiment rather than blind dogma, I hooked up my Roland *U20* keyboard to a spectrum analyser to check the ultrasonic output. Apart from a –80dBm spur at 27kHz, presumably the sampling frequency, the output was commendably clean, with nothing above the –90dBm rf noise floor. No manipulation of the controls or programming could produce any output above 27kHz, which did not come as a total surprise. A moment's thought shows that in fact you are much more likely to get ultrasonics from real instruments which have no inherent bandwidth limitations; but with the possible exception of dog whistles, ultrasonic output is likely to be low.

No input rf filtering was used in my simulations or real measurements, because the aim is to measure the slew-rate of the amplifier alone. The upper bandwidth limit of an audio system must be defined somewhere and I must admit I thought this was too obvious to need further repetition. As I have explained several times before, it cannot be done properly by just slapping an *RC* network on the amplifier input unless you know that it will be driven from a defined source impedance.

I am afraid Duncan's slew-rate simulations of my circuitry are completely worthless because:

1) The VAS beta-enhancer emitter resistor has been omitted from Fig. 4. This component is essential to pull charge carriers out of the base of the VAS transistor and therefore has a major effect on slewing behaviour.

2) The test signal is already slewrate limited before it reaches the amplifier. Mine is not.

3) The top output emitter resistor is the wrong value, though this mistake may not affect slewing much.

4) For reasons we can only guess at, all the transistors have been changed for different types. Since slew behaviour depends on the magnitude of currents, transistor beta may affect it significantly. For comparative purposes, this change alone renders the results meaningless

After this, to be told that the simulated circuit 'precisely follows' the one that I published can only be described as hilarious. In view of all the discrepancies there seems no point in quibbling about the numerical results, but the long settling tails on Duncan's outputs are definitely erroneous, and do not exist either in competent simulation or real life. I suppose the probable cause is overloading of some internal circuit node.

I think Duncan may be under the impression that I am advocating the generic/Lin configuration as the best possible for all parameters under all circumstances. This is not and has never been the case. However, the generic circuit is unquestionably the basis for 95% or more of the amplifiers that have ever been built, and so is the obvious place to start enquiries into amplifier design.

Unexpectedly, my investigations into the linearity of this architecture revealed that it was capable of much lower distortion than is normally believed possible, such as 0.0015% at 1kHz, while still using safe and modest amounts of negative feedback. There is still no deep-laid secret to this, (unlike Duncan's preferred circuitry) but it does require a clear appreciation of the various distortion mechanisms, and a knowledge of the cheap and simple ways in which they can be minimised. It is perfectly possible, and even likely, as my writings have already said, that faster slewing can result from different amplifier configurations.

Having disposed of this not-so-bracing shower of destructive but fallacious criticism, I thought we would at least find out the superior but secret circuit methods that Duncan has trailed before us for so long. I was astonished to see that the relevant circuitry was concealed in an op-amp sub-circuit in his simulation, and only described as a "discrete op-amp".

It strikes me that to continue to protest that you know a better way, and then after everything refuse to reveal it, can only invite ridicule.

The diatribe on differential amplifiers is also deeply depressing. Duncan does not choose to disclose the details of the circuit he is simulating, but the quantity of high-order harmonics seems to indicate that there is an output stage present generating crossover distortion. As far as I can follow it, his contention seems to be that a carefully-tuned amount of input-pair imbalance allows some harmonics of the crossover distortion to be manipulated in amplitude by partial cancellation.

This sort of tuning appears to introduce another trim control, which will be deeply unwelcome on production lines, and also assumes that the quiescent current is exactly set and exactly maintained to ensure that the output stage generates precisely the right amount of each harmonic. This does not appear to me to be the way forward.

A more serious objection is that an unbalanced pair generates substantial second-harmonic distortion, going up at 12dB/octave with frequency. Whatever may be happening at 1kHz, I suggest that thd at 20 kHz will increase to diabolical levels. Another problem with an unbalanced input pair is much increased dc offset.

There is no evidence that the relative levels of harmonics have a complicated and subtle effect on the perception of distortion, though there has occasionally been speculation that the rate of fall of harmonics or the balance between odd and even has some kind of special significance.

These delicate cancellations completely miss the point; the whole aim of reducing distortion to sub-0.001% levels is to ensure that it is below the level at which anyone rational could claim it to be audible. The seamy details of its character can then be blissfully ignored, instead of agonising over whether it is better or worse to have the 7th harmonic above the 9th on a wet Tuesday with the wind in the NE.

There is a lot more I could say, but I don't think it's necessary.

...and as for delays

Readers of last month's *EW+WW* will have noticed that Ben Duncan appears to have given up technical writing and now only produces wholly negative knocking copy. Why he should feel it is appropriate to wage some sort of personal vendetta is a mystery to me.

I wish to state forcibly that his last article (Delayed Audio Signals) is a piece of misrepresentation. Throughout it, reference is made to a circuit fragment which is attributed to me. However, in scaling the component values, he has seen fit to alter the *RC* time-constant I used in the reference he quotes.

My original values were 10k for the upper feedback arm, 500Ω for the lower and 220µF for the capacitor. This gives a –3dB point at 1.4Hz, which is appropriate for avoiding capacitor distortion at low frequencies. The altered values (6.66k, 333Ω, 150µF) give a roll-off at 3.2Hz – over an octave higher.

Since the whole article is based on the values of these components, this alteration renders most of it null and void in the same way as in his previous hatchet job on slew rates. Since the alterations are aimed at making Duncan's case good, I find it difficult to accept that they were made by mistake and have to consider the possibility that this is a deliberate piece of misrepresentation.

Douglas Self

Ben Duncan replies:

I am surprised that Douglas is upset. Although my work has been criticised and even described as 'putative' in his series of articles, I have praised in print what I have found to be good (a substantial proportion) and, in my most recent piece, tried hard in the limited space to set matters in perspective.

Even if I have accidentally misrepresented his RC values (easily done when scaling three circuits back and forth), the octave they are out by is a negligible misrepresentation in the scheme of bass delay. In Self's circuit in *EW+WW*, Sept.1994 (p.761), there are *two* HP filters, both –3dB at about 1.5Hz, which largely cancels the minute misrepresentation of which I am accused.

As my whole article is based on HP filtration in the total audio path and how it stacks up, I cannot see that the tiny differences being argued over affect my conclusions.

Self on Hawtin

In his latest letter, Mr Hawtin seems to be trying to establish that an amplifier with fets in can have low distortion. Of course this is true and was never in dispute. What I have said is that for a given amplifier architecture, fets would always distort more than the equivalent circuit using bipolars.

I really can't see how this can be disputed; the V_{gs}/I_d law of fets compels the crossover distortion to be much worse than for bipolars in any straightforward output stage. This does not mean that it is not possible to add complications that make the overall performance good; an example of this is Robert Cordell's design[1] which includes extensive error-correction circuitry in the output stage to linearise the fets. By the way, his output stage gain plots look much like mine, sharp corners and all.

In the selected data given, (which I do not accept is a representative statistical sample) I assume that 'hybrid' means bipolar drivers combined with fet output devices. The purpose of this is of course to make the power fets behave more like power bipolars. As I have previously written[2,3], the hybrid combination is a good deal better than fets alone, though nothing like as good as the purely bipolar equivalent. because the sharp gain changes that always seem to appear in fet outputs still persist.

One difficulty with the second-hand test reports that are referred to is that no test conditions are given; measurement bandwidth can make a major difference to the numerical results. In particular, the figure of 0.0002% needs a good deal of explanation, because this would be below the noise floor of even a quiet amplifier, and impossible to measure with any thd equipment I have ever come across.

I have no intention of commenting on the rest of the examples given; without details of the circuitry a raw thd figure teaches us nothing.

I find it unrewarding to reiterate the basics of electronic theory in EW+WW, *particularly to those who seem to have no interest in learning it. If Mr Hawtin has difficulty in disentangling bandwidth and slew rate then any elementary textbook would put him straight.*

It is true that as frequency is increased and an amplifier goes into slew-limiting, the output waveform will become triangular, and eventually its rms level will fall by 3dB; to call this "bandwidth" would be madness; apart from anything else it would be level-dependent. The word has a precisely defined meaning which is not going to change just because Mr Hawtin wishes to use it in an idiosyncratic way. A linear system may have a bandwidth limit, but it cannot exhibit a slew-rate limit because this is a non-linear effect. The distinction is fundamental, and surely not beyond the grasp of someone who feels qualified to lecture us all on amplifier design.

Mr Hawtin's appreciation of fet outputs is also in error. Bipolar transistor beta certainly varies with collector current, though as I explained at some length [in 3] (which I can only assume Mr Hawtin has never read), beta variations only affect linearity significantly for loads up to 4Ω. Whatever the load resistance, the stage remains an order of magnitude more linear than its fet equivalent. I have simulated and measured it. Has Mr Hawtin done either?

Mr Hawtin's thoughts on bipolar output stages might be more valuable if he appreciated that they do not have a gain of 100×; unity gain is almost universal, for very good reasons. Similarly, the V_{gs}/I_d *law is not very linear, and repetition will not make it so. This claim will be ruthlessly explored in a future article. You have been warned...*

Likewise, relentless repetition will not make all transistor amplifiers rolloff at 15kHz. They just don't. I have a production-line making bipolar power amps that are flat to 0.1dB at 20kHz. And, just for the record, the first hi-fi mag I opened today reviewed a bipolar amplifier that was –0.5 dB at 220kHz.

Can we stop this now, please? I really do have better things to do.

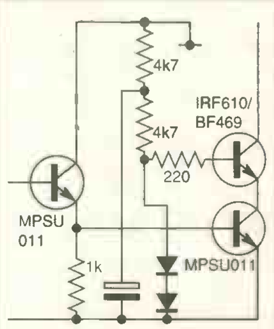
References

1. R Cordell 'MOSFET Power Amp with Error Correction', *JAES* Jan/Feb 1984.
2. D Self 'Sound MOSFET Design', *EW+WW* Sept 1990, p760.
3. D Self 'Distortion In Power Amplifiers' Part 5, *EW+WW* Nov 1993, p934.

Self questioning

First I wish to congratulate Mr Self on his design procedure. I followed the series and subsequent debate with high interest. I am not a technician, and I think that Mr Self was didactic enough for me to follow his basic ideas.

Some minor queries, however, did arise. Why has Mr Self not used a cfp input stage in his final design? Would using *MAT02/03*s here be an improvement?



Combined buffer/cascode amp.

In the voltage amplifier, why not combine the cascoding and buffering to get the best of both worlds? (Figs 4d,f) My suggestion is shown in the diagram.

Good linearity is claimed for 'ring-emitter' power transistors. Would there be any benefit from using, say, the *2SA1095* and *2SC2565* in his design?

Referring to 'Distortion off the rails', why not use a separate supply for the input and voltage amplifier stages, and/or stabilisation?

These questions may sound naive, but then I am a psychiatrist, not an electronics engineer. Thank you for your splendid work – it has helped my understanding.

Simon Rambert
Bern, Germany

...and to Erik Margan (Follow the leader, letters, EW+WW, April 1995)

Mr Margan says that Subjectivism has been around long enough, without much concrete progress. I think it would be truer to say that it has been around more than long enough, without making any progress at all; if anyone feels that they have made genuine headway in comprehending how the unmeasurable avoids being inaudible, then they are keeping awfully quiet about it.

Talking of progress, just how much should we expect when dealing with a 'subject' that claims to be so ineffably subtle as subjective audio, and which has been around about twenty years?

A comparison may be instructive. At the end of the last century, atoms *were still regarded as the indestructible billiard-balls of Dalton; then in 1896 Becquerel discovered that radioactivity existed. This was certainly a subtle phenomenon, undetectable by human senses, and it was also a radical one because it revolutionised classical physics.*

In the succeeding twenty years, physicists discovered alpha and beta emission, showed that atoms were composed of a flimsy electron shell with a massive nucleus, measured the energy levels within that nucleus, and went off to demonstrate the transmission of elements through nuclear decay – from scratch.

This is an impressive record; in the case of Subjectivism, however, two decades of hand-waving seem to have brought no progress at all, and this gives a strong indication that the effects 'studied' do not in most cases exist...

I also differ mildly on the diagnosis; Mr Margan feels that the phenomenon of subjectivism is a consequence of failure of communication between two groups of people. I would say that there are three groups of people here: engineers, subjectivists and musicians. This is an obvious simplification, with overlap between the categories, but perhaps it is nearer the truth.

While the attitudes of engineers and Subjectivists have been examined at interminable length in these pages, I find musicians (and I accept that is a broad category) have a distinct approach of their own. A musician is interested in the sound, by which he means the sound of a dominant seventh versus a flattened fifth, or pwm versus fm in the oscillator of a digital synthesiser. He doesn't mean the doubtful perceptibility of 0.01% of crossover distortion, nor indeed the undoubted inaudibility of rhodium permanganate speaker cables. He is talking of differences that are audible to – and measurable by – everyone, if they care to measure the thd of a fuzz-box or a skilfully overdriven valve guitar amp.

Obviously we can't do without the engineers, and we can't do without the musicians. However, the remaining group seems less immediately useful.

As for Mr Margan's examples of audio perception, I am afraid I must decline to believe the first one. The resistor values he gives produce a level mismatch of 0.025dB, and this is at least fifty times better than the accepted capabilities of human hearing. As for being able to detect this difference after a delay of several minutes, I just can't see it.

The other two scenarios show that human hearing works OK, but who said it didn't? The sound of a functionally-challenged automotive wheel bearing is not hard to identify, and a relatively simple dsp system could do it – acoustic signature analysis by computer is not a new idea, though no-one would pretend that the algorithms mimic human perception.

The Subjectivist/reality debate has had a long run but ultimately it is sterile because we are usually arguing about non-existent things.

Slewing the bandwidth

I congratulate Ivor Brown on getting to grips with the somewhat enigmatic networks usually found on the outputs of power amplifiers ('Between Amplifier and Speaker' Feb. 1995).

I was particularly glad to see him emphasise that the damped ringing that is almost universally seen during capacitive loading tests is due to the output inductor resonating with the load capacitance. It has nothing to do with amplifier stability as such. The ringing is usually around 40kHz or so, and this is much too slow to be laid at the door of a normally compensated amp.

If a power amp is deliberately provoked, by shorting the output inductor and applying a capacitive load, the worst values are usually around 100nF, rather than the 2μF intended to roughly simulate an electrostatic speaker. The oscillation is usually around 100-200kHz. If allowed to persist, this can be destructive of the output transistors. In this case there is no such thing as 'nicely damped ringing' because damped oscillation at 200kHz means you are one bare step away from disaster.

I felt less enthusiastic about one statement that Mr Brown made; "Most importantly, open-loop bandwidth of the amplifier must cover the whole audio frequency range." I can see no reason why this needs to be the case, and I suspect it is true of very few amplifiers indeed. In my view, the lesser condition that "There must be an adequate negative-feedback factor over at least the audio range" might be nearer the truth.

As I have written in the past, an open-loop bandwidth of 10Hz may appear to be intolerably sluggish, but once the negative-feedback is applied, the bandwidth is extended in the time-honoured manner.

As recent letters have shown, some confusion still exists between open-loop bandwidth and slew rate, although in truth these parameters have very little to do with each other. In the typical amplifier they are determined by different mechanisms occurring in different parts of the circuit, and can to a large degree be altered independently. It is easy to find op-amps where a 1Hz open-loop bandwidth coexists happily with a 15V/μs slew rate. One example is the *OP237*.

In view of the conceptual difficulties that always seem to arise in discussions on negative feedback, I wonder if Mr Brown would be prepared to expand on, and clarify, the statement I have quoted?

Douglas Self
London

Ivor replies:

Thank you to Douglas Self for his comments. I must plead guilty to having been careless in my choice of words. To say that the open-loop bandwidth of audio power amplifiers must *cover the whole audio frequency range is incorrect. I should have written that it* should *cover the whole range, but I do not think this simple change will satisfy.*

I offer three reasons for my opinion. Note that I have used the word opinion. As explained below, the relevance of many points in the design of audio systems to subjective assessment of musical performance is not fully understood.

Compare the analytical performance of laboratory instruments with that of the human ear and brain. We respond to two pressure waveforms that may contain components from many sources.

Analysis reveals the components that come from the individual sources. With monophonic reproduction this is reduced to one waveform, yet we are able to detect the individual instruments in an orchestra. Laboratory equipment does not begin to approach such a level of performance. We do not know which features of an amplifier may be critical to prevent our analytical ability being impaired by the existence of electronic devices in the signal path.

Negative feedback is used in amplifiers to obtain a more linear system, which is generally accepted as a desirable feature. It seems sensible to maximise this advantage at all audio frequencies.

The most common method of assessing amplifier performance is to use steady-state sinusoidal inputs. Music is not a steady-state signal, and while such assessments are useful they have limited value. Consider two feedback systems with closed-loop gains of 100; one includes an amplifier with open-loop gain of 1000 and the other amplifier's gain is 10,000. The table shows the signal levels at various places in the circuits for 10V output.

Open-loop gain	1000	10,000
Output signal	10V	10V
System input signal	100mV	100mV
Feedback signal	90mV	99mV
Net amplifier input	10mV	1mV

Feedback voltage is subtracted from the system input to give the net input to the amplifier. Consider an extreme case where the input is a fast transient and limited open-loop bandwidth prevents the feedback signal from immediately following the system input. For a short time the lower gain amplifier has a net input some ten times the normal; the higher gain one ten times that.

