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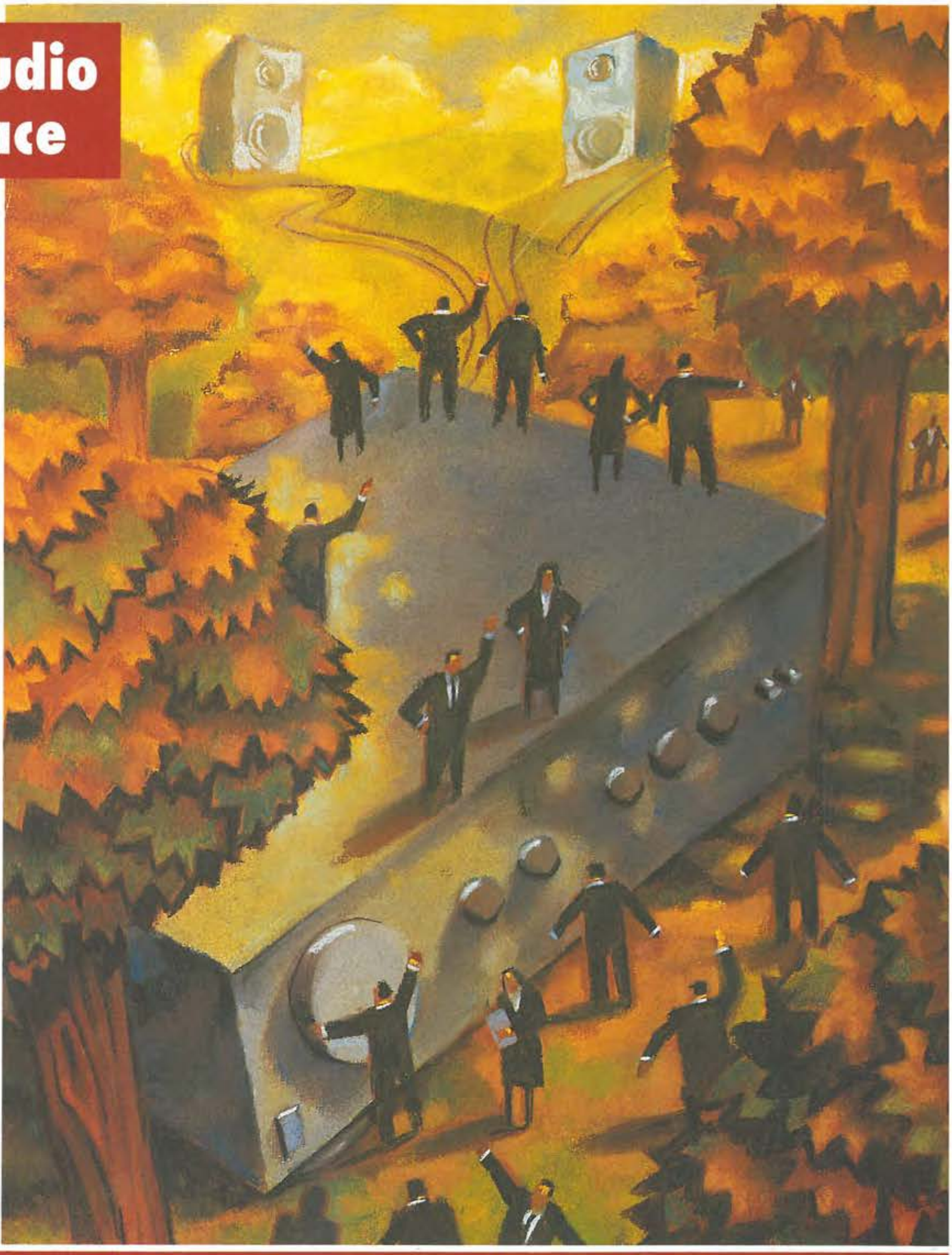

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Switched capacitor psus

Microammeter

Digital tv explained

MSF clock interface

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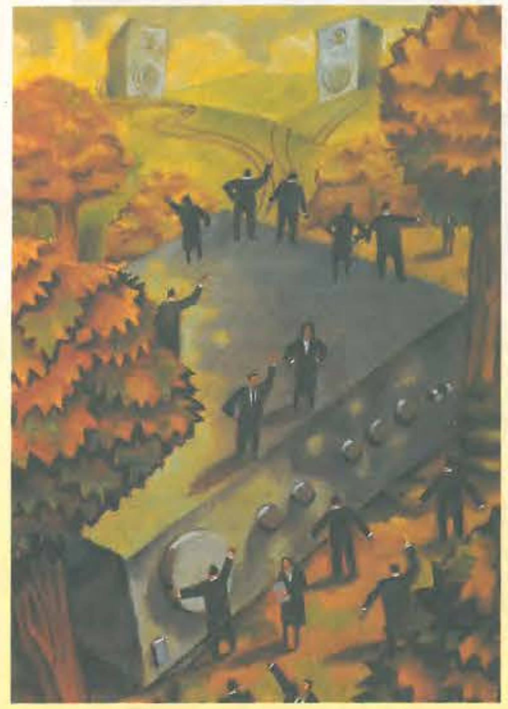

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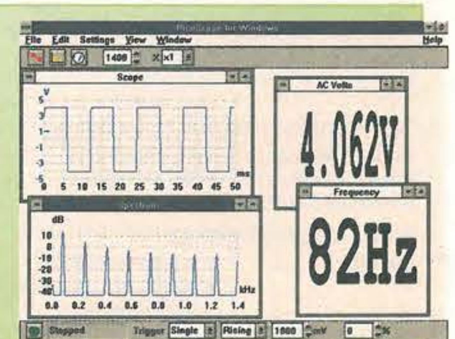

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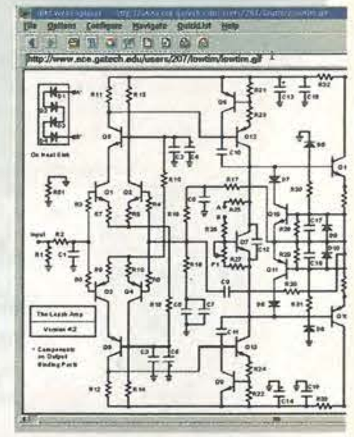

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OCTOBER ISSUE ON SALE 4 SEPTEMBER

History: who needs it?

BT is justifiably proud of its sensitive attitude towards ecological issues, sponsorship of the arts and its community programmes. All the more remarkable then that the company's regard for its own heritage is zero. For by the time you read these words, its highly regarded museum in London, The Story of Telecommunications, will have closed.

In a year that marks the 150th anniversary of the birth of Alexander Graham Bell, this seems a strange kind of celebration.

The collection will remain, but will no longer be open to visitors, and the same applies to the resource centre used by many researchers. BT says the closure is made pending a decision on its heritage policy – a subject of internal debate for the past 12 years without decision. Consultants have come, consultants have gone and still BT has no firm line on how to handle its history.

In reality, the closure is a simple matter of cash. Not a single department within BT is prepared individually to fund the cost of running what could – and should – be a showpiece facility. Hence the museum has been forced to shut its doors, simply to save five clerical salaries amounting to around £100,000 a year – or more graphically, 16 minutes of BT's annual profit.

For this trifling amount, BT is prepared to close down a facility that attracts 21 000 visitors annually – many of these youngsters on educational trips – and disperse to the four winds the public goodwill and accumulated knowledge and expertise of its staff.

It seems Philistine, but this is the same company which, notwithstanding the valiant efforts of the Computer Conservation Society in reconstructing Colossus at the wartime codebreaking headquarters at Bletchley Park, is intent on selling the site for development rather than dedicate it as a national museum of information technology. BT needs to make friends – not enemies – yet the company appears oblivious to the way it is alienating public opinion.

Earlier this year, the GEC company was forced to back down on its planned auction sale of historic Marconi artefacts, when the weight of informed opinion made this policy untenable.

BT's own heritage of what is effectively the national collection of communications is far larger in extent, and embraces an even wider rapport with everyday life, yet it is to be locked up and withdrawn from public gaze. Can this really be termed a responsible action for a major public company?

Had BT offered to donate its collection to the Science Museum, this would have been another matter. If it argued that a national museum of communications should be established on an independent basis – with the offer of some funding – this would be greeted with sympathy and understanding.