This momentary large input must cause an increase in the distortion products produced in the amplifier, with more produced in the higher gain amplifier. A large short-term signal will not appear if the open loop bandwidth of the amplifier is of the same order as the bandwidth of the input signal to the system. This requirement becomes more critical as the ratio of closed-loop to open-loop gain is reduced. The extreme case described is hardly likely to be found in practice, but it serves to illustrate the principle.

Finally, the fet amplifier described in the April 1990 issue started as an exercise to design a power amplifier with what seemed to be all the right features, including wide open-loop bandwidth. Communications from readers and others who have built the design have, without exception, been favourable, In particular, comments have been made about clarity, definition and separation of sources within the stereo source.

I agree with Douglas that slew-rate should not be a problem in a reasonably-designed amplifier. Slew-rate and bandwidth are not directly related, although low figures of each tend to go together.

The fascination of audio circuits is that laboratory measurements are not the final arbiter of performance. One day this may not be the situation and it may be established what is, and what is not, important. Until then I consider that wide open-loop bandwidth should be regarded as desirable in audio amplifiers.

Ivor Brown
Uxbridge, Middlesex

Increasing momentum

Re R Lerwill's letter (*EW+WW* Apr 95) on the uncompensated increase in momentum of cathode rays, the problem does not arise if, instead of mass being regarded as a scalar quantity, it is regarded as a vector quantity. This is counter-intuitive, but not absurd when one considers the ways in which mass may be measured. If it is measured as a body's resistance to force, then the force is a vector and so must be the resistance. If it is measured as the source of gravitational attraction, then this, although exerted in all directions when measured, must be associated with one in particular.

Lerwill's accelerated electrons

may offer increased resistance to an accelerating or retarding force in the direction of their high speed, but perhaps they offer only their standard resistance in a direction at right angles. It should be possible to test this in a particle accelerator.

K A. Stagg,
Waterlooville, Hants

Quad speed reduced

The article by Guruprasanna and Lanka Kumar in *EW+WW*, March 1995, Quad speed RS232, contains several interesting ideas but the suggested implementation is flawed because the issue of discontinuous phase-shift at the symbol boundaries has been completely ignored (at least within the published text).

While the system described does indeed reduce the baud rate required, it fails to achieve the underlying objective, which is a reduction in the bandwidth needed to carry the signal.

Studying figure 4 shows that if any of the phase signals 0 to 7 (which all end at a high level after 8 cycles of the main clock) is followed by a symbol encoded by phase 9 (which is low after main clock period 8 but goes high at the end of clock period 9), a very narrow negative pulse is transmitted. This pulse is only half as wide as one bit-period of the original data stream, so is more likely to suffer corruption due to noise than the original data would have been, had it been transmitted at the 'raw' bit-rate.

In other words, the bandwidth required on the RS-232 link has been doubled, rather than reduced by a factor of four as desired. This is the opposite of what the authors of the article intended to achieve, illustrating that when extending concepts from modulation theory down to baseband, one must be careful not to get caught out.

That the frequency content of the datastream can be reduced at the expense of allowing data transitions to occur during a larger number of more tightly defined time windows is an interesting idea. Practically, it would require a more subtle coding scheme than that shown.

Duncan Learmonth
Chelmsford, Essex

Making the point

Many of your readers will know that the transistor was discovered by Bardeen and Brattain of Bell Laboratories in late 1947. I use the word 'discovered' rather than 'invented' because the device which they accidentally created, the point-contact transistor, was nothing like what they were looking for! Working junction transistors, which is what Schockley, the Bell Labs theorist, was really seeking, were made some four years later.

Bell's point-contact technology was licensed to many commercial firms, and millions of point-contact transistors were made although operation of the device was very poorly understood in theoretical terms, and its production employed a highly empirical technique: 'forming'.

It involved fusing the point contacts to the germanium die using current pulses. The resulting structure, usually of pnp polarity, had a common-base current gain ('alpha') considerably more than 1!

While the majority of point-contact transistors were made in the USA, a number were made in England by the General Electric Company, Mullard, and Standard Telephones and Cables. Many of them were experimental types which were only offered to government laboratories such as Harwell, but a number ultimately became commercially available. A few devices were also made in France and Germany.

In the USA some types were sold by Bell Laboratories and its commercial arm, the Western Electric Company. Many small companies also sprang into the new market, some destined to become giants, such as Texas Instruments.

Today, the very existence of the point-contact transistor is known only to historians, and it is lucky to rate even a sentence in general electronics textbooks. Few specimens survive, and the demise of the electronic 'junk shop' makes them even harder to find.

As well as the point contact transistor, I am interested in the development of the semiconductor industry, in the UK and Europe in particular, from 1947 up to about 1960. If any *EW+WW* readers have interesting information or anecdotes from that period, data books or sheets, circuit cards or early semiconductor devices (particularly point-contact transistors), I would like to hear from them. All letters will be acknowledged.

Dr. Andrew Wylie
Purley, Surrey

Safe discharge

In my circuit idea, Safe NiCd battery pack discharger, in the April issue, an error has crept into the text at some point. The final sentence should read ... 'To take any number of cells up to a maximum of 12, the zener voltage should be 2/3 the final terminal voltage and discharge cumnt adjusted by *R2* to O.5A'.

Bill Hume
Newmilns, Ayrshire

Tesla driven

Re the article I recently wrote on Tesla Coils, please note that the voltage equation I included is not valid for pulse-driven coils. The correct equation observes conservation of energy and is :

$$V_0 <= V_{cp} \times \sqrt{\frac{C_p}{C_s}}$$

$$\left(\text{or } \sqrt{\frac{L_s}{L_p}} \text{ or } \sqrt{\frac{Z_s}{Z_p}}\right)$$

For this ideal to be reached, secondary loading must be minimal. It is also obvious that a lot of power is needed to reach voltages much higher than half a million or so, even if the Q of a coil reached 300 (most coils would get between 150 and 250). An analysis of the system[1] shows that with a pulse repetition frequency of 100Hz (number of mains half-cycles/s), the energy has dissipated by the next capacitor discharge.The spark gap sets primary capacitor voltage and also capacitor energy storage. Maximum voltage must therefore depend on the amount of energy available to charge the secondary capacitance from each primary capacitor charge. In practice, spark formation would prevent this ideal being reached.

The coupling constant recommended ensured that voltage peaks (caused by impedances reflected from secondary to primary and vice versa) would not occur too far down the coil, overstressing secondary insulation. The height at which a peak will occur may be crudely described as:

$$h_{(pk)} = h_{sec} \times (1-k)$$

where k is the coupling constant.

Running a coil with a well regulated transformer (e.g. a microwave transformer) will necessitate current limiting effective at 50Hz (the rf chokes are quite ineffective at mains frequency). The example coil used a high leakage inductance C-core demonstration transformer made by German firm, Leybolds.

Limiting means that the simple sparkgap arrangement shown in the article shorts the transformer when it fires. A better arrangement would alternate between charging the primary capacitor and discharging into the coil, eliminating the need for any current limiting. ■

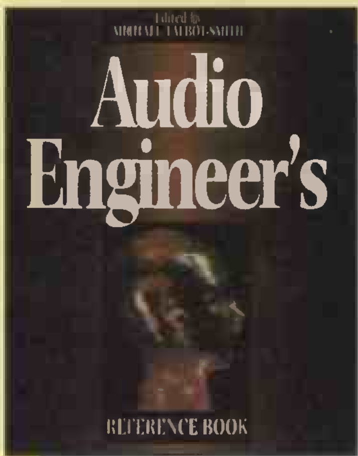
M.J. Watts,
Wellington, New Zealand.

Reference

"Q", K.L. Smith pp51-53, *EW&WW*, July 1986.

Reference books to buy

For Audio Engineers



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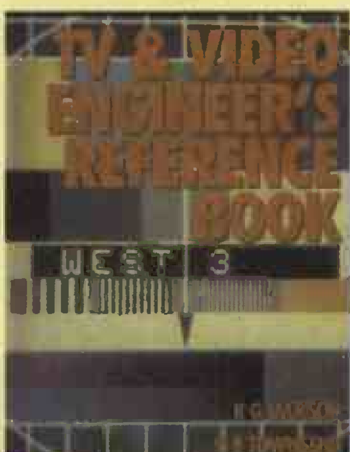

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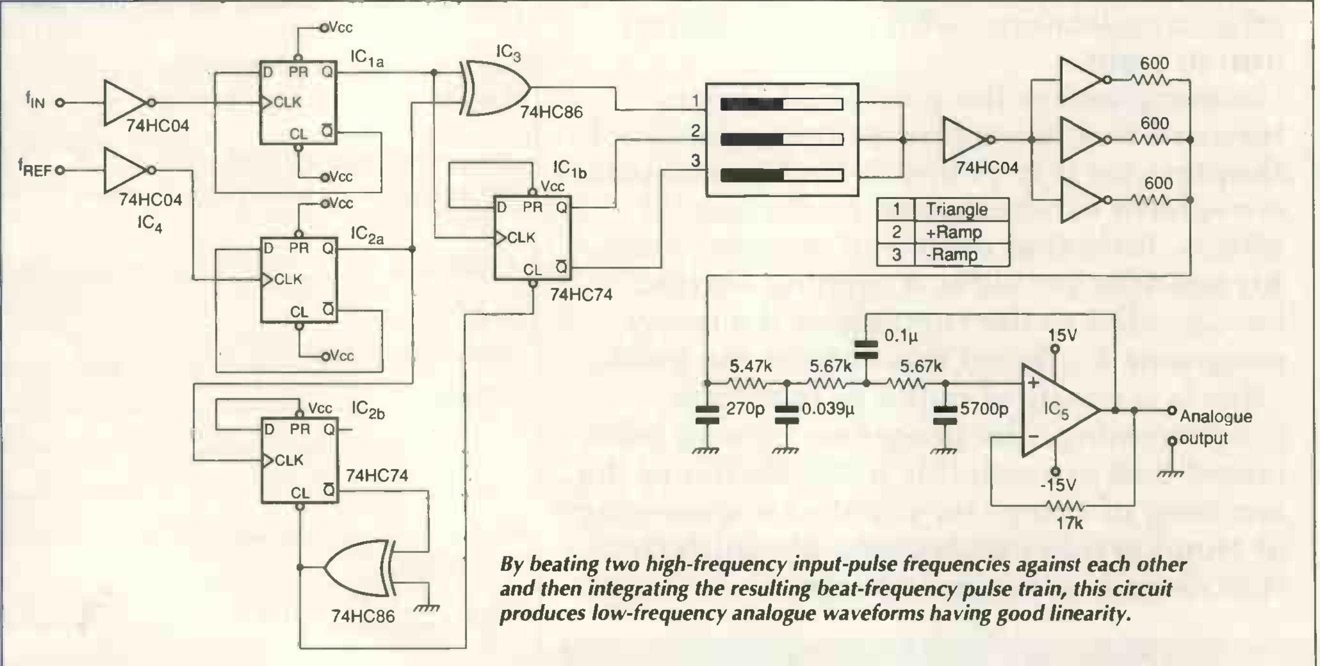
This circuit generates three different waveforms having frequencies less than lHz: triangle waves, positive ramps, and negative ramps. At very low output frequencies, stability of the circuit's input frequencies almost completely determines the output waveform's linearity.

Gate IC_3, exclusive-or logic, beats input frequency f_{in} against reference frequency f_{REF}, thus producing a train of pulses whose periods increase gradually until the frequency sources are completely out of phase. Then, the pulses' periods decrease until the sources are again in phase. Flip-flops IC_{1A} and IC_{1B} produce 50% duty cycle inputs for exclusive-or gate IC_3.

The op-amp and its surrounding components form a third-order, low-pass filter, whose f_c is 1kHz. This filter averages the output of pulse buffer IC_4 to produce a triangle waveform having a peak amplitude of V_{cc} and a frequency of $|f_{IN}-f_{REF}|\div 2$. Be sure to select low-dielectric-absorption capacitors for the filter circuit.

Ramps are generated by the circuit in a similar manner, except that the phase comparator of the set-reset flip-flop, formed by IC_{2A} and IC_{2B}, replaces the exclusive-or gate. The phase comparator sets on every other negative transition of f_{IN} and subsequently resets on every other negative transition of f_{REF}. If f_{IN}'s frequency is greater than that of f_{REF}, then the width of the Q output pulse of IC_{2B} will gradually increase. This increase produces a positive-going ramp at the circuit's output. If frequency of f_{IN} is less than f_{REF}, the output will be a negative-going ramp. Note that the filter's step response controls the ramp's reset time. Selecting a frequency greater than 100kHz for f_{IN} and f_{REF} attenuates the pulse's ripple. This relaxes the reset-time restrictions.

Michael A Wyatt
Honeywell SSO, Clearwater, FL



By beating two high-frequency input-pulse frequencies against each other and then integrating the resulting beat-frequency pulse train, this circuit produces low-frequency analogue waveforms having good linearity.

Current sink widens vco frequency range

Output frequency span of the familiar *HC4046* voltage-controlled oscillator, vco, is about one decade, and the device exhibits fairly good linearity over an input voltage range of 1 to 4.75V.

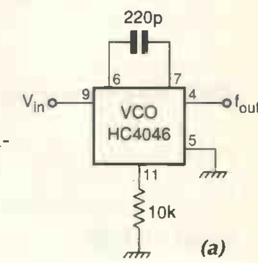
This circuit widens this frequency span to three decades. It replaces the single frequency-determining resistor from pin 11 to ground with a precision voltage-controlled current sink comprising an *LM358* op-amp and transistor Tr_1. The current sink overcomes the limitations of the integrated current sources normally responsible for charging and discharging the timing capacitor.

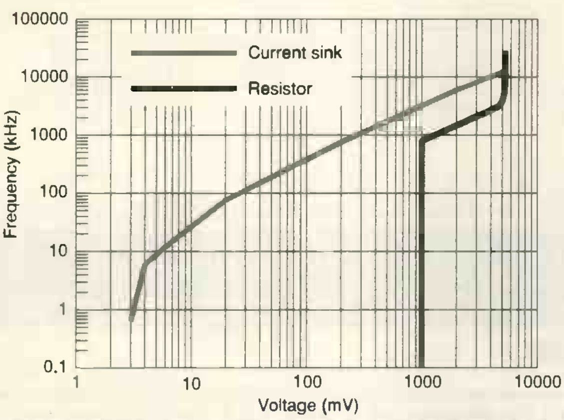
A fixed level of 2.5V is applied to the vco input at pin 9. Because the voltage on pin 11 cannot exceed 2V, the current sink must operate below this level. To meet this requirement, resistors R_1 and R_2 divide the input signal before it reaches the current sink's input.

The graph compares the frequency-versus-voltage characteristics of the standard circuit with those of the new circuit. By using the voltage-controlled current sink, the linear tuning range spans three decades.

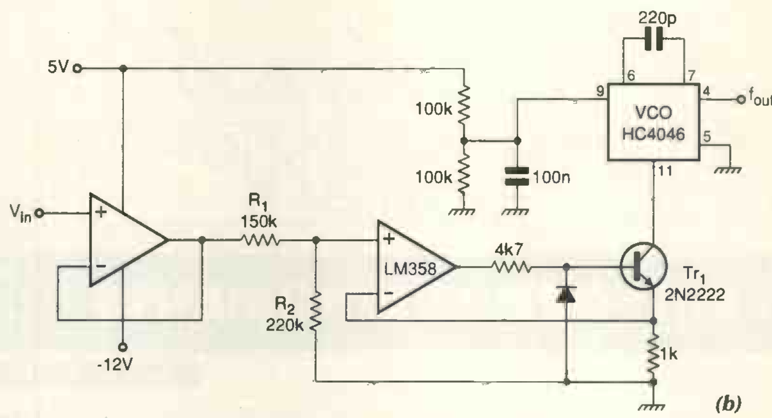
At the low-frequency end, output phase noise is quite noticeable because the current sink operates at very low current levels. When the loop is locked to a clean reference, the feedback reduces this noise. A better way to remove this output phase noise is to operate the vco near its maximum frequency and then divide the output digitally. This technique reduces the phase noise by the amount of the division.

Antonio Tagliavini
Applicazioni Digitali e Analogiche, Bologna, Italy





The current-sink circuit lengthens the frequency span of the standard circuit from about one decade to three decades.



By substituting a voltage-controlled current sink for the standard circuit's fixed 10kΩ resistor (a), the circuit in (b) extends the HC4046 vco's frequency range.

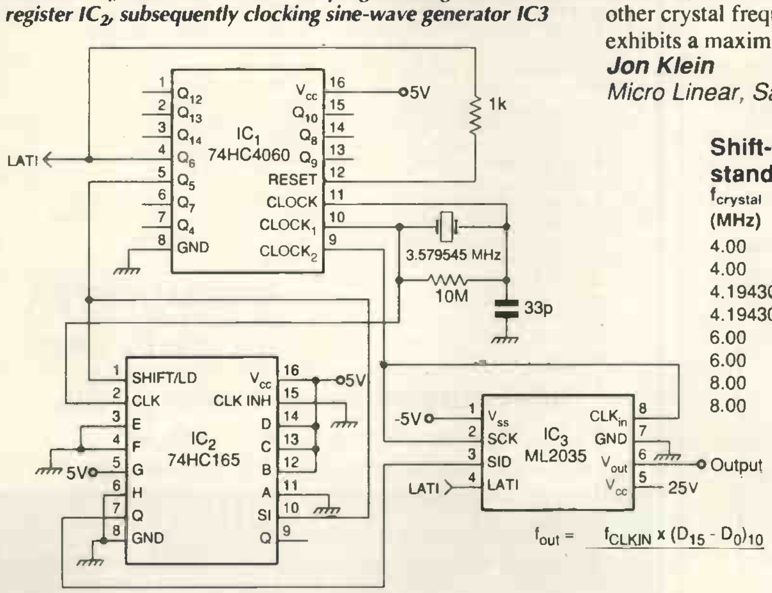
Programmable oscillator runs without a micro

Using a clever scheme adaptable to other programmable devices, this circuit allows you to operate the *ML2035* programmable sinewave generator, IC_3, without a controlling microprocessor.

A *74HC4060* counter, IC_1, provides the sinewave generator's clock and a gating pulse to shift register IC_2. When IC_1's Q_5 output, on pin 5, goes high, IC_2 begins shifting eight hard-wired bits into the sine-wave generator to program it. After IC_2 shifts the eight bits, Q_5 goes low, enabling normal operation. The circuit can produce both 50 and 60Hz outputs from an NTSC colour-burst crystal operating at 3.579545MHz. The table lists binary codes for other crystal frequencies. The sine-wave generator's output exhibits a maximum of 0.5% thd.

Jon Klein
Micro Linear, San Jose, CA

Counter IC_1, first clocks in an 8-bit programming code via shift register IC_2, subsequently clocking sine-wave generator IC3



Shift-register values and frequency errors for standard crystal values.

$f_{crystal}$ (MHz)	f_{out}	D_{10}	D_{16}	*HC165* code ABCD	EFGH	Error
4.00	50	105	69	1001	0110	0.14%
4.00	60	126	7E	1000	0001	0.14%
4.194304	50	100	64	1001	1011	0%
4.194304	60	120	78	1000	0111	0%
6.00	50	70	46	1011	1001	0.14%
6.00	60	84	54	1010	1011	0.14%
8.00	50	52	34	1100	1011	–0.82%
8.00	60	63	3F	1100	0000	0.14%

PLC on a chip

One chip – programmable in simple language via a pc serial link – interfaces both analogue and digital i/o subsystems.

There are two main drawbacks to designing a microcontroller into a control application. One is the need to learn a high or low-level language in order to program the device. The other is finding a means of getting the code from the platform it was written on into the controller.

A new controller from Timely Technology – *the ITC232-A* – is intended to overcome both these problems, in addtion to being highly integrated. According to the device's distributors, it can reduce the implementation of many stanmdrad programmable logic controller applications from days to hours.

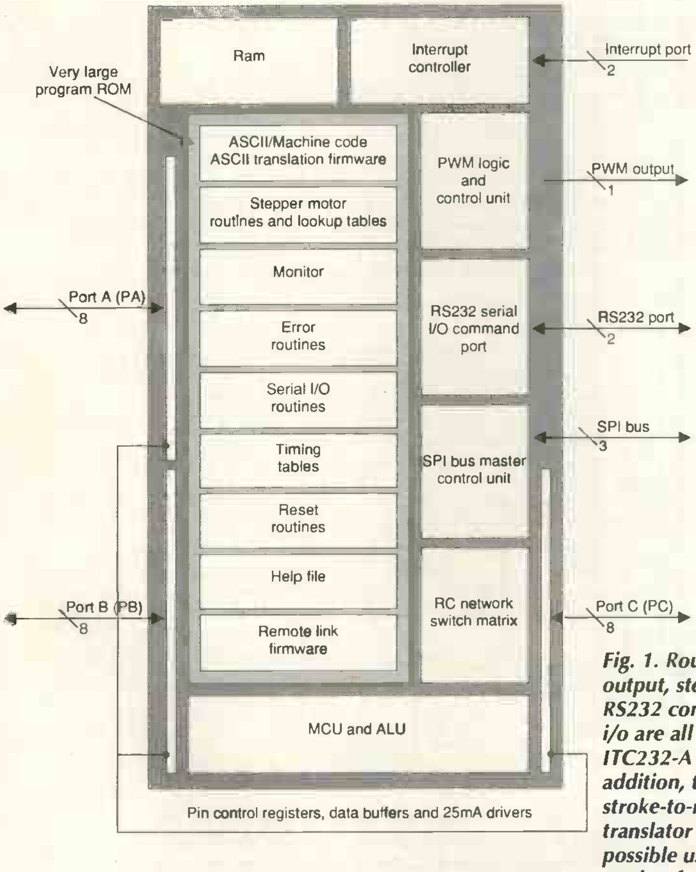


Fig. 1. Routines for pwm output, stepper motor driving, RS232 comms and SPI bus i/o are all built into the ITC232-A controller. In addition, the device has a key-stroke-to-machine-code translator so programming is possible using simple key storkes from a PC.

Programming is carried out via a simple serial link to a PC running a low-cost terminal-comms package. As you will see from the panel, the *ITC232* is programmed via simple, user-friendly, commands since the chip has its own key-stroke to machine-code translator.

Although the device can be programmed for wide variety of control tasks, a number of fully worked specific applications have already been developed. In addition to the three applications outlined in this article, there are notes describing how to analyse active filters, control remotely via modem and implement an optical fibre link. Further notes discuss error processing, pulse-data handling, multiple addressing, frequency counting and data conversion. **Figure 1** gives you an idea of the device's level of integration.

The device only has 40 pins, most of which are dedicated to i/o. But should the number of i/o lines be a restriction, additional control functions can easily be added via the device's SPI bus, for which there is already a wide varioety of compatile i/o chips.

Keystrokes – not compilers

Commands to the controller are typed at the command prompt of any terminal program, or stored in batch files to be sent to the board when needed. Complex command structures can be achieved without software via 'dial-up scripting utilities' found in most terminal and communications packages.

Many of you will find that you already have such a package, even if you have never used it. Most of them offer commands including decrement, getstring, if, jump, string monitor and search.

The scripting utility in *PFS:WindowWorks* offers all of these features, and more within its terminal program, for around £60. Others are more sophisticated, but all of them should save you a lot of time and money in getting complex applications up and running quickly without resorting to compilers and debuggers.

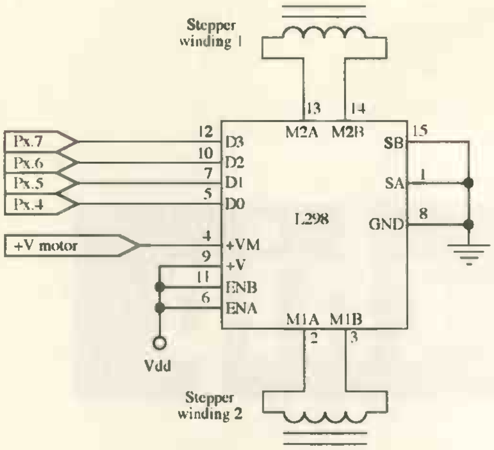


Fig. 2. Three of these stepper motor drive circuit can be connected to the controller simultaneously via ports A, B and C. Maximum permissible motor rail, i.e. +V motor, is 48V.

Software writers should not feel left out, however, as many comms utilities like Delrina Wincomm Pro (also around £60), offer 'C' within their Scripting Utility, which may be more to their liking. Your routines will simply batch up ascii keystrokes, then transmit, monitor and process as required.

Triple stepper motor control

It is possible to be rotating up to three stepper motors within an hour of connecting the board to COM1. This capability is depicted in slow motion by a software simulator supplied with each evaluation board. The simulator is supplemented by a tutorial on stepper motor basics, including monophase, biphase and half-stepping modes.

By simply typing *SAL100* at the prompt on your pc, for example, you will turn the stepper motor on port A 100 steps to the left. The chip takes care of all housekeeping, including the provision of automatic and programmable last-pulse braking.

A power driver and current controller like the *L298* and *L297* respectively, are all that you will need to complete your hardware design. Such a design, one port of which is outlined in **Fig. 2**, is covered in detail in an existing application note.

The digital i/o ports are arranged as three eight bit ports whose pins are individually programmable as inputs or outputs and each capable of sinking 25mA. This makes them just as suitable for controlling relays and leds as reading the status of switches, counters and encoders. These pins and ports can be written to as easily as read – with single keystrokes. Typing PWA254 on your keyboard for example, will write the decimal value 254 to Port A. Binary and hexadecimal values can be read or written just as easily.

The board can be programmed and left to run as a stand-alone reactive controller, configured to raise an alarm if conditions change beyond its ability to suppress them. However, some applications may demand that a host pc is alerted. This too has been accommodated in the chip design via two interrupts which send a single ascii identification back to the host via the three-wire RS232 command interface, at 300 to 115,200 baud.

Flexible, interrupt driven, pwm

The chip has a pulse-width modulation output on pin 35. Properties of this function are also detailed in the software simulator. Frequency limits are 10Hz and 10kHz and the duty-cycle range is 0 to 100% in 1% intervals.

The pwm signal is interrupt driven; that is the *ITC232* can do other things while the pwm is on, except that when the stepper motor is on; in this case the pwm remains high or pulled low while stepping takes place.

At the simulator prompt, typing W1000 followed by the enter key causes an audible 1kHz tone, produced by the simulator. In addition, you will see an 'oscilloscope' on the screen. Alt-S toggles the sound while Alt-O turns the scope on and off.

The default duty cycle is 50% and a mes-

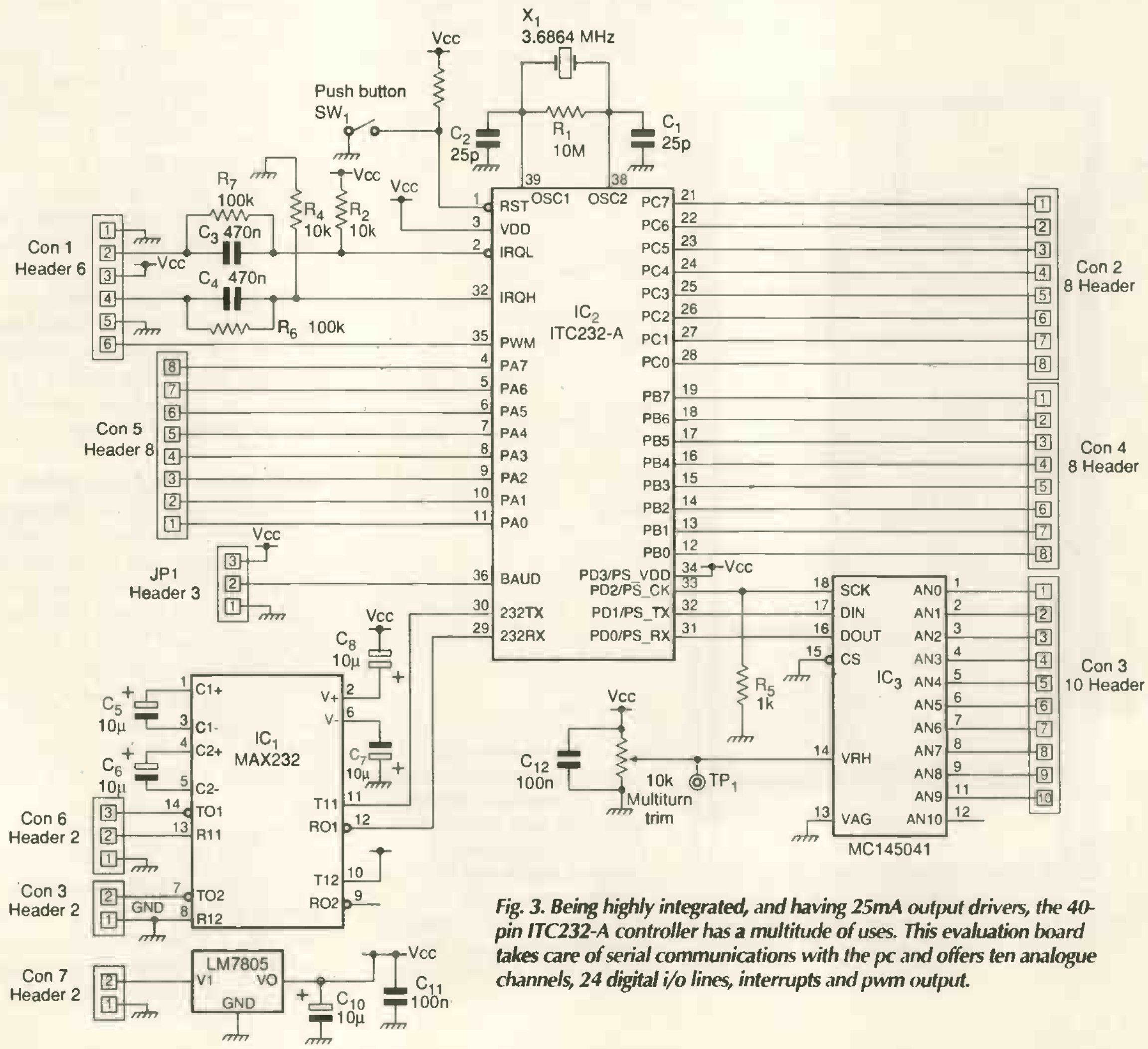


Fig. 3. Being highly integrated, and having 25mA output drivers, the 40-pin ITC232-A controller has a multitude of uses. This evaluation board takes care of serial communications with the pc and offers ten analogue channels, 24 digital i/o lines, interrupts and pwm output.

sage f=00999 is returned by the *ITC232*. This is the actual frequency resulting from rounding errors and crystal resolution.

Three main uses for this feature are:

- Generating an analogue voltage by integration with an *RC* network.
- Varying the speed of a dc motor.
- Producing a given number of pulses by feeding the pwm pin output to an interrupt request pin and counting the lows or highs received by the computer.

Reading resistance or capacitance

Time constant of a series *RC* network can be read directly by the *ITC232*. One end of the resistor connects to V_{CC}, and one end of the capacitor to ground. Pins PC.0 to PC.3 connect to the junction between the capacitor and the resistor.

Command <R>, for resistance, is sent to the device, followed by a <0> or <1> or <2> or <3> for each bit, and finally the eneter key. The controller pin is turned into an output and brought low, discharging the capacitor. Next, the pin is turned back into an input and the time to reach the low to high transition is smapled and sent back to the terminal as a value in the range 0-32767. Units are arbitrary. Should the value be larger than 32767, a time out error is returned.

Further application notes explain how you might read the conductance of a solution, as well as measure various sources of capacitance ■

Control command summary

These are the single-key commands needed to control the ITC i/o control processor. Items within <> symbols are mandatory while items within {} are optional.

<B> *n* sets serial bit rate to *n*, which is between 300 and 115200baud.
<H> calls the help function.
Interrupts: on interrupts, L or H is sent to terminal.
<OFF> returns DISCONNECTING ASCII(#7) > and makes PA.0 an input (to hang up the phone). Only available if in phone mode (baud pin is low and IRQL asserted before a command is received after reset or power-up).
<P>ort <C>onfigure <A> or <B> or <C> {B,%,D,H,$} <*value*> sends *value* to the port specified.
<P> <C> <S>erial <R>ead or <W>rite or <A> {B,%,D,H,$} <V>alue configures serial i/o. PCS0 disables the serial port.
<P> <C> <A>,<B> or <C> or <S> <?> {B,%,D,H or $} Returns the port configuration.
<P> <R> <A> or <B> or <C> or <D> or <S> {B,%,D,H,$} Reading PS sends previously written value out the PD1/SP_TX pin using the Read configuration.
<P> <W> <A> or <B> or <C> or <S> {B,%,D,H,$} <*value*>.
<RESET> is equivalent to a hardware reset.
<R> <0> or <1> or <2> or <3> reads resistance on port C pins 0-3.
<W> followed by H or L sets the pwm line high or low. Decimal suffix between 10 and 10000Hz instead of H or L determines pwm frequency. Duty cycle is 1:1 unless the frequency is followed by an integer between 0 and 100.
<W> <?> returns the last <W> width command.
<S> <E> <A> or <B> or <C> <M>onophasic or <B>iphasic or <H>alf step <Speed> <;> <Stop delay> initiates the stepper procedure on A, B or C ports. <Speed> is in steps/s (10-4000). <Stop delay> is in steps (0-255).
<S>tepper <D>isable <A> or <B> or <C>
<S>tepper <?> {B or % or D or H or $} or <S>tepper <E>nable <?> {B or % or D or H or $} returns the configuration, the active steppers and the last value written to each active stepper in the requested format.
<S>tep <A> or <B> or <C> <L>eft or <R>ight <Number of steps> makes the motor step.
<@> repeats the last command.

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***ITC232* – a new concept in i/o control that adds a versatile and easily programmable logic i/o controller to your pc or terminal.**

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Each board is supplied with applications notes, software simulator, manuals and diagrams, power supply and an RS232 COM1 cable.

The ITC232 incorporates an on chip keyboard to machine-code transalator, which makes programming easy. Via a ready-implemented RS232 links, the i/o232 can connect your application to windows in minutes rather than weeks. Standard routines and simple in-built keystroke-to-machine-code software shortens design cycles, resulting in cost savings.