> ***...the museum has been forced to shut its doors, simply to save 16 minutes worth of BT's annual profit.***

But the abrupt closure of an established facility purely to satisfy the accountants smacks of putting shareholders' gain before public responsibility.

There is now growing support for creating a National Museum of Information Technology, embracing communications, electronics and computing, with the Bletchley Park campus as the favoured site. Setting past indecision aside, now is the time to build on this momentum and create something for the new millennium. Otherwise our future generations will have to be content with photographs and fading memories.

As the descendant of the nineteenth-century telegraph companies, the old GPO telephones department and more, BT has a clear responsibility to the country to provide a facility for conserving its heritage. This task can be shared with other operators and manufacturers in this booming industry – and the time to start is now.

Ironically, it is one of Britain's newest and smallest telecomms operators which is showing the way. This year ScottishTelecom opened its World of Telecommunications museum in Edinburgh. It is ironic that BT's closure of its London museum has made this new venture the country's premier telecomms visitor centre. ■

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GPS is on the market

GEC-Plessey Semiconductors (GPS) is up for sale. GEC managing director, George Simpson, told a stockbrokers' analysts' meeting recently that he was looking to dispose of the UK's last fully integrated microelectronics manufacturer.

It is assumed that GPS and GEC were implementing the disposal before Simpson told analysts. However, GEC refused to make any official comment on the situation other than to say: "We have not announced any disposals."

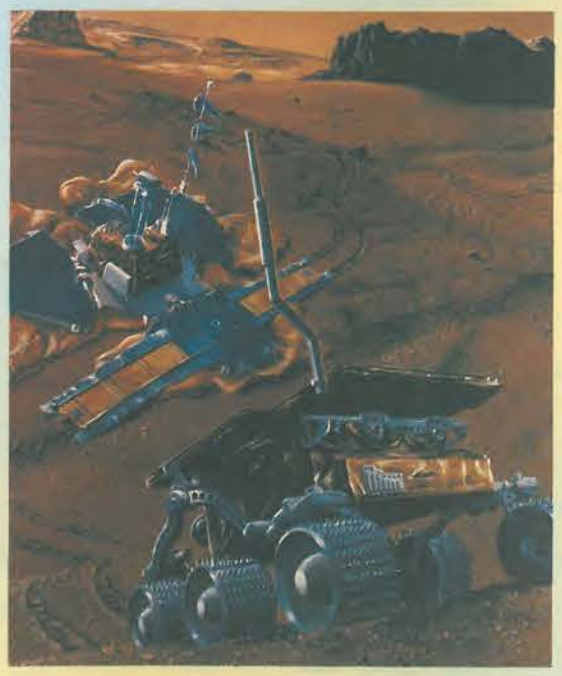
In a written statement, GPS managing director Tom Urwin said: "GPS is now actively seeking a suitable partner." GPS appears to look on the move as an opportunity to find an owner which will invest in it appropriately. Under GEC's ownership, investment constraints prevented GPS from maintaining leading edge production process technology.

Obvious candidates for ownership of GPS are Rockwell Semiconductor of the US and Siemens of Germany. "GEC-Plessey has some attractive positions – they have a very good bipolar capability," Dr Dwight Decker, president of Rockwell, said last autumn. "What we're looking for is mixed signal technology with a broad application range. There are good reasons why we'd like to be geographically distributed." Rockwell talked to GEC last year about buying GPS.

Siemens Semiconductor owned briefly a share of GPS following the GEC-Siemens takeover bid for Plessey. Jurgen Knorr, president of Siemens Semiconductor said at the time: "We wanted GPS; we were interested in it years before."

If those two fail to buy GPS, then there is a possibility that one of Korean Big Three – which want to diversify out of dynamic ram – could buy GPS for its logic and system asic capabilities.

GPS is ninth in the world for gate array logic and has an unrivalled capability for putting systems-on-a-chip – expected to be the hottest growth area in the semiconductor business for the next decade.



Independence Day... Nasa's Jet Propulsion Laboratories' (JPL) latest probe, the Mars Pathfinder, is set to land on the planet on 4 July. It comprises the Lander which will oversee the various experiments and communicate the results back to Earth, and the Rover vehicle which will explore the planet surface, taking pictures as it goes. To cut the project's cost, the Mars Pathfinder has made use of commercial-off-the-shelf components including VMEbus cards and Wind River's VxWorks real time operating system (RTOS). The RTOS is used to control the probe's flight and planet descent – triggering the airbags before it hits the planet's surface – and the Rover vehicle.

OTP memory has 60ns access

Philips Semiconductors has introduced a fast 60ns one-time-programmable, or otp, memory process for its microcontroller products. One-time-programmable memory made by this process is now included in Philips *XA* 16-bit and *80C51* 8-bit microcontrollers.

Developed for Philips Semiconductors by Philips Research Labs, the new OTP is claimed to incorporate three innovations.

The first is that it takes only two additional mask steps compared with its standard 0.5µm c-mos process.

Secondly, it avoids the use of mask-intensive high voltage (13V) transistors to switch the programming voltage. Instead, it exploits the fact that the transistors only need to survive long enough to program the memory matrix once and uses low voltage transistors that need no extra masks.

Including a conductivity-improving titanium salicide layer between the metallisation and control gate is the third development.

Philips has extended the control gate sideways so that it forms a top hat over the floating gate. This shields the control gate from contact with the salicide.

Each memory cell measures 7µm^2 and can operate between 1.8 and 3.3V. Blocks of up to 4MByte can be created at a time.

The company claims with the conventional vertical stack of equally sized floating gate-insulator-control gate, any salicide deposited on top of the stack also gets on to the stack sides and shorts out the two gates.

UK research spending worsens

The proportion of UK electronics companies' sales turnover spent on research and development fell last year, according to the DTI's *UK R&D Scoreboard 1997*.

Although r&d expenditure went up two per cent to £1.204bn, as a percentage of sales it fell by 0.1 per cent to just 3.1 per cent of sales turnover. This fall was not as great as that in the previous year, however, when it fell from 4.2 to 3.2 per cent.

"Too many of our companies appear to lag behind the competition," said science, energy and industry minister John Battle, commenting on the scoreboard. "Evidently those companies may be under-investing in absolute amounts and in terms of research and development intensity [percentage of sales spent on r&d] when compared with their top international competitors. This could have serious implications for the long term."

1996 Rank	Company	R&D spend	% of sales	1995 Rank	Company	R&D spend	% of sales
1	GEC	£432m	6.9	1	GEC	£412m	7.1
2	Siemens	£133.6m	8.9	2	ICL	£149.6m	4.8
3	IBM	£112m	2.2	3	Siemens	£132.6m	10.3
4	ICL	£101.3m	3.5	4	IBM	£100.7m	2.2
5	Racal Elec	£70.8m	6.7	5	Racal Elec.	£64.7m	6.8
6	Motorola	£36.7m	1.2	6	Motorola	£36.7m	1.2
7	HP	£31.9m	1.7	7	BICC	£31m	0.7
8	BICC	£30m	0.7	8	HP	£29.8m	1.7
9	Bowthorpe	£23.5m	4.5	9	Rank Xerox	£21m	0.5
10	Rank Xerox	£20m	0.6	10	Cray Elec.	£20.3m	7.7

GEC maintained its position at number one spending £432m, increasing its expenditure by five per cent on the previous year. Siemens unseated ICL, whose research and development expenditure dropped 32 per cent, and moved to number two with a spend of £133.6m. IBM, ICL and Racal came third, fourth and fifth respectively.

Despite wanting more investment, Battle called for greater focus on the management of r&d. "Of course, devoting simply more money to r&d is not the only answer to ensuring the long term competitiveness of UK companies," said Battle. "Investment needs to be managed carefully so that it leads to successful development of new ideas."

CDs allow audio recording

Philips Electronics says it will introduce one of the first audio cd recorders by the year end. The cd recorder is based on Philips' recordable cd technology and uses cd recordable disks that allow only one recording or cd rewritable disks that can be recorded and erased several times.

The cd recorder will be introduced in Europe first and then into markets worldwide. It is expected to cost £450. However, the market for the recorder may be limited to just two or three years especially once recordable DVD systems become available in 1999.