Technical features

- Serial command and control interface, 300 to 115,200 baud
- 10 channels of analogue i/o
- 24 digital i/o individually configurable as input or output and organised as three ports.
- Pulse-width modulated output 10 to 10,000p/s, 0 to 100% duty cycle in 1% steps.
- 3 stepper-motor outputs, 10 to 4000 steps per second, monophasic, biphasic and half step
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The features of both analogue and digital filters have been used together to improve the bandwidth of samplers. Erik Margan illustrates by example the improvements to be obtained by treating the combination as a single filter.

Antialiasing with mixed-mode filters

Analogue and digital filtering in combination can be used in sampling systems to improve system bandwidth, while retaining high out-of-band signal and noise rejection for effective antialiasing, without the need to increase the sampling frequency. Alternatively, less complicated, lower order filters can be used for attaining the same performance. A method of optimising the filter requirements is discussed.

As an example, suppose the input signal is to be sampled to 12-bit accuracy with a sampling frequency of 2MHz. In this case, frequencies above the Nyquist frequency (1MHz) should be attenuated by at least 2^{12}, or about 72dB. Assume also that constraints such as amplifier bandwidth and phase margin, component tolerances, layout parasitics, thermal effects, etc, limit the filter design to a 6th-order type.

Normally, Chebyshev or elliptic (Cauer) filter types are used for effective antialiasing, since these provide sharp cut-off and the procedure described here is not required. However, for a perfect transient performance or to preserve a high degree of phase coherence in complex signals, the filter must be of the linear-phase type, leading to a Bessel-type filter[2], an all-pole equi-ripple phase filter (±0.05°) or other filter types that can be compensated via phase equalisers.

The use of phase equalisers is limited to band-pass filters, since it is difficult to match the filter phase in wide bandwidth. Bessel filters have a smooth knee in the frequency domain, which makes them a poor choice for anti-aliasing applications. On the other hand, in contrast to the equi-ripple phase types, they can be built from a cascade of relatively low-Q sections, which makes them relatively insensitive to component tolerances. Most importantly, their time-domain performance is ideal.

Although a Bessel filter will be used in the example, calculating the stop-band asymptote of a 6th order Butterworth filter that satisfies the no-alias requirement gives a simple relation from which the required system asymptotes can easily be calculated. The frequency f_A at which the nth order Butterworth system reaches the required attenuation A can be calculated from,

$$f_A = 10^{\frac{\log_{10}\left(A^2-1\right)}{2n}} \qquad (1)$$

Equation 1 assumes a normalised system, with its –3dB cut-off frequency f_C=1 and the response at zero frequency A_0=1. Taking $A=2^{12}$ and n=6 results in f_A=4.

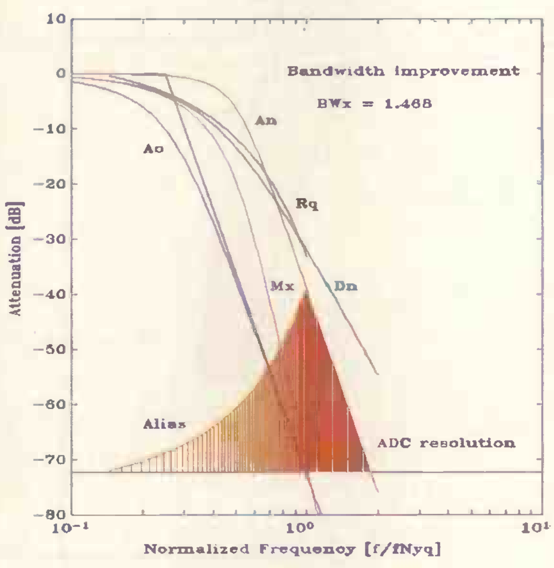

Fig. 1. Mixed-mode filter bandwidth improvement. Frequency scale normalised to the Nyquist frequency (0.5 of the sampling frequency). Attenuation scale normalised to the system gain at dc. Dotted curve A_o is the response of the original 6th-order analogue-only filter, reaching the 12-bit a-to-d converter resolution limit of –72dB at the Nyquist frequency. If the analogue filter bandwidth is moved upward (A_x), so that the converter resolution limit will be reached at $1.87f_{Nyq}$, the dark-shaded part area from f_{Nyq} to $1.87f_{Nyq}$ will generate an alias spectrum from f_{Nyq} to $0.13f_{Nyq}$ (light-shaded). The alias spectrum envelope, flipped about the frequency axis, determines the minimum required attenuation dashed line R_q of the digital filter D_x, which would make the alias spectrum envelope equal to the a-to-d converter resolution limit. The resulting mixed-mode filter response M_x will have its –3dB cut-off frequency 1.468 times higher than A_o.

Now calculate the 6th-order Bessel system polynomial coefficients (see the Bessel panel), divide them by $\sqrt[n]{d_0}$ to normalise the system to have the same stop-band asymptote as the Butterworth filter and extract the polynomial roots[3] to get the poles.

Since f_A must be equal to the Nyquist frequency, denormalise the system by taking the inverse value of f_A, which gives the Butterworth bandwidth relative to the Nyquist frequency f_{Nyq}, equal to 250kHz. The poles of the Bessel filter must also be divided by f_A, resulting in a –3dB bandwidth of 144kHz. This is the reference figure for the analogue-only antialiasing filter. If this figure is not high enough and if the choice of the analogue-to-digital converter limits the maximum sampling frequency, use mixed-mode filtering to expand the system bandwidth.

Analogue/digital filters

The idea of using mixed-mode filtering comes from the fact that the total system frequency response is a simple multiplication of the analogue and digital filter frequency responses. Transforming the digital z-domain response is trans-

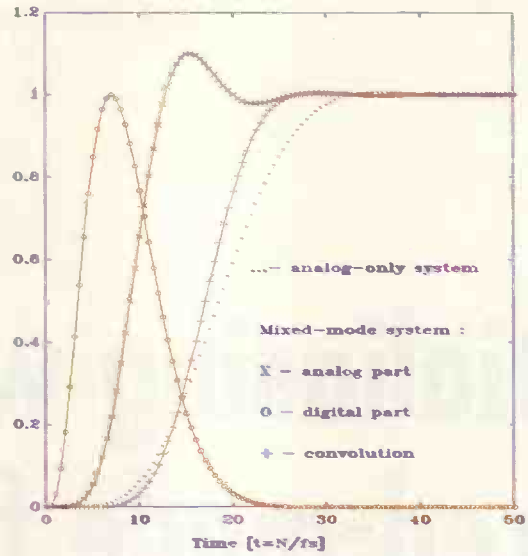

Fig. 2. Time-domain representation of the mixed-mode filter performance. Convolving the analogue filter step response with the digital filter impulse response gives the perfect step response with a rise time shorter than the analogue-only filter.

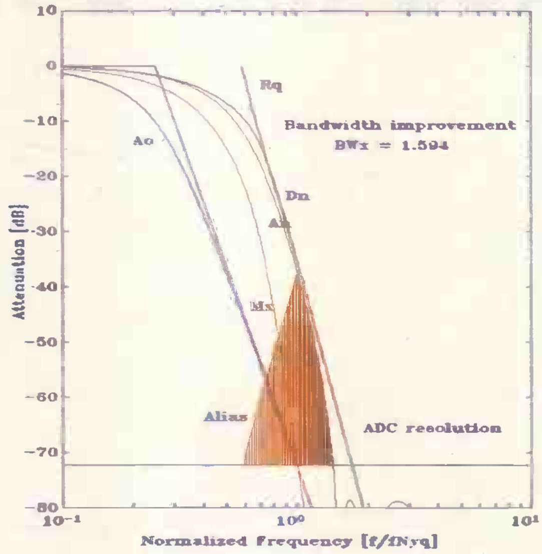

Fig. 3. Example of mixed-mode filtering, using zeros with analogue filter. Zeros are at 1.5,2.0 and 4.0 times f_{Nyk}. Analogue/digital filter can be now moved up by 2.37, while still resulting in a relatively narrow alias spectrum and giving total bandwidth improvement of nearly 1.6.

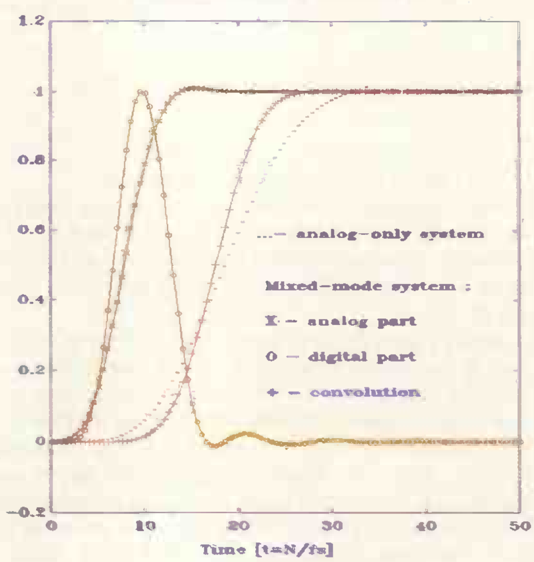

Fig. 4. Time-domain performance of the Fig. 3 mixed-mode filter, using zeros and poles in the analogue section. Note better rise-time of the mixed-mode step response.

formed into its s-domain equivalent gives,

$$H(s) = A(s) \times D(s) \qquad (2)$$

That is also true for the reverse case (i.e. a system formed from a digital filter, a d-to-a converter and analogue filter). In the time-domain, Eq. 2 becomes the convolution integral of

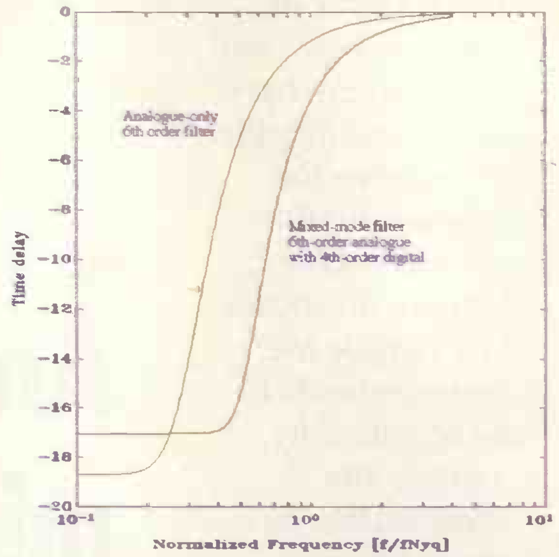

Fig. 5. Time-delay (phase vs frequency derivative) of all-pole mixed-mode filter is constant up to a frequency more than double that of analogue-only filter.

the analogue signal with the digital filter impulse response and convolution is exactly the process performed by digital filtering, the digital filter coefficients representing the sampled equivalent of the impulse response[3].

However, as is well known from analogue filters, cascading two separately optimised filters reduces the total system bandwidth more than one would like. It is thus better to use a single filter system but of higher order. Since the limit is a 6th-order analogue filter, calculate a 10th-order filter, assign six of its poles to the analogue part and the remaining four to the digital part. A higher order filter has a steeper stop-band and so its bandwidth can be higher while still satisfying the antialiasing condition, but how much higher is not yet known. **Figure 1** shows the optimisation criterion.

Dotted curve A_o is the 6th-order analogue-only reference system, shown along with its pass-band and stop-band asymptotes. A_x and D_x are the analogue and digital part of the mixed-mode filter M_x, which is a 10th-order Bessel filter. Of its ten poles (arranged as five complex-conjugate pairs), six of them, in three pairs, have been assigned to the analogue filter A_x and the remaining four in two pairs to D_x.

Since A_x is of the same order as A_o, its stop-band slope is the same as the reference, allowing easy calculation of the effect of increasing its bandwidth. In Fig. 1, it has been increased by 1.87 and the line-shaded frequency band between the Nyquist frequency f_{Nyq} and $1.87f_{Nyq}$ will, when sampled, be reflected into the dot-shaded alias spectrum between f_{Nyq} and $(2\text{-}1.87)f_{Nyq}$. The difference, in dB, between the a-to-d converter resolution level and the alias spectral envelope gives the minimum required attenuation (shown as the dashed line R_q) that the digital filter must have to suppress the alias spectrum below the ADC resolution level.

From Fig. 1, one could conclude that optimal performance is reached whenever the mixed-mode response reaches the a-to-d converter resolution level at the Nyquist frequency, but be warned that this will not be so in the majority of cases. Instead, the optimum is achieved by iteration – first, shift upward the analogue and digital frequency responses (the poles multiplied by a factor between 1 and 2), then calculate the alias spectral envelope, take the difference between the a-to-d converter resolution level and the alias envelope and finally compare it to the frequency response of the digital filter. If the filter is much below the required level, repeat the process; if it is above the required level, multiply the poles by

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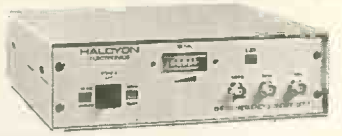

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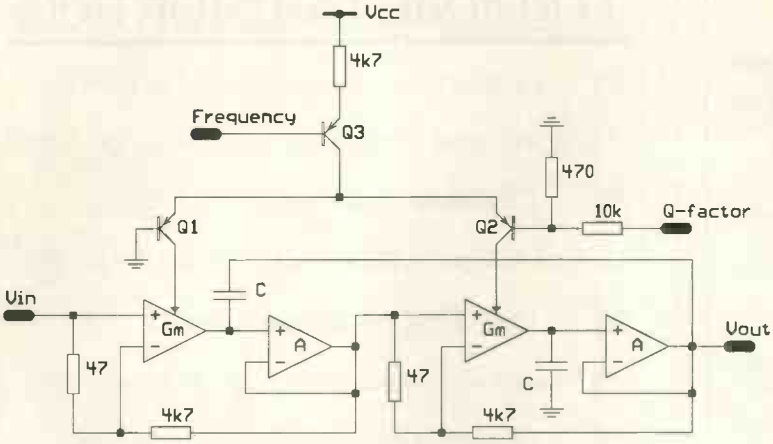

Fig. 6. Two-pole, voltage-controlled filter example. Cascade of three such sections needed for the six-pole example of Figs 1 and 2. This is a classic Sallen-Key configuration in which the resistors have been replaced by transconductance amplifier's g_m and each g_m-C pair buffered. Buffer op-amps ACFB must be of the wide-band type (i.e., with current feedback) to prevent parasitic transfer function zeros. Q and frequency of each two-pole section must be adjusted separately, in accordance to the poles selected. Resistive dividers of 4.7kΩ and 47Ω keep the OTAs in the linear range and prevent slew-rate limiting for large signals.

a factor lower than 1 and test the result again.

From the shape of the alias spectral envelope it is clear that there is no point in making the digital filter of high order. Likewise, it is advantageous to choose the poles having smaller imaginary part for the digital filter, since this results in a smoother response and consequently greater bandwidth improvement factor. In this example, the mixed-mode system has its –3dB cut-off frequency at 211.5kHz, which is 1.468 times the all-analogue filter bandwidth.

Splitting the filter poles between the analogue and digital part may also be taken into consideration; designers of systems that must operate in real time will look for the pole selection that gives the digital filter a more symmetrical impulse response – every other complex-conjugate pole pair is assigned to the digital filter. This property of symmetry can then be exploited to reduce the required filter coefficients (and consequently the number of multiplications) by half, speeding-up the digital filtering process.

On the other hand, when the available analogue gain-bandwidth product is critical, the designer may prefer to assign the poles with the lower imaginary part to the analogue filter, but at some expense to the bandwidth improvement.

Figure 2 shows the time-domain behavior of the same filters used to produce Fig. 1, with the time scale normalised to the sampling period and the markers on the curves corresponding to actual samples. Analogue step response, with its notable overshoot, convolved with the digital impulse response gives a perfect step response with a rise time shorter than that of the analogue-only filter (the dotted curve).

From Fig. 1 it is also obvious that all-pole filters can not achieve a bandwidth improvement greater than about 1.5, since this would require the analogue filter asymptote to approach the sampling frequency at the a-to-d converter resolution level, extending the alias spectrum towards dc, where it would be hard to eliminate. If the analogue filter is designed to have some stop-band zeros at the sampling frequency and its first few multiples, a greater bandwidth improvement will be possible. One such case is shown in **Fig. 3** and **Fig. 4**, where a six-pole, six-zero analogue filter is combined with an eight-pole equivalent digital filter. Zeros are at 1.5 , 2.0 and 4.0 times f_{Nyq}, which were not chosen for optimum pass-to-stop band transition, but for narrowing the alias band.

While the bandwidth improvement in both cases may seem small, it will be appreciated by those who use spectrum analysis daily. It must be noted that the resulting improvement in phase linearity is even greater than in bandwidth, since the additional extension comes from the use of a higher order filter. **Figure 5** shows how the all-pole, mixed-mode system time-delay, i.e. the phase *vs* frequency derivative,

$$t_D = \frac{d\varphi}{d\varpi}$$

remains constant up to a frequency more than double that in the analogue-only filter.

If the a-to-d converter system is to be used with different sampling frequencies, the digital filter part can be left unchanged, but the analogue filter must be frequency-shifted accordingly; transconductance operational amplifiers used for frequency control offer the best way of doing this[4]. **Figure 6** shows an example of a two-pole filter section, with separately adjustable frequency and Q.

Voltage at the base of Q_2 of about ±50mV dc sets the Q (the imaginary components of the pole pair) and the control voltage at the base of Q_3 (ranging from V_{CC}–0.7V to about +0.7V) sets the frequency; the magnitude of the pole pair – the ratio of the imaginary to the real component remains unchanged. A cascade of three such sections is needed for the six-pole analogue filter, each section being adjusted separately and the adjustments remaining in fixed proportions as the frequency control voltage is changed. A simpler, but less flexible, solution is to make all the transconductances equal and select the values of capacitors as required by the poles.

I built my experimental filter using RCA *CA 3080* operational transconductance apmlifiers and Comlinear *CLC 400* current-feedback devices. However, the Linear Technology *LT 1228*[5], which is a single-chip OTA with current feedback, is the natural choice. Transfer function of the filter in Fig. 6 is,

$$\frac{V_{out}}{V_{in}} = \frac{g_{m1}g_{m2}/(k^2C_1C_2)}{s^2 + sg_{m1}/(kC_1) + g_{m1}g_{m2}/(k^2C_1C_2)} \qquad (3)$$

where k is the attenuation of the OTA input resistive divider (1/101), and g_m is the OTA transconductance, set by the bias currents from the collectors of Q_1 and Q_2 . Comparing Eq. 3 with the general two-pole transfer function :

$$H(s) = \frac{p_1p_2}{(s-p_1)(s-p2)}$$

$$= \frac{p_1p_2}{s^2 + s(-p_1 - p2) + p_1p_2} \qquad (4)$$

and normalising $g_{m1}=g_{m2}=1$ produces,

$$C_1 = \frac{1}{k(-p_1 - p_2)}$$

and

$$C_2 = \frac{1}{k^2 p_1p_2C_1} \qquad (5)$$

Aliasing

In theory, the bandwidth of the sampling system is equal to the Nyquist frequency, which is one-half of the a-to-d converter's sampling frequency. In practice, however, correct waveform spectrum can be found only if the input signal frequencies above the Nyquist frequency are attenuated to levels lower than the a-to-d converter resolution, to avoid 'aliasing' (if the signal contains discrete frequency components above the Nyquist frequency, or broadband noise). This is known in literature as the Shannon's sampling theorem (see Further Reading).

Aliasing can be best understood if the reader remembers the scene from Western movies, where the wheels of the stage coach seem to be rotating backwards, while the horses are running wild to escape from the desperados behind. What is perceived, is as if the wheels rotate with a frequency equal to the difference between the frequency at which the pictures were taken and the actual wheel rotation frequency.

A wheel, rotating at exactly the same frequency (or its integer multiple or sub-multiple) as the picture rate, would be perceived as stationary (remember the stroboscope effect). This is the same as if an a-to-d converter is sampling a signal of a frequency equal to its sampling frequency – such a signal can not be distinguished from a d.c. level. Likewise, a signal with a frequency slightly lower than the sampling frequency, could not be distinguished from a low frequency, equal to the difference of the two.

Table 1. Poles used in the example of Fig. 1 and 2.

Analogue-only system	Mixed-mode system Analogue	Digital
–0.1346 ±0.2494*i*	–0.3886 ±0.1534*i*	–0.4066 ±0.0510*i*
–0.1999 ±0.1405*i*	–0.2870 ±0.3657*i*	–0.3506 ± 0.2576*i*
–0.2273 ±0.0464*i*	–0.1826 ±0.4836*i*	

Alternatively, normalising $C_1=C_2=1$ produces,

$$g_{m1}=k(-p_1-p_2)$$

and

$$g_{m2}=\frac{k^2 p_1 p_2}{g_{m1}} \qquad (6)$$

Poles p_1 and p_2 are the suitable complex-conjugate pair of the mixed-mode filter poles.

Bessel filters

Bessel filters[2] are optimum in the sense that all the derivatives of the envelope (group) delay response are zero at origin, which results in a maximally flat envelope delay. This means that all the relevant frequencies pass through the system with equal time delay, resulting in a transient response with a minimal overshoot. In the complex frequency plane, a system with pure time delay may be represented by

$$H(s)=e^{-sT} \qquad (7)$$

First, normalise this by making T=1; then expand e^{-s} as a polynomial. However, if this is done using the Taylor series expression for e^x and if the polynomial degree exceeds 4, the resulting polynomial would not be of the Hurwitz type, since some of the poles would be in the right-half of the complex plane, making the system unstable. But there is another expression for e^{-s} that we can use:

$$e^{-s}=\frac{1}{\sinh s+\cosh s}=\frac{1/\sinh s}{1+\cosh s/\sinh s} \qquad (8)$$

The series for hyperbolic sine function has even powers of s and the hyperbolic cosine odd powers of s. When these polynomials are divided using long division, the poles of the resulting polynomial meet the stability requirement. Expressing this as a partial fraction expansion truncated at the nth fraction gives an nth-order Bessel system. This can be expressed as

$$H(s)=\frac{d_o}{B_n(s)} \qquad (9)$$

where

$$B_n(s)=\sum_{k=0}^{n} d_k s^k.$$

$Bn(s)$ is an nth order Bessel polynomial which, for different n, satisfies the relations,

$$B_0(s)=1$$
$$B_1(s)=s+1$$
$$B_n(s)=(2n-1)B_{n-1}(s)+s^2 B_{n-2}(s) \qquad (10)$$

The coefficients d_k of the resulting polynomial can be calculated as,

$$d_k=\frac{(2n-k)!}{2^{(n-k)}k!(n-k)!)},\ \text{for } k=0,1,2...n \qquad (11)$$

Roots of $B_n(s)$ are the poles of $H(s)$. Calculated in this way, the system is normalised to a time-delay of 1 for any n, which results in a bandwidth increasing with n. In these calculations, a different normalisation is used: the asymptote of the filter stop-band is made equal to that of the Butterworth filter of equal order, by dividing the polynomial coefficients d_k by $\sqrt[n]{d_0}$.

Bessel filter poles are found in the left-half of the complex plane, on a family of ellipses with one focus at the origin 0+0i and the other on the positive part of the real axis. **Table 1** shows the poles used in the example of Fig. 1 and Fig. 2. These values are given relative to the Nyquist frequency – to get the true values, multiply them by 1MHz.

Filter response calculation

In the frequency domain:

$$H(s)=\frac{\prod_{i=1}^{n}(-p_i)}{\prod_{i=1}^{n}(s-p_j)}\cdot\frac{\prod_{j=1}^{m}(s-zj)}{\prod_{j=1}^{m}(-zj)} \qquad (12)$$

where $s=j\omega$ and p_i are the poles and z_j are the zeros (if any).

Magnitude in decibels is

$$M(\omega)=20\log_{10}\sqrt{H(j\omega).H(-j\omega)} \qquad (13)$$

In the time domain, calculate the residue of each pole and sum the residues at each time point to get the impulse response. For the step response, each residue is multiplied by $1/s$ the Laplace transform of the input unit-step. The residue of the kth pole can be calculated as,

$$R_k(t)=\lim_{s\to p_k}(s-p_k).\frac{\prod_{i=1}^{n}(-p_i)}{\prod_{i=1}^{n}(s-p_i)}\cdot\frac{\prod_{j=1}^{m}(s-z_j)}{\prod_{j=1}^{m}(-z_j)}.e^{p_k t} \qquad (14)$$

Terms $(s-p_k)$ cancel for $i=k$ before limiting. Next, make $s=p_k$, without using the limiting process. By doing so, the general applicability of Eq.14 is lost – it does not hold for systems containing coincident poles, but for all optimised system families the result is still valid. The time t can be chosen to start from 0 up to any desired time, in sampling period increments. Then:

$$f(t)=\sum_{k=1}^{n}R_k(t) \qquad (15)$$

In summary

From all this, one can see that mixed-mode (analogue plus digital) linear-phase filtering can be used effectively to extend the usable spectral bandwidth of sampled signals by about 50% and the phase coherence by more than 100%, while keeping the signal spectral resolution, the sampling frequency and the number of samples unchanged. ■

Further reading

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4. Azadet, K. Linear-phase, continuous-time video filters based on a mixed analogue/digital structure. *ECCTD '93-Circuit Theory and Design*, pp.73-78. Elsevier Sci. Publ.
5. Hickman, I. Versatile twin amplifier has many uses. *Electronics World +Wireless World*, December 1993, pp.1044-1048.

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WINNER

First published in the March 1995 issue, this meter is the latest design to win the Thurlby Thandar 1GHz spectrum analyser detailed on the right.

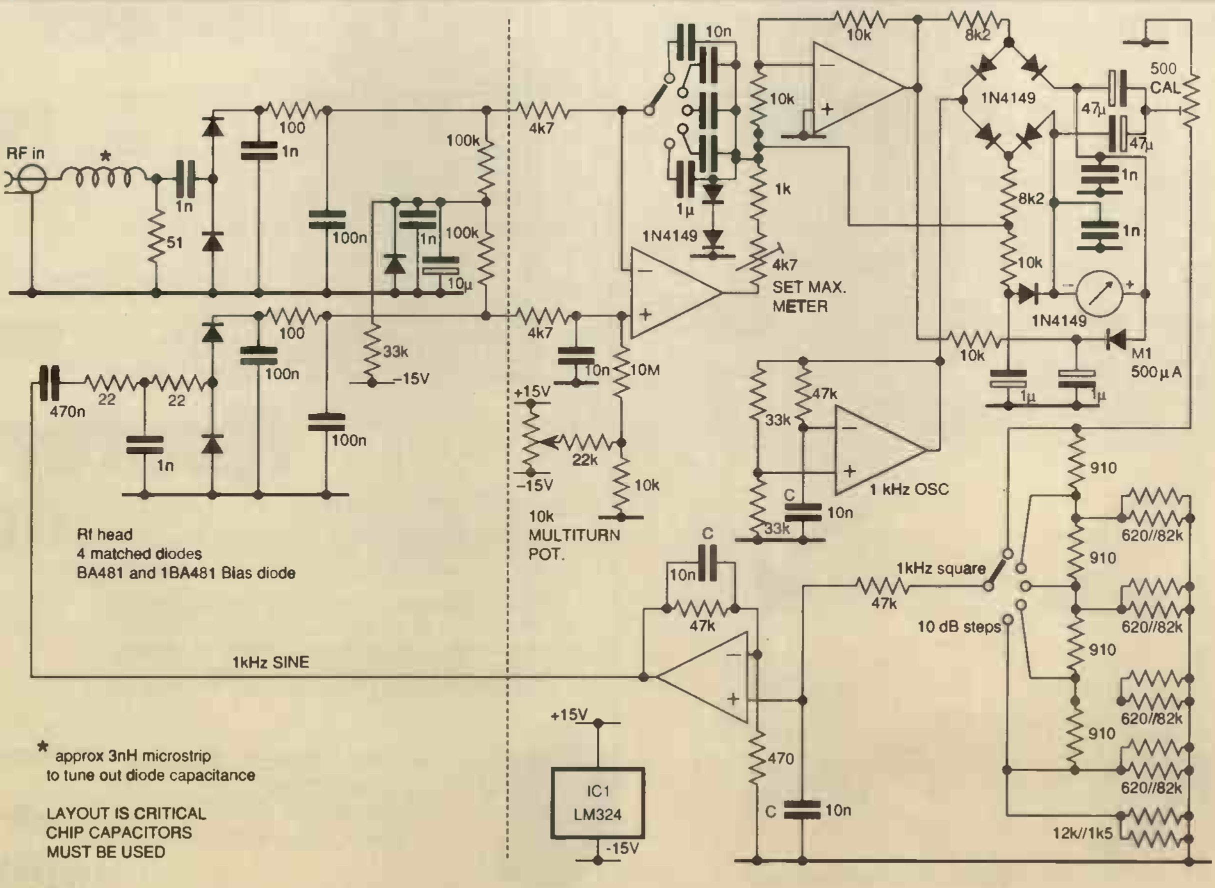
Vhf meter is accurate to 0.1dB

Mixer-type Schottky diodes can detect rf signals between around –35dBm and 20dBm. Response flatness is good to about 1GHz with a cheap *BA481* and better with higher frequency-response types; a general-purpose Schottky measures higher powers in a 50Ω system but with a reduced frequency response.

Since the diode output is temperature-variant and somewhat variable between batches, the circuit shown uses two matched detectors, the second one fed with a 1kHz sinewave, adjusted in amplitude by the error amplifier circuit until the outputs are balanced. A chopper composed of four *1N4149* diodes provides the 1kHz square, which is then formed into a sinewave by the op-amp, and also provides a dc level proportional to the amplitude. This is a measure of the rf input level and is read, in this case, on a moving-coil meter calibrated in decibels, though it could easily be digitised. Ranges of 10dB are selected in the 1kHz drive to cover the 55dB dynamic range, and the error integrator capacitors are switched to provide the relevant time constant. The circuit is inherently linear.

The 10kΩ multi-turn pot zeroes the meter. Additional diodes on the meter cope with reverse drive if the integrator is zeroed too low, since the chopper provides only forward drive to the meter. On the integrator output, the 4.7kΩ pot sets maximum meter current. Accuracy to several hundred MHz is about 0.1dB, without the need for stable supplies. Capacitors marked *C* should be of the same type to match temperature coefficients.

P D Brooking
Ryde, Isle of Wight



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Half-duplex-to-RS232 converter

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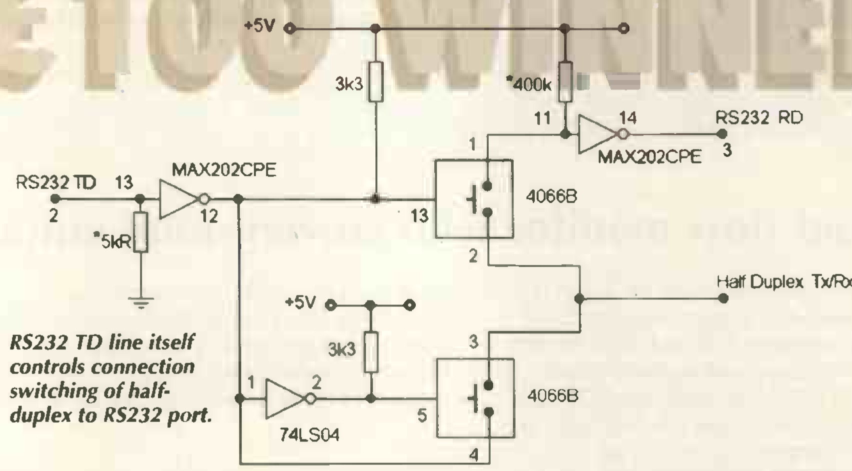
Connecting a half-duplex line to a full-duplex RS232 port requires a decision on whether the half-duplex goes to the TD transmit line or the receive line RD.

A truth table for such a converter is constructed as follows. When TD is at logic 1, it is either at rest, 'marking' or transmitting a 1 data bit. If it is in the first state, data might be passing from the half-duplex line to the port and the half-duplex should go to RD. If it is transmitting, RD will be marking at logic 1 and the half-duplex line should go to either RD or TD.

TD	RD	Half-duplex
0	1	TD
1	1/data	RD

As the table shows, TD is usable to control switching. The *MAX202CPE* converts and inverts RS232 levels to ttl levels; the pull-up/down resistors are internal and are shown to illustrate the requirement. For ttl compatibility, the *4066B* is run from 5V.



RS232 TD line itself controls connection switching of half-duplex to RS232 port.

Bill Geake
University of Edinburgh
Royal Infirmary of Edinburgh

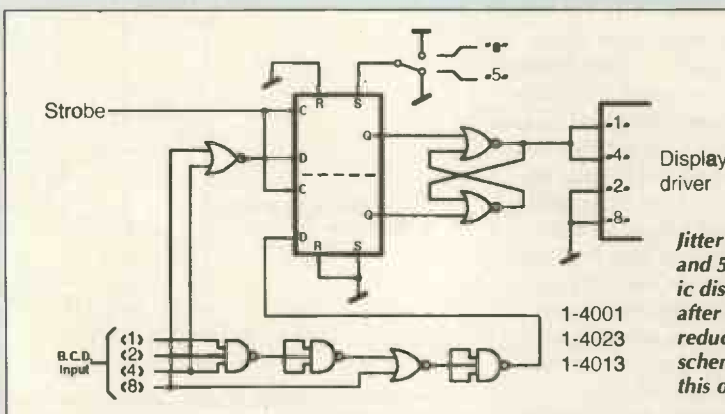
Calming a digital display

Annoying ±1 jitter in a numeric display can be reduced considerably by allowing only 0 and 5 to show when the input is under or over 5, but this still leaves the jitter when the input is between 9 and 0 or 4 and 5.

In the arrangement shown, a hysteresis of four counts switches the display to read 5 when the input is greater than 7 and, in the reverse direction, 0 when input is less than 3.

The flip-flop allows clocking of strobed displays and the switch allows '0' to be permanently displayed.