Recordable DVD will be able to record audio and video sources and could also be used as a computer peripheral, recording computer data.

The recorder will implement copy protection agreements such as the Serial Copy Management System.

DVD – slow take off?

DVD players for television sets will penetrate less than two per cent of homes during the next five years, says a study from Strategy Analytics, an electronics and IT consultancy. The report claims that less than half a per cent of European homes will have DVD by the year 2000. Worldwide, DVD will only become accepted when recordable DVD becomes available, thereby replacing VHS video recorders. Annual sales are expected to reach 10m units by 2010.

Actel £1000 fpga design tools is now free on a Web page

Actel has made a version of its fpga design tools free for distribution over the Web, saving designers over £1000.

"Customers have asked for many years why they have to pay for this software," said Vaughan Price, European sales director. "We've taken the big step. This is something that has to be done."

Actel is the first company to make its design software for fpgas free, claimed Price. Altera is expected to follow suit, later this year.

The Designer Lite software includes all the features needed for fpga design. The only drawback is the tool is limited to designs of 8000 gate or fewer.

"Below 8000 gates, people look at fpgas for first choice in logic integration. We want to remove all the obstacles," said Price. At these gate counts, designers will save up to $2500 on the cost of tools. "But the real saving," he said, "is in the time and the convenience."

Included in the software is place and route, timing driven layout and analysis, and Verilog and Vital libraries. By the month end a VHDL synthesis tool will be added to the kit.

Designer Lite is available from www.actel.com.

Europeans show interest in buying digital radio

Over a third of households in Europe are interested in buying digital radios. This is one of the results of a consumer study conducted by RSL/Ipsos for the World DAB Forum.

The survey was undertaken across six European countries with the participation of 5000 people. More than a third of those sampled who are also car drivers confirmed their interest in buying a DAB car radio due to its better reception.

A similar level of interest was shown for hi-fi sets and portable radios. pc-linked DAB services also appealed to special interest groups who use IT systems and pcs on a regular basis. The respondents were prepared to pay up to 50 per cent more for car radios and hi-fi receivers featuring DAB features, and twice as much for portables.

This was the largest ever consumer study of potential DAB penetration in Europe.

SiGe device costs could be cut by new process

An Italian research group is attempting to reduce the cost of silicon-germanium devices by using ion implantation for their fabrication.

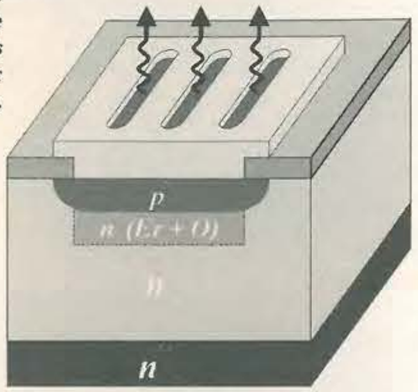
SiGe heterojunction bipolar transistors, or hbts, are faster than silicon-based ones, letting devices operate at 5GHz without resorting to GaAs technology.

The group, called IMETEM, comprises 15 people from the University of Catania. It resides at SGS-Thomson Microelectronics' wafer fab and research centre in Catania.

The group claims that existing techniques for producing SiGe structures, like molecular beam epitaxy (MBE) and chemical vapour deposition (CVD), although successful, are difficult to incorporate into current ULSI technology. This, it says, is delaying the adoption of SiGe commercially.

Its proposed approach for fabricating SiGe heterojunctions is a technique based on high dose germanium ion implantation into silicon. Ion implantation offers a high wafer throughput and can be confined to particular regions of the wafer.

Other IMETEM devices under development include an infra-red silicon led (see diagram). Silicon is not normally suitable for leds but, by including erbium-oxygen complexes in the material lattice near the PN junction, IMETEM has created an emitter that operates at 1.54μm, the fibre-optic communication wavelength. The group is also starting to look at silicon carbide-based semiconductors for high voltage use.



Silicon is not normally suitable for leds, but a new SiGe process from Italy could be used to produce leds for fibre-optic communications.

New ideas for multi-media lans

US firms Microsoft, 3Com and videoconferencing specialist PictureTel are to collaborate to enable local area networks (LANs) to carry moving pictures and voice as well as data.

This, they believe, will "crack the technological barriers" which so far have prevented video communications to every office desktop.

To do this they have brought together PictureTel software – built around LiveManager 3.0, which is designed to handle continuous video and voice traffic without affecting other critical LANtraffic – with Microsoft's NetMeeting 2.0 software and 3Com's PACE high-performance LAN technology.

Meanwhile, Sony Broadcast & Professional UK has signed an ISDN reseller agreement with BT.

This is aimed at reducing the cost of purchasing systems by allowing Sony to sell a combined package of videoconferencing equipment and ISDN.

New c-mos process promises 2.4GHz devices

Texas Instruments has detailed a 0.18μm c-mos process at a recent technology symposium in Japan. This is a follow on technology to its 0.21μm TImeline process announced last year.

With 25 picosecond delays and 2.4GHz analogue speeds, the company is targeting the 0.18μm process at wireless and optical communications. Radio and intermediate-frequency circuits, along with dsp baseband processing could be placed on the same device, claims Richard Kerslake, TI's Asic marketing manager. The 2.4Gbit/s data rate is aimed at the new STM16 fibre optics communications standard.

"What would be really beautiful," said Kerslake, "would be to run this [STM16] straight into c-mos without using anything horrible like GaAs."

For the process, TI has switched to a shallow trench isolation. This, coupled with the reduced process dimensions doubles the available transistors.

While few, if any, designs will need 250m transistors, halving the die size for a given design could bring significant cost reductions.

The as yet experimental process is to move to a production standard in the near future. Commercial use for the process is not expected for two years, but Kerslake expects design starts by the end of next year.

Most manufacturers – TI included – have yet to move to full production on 0.25μm geometries.

TI's own TImeline process, actually 0.21μm drawn, is only now ramping to production on microprocessors. Asics will follow shortly, said Kerslake.

Traffic info for mobile phones

Trafficmaster, the traffic-guidance system firm, is to launch a service in the autumn which will provide mobile phone users with details of road hold-ups and delays.

The traffic firm would not disclose details of its proposed system.

However, it is expected that small in-car receivers will warn drivers of nearby traffic problems, prompting them to call a hotline for further details.

A spokesperson for Trafficmaster said:"We have recently concluded a three-year deal with a mobile phone operator." However, she could not confirm the operator's identity

Reports that the deal is with Cellnet were branded as hearsay by both companies. Dave Massey of Cellnet's corporate affairs said:"From our point of view this is speculation."

Altera and Jam

Jam, a programming language for in-system programmable logic devices, has been unveiled by Altera.

The company claims JAM more than halves file sizes, while reducing the average device programming time by a factor of ten. Altera and its supporters are proposing the language goes to JEDEC for adoption as an industry standard.

Programmable logic manufacturers Cypress Semiconductor and Vantis, test equipment companies GenRad, JTAG and Teradyne, and device programmer manufacturers Data I/O and BP Microsystems all support the language. ■

RESEARCH NOTES

Jonathan Campbell

"Second silicon revolution" gets commercial fillip

Traditionally, the focus of the microelectronics industry has been to pack more transistors onto a chip, leading to more powerful computers. But micromachine technology developed by Sandia National Laboratories and being commercialised by Analog Devices could change that strategy by working to integrate electronics with small machines into usable devices.

Such chips will have the ability to sense where they are and what is going on around them, with an early application likely to be in automobile air bag sensors.

A three-axis accelerometer that can detect changes in velocity along three planes of motion has already been built using Sandia technology and designs developed by the University of California at Berkeley's Sensor and Actuator Lab.