Des Keppel
Ballon
County Carlow, Ireland



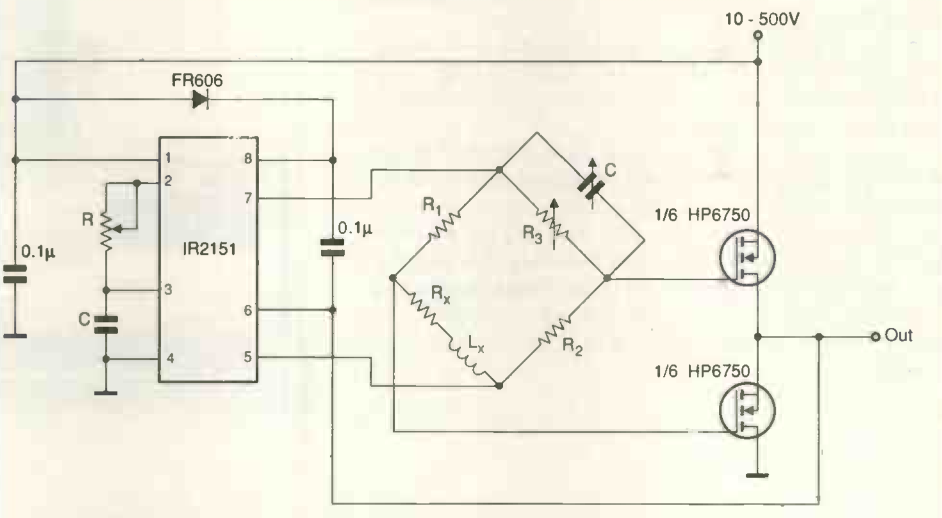
Jitter between 4 and 5 in a numeric display remains after other jitter-reduction schemes are used; this one stops it.

High-voltage, frequency-controlled Maxwell bridge

International Rectifier's *IR2151* is a bridge driver consisting of an oscillator, with external timing components, and mosfet output. It is used here to drive a Maxwell bridge to measure inductance in the L_x position at a variable frequency and at voltages up to 600V; as a metal detector, it will monitor changes in the value of the inductance when it is brought near a ferromagnetic object.

Frequency is determined by the value of R and lies in the kilohertz range. Maxwell bridges are used in the measurement of magnetic permeability at various frequencies, a process in which the power mosfet output can be an advantage.

Kamil Kraus
Rokycany
Czechoslovak Republic



Oscillator with mosfet output power drives a Maxwell bridge, not only for inductance and permeability measurement, but also to function as a metal detector.

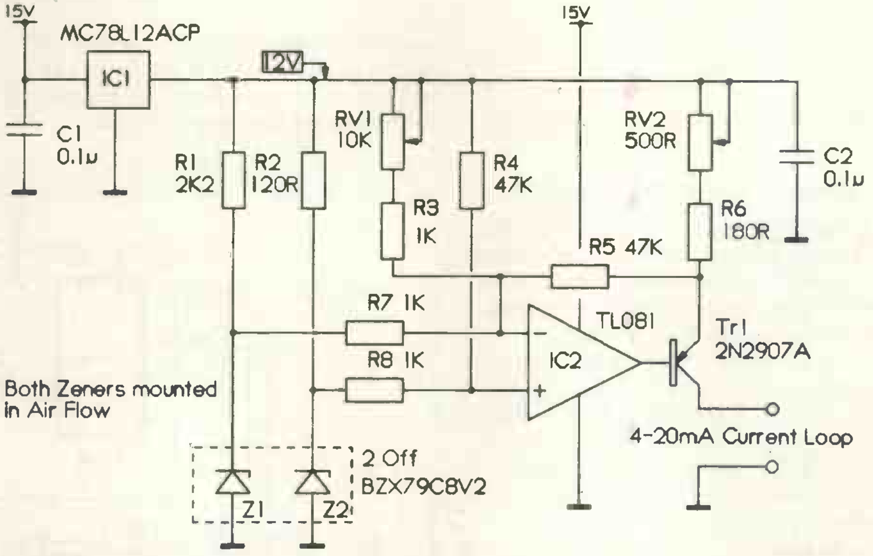
Fluid-flow monitor with current-loop output

Although zener diodes are linked in the mind with stability, those having a zener voltage of 5.6V and above do show a positive temperature coefficient; the *BZX79* 8.2V zener, for example, has a 4.6mV/°C coefficient which is linear over a 30°C temperature span. Power dissipation in the device causes self-heating, which makes it difficult to use as a thermometer but, used as a 'hot-wire' anemometer, the effect is useful. Fluid flow across the zener removes heat and its temperature falls, while dissipation is constant due to the reasonably constant breakdown voltage.

In this circuit, two zeners each have a current-limiting resistor $R_{1,2}$, with different values to give different power dissipation in each zener. The one with low dissipation loses all its heat to the fluid at low flow rates, so that its temperature is about the same as that of the fluid; the other takes a higher current, so that a faster flow is needed to remove heat. At high flow rates, both temperatures are the same, but at lower rates, the high-dissipation zener warms and the difference in voltages is sensed by the op-amp. Since the devices are in the flow, response is rapid.

For a remote indication, the op-amp is made to drive a 4-20mA current source, provided by Tr_1. Trimmer RV_1 provides offset adjustment and RV_2 adjusts gain. Component values in the circuit shown are for air through an 8mm orifice, flowing at around 2m/s, both zeners being laid across the orifice.

P Harley
Newcastle-upon-Tyne
Tyne and Wear



Temperature coefficient of zener diodes – normally a disadvantage – is used here to provide fluid-flow sensing. Output is a 4-20mA current loop.

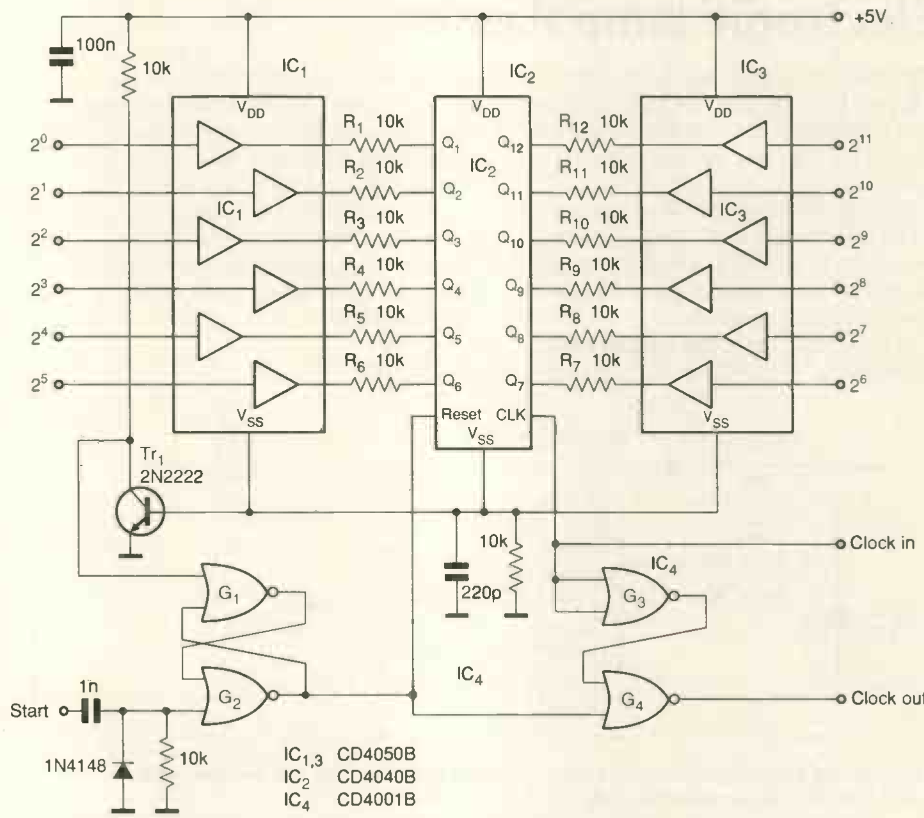
Programmable stepper-motor pulse generator

As an alternative to previous circuits for programming the number of pulses for a stepper motor, this is rather simpler.

Outputs of a 12-stage binary counter IC_2 are taken to the outputs of twelve non-inverting buffers, $IC_{1,3}$, via resistors $R_{1\text{-}12}$; inputs to the buffers determine the count and, therefore, the number of pulses at the output. Gates 1,2 form an RS flip-flop and gates 3,4 an And gate.

A start signal sets the flip-flop, which in turn resets the counter, which starts to count. While the counter outputs differ from the gate outputs, current flows in one or more of the resistors $R_{1\text{-}12}$, this current flowing to ground via the V_{ss} terminals of the counter and buffers and the base of Tr_1, maintaining saturation of Tr_1. As counter and buffer outputs equalise, the current stops, Tr_1 cuts off and the RS flip-flop resets. This closes the And gate and stops clock pulses at the output.

M S Nagaraj
ISRO Satellite Centre
Bangalore
India



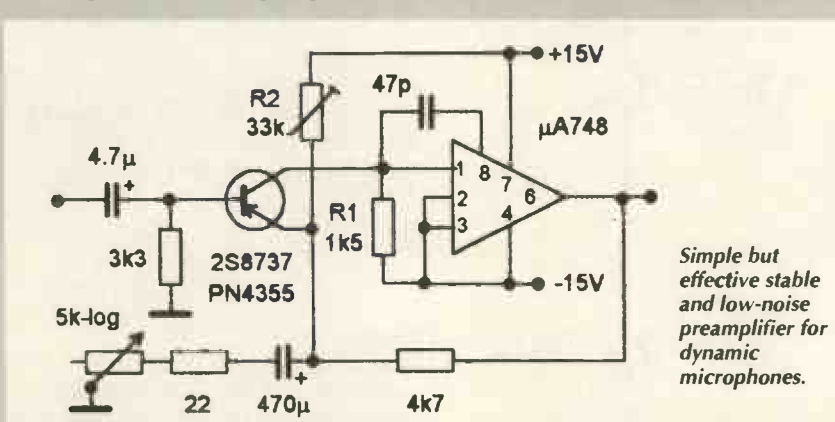
Pulse generator provides predetermined number of pulses to a stepper motor driver, the circuit using only four ICs.

Stable microphone preamplifier

In audio equipment preamplifiers, low noise must often be combined with variable gain. Low-noise op-amps do not, in general, suit low-impedance inputs such as a dynamic microphone and it is therefore common to use either input transformers or low-noise transistors in a feedback loop, an arrangement which can produce instability with large variations in gain setting.

By-passing the input differential pair of the *748*, as shown, removes potential instability and the transistor is selected for a low noise characteristic when driven from a 200Ω source, its collector current being adjustable for the same reason. The circuit is remarkable for its low noise, stability and low distortion, in spite of the fairly basic nature of the circuit and the use of the general-purpose op-amp. Resistor R_2 adjusts the circuit for symmetry.

Michal Dolezal
Ostrava
Czechoslovak Republic



Simple but effective stable and low-noise preamplifier for dynamic microphones.

Electronic lamp flasher

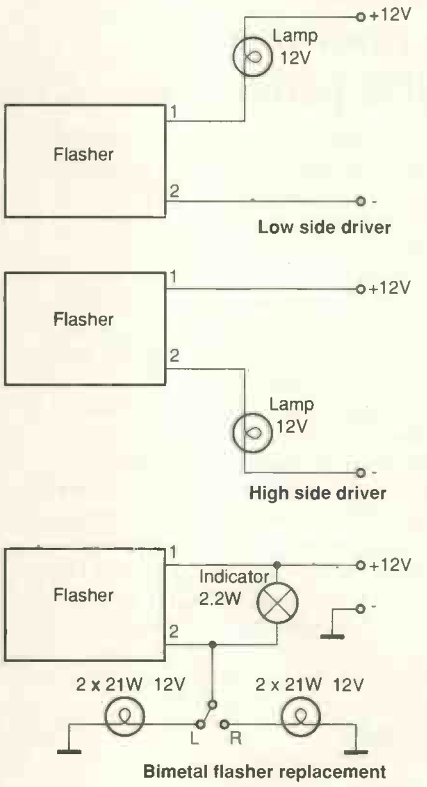
Replacing the usual bimetallic-strip switch in lamp flashers, this circuit only uses power when the lamp flashes and can be used for high-side and low-side switching.

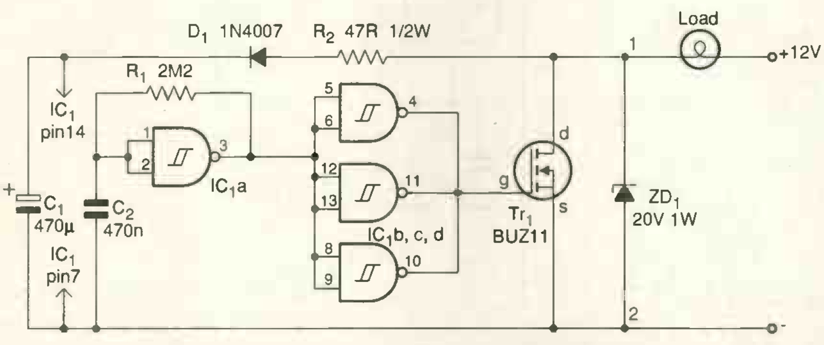
One gate of the Schmitt Nand operates as a 1Hz free-running multivibrator and the other three in parallel as a buffer to switch the *BUZ11* power mosfet. Zener 1 isolates the circuit from transients.

At power up, circuit states are such that the mosfet is off and C_1 is charging via D_1, R_2 and the lamp. During this time, C_2 charges through R_1, eventually toggling the multivibrator, switching the mosfet on and lighting the lamp. Power for the IC is supplied all this time by the charged C_1.

Capacitor C_2 is now discharging through R_1; the reverse process takes place, the mosfet goes off, allowing C_1 to charge up again and the cycle repeats.

NK Goodman
Hastings
East Sussex





Two-terminal bimetal flasher replacement, usable in both high-side and low-side application and using no permanent supply. IC_1 is a cmos device.

Adjustable output-resistance amplifier

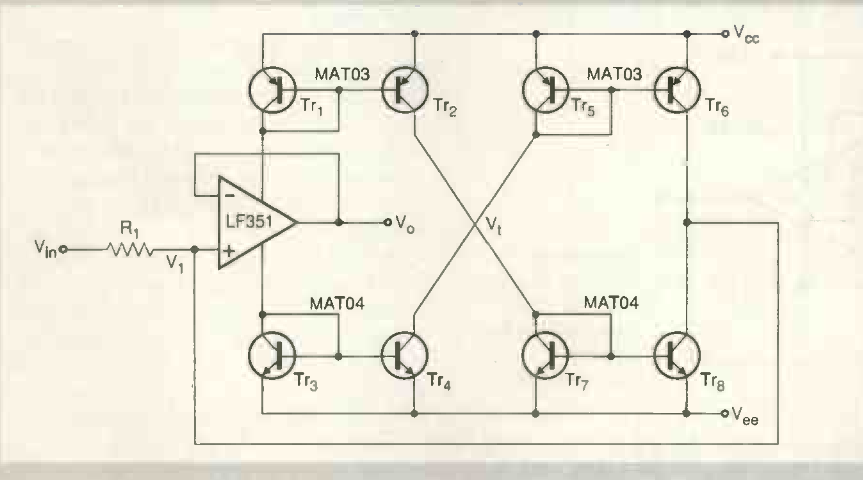
Line-driving amplifiers conventionally require a series output resistor to pad the resistance which, if the amplifier output impedance is low, causes half the output to be lost and power dissipated. This circuit synthesises the required impedance without these drawbacks.

With the input shorted to ground, a test current i_t forced into the op-amp output flows in one of the power-supply lines, where it is sensed by one of the left-hand current mirrors, depending on its sign. It is then inverted by the relevant right-hand current mirror and forced in R_1, where it causes a voltage drop $v_t = i_t R_1$ to be developed. Since, due to feedback round the op-amp, $v_t = v_1$, the output resistance is

$$R_o = v_t / i_t = R_1$$

The impedance is not exact because of the realistic current-mirror gain, but trimming R_1 corrects the error.

Dan Stiurca
Iasi
Romania



Line driver has the required output impedance without loss of amplitude, without an excess of power dissipation and with a rail-to-rail swing.

NEW PRODUCTS CLASSIFIED

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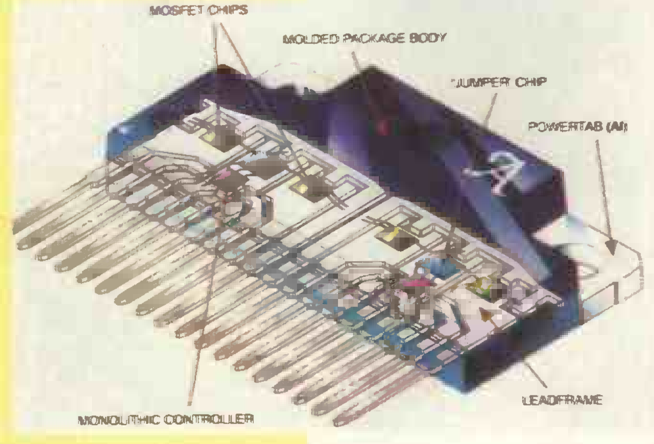
ACTIVE

Memory chips

Synchronous flash simms. New memory modules from Smart Modular Technologies allow operation with zero wait states at frequencies to 33MHz, using a clock signal to work synchronously, outperforming asynchronous simms and at least equalling dram simms. An on-board chip arranges the optional active reset control so that the simm always powers up in the right state during hot insertion. Since there is no standard pinout for these devices, they use the arrangement for asynchronous types, slightly modified to affect only read. Smart Modular Technologies. Tel., 01604 497735; fax, 01604 497739

Mixed-signal ICs

Pwm stepper controller/driver. Three multi-chip modules by Allegro, the *SLA7024M/26M/29M* are pwm controller/drivers for two-phase unipolar stepper motors. Each uses four nmos fets for the driver output and, in most cases, external heat sinks are not needed; in case they are, the *SLA7026M* has an electrically isolated power tab to transfer heat. Inputs are compatible with 5V logic and micro output. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.