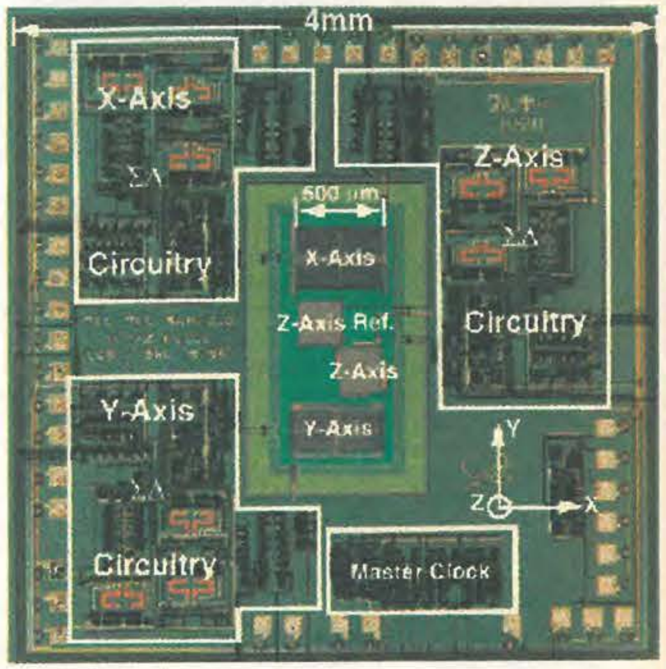
Analog Devices, an industry leader in the manufacture of air bag micromachine sensors, is taking the technology to make integrated micro machines – a tiny, smart machine that combines microcircuits, sensors, and actuators on a single computer chip. In the air bag sensor, Analog uses micro machines to signal when a vehicle is undergoing sufficiently rapid and sustained deceleration for the air bag to deploy. Sandia is already a world leader in this field and the work is expected to help stimulate production of a new generation of very small consumer and military devices, such as anti-theft, anti-skid, and active-vibration-control systems.

Sandia estimates the specific market for micromachine-based inertial sensors worldwide is estimated to be $3.8 billion, and talks in terms of these new devices heralding a "second silicon revolution."

Researchers at the University of California at Berkeley's Sensor and Actuator Lab, credited with making some of the earliest known micro machines, also will be involved in designing new, smarter products.

One advantage of the technology is that batches of silicon micro machines can be fabricated through manufacturing techniques already widely used to make integrated circuits. So micro machines are far cheaper than the complicated multi-metal constructions originally necessary to signal an air bag to inflate. Rather than being made individually, micro machines can be fabricated quickly and cheaply by the thousands.

Also, because the machines have so little weight, they are less likely to be damaged by sudden deceleration, because force is proportional to mass – which, in this case, is almost non-existent.



Micromachines integrated into a chip promise the next silicon revolution

DNA computers – made on the workbench

Manufacture of biological logic gates has brought forward the possibility of building DNA computers able to process in just a few hours complex information that would take today's conventional computers hundreds of years to solve.

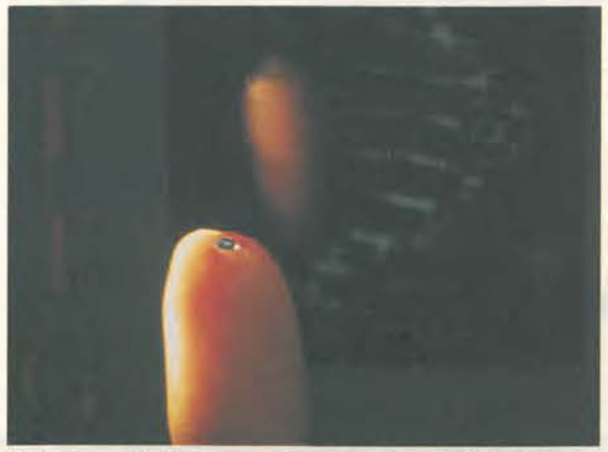
One of the most surprising aspects of this recent breakthrough at University of Rochester is that the DNA logic gates were made using only the most common-place biological laboratory techniques, such as DNA ligation and gel electrophoresis

Conventional logic gates convert the endless series of binary data coursing through the machine into a series of signals that the computer uses to perform its operations. They process electronic signals from transistors, converting two input signals into one output signal in a way that allows a computer to perform complex operations.

Up to now, the only logic gates used for computing have been electronic structures that detect signals coming from transistors.

But the logic gates, developed by Animesh Ray and Mitsu Ogihara at Rochester, rely not on electrical signals to perform logical operations, but rather on DNA codes.

The Rochester gates – actually tiny DNA processing centres – detect specific fragments of the genetic blueprint as input, then splice together the fragments to form a single output. For instance, a genetic And gate links two DNA inputs by chemically binding them so they are locked in an end-to-end orientation, much as two



Scientists at the University of Rochester have built logic gates that rely not on electrical signals but rather on DNA codes to perform operations. Picture shows a sample of DNA against a computer monitor displaying a model of the DNA helix.

Legos might be fastened end-to-end by a third Lego stuck on top.

An enzyme called DNA ligase seals the gap between the ends of the two input strands, yielding a single new strand.

Using regular gel electrophoresis, the length of this new strand can be precisely measured, providing the DNA computer's answer or output to the two input strands.

Ogihara recently showed mathematically that a computer consisting of a series of DNA-filled test tubes can work more efficiently than a digital computer in analysing the information cascading in from a tangled web of logic gates. This includes the type of calculations now done every day, as well as more complex arrangements.

Ray and Ogihara are among a growing group of scientists who believe that DNA could serve as a very compact, efficient, and accurate form of memory in computers – just as it does in the cells of the human body. The potential benefits of a DNA computer are astounding: 500g of DNA has the capacity to store more information than all the electronic computers ever built, and the computing power of a teardrop-sized DNA computer using the new DNA logic gates could dwarf that of the world's most powerful supercomputer, which even now is the size of a house.

Contact: Animesh Ray, University of Rochester, Tel: 00 1 716 275 8986.

Honey I shrunk the lab!

Silicon laboratories, where the functionality of specialised chemical analysis instruments has been shrunk down to fit on a computer chip, have come closer to reality following an announcement from a researcher at Purdue University.

Technology has been patented by Purdue, and PerSeptive Biosystems of Boston, that will allow scientists to pack hundreds of laboratories – each fully capable of carrying out complex chemical analyses – on a single silicon chip, reducing the cost and boosting the efficiency of many chemical and medical analyses.

The laboratory chips should be available in three to five years, according to Fred Regnier, professor of chemistry at Purdue and co-founder and chief technical officer of PerSeptive Biosystems. The aim is to allow physicians and medical professionals to perform chemical analyses using the chips that currently are done at specialised laboratories.

For instance, in standard chromatography, a solution to be separated is poured through a tube or column packed with various particles that are coated with a chemical compound. The different components of the solution are attracted to the particles with different affinity. As the mixture flows through the column, it separates into a series of zones, each containing a pure substance.

The miniature laboratories employ the same principle. The difference is in their size and the way they are made. Channels and microscopic 'particles' are created, this time using photolithography and chemical etching – the same technologies that are used to build semiconductors. The entire laboratory – with chemical reaction vessels the size of a speck of dust and chromatography columns the size of a human hair – is cut from a single piece of silicon, similar to the creation of a sculpture. Liquids are moved on the chip by voltage applied at the ends of the channels.

What makes this device different to similar devices under development is that Regnier has found a way to create tiny, rectangular particles within the channels. These monolith structures, etched into the column as a single unit, serve the same purpose as the packing materials used in conventional chromatography columns, and they allow the miniature laboratory to perform more complex procedures.

Despite their diminutive size, the laboratories on a chip can obtain accurate measurements using only a fraction of a drop of liquid. Instead of working with microlitres of liquid, the chips need only picolitres. Yet measurements can still be made to within a few percent accuracy.

The mini-laboratories also differ from standard chromatographs in that they contain no moving parts. So they are much simpler device and are much less expensive to build than conventional laboratory equipment. A standard liquid pumping system and column, for example, may cost £10000, but a chip can be fabricated for £300, and up 100 mini-laboratories can be lined up on a single chip.

The ability to fabricate specialised instruments at low cost and to connect large numbers of them together could have a major impact on areas of science such as clinical analysis and drug discovery. The new technology may be particularly useful in pharmaceutical laboratories where scientists analyse thousands of natural and synthetic compounds in search of new drug candidates.

Other applications could include clinical settings such as a doctor's office, where the miniature laboratory could be used by medical professionals to perform diagnostic procedures. For simple diagnostic procedures, laboratories could be designed to work in a fashion similar to pregnancy test kits.

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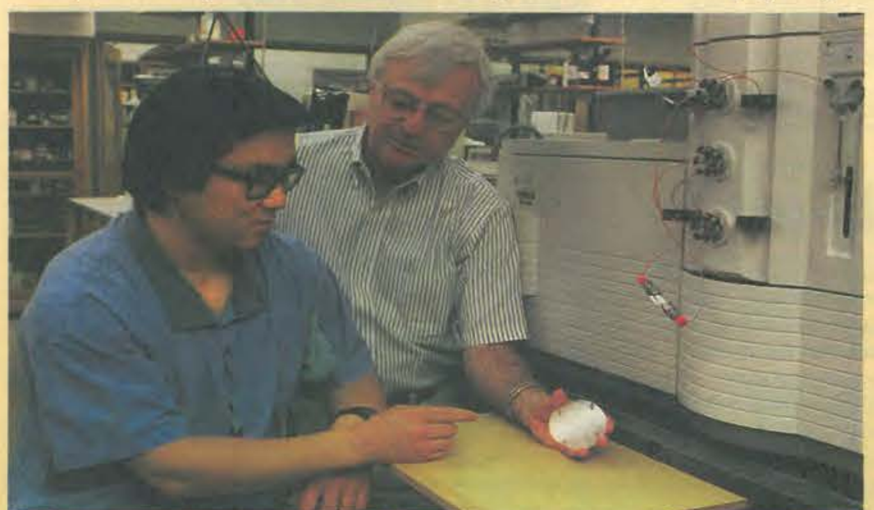


Purdue researcher Fred Regnier (right) and colleague stand next to a liquid chromatograph while examining a silicon wafer that contains a scaled-down version of the device. The mini-laboratory is capable of carrying out many of the same types of chemical separations as the full-sized instrument.(Purdue News Service Photo by David Umberger)

Process adds to coating development

A new coating process that enables novel compounds to be formed and also solves some of temperature problems currently inherent in coating chips has been developed at the University of Buffalo.

By using such extremely high temperatures and then quenching the heat, the new technique solves one of the trickier problems in computer-chip fabrication: how to coat them while avoiding high temperatures that can cause computer chip samples to fail. This has been a serious drawback for fabricators of expensive chips for research-grade supercomputers, such as the Cray.

The new hybrid technique marries the advantages of the two conventional fabrication methods – laser ablation and molecular beam epitaxy – while overcoming their disadvantages. Just as importantly, the method, called laser assisted molecular beam deposition, OR Lambd, can also manufacture new coatings for electronic devices in the same stage.

"Instead of simply sputtering a target material from point A to point B, we're chemically modifying it at the same time," says James Garvey, professor of chemistry and principal investigator.

The key to Buffalo's fabrication method is that it causes chemical reactions that would be impossible to generate otherwise, and it does it all in one step.

Like laser ablation and molecular beam epitaxy, Lambd uses extremely high heat to remove particles of a target material from one surface and transport it to another, creating a thin film.

What makes Lambd different is that those molecules then collide with a pulse of gas, the identity of which is determined by the type of chemical product desired. The technique also seems to remove from the fabrication process certain toxic precursors that otherwise are necessary in generating thin films.

For example, electronic wafers often must be coated with titanium nitride to act as a diffusion barrier – a process that requires a very toxic precursor, requiring extensive safety and disposal equipment.

With the new technique, nitrogen gas is simply blown over a rod of titanium, depositing titanium nitride directly on the substrate. The process results in a thin film of protective titanium nitrate without the use of toxic precursors.

So far, several unique materials have been formed using Lambd, such as hybrid organic/inorganic films where an organic material with good optical characteristics has been encapsulated in silica glass. These new materials could have valuable photonics applications for new computer devices.

Laser assisted molecular beam deposition, or Lambd, could open up new chip coating opportunities

Explosion packs punch into magnetic pulse

In a remote corner of Los Alamos National Laboratory, scientists from around the world spent weeks setting up an apparatus designed to measure what happens to certain compounds under extremely high magnetic fields – then they blew it to bits.

The Dirac experimental campaign is an international collaboration begun in 1996 to investigate the atomic structure and chemistry of materials by subjecting them to an exceedingly strong magnetic pulse driven by an explosive charge. Scientists are probing how materials conduct electricity in extreme conditions. The magnetic forces created briefly in the experiment by the explosion make it possible to investigate aspects of the structure of condensed matter that are otherwise impossible to study.

To attain the intense magnetic fields required, the research team placed sample materials inside an electromagnet, with the samples cooled to a few degrees above absolute zero. The magnet was surrounded with about 18kg of explosives, arranged to produce a perfectly uniform implosion.

In the millionths of a second before the equipment turned to dust, sensors captured measurements of how the sample's electrical resistance changed as the magnetic field was squeezed and concentrated by the blast.

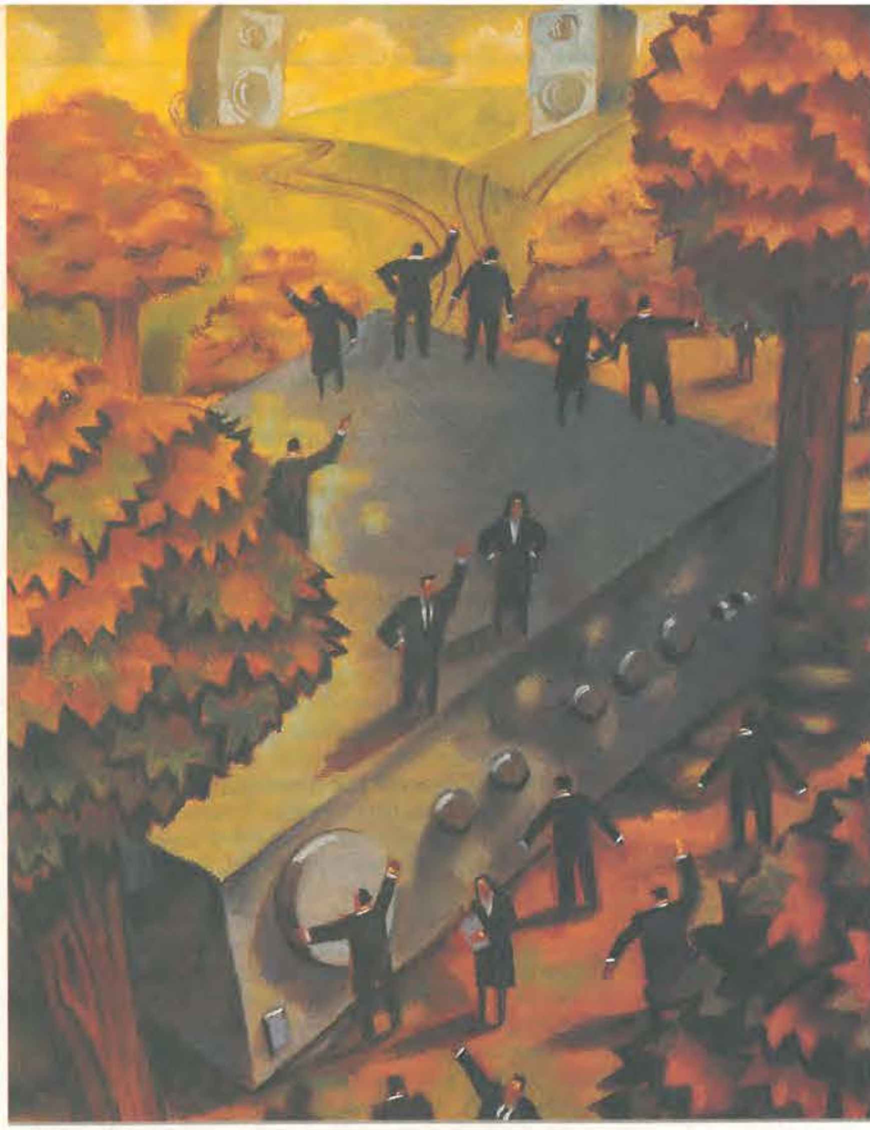
Magnetic field strength is measured in teslas.

Sustained fields in research magnets are generally around 60 tesla – one tesla is about 20000 times stronger than Earth's field. The Dirac experiments have achieved momentary pulsed fields reaching 850 tesla. Insights from these experiments may help in the design of superconductors and better semiconductors.

But for the scientists, the pay-off is the glimpse at secrets that are normally hidden from observation. ■

Contact: Los Alamos National Laboratory, New Mexico, USA.