Microprocessors and controllers

16-bit, 3V controller. Microcontrollers in NEC's *78K4* family carry out 16×16 multiplications in under 1.2µs. They are single-chip devices, four versions having no rom, with 128K rom in mask, one-time programmable and uv erasable. Integrated peripherals include uarts, high-drive parallel and pwm outputs and a timer unit for stepper-motor control as well as data conversion. Of the 64 i/o lines, 24 sink up to 8mA and eight transistor drive outputs will source 5mA. Source code is compatible with that used in the company's *78K0* and *78K3* devices. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

8-bit, 5mips. Microchip's *PIC16C73* field-programmable, risc-based microcontroller has 4096 words of one-time-programmable program memory, a low-power, 5-channel, 8-bit a-to-d converter and operates at up to 5mips. Its Harvard-architecture risc processor has a 200ns cycle time and peripherals include a timer subsystem. I/o functions include a synchronous serial port supporting SPI or I^{2}C/Access bus protocols, a 5Mb/s usart with a baud-rate generator, two 80kHz pwm outputs and a 16-bit capture/compare feature. The device takes less than 15µA from 3V at 32kHz and under 1µA when asleep. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

CTV micro. Toshiba has a new member of the *TLCS-870* family of 8-bit microcontrollers: the *TMP87PM36N* one-time-programmable device for colour television receivers and other consumer products; it is programmable by means of a standard eprom programmer and otp adaptor. Features include a four-channel, 6-bit a-to-d converter for afc, an I^{2}C bus with multi-master control, pwm outputs to give 7 or 14-bit resolution and a remote pre-processor. An on-screen display function provides 128-character, 24-column by 12-line output with variable positioning. Operating speed is over 8MHz in the voltage range 2.7V-5.5V. Toshiba Electronics (UK) Ltd. Tel., 01276 694600; fax, 01276 691583.

Shrinking micro. NEC's *V53A* microprocessor continues its microscopic tendencies with a further reduction in size from 20mm^2, itself a reduction from 28mm^2, to 14mm^2, with a height of 1mm. Everything else remains the same in the chip, which is used for cpu-intensive work like number-crunching and data sorting and now, probably, for cellphones. Package is a 120-pin TQFP with a pin pitch of 0.4mm. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

68040-based multiprocessor. BVM offers the *RAMnet* software package to enable 16 *68040*-based *BVME4000* cpus to be used on one 32-bit backplane, each being able to use all installed memory, which amounts to 512Mbyte if all cpus have the maximum 32Mbyte. Operating system is OS-9. One of the cpus acts as system controller and another looks after i/o capture, signal conditioning and data processing; others can be committed to other functions such as disk control. Since *RAMnet* is compatible with common real-time comms protocols, the system becomes a super-processor interfacing with other remote systems. BVM Ltd. Tel., 01489 783589; fax, 01489 780144.

Optical devices

Bi-colour led. Dialight's *5551-3508* is a flush-mounted two-colour led, providing red and yellow or a mixture to give green. It is in a three-lead, in-line package, the 3mm flat-topped led having a viewing angle of ±50°. Luminous intensity is 4mcd at 20mA. Dialight. Tel., 01638 662317; fax, 01638 560455.

Oscillators

VCXOs. Voltage-controlled crystal oscillators in IQD's *IQVCXO-173* range are designed for use in phase-locked loops working at frequencies in the 1-45MHz range. Pulling is a minimum of ±100ppm for a voltage swing of 4V around 2.5V. Power required is 40mA from 5V and the output of the 14-pin dil devices is compatible with hcmos/ls/ttl devices. IQD Ltd. Tel., 01460 74433; fax, 01460 72578.

PCMCIA oscillator. With a height of under 1.3mm and a footprint of less than 35mm^2, Statek's CX0-M crystal oscillator is designed for use in PCMCIA cards, working from 3V or 5V. Frequency range is 1.25-70MHz, stability is ±100ppm between −40°C and 85° and calibration tolerance options ±0.01%, ±0.1% and ±1%. Tighter specification are available. Advanced Crystal Technology. Tel., 01635 528520; fax, 01635 528443.

Stepper driver. From Nanotec-Electronic, the *IMT 901*, is a driver IC for bipolar, constant-current stepper motors. Supply is 12-40V dc and phase current is settable using fixed resistors up to 2.5A/phase. Switching allows full, half, quarter and eighth-step stepping to give quasi-sinusoidal output, with automatic current boosting in the half-step mode to give about 15% more power. The 56mm diameter, 50mm high package contains the driver, optional oscillator for low and high frequency, motor connector and screened 14-lead cable for power and data. Nanotec-Electronic GmbH. Tel., 00 49 8121 79992; fax, 00 49 8121 79991.

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Digital signal processors

Pixel processors. Using field-programmable gate arrays for flexibility, Sundance has produced configurable, high-performance pixel processors, *SMT308/9*, these first two in the family being meant for use with the TI *TMS320C40* general-purpose dsp, conforming to the *TIM-40* board module standard. They can be configured to carry out low-level video operations, leaving higher-level functions to dsp software. The processors boost the throughput of the *C40* by 'an order of magnitude', since a few gate delays take up much less time than software instructions. *SMT308* digital video board is for high-resolution digital cameras to identify areas of interest in the frame, so that data rates to the video processor are reduced. *SMT309* is a run-length encoder for use as a co-processor to the *C40*, settable upper and lower limits cutting out unwanted data for a higher processor operating speed. Sundance Multiprocessor Technology Ltd. Tel., 01494 431203; fax, 01494 726363.

Programmable logic arrays

40,000-gate fpga. From AT&T comes the *ATT2C40* field-programmable gate array containing 40,000 gates and claimed to be the highest-density fpga on the market. It is a 0.5μm, three-level metal device, available in two speeds having logic cell delays of 5.1ns or 3.8ns, the latter exhibiting a clock-to-output delay of under 11.3ns, a setup time of under 5ns and a zero hold time. In total, 900 programmable logic cells are combined with 3600 registers and 480 user i/os. Up to 57,600bit of user ram and rom are available. Packages include pqfps with 208, 240, 304 and 428 pins and a 428-pin ceramic pga type. AT&T Microelectronics. Tel., 01734 324299; fax, 01734 328148.

COMPUTER

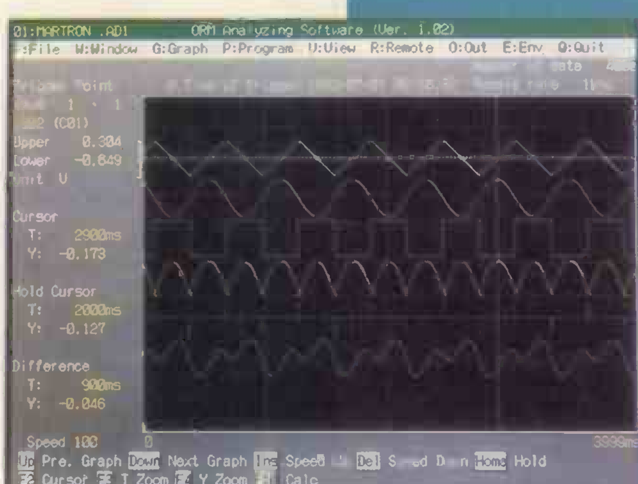
Data acquisition

16 channels for audio and telecomms. LSI has a 16-channel data acquisition board for use in fast processing of audio signals in test equipment, audio compression and telecomms. Carrier board *DBV/DMCB* fitted with up to four *AM/D16QS* daughter modules provides 16, 50kHz, 16-bit-wide channels, data channels being mapped into the system processor memory. The carrier board complies with dBeX32 and can be daisy-chained with up to four other i/o boards. Loughborough Sound Images Ltd. Tel., 01509 634300; fax, 01509 634333.

Data communications

Voice/data multiplexer. *SwitchIT* is a voice and data switching multiplexer by ML Electro-Optics that integrates voice, fax data and lan traffic to be transmitted on one digital line, the voice switching feature allowing a high-quality, multi-site, private voice network to use 64Kb circuits instead of 256Kb or more. *SwitchIT* operates on leased-line, dial-up and frame relay services. M L Electro-Optics Ltd. Tel., 0161 627 1100; fax, 0161 678 0124.

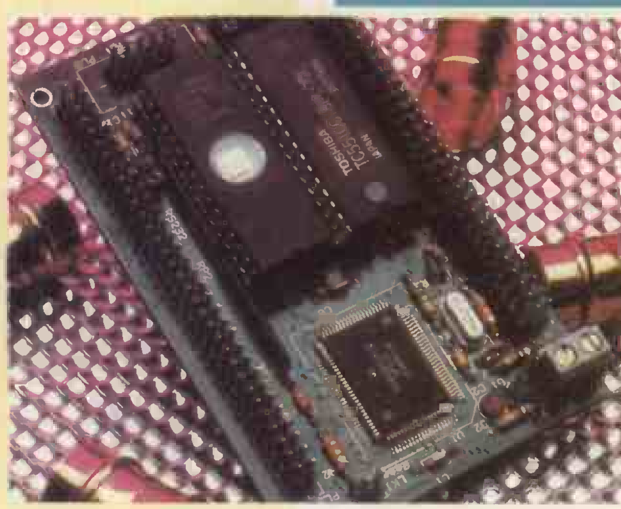
Waveform analysis. *ACRA View* is a software package to enable the analysis of waveforms from Yokogawa's range of oscillographic recorders. Waveforms can be examined and analysed, manipulated, converted to other file formats such as Ascii or Lotus 1-2-3, presented as colour graphics, formats including X/Y, multiframe, overlap and trend, or as digital data. Up to 32 channels can be handled simultaneously. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.



Computer board-level products

16/32-bit controller. CMS's *Micro-Midget* is a 16/32-bit microcontroller for use in 'intelligent' control systems, using an advanced, real-time operating system supporting high-level languages including C. It has up to 22 digital I/o lines, configurable for input or output, a single serial port operating at 38400baud and driving RS-232/485, and two 16-bit timer counters. Its peripheral expansion bus is usable with 68000-compatible devices, 8051 or I²C bus peripherals. Cambridge Microprocessor Systems Ltd. Tel., 01371 875644; fax, 01371 876077.

Thermal modelling. Flomerics announces a thermal modelling service based on its thermal management software package, *Flotherm*, the new service being intended for small companies who need to use the process infrequently to find the cooling needs of a new piece of equipment at the design stage. Simulations produced by *Flotherm* predict the radiative, convective and conductive heat transfer characteristics in a system or component and highlight hot spots, showing air flow, temperature and pressure in the system. Possible problems shown by the simulation are discussed by the customer and modelling team, who will advise on solutions. *Flotherm* models come on disk and allow customers to use their own computers to rotate and analyse the simulation, or can be supplied as animated video recordings. Flomerics Ltd. Tel., 0181 9418810; fax, 0181 9418730.



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PASSIVE

Passive components

Encapsulated transformers. Clairtronic introduces, in its 1995 brochure, a new family of transformers working on the Eurovoltage 230V. Types include thermally protected low-profile units and inherently short-circuit-proof miniature models. All use flame-retardant materials to UL94VO and can be used in products to meet EN60950. Clairtronic Ltd. Tel., 01753 692022; fax, 01753 535096.

SM capacitor arrays. Surface-mounted capacitor arrays in the *MNA* series by Rohm contain two or four components in standard 0805 or 1206 packages, taking up to 45% less space than discrete SM devices. Dielectrics are COG, X7R and Y5V and the capacitors come in values from 11pF to 680nF. Convex terminations allow inspection of the mounting. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

Connectors and cabling

Pcb terminal blocks. Bulgin pcb-mounted terminal blocks are available in two-piece pluggable form or as a single-piece fixed type. They incorporate a rising-cage pressure-clamping feature for low contact resistance and reliability, in which clamps compress the wires and serrations form a gas-tight interface. Bodies are in UL94V-O rated polyamide and contacts in tin-plated phosphor bronze. Ratings for the connectors, in 2 to 24-way form, are 15A at 250Vac for the pluggable type and 210A at 250Vac for the fixed variety. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

Crystals

Ceramic resonators. Coaxial ceramic resonators by Siemens Matsushita are made in standard and miniature versions and are all based on a quarter-wavelength design in high-permittivity material. Frequency coverage is 450MHz-3.5GHz for the standard type, with Q between 250 and 400; the miniature models, which measure 16 by 4 by 4mm, cover 500MHz-4.5GHz at Qs of 200-350. Quantelec Ltd. Tel., 01993 776488; fax, 01993 705415.

Filters

Mains filters. Power filters in Schaffner's *FN402* series are available in both pcb and chassis-mounting form in ratings from 0.5A to 6.5A ac at line voltages to 250V, package size being 28 by 48mm, 16.5mm off the board. Filters comply with American and European safety standards and they are suitable for use in equipment meeting IEC950, with versions available for medical use. Leakage current is 2μA maximum. Schaffner EMC Ltd. Tel., 01734 770070; fax, 01734 792969.

Analogue oscilloscopes. Tektronix's *TekBench* family of instruments is a collection of oscilloscopes, counters, multimeters, power supplies etc. for use in training, production and servicing, the two newest members of which are the *TAS 200* series 20MHz and 50MHz, dual-channel, analogue oscilloscopes having broad similarities to the *TDS* family. A 'user interface' approach to operation is replaced by knobs and switches, although there is on-screen indication of control settings. A useful feature is automatic 50%-level triggering to eliminate trigger adjustment. Tektronix UK Ltd. Tel., 01628 486000; fax, 01628 474799.

Hardware

Fan trays. Intelligent fan trays, as opposed to the dumb kind, are made by Vero to fit into the top of *IMRAK* 19-in racks without reducing space available to equipment. They have four dc fans, an autoranging, universal power supply and a control unit, which uses a thermistor to activate the fans if temperature exceeds 35°C, increasing power until it reaches 55°C. If a fan fails or is blocked, the others speed up, an alarm alarms and an indicator indicates, while a ttl signal initiates a controlled shut-down. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

Screened enclosures.

Measurements on Vero's 3/4U caseframes in the *KM6-EC* range show excellent attenuation of both E and H fields. With no additional screening, the units exhibit H field attenuation of over 30dB from 10kHz to 10MHz and more than 90dB in the E field at 1MHz down to 40dB at 100MHz. Fitting beryllium copper fingers round the edges of front and rear apertures improved the performance to 60dB at 100kHz in the H field and in the E field over 100dB at 1MHz. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

State detector for inductors. Offered by Jensen is the *Mag-Probe*, a completely passive device that indicates whether a solenoid, motor, reed relay or virtually any other component containing a coil is energised or not, with no connection or physical contact. One benefit of this is that equipment might not need to be shut down for test, and another is that there is no magnet present to inflict physical damage on the equipment. *Mag-Probe* is available in standard and high-sensitivity versions, one of them working in almost any case, with either ac or dc. Operation is simple: put the tip of the detector near the coil or a shaft, whereupon a led on the other end lights up if the coil is energised. Jensen's 1995 catalogue has a full description and is now available free. Jensen Tools. Tel., 0800 833246 (free); fax, 01604 785573.

Hand-held enclosures. *Veronex* high-impact polystyrene enclosures come in four plan sizes (90 by 50 by 16mm to 190 by 100 by 60mm) and five heights and are finished internally with nickel acrylic paint to give attenuations of 110dB at 200MHz and 45dB at 1GHz. Belt clips and feet are available. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

Burn-in IC sockets. Yamaichi introduces *IC51*, a family of IC sockets for test and burn-in, accommodating Ics with pin pitches from 1.27mm to 0.5mm with a parallel clamping device to eliminate strain on the pins. Over 10,000 variants handle most types of package, including custom types. Temperature range is –55°C to 170°C. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Equipment enclosures. Cases for equipment used in a wide range of industries are made by Manitron Enclosures in steel, stainless steel, aluminium, polycarbonate and ABS, with interference screening and silk-screen printing if required. Options include machining, painting, anodising, plating and engraving and there is a range of accessories. Specially made enclosures can be made in steel and aluminium. Manitron Enclosures Ltd. Tel., 01270 764171; fax, 01270 763160.

Instrumentation

Earth tester. Avo's *DET6D* is an automatic, three-terminal earth tester for outdoor use. It checks for excessive current and voltage spike resistances and noise to eliminate false results and presents readings on a large lcd. In two automatic ranges, the instrument measures from 100mΩ to 2kΩ. Spike kits for use with two or three terminal, complete with cables, cable winders and clips, are offered. Avo International Ltd. Tel., 01304 202620; fax, 01304 207342.

Frequency standard. *FS 700* frequency standard from Stanford Research Systems uses Loran signals, transmitted for navigation and traceable to caesium clocks, to give a long-term stability of 1 part in 10^{12}. *FS 700* uses timing data from the signals to lock its own oscillator to provide a 10MHz output in the form of a ttl-compatible signal adjustable in frequency between 0.01Hz and 10MHz in a 1:2.5:5 sequence. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Oscilloscope isolation amplifier. Vann Draper's *H5001* single-channel isolation amplifier allows oscilloscopes to look at equipment such as motor controls, switched-mode power supplies and power semiconductor circuitry without the need to remove the instrument's earth connection. It employs optical and transformer techniques to handle signals up to 2kV from earth, using standard oscilloscope probes. Plastics are used in the casing and controls of the amplifier and emi and rfi protection is provided. Vann Draper Electronics Ltd. Tel., 0116 2813091; fax, 0116 2570893

Temperature controller. Brainchild announces the *BTC-404* 1/4DIN analogue temperature controller, which works with J or K thermocouples to cover the 0-1200°C range. Proportional or on-off control outputs are offered, to an accuracy of ±1% of span, the on-off control coming from a 10A relay. Alternatively, a pulsed voltage output drives a solid-state relay, or there is a 4-20mA linear loop or a 0-10V linear output. Indication is by led digital readout and the unit occupies only 53mm behind the panel. Brainchild Temperature Controllers Ltd. Tel., 01903 216514; fax, 01903 219662.