Explosions have been used at Los Alamos to generate massive magnetic pulses.



The interface between an audio power amplifier and its loudspeakers is important, but not for the reasons you might think, reveals Doug Self.

The audio power interface

A good deal has been written in recent years about speaker cables. Much of this has been raving nonsense – perpetrated by those drawn to audio as it appears to be a subject in which one can profess expertise without the trouble of actually learning anything.

Some approaches are well-intentioned but misguided, such as worrying about how cables perform at 1GHz. This article attempts something most unusual – it looks at what cables actually do in their real situation.

Figure 1 shows the most important elements in the amplifier-cable-speaker system. Their basic interactions must be understood before worrying about possible second-order effects at radio frequencies.

Putting a damper on damping factor

One of the most misunderstood features of this system is the effect of amplifier output impedance on speaker dynamics. The parameter 'damping factor' purports to measure how much this output impedance degrades the electromagnetic damping of loudspeaker units. In fact 'damping factor' must be the most misnamed parameter in existence, for an amplifier's ability to damp loudspeaker resonances is effectively NIL. This is so important and apparently so little understood that the next section deals with this first.

Damping factor is defined as the ratio of the load impedance R_{load} to the amplifier output resistance R_{out},

$$damping\ factor = \frac{R_{load}}{R_{out}}$$

Solid-state amplifiers often have output resistance around 50 to 100mΩ so an 8Ω speaker means a damping factor of at best 160 times. This simple definition ignores the fact that amplifier output impedance usually varies considerably across the audio band, increasing with frequency as negative feedback factor falls; the output 'resistance' is actually more like an inductive reactance at high audio frequencies. The presence of an output inductor to give stability with capacitive loads increases this reactance.

A further condemnation of damping factor is that the bass

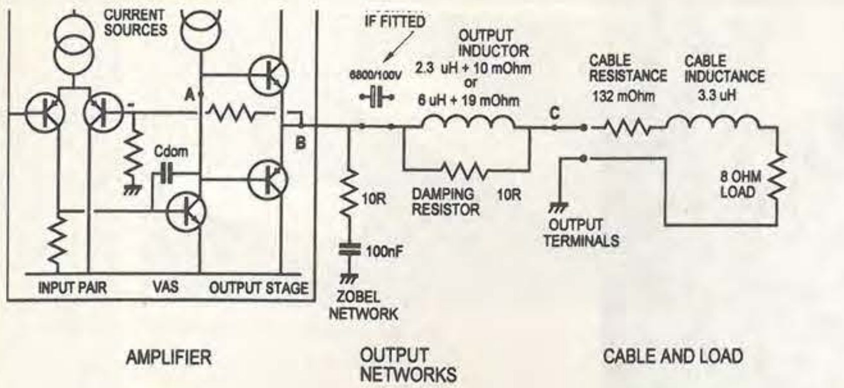


Fig. 1. Amplifier-cable-speaker system. Simplified amplifier with Zobel network and output inductor, 5m of cable with series resistance and inductance, and a resistive load. Values are typical.

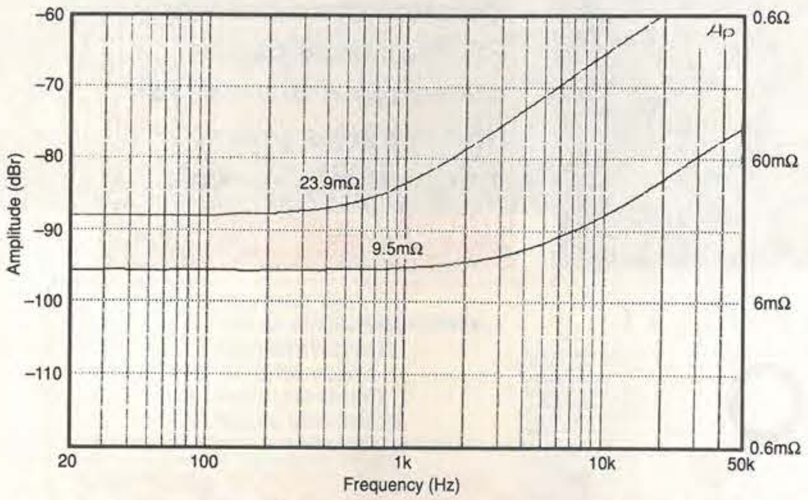


Fig. 2. Output impedance of a 'blameless' amplifier, with and without 6μH output inductor. Adding the inductor, upper trace, increases both the flat low-frequency output impedance, due to its series resistance, and the rising high-frequency impedance.

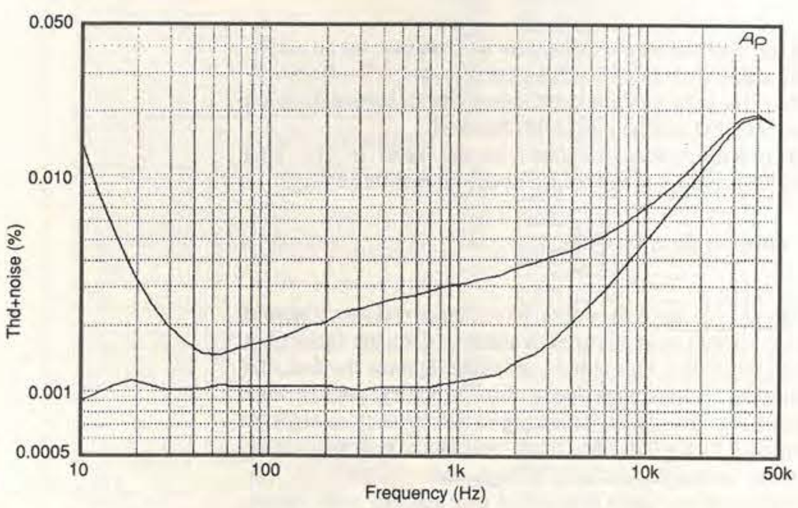


Fig. 3. Distortion with and without a standard 6800μF/100V electrolytic at 40W into 8Ω. There is extra distortion across the audio band, as well as a sharp rise at lf.

resonance is the one frequency where a real loudspeaker can be pretty much guaranteed to have an impedance nothing like 8Ω. With a sealed box, peak resonance impedance is often 30Ω or more.

In the face of these potential complications, it is a mercy that damping factor as such has very little effect on loudspeaker behaviour. A factor of 160 times, as derived above, promises a truly radical effect on cone response – implying that resonances and such have been reduced by 160 times as the amplifier output takes an iron grip on cone movement. Nothing could be further from the truth.

The electrical damping of a loudspeaker unit depends on the total series resistance in the voice-coil circuit. This is the sum of speaker coil resistance, resistance in the crossover, speaker cabling, and, last of all, amplifier output impedance. Values for these will be typically 7Ω, 1Ω, 0.2Ω and 0.05Ω, so the amplifier only contributes 0.6% of the total, and its effect on speaker dynamics must be negligible.

Does amplifier output impedance matter?

Setting guidelines for crossover inductors, Neville Thiele states that a winding resistance of 5% of the voice-coil resistance was low enough to have no significant effect on drive unit dynamics.[1,2]

If it was possible to make an amplifier change the speaker low-frequency resonance, this would be a really poor idea; the bass resonance is an integral part of the speaker design, and tampering with its frequency and Q is unlikely to improve anything.

The irrelevance of 'damping factor' was practically investigated and fully confirmed by James Moir as far back as 1950[3], though this has not prevented periodic resurgences of controversy, nor stopped writers from claiming that inadequate loudspeaker cables have a direct effect on bass transient response, with 'bad' cables usually alleged to yield 'flabby' or 'woolly' bass. In contrast, Martin Colloms deals very sensibly with damping factor in reference 4.

So 'damping factor' is almost an arbitrary ratio, with little physical reality. Nonetheless, one reason to strive for a high figure – which after all, can do no harm – is the numbers game of impressing potential customers with specifications figures. It is as certain as anything can be that the subjective difference between two amplifiers, one with a damping factor of 100, and the other boasting 2000, is undetectable by human perception. Nonetheless, the two specifications look very different in the brochure, so minimising output impedance to maximise the damping factor may be of commercial importance. See the end of this article for more details.

Although amplifier output impedance is of little interest in terms of speaker dynamics, it is still worth minimising. Too high an output resistance causes power losses into low-value loads. These look more significant than they are when expressed in watts.

A potentially more serious problem is that non-zero output resistance causes output voltage to vary as speaker impedance varies with frequency, introducing unwanted response irregularities.

Despite this demonstration of its irrelevance, the rest of this article continues to refer to damping factor, to show how an apparently impressive figure shrivels away as more parts of the system are taken into account.

Amplifier output impedance

Audio amplifiers, with a few very special exceptions,[5] approximate to perfect voltage sources and aspire to zero output impedance across the audio band. The amplifier output is essentially unaffected by loading, so that the frequency-variable impedance of loudspeakers does not give an equally variable frequency response.

While a true zero impedance is impossible with simple negative feedback, a very close approximation is possible if even modest feedback factors are used. A judicious mixture of voltage and current feedback can make the output impedance zero, or even negative – in other words, heavier loading makes the output voltage *increase*. This is clever, but usually pointless, as you will see. Solid-state amplifiers are quite happy with lots of feedback, but it is usually impractical in valve designs.

The highest output impedances are found in valve equipment, where global feedback inclusive of the output transformer is low or non-existent; impedances around 0.5Ω are usual. Some idiosyncratic semiconductor designs also have high output resistances[6] with 0.6Ω; far too high.

Figure 1 shows a simplified amplifier plus output networks – Zobel network and output inductor – and simple representations of the speaker cable and load. The output impedance of a real solid-state amplifier is very low indeed if it has a reasonable amount of global negative feedback.

Using a 'blameless' Class-B amplifier with a complementary feedback pair output stage[7] the negative feedback factor at 20kHz is 29dB as usual, increasing at 6dB/octave as frequency falls. **Figure 2** shows the resulting output impedance of this 'blameless' design in its 'naked' state, ie measured at point B before the output inductor, by injecting a 10mA signal current into the output via a 600Ω resistance.

At low frequencies, the output impedance is approximately 9mΩ, corresponding to an 8Ω damping factor of 890. To put this into perspective, one metre of 32/0.2 equipment cable, i.e. 32 strands of 0.2mm diameter, has a resistance of 16.9mΩ. It is thus quite possible for the internal cabling resistance in an amplifier to equal or exceed the output impedance of the amplifier itself at low frequencies.

Output impedance rises at 6dB/octave above 3kHz, as global negative feedback reduces, reaching 36mΩ at 20kHz. The 3kHz break frequency does not correspond with the amplifier dominant pole frequency, which is much lower, at around 10Hz.

More to output impedance than meets the eye

The closed-loop output impedance of an amplifier is determined by its open-loop output impedance and the negative feedback factor. The former is not simply the output impedance of the output stage, because it is driven from the voltage amplifier stage, giving a significant and frequency-variable source impedance at point A, Fig. 1.

For the standard emitter follower and complementary feedback pair stages I have considered before, *driven from a zero-impedance source*, the raw output impedance is in the region of 150 to 180mΩ for both, if the emitter resistors R_e are my usual 0.1Ω. Increasing R_e to 0.22Ω increases the output impedance to 230 to 280mΩ, demonstrating that these resistors form the greater part of the output impedance.

Taking the average open-loop output impedance as 200mΩ, and the negative feedback factor at 20kHz as 29dB, or 28 times, you would expect the closed-loop output impedance to be 200/28, ie 7mΩ. Since it is actually about 33mΩ at this frequency, there is clearly more going on than simple theory suggests.

In a real amplifier the output stage is not driven from a zero impedance, but a high one that falls proportionally with frequency; for my 'blameless' Class-B design it falls from 3kΩ at 1kHz to about 220Ω at 20kHz[8]. A 220Ω source impedance gives an open-loop output impedance of about 1Ω, which divided by 28 gives 35mΩ – very close to the measured value.

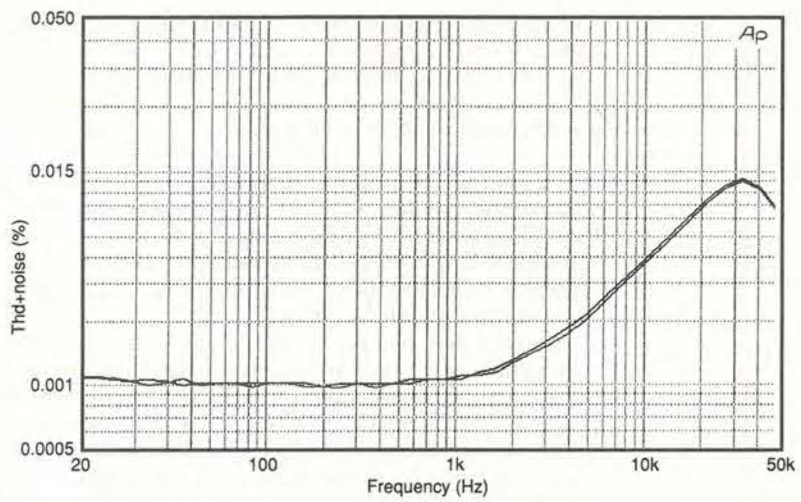


Fig. 4. With and without a very large output capacitor, the BHC Aerovox 100 000μF/40V driving 40W into 8Ω. Capacitor distortion is eliminated.

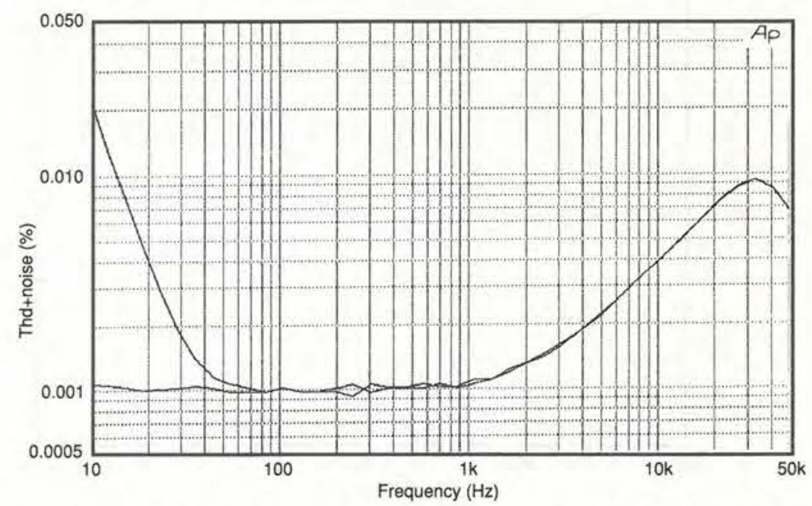


Fig. 5. Distortion with and without an 'audiophile' Cerafine 4700μF/63V capacitor. Midband distortion is eliminated but the low-frequency rise is much the same as with the standard electrolytic.

Fortunately, all these measured closed-loop values are so low compared with other impedances in the amplifier-cable-speaker system that they can be ignored. There seems little point in deeper investigation at this stage.

In practice, my 'blameless' design has an output inductor of 6μH. This is at the high end of the permissible range of inductance, the aim being absolutely ensured stability into capacitive loads.

My version has 20 turns of 1.5mm diameter copper wire, with a dc resistance of 19mΩ. This is a heavyweight component, designed to minimise resistive losses into a 4Ω load, but even so its extra resistance pushes the flat part of the impedance curve up to 24mΩ, so inductor resistance dominates the low-frequency output impedance as measured at the amplifier terminals, point C.

Damping factor is abruptly reduced from 890 to 330. If low output impedance is a priority, the coil wire must be as thick as cost/quality trade-offs allow. Naturally the coil inductance makes the rising portion of the impedance curve higher. The output impedance now rises from 700Hz, still at 6dB/octave, and reaches 0.6Ω at 20kHz.

Since it is clear that the output networks of an amplifier have important effects on its output impedance, their role and component values are worth a look.

The role of the output capacitor

AC-coupled amplifiers have the advantage that there are no dc-offset problems, yet they remain unfashionable. The output capacitor used to be condemned for reducing the low-frequency damping factor, but as explained above, this is actually the least of our worries.