Go/no-go tester for GSM. *CTD 52* by Rohde & Schwarz carries out rapid testing on GSM mobiles, offering all the characteristics of a GSM base station. On one keystroke, it will indicate pass or fail on selectable rf channels, measuring power and sensitivity and carrying out an echo test to verify the mobile's operation. It should appeal to retailers, who are enabled to perform fast checks on cellphones to determine whether the mobile is at fault or malfunction is due to another cause. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

Cameras

Oem ccd. Sony has produced the *NDP-40BYE*, a low-cost, monochrome, ccd camera in board form, designed for conferencing, cashpoints and machine vision. It measures 40 by 40 by 26.5mm and gives a resolution of 380 tv lines from a 500 by 582 pixel format, with automatic exposure control and a 0.02s to 10µs shutter speed. Power needed is 120mA at 12Vdc; minimum sensitivity 0.3lux; and composite video output 1V pk-pk into 75Ω. Standard lens has a 3.6mm focal length, although other types are available. Sony Computer Peripherals & Components. Tel., 01932 816000; fax, 01932 817001.

Energy monitoring. Seltek's *UPM6000* portable energy monitor, which is supplied with software, has a 80-260V AV input range, is usable with three-wire or four-wire systems and reads V, A, VAR, kW, PF, Hz, kVArh and min/max values; it also gives harmonic analysis up to the 25th for voltage and current. Software allows 12 channels to be configured with simultaneous voltage and current harmonics and up to 30 trend graphs of any of the measured parameters can be displayed on a pc. Seltek Instruments Ltd. Tel., 01920 871094; fax, 01920 871853.

Literature

Livingston Hire. For 1995, Livingston is computerised. Its catalogue, that is. It is free, comes on disk, runs under Windows and asks the relevant questions to lead one to the very instrument for the job in hand; there is even an order form to fill in and print out. Provided you can answer the questions honestly and not say that the job will be done by next Tuesday, when the rented instrument will probably be gathering dust for at least three months before the job starts, the Decision maker will tell you whether renting or buying is a better bet. Call free on 0800 88 6000. Livingston Hire Ltd. Tel., 0181 943 5151; fax, 0181 977 6431.

Enclosure catalogue. Vero offers a 68-page technical catalogue on *IMRAK 400* wall-mounted enclosures and *IMRAK 1400* screened, free-standing racks and cabinets, together with all the bits and pieces such as chassis supports, panels and other accessories. Also in the catalogue are descriptions of pre-configured telecomms wiring closets and racks, patch panels, cable management products and specialised products for use with optical fibres. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

Tools and test gear. A new 66-page catalogue from Jensen of Phoenix, Arizona, has tools and test gear for work with all types of electronic equipment, including telephones, lans and computers. Tools range from tweezers to power tools and comprehensive toolkits and test gear from people such as Fluke, Tektronix, Wavetek and Microtest, ranging from pocket test meters to digital storage oscilloscopes. Free from Jensen Tools. Tel., 0800 833246 (free); fax, 01604 785573.

Materials

Static-shielding bag. 3M's *Model 2750* re-usable shielding bag consists of stainless steel microfibres suspended in 7mil of antistatic LDPE which, although much thicker than the common metallised gossamer layers, is more transparent. The bags are effective in relative humidities up to 10% and retain their properties for up to five years. 3M United Kingdom plc. Tel., 01344 858000; fax, 01344 858758.

Printers and controllers

Printer sharing. Swiss firm Rotronic has introduced their ICE printer-sharing switchers through Interconnections Ltd, recently taken over. Four models allow either printer sharing between two or four computers or computer sharing between two or four printers. Switching is either manual or by a tsr program, which scans the system for requests after eight inactive seconds. Power comes from the computer. Interconnections Ltd. Tel., 01293 822781; fax, 01293 822786.

Thermal line printers. Two new battery-powered printers by Fujitsu, the *FTP-623/633*, are for paper widths of 2in and 3in respectively, both being powered by NiCd, Ni-MH or lithium-ion packs at voltages between 4.2V and 8.5V. The thermal line dot system, said to be an improvement on the wire dot type, allows printing speeds to 480 dotlines/s with a density of 8dot/mm. They are available as mechanisms or complete with interface board or with microcontroller and gate array for complete driver control. Fujitsu Microelectronics Ltd. Tel., 01628 76100; fax, 01628 781484.

Production equipment

Programming in production. Data I/O has introduced an automated handling and programming equipment for production runs. The *ProMaster 7500* uses two of the company's *AutoSite* programmers in an automatic handler, the whole turning out tested, sorted and laser-marked devices in both surface-mounting and dil packages with up to 84 pins. Pick-and-place heads in the handler can rotate devices, so that work goes on without regard to orientation. Data I/O Ltd. Tel., 01734 440011; fax, 01734 448700.

Time-code receiver. Temic *U4224B* and *U4221B* bipolar, straight-through time-code receivers are intended for use in radio-controlled clocks, receiving the time-code data from the National Physical Laboratory at Rugby. Data produced by the receiver can be used as a real-time clock for cordless equipment such as portable telephones, camcorders and equipment for navigation and security. Signals from the 50kW NPL transmitter are at 60kHz and are binary-coded decimal in form to give time, date and year(!) information. *U4224B* accepts 1.2-5.25V at 30µA (1µA asleep) and a setup time of 2s, while *U4221B* takes 2.4-5.5V at 40µA (0.2µA) and sets up in 2.5s. Macro Group. Tel., 01628 604383; fax, 01628 666873/668071.

Plastics assembly. *Miniprobe* by Kelly Ultrasonics is a lightweight, ultrasonic plastics assembly tool for use in swaging, staking and welding. It looks a little like a small soldering iron and, since the tool itself is not heated, heating of the workpiece is very localised, the operation and cooling being so fast that little thermal energy is transferred. Welwyn Tool Co. Ltd. Tel., 01707 331111; fax, 01707 372175.

Power supplies

Ac voltage stabiliser. An automatic ac voltage stabiliser, the Gardners *AVSU*, smooths out extended voltage fluctuations and brown-outs in the mains supply. It copes with inputs up to 20% down on nominals of 220V or 110V, automatically selected, and provides over-voltage suppression and rfi filtering. Rated power of 2kVA can be taken from either the 220V or 110V outputs, which are both present, or shared between them. Further units offer ratings of 500VA, 1kVA and 4kVA. Gardners Ltd. Tel., 01202 482284; fax, 01202 470805.

Battery monitor. Opalport Electronics offers a software-based battery monitor, the *BA319* in a 3U 19in rack, which observes the current health and predicted behaviour of up to 400 cells, with an option up to 700. It presents a detailed review of every aspect of cell performance and prospects for the future. Opalport Electronics Ltd. Tel., 01249 758161; fax, 01249 750626.

Wide range – two sizes. Instead of the usual afterthought approach to psu design, in which the space left available is invariably far too small, the method adopted in the Vicor range is to offer over 11,000 different combinations of ac/dc input and dc output in only two sizes of baseplate-cooled psu – 25-100W in 58 by 61 by 13mm and 50-200W in 117 by 61 by 13mm. Requirement changes can be coped with by simply using a different module or trimming the output for smaller changes. XP plc. Tel., 01734 841010; fax, 01734 843423.

High-power converter. *HPDU40* by Gardners is a 5V, 8A, low-profile dc-to-dc converter for mounting in a standard Euro-rack or similar enclosure. A typical use, says Gardners, would be as a standby to maintain power to a bank of memory cards during maintenance work, in which case any dc input from batteries to 24V/48V/72V dc supplies and rectified ac can be used, without too much emphasis placed on stability of the ac. Non-polarised input terminals avoid disaster from reversed connections. Shielding and filtering against rfi at both ends is provided. Gardners Ltd. Tel., 01202 482284; fax, 01202 470805.

European power supply. EAO Highland's new *EcoPower* psu is, the company believes, the first to carry the European EMC mark, CE. It is an industrial unit, producing 24V, 5A dc from a 115-230V 50Hz ac lightning-protected input and also a safety extra low voltage (selv) output. The unit functions as mains filter, meeting Class B specifications on rf interference. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Radio communications products

Power amplifiers. ENI announces the introduction of the *Models 604L/607L* rf broadband power amplifiers covering the frequency ranges 500kHz-1GHz and 800kHz-1GHz respectively, producing linear outputs of 4W and 7W. Any load vswr from open-circuit to short-circuit is acceptable without damage and the units meet the usual rfi/emi standards. Holaday Industries. Tel., 01628 478155; fax, 01628 476871.

QSPK demodulator. *TDA8040* and *TDA8041* from Philips form a two-chip, fully integrated demodulator for quadrature phase-shift keyed signals used in digital video broadcast and digital telephony and performs all analogue and digital functions. The *8040* demodulator takes if up to 150MHz and outputs I and Q signals, which are then digitised in the *8041* controller to recover symbol clock and decode the symbols, obtain afc and agc. Phase error is less than 0.5°. Only a tank circuit for the vco and a pair of varicaps are needed externally, an on-chip voltage stabiliser being provided. Philips Semiconductors (Eindhoven). Tel., 00 31 40 722091; fax, 00 31 40 724825.

Switches and relays

Power switches. From BLP Components, the *Series 32 Powerpulse* relay, designed to meet BS3676 Part 1 on lighting requirements and now confirmed in its suitability for use with fluorescent lighting up to 30A, part of the testing being equivalent to switching a 5ft tube 50,000 times. The switch handles relative humidity to 95%, temperatures to over 70°C and voltage to 758V ac. BLP Components Ltd. Tel., 01638 665161; fax, 01638 660718.

Black keyboard. *Model 1800 Black Magic* by Cherry is a 19-in keyboard in the 'in' colours for computing equipment, black and grey, joining the standard 19-in model *3000*. It is intended for banks and checkouts, or anywhere where size is important, and can be mounted in a 19-in rack. Models with 101 or 102 keys are available. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

Sealed keypads. Silicone-rubber, sealed keypads in the *Series 84LS* range by Grayhill are but 9.07mm thick overall and come in 12-key and 16-key versions. Contact resistance is 10Ω maximum and is compatible with mos, ttl and dtl circuitry. A choice of matrix or single-pole/common-bus circuits is on offer and shielding is optional. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Transducers and sensors

Control-shaft encoder. For the accurate adjustment of medical and measuring equipment, Panasonic's optical encoders offer extremely long life and small size, having a height of 8mm. The encoders simply push onto the shaft, one version being provided with 31 leds to indicate position. The devices provide 40 pulses per revolution and operate from 5V. Panasonic Industrial (Europe) Ltd. Tel., 01344 853827; fax, 01344 853803.

Tilt sensors. Absolute inclinometers by Control Transducers in the *A-ID* range use optical encoders with a unique, rather than incremental code. From one to 15 encoders may be networked on a six-wire cable at distances of 330m, to an RS-232 port, for which an interface is provided. From 2 to 65536 codes per revolution are available with 9-bit or 12-bit accuracy over 360° at rates from 38.4 to 115.2kbaud. Control Transducers. Tel., 01234 217704; fax, 01234 217083. ■

DATA RATER Continued from page 482

Spectral behaviour

LM4861's spectra into 8Ω are the highest **(Fig. 21)**. The second harmonic dominates numerically as might be expected from the thd residue. There are plenty of high level, high order harmonics too, but at least the evens dominate.

Spectra for the *4861* at 18dB below clip is quite different **(Fig. 22)**. The second harmonic has not reduced in proportion though the higher harmonics have increased – suggesting fatiguing sonics.

Unfortunately, the *LM4860* spectra demonstrated an unexplained noise problem **(Fig. 23)**: a faulty ic is assumed, although %thd is within spec. Only one sample was provided and a replacement was not available in time, but the spectral pattern is recognisably as *LM4861*.

At the onset of clip, most of the *LM12* products are just below 100dB **(Fig. 24)** – note the dominant third harmonic, while 25dB below clip **(Fig. 25)**, the noise floor has increased. Harmonics poking above the averaged noise are just 2nd, 3rd, 4th and 6th – not unpleasant. Harmonics do not change outside the certainty limits (±2.5dB) when ripple is considerably raised from about 15mV to 350mV by resistive upstream rail loading.

LM3875 spectra just below clip are similar **(Fig. 26)**, yet not quite like LM12: the high order even harmonics are consistently slightly higher. At –25dB below clip, the spectra **(Fig. 27)** are very like *LM12* under the similar conditions. *LM3886* performance was also very similar, but more like the *LM12* at high levels, while at –25dB down, the 3rd, 5th and 7th are stronger and replace the 4th and 6th. This small but possibly crucial difference is conceivably a consequence of the VCA type elements employed for muting.

At just below clip, the spectra of the *PA42* and added mos output stage is ragged though nearly all below –90dBr **(Fig. 28)**. But at 26dB down, affecting most listening, the only harmonic readable above the averaged noise floor is a tiny amount of second **(Fig. 29)**.

I suggest this amplifier will sound very natural when used with efficient loudspeakers. 800mV of ripple was provoked and had no effect, demonstrating solid psr from the harmonic perspective.

However, the *PA45* mainly makes odd harmonics **(Fig. 30)**. Worse, the high harmonics are almost level with the low order ones and will be very prominent by ear. At –25dB below clip, while cleaner, odd harmonics still dominate the evens.

With the supply raised to +62V for the *PA45* (still 20% below maximum) all the harmonics are reduced **(Fig. 31)** and the higher harmonics are particularly suppressed. Under these conditions, excellent sonics should result.

Moral: Those who believe mosfets have high %thd may be using them wrongly.

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Acknowledgment

The author would like to acknowledge the assistance given by Audio Synthesis.

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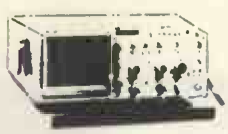

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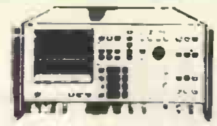

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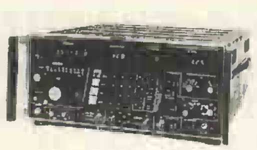

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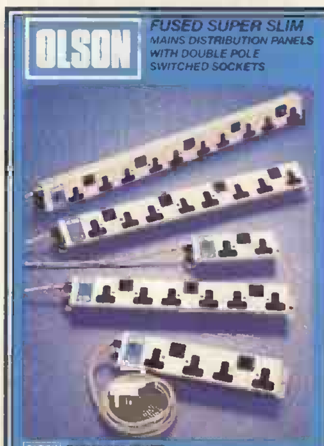

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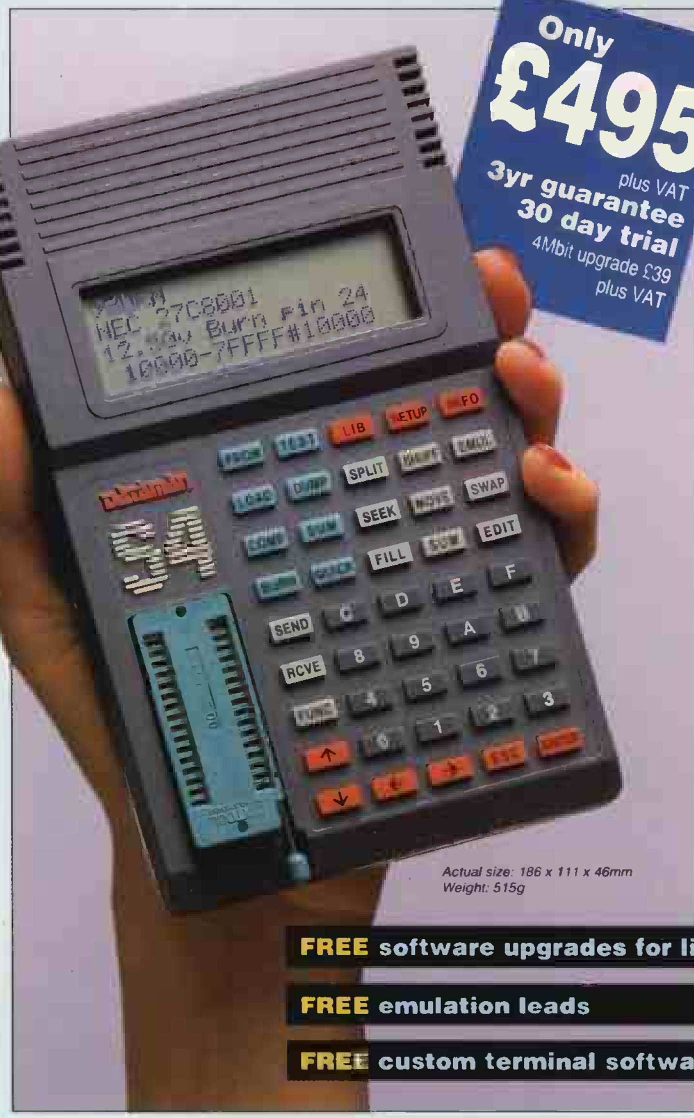

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