Large output electrolytics have significant equivalent series resistance, or esr, and some series inductance. For typical amplifier-output sizes esr will be of the order of 100mΩ; this is probably why ac-coupled amplifiers rarely had output inductors, as it is enough resistance to provide isolation from capacitive loading, as discussed later.

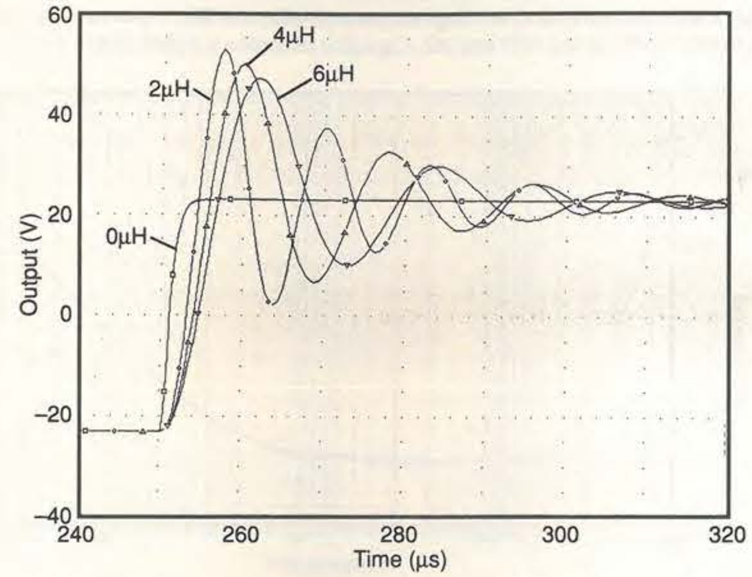


Fig. 6. The output inductor value has little effect on ringing amplitude.

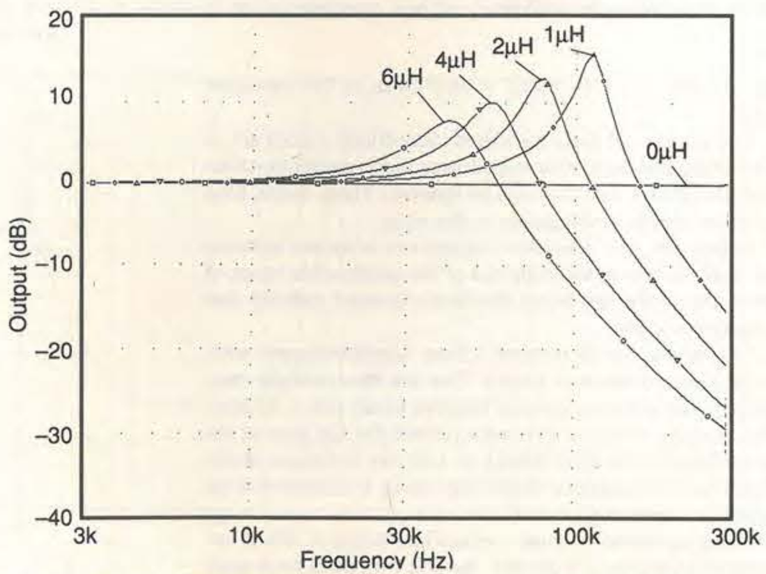


Fig. 7. Effect of output inductor value on frequency response. Higher values cause more invasion of the audio band.

The series inductance is very low and probably irrelevant, being quoted by one manufacturer as "...a few tens of nanohenries". Capacitor esr alone is enough to reduce the damping factor to 80.

A more serious problem with amplifier output capacitors is the distortion they cause. Like other forms of capacitor distortion, this seems to be something people just don't like to talk about.

Output capacitor distortion is a serious problem; it is not confined to low frequencies, as is the case in small-signal circuitry. For a 6800µF/100V output capacitor driving 40W into 8Ω, there is 0.0025% mid-band third-harmonic distortion, as in **Fig. 3**. This is five times more than the amplifier generates in this frequency range.

The harmonic distortion rise at the low-frequency end is much steeper than for small-signal capacitor distortion, for reasons that are not yet clear. A load of 4Ω doubles the distortion at a given frequency.

I know only two cures for this problem. The head-down no-messing approach uses a huge output capacitor, far larger than required for a good low frequency response. A 100mF/40V *Aerovox* from BHC gave **Fig. 4**. Alternatively, an allegedly 'audiophile' capacitor gives some interesting results; a *Cerafine Supercap* of only moderate size, at 4700µF/63V, gave **Fig. 5**, where the midband distortion is gone, but the low-frequency rise remains.

What special audio properties this component is supposed to have is unknown; electrolytics are never advertised as 'low midband thd', but that seems to be true here. The case volume is about twice as great as conventional electrolytics of the same value, so the difference may be a thicker dielectric film than normally used for this voltage rating.

Either of these special capacitors is more expensive than the rest of the amplifier electronics put together. A dc-coupled amplifier with protective output relay looks like the more economical option.

The Zobel network

Almost all power amplifiers have a Zobel network in their arrangements for stability. This simple but enigmatic network comprises a resistor and capacitor in series from the amplifier output rail to ground, on the amplifier side of the output inductor.

The resistor is usually 10Ω, and the capacitor almost invariably 100nF. These convenient values and their constancy in differing amplifier designs suggests they are not critical, but not so; my experiments suggest that the traditional values really are optimal for stability.

The function of the Zobel network – also called a Boucherot cell – is rarely discussed, but it is sometimes said to prevent an excessively inductive reactance being presented to the amplifier output by a loudspeaker voice-coil. The implication here is that this could cause high-frequency instability.

It is intuitively easy to see why a capacitive load on an amplifier with a finite output resistance might cause global hf instability by introducing extra lagging phase-shift into the global negative feedback loop. But it is less clear why an inductive load should be a problem; if a capacitive load reduces stability margins, surely an inductive load would increase them.

Some experiments were clearly necessary. I removed the standard 10Ω+0.1µF Zobel from a 'blameless' Class-B amplifier with my usual negative feedback factor of 29dB at 20kHz. This amplifier had a complementary feedback pair output.

With an 8Ω resistive load, the thd performance and stability were unchanged. However, a 0.47mH inductor in series with 8Ω, roughly simulating a loudspeaker unit, provoked what appeared to be local vhf instability in the output stage. There was definitely no Nyquist instability of the global negative feedback loop, and no evidence that an inductive load presented difficulties in the audio band.

Increasing the Zobel resistance from 10 to 22Ω gave some evidence of stability problems, and I had to conclude that the standard values are just about right. Note that the Zobel network is at much too high an impedance to have any shunting effect on the amplifier output impedance.

A few designs have a second Zobel network after the output inductor; the thinking behind this is obscure. Reference 9 shows that a second Zobel network can be useful – if it is assumed that cable capacitance is at least 50nF. Since 500pF is a more usual value, the relevance of this is unclear.

The output inductor

The function of an output inductor is to prevent instability when a capacitive load is connected. It does this by isolating the amplifier from the shunt capacitance without causing significant losses at audio frequencies.

Low-value resistors have been used instead, but their efficiency is questionable, and they seriously degrade damping factor. Direct connection is only safe for simple amplifiers with very low feedback factors.

The resistance of even thick inductor wire causes an apparently disastrous collapse in the damping factor from 890 to 330. As such, this is unimportant. The worst effect of resistance is probably unwanted response irregularities due to speaker impedance variations with frequency.

An output inductor must be air-cored to eliminate the possibility of distortion due to magnetic saturation. Ferrite vhf chokes give stability, but their linearity must be regarded as dubious. In the seventies, there was a fashion for using one of the big power-supply electrolytics as a coil-former; this is not a good idea. The magnetic characteristics of the capacitor are unknown, and its lifetime will be reduced by heat dissipated in the coil resistance.

Usually, the output inductor has a value of between 1 and 7µH, the upper limit being set by the need to avoid significant roll-off at 20kHz into a 4Ω load. Great caution is required when designing an amplifier for home construction, and so when I designed the Class-B 'blameless' amplifier,[10] the inductor value was set near the upper limit at 6µH to ensure stability. If 2Ω loads are contemplated this upper limit must be halved.

What is the least inductance needed?

Further investigations into the minimum inductance required for stability showed that a 'blameless' amplifier without its inductor is still fairly immune to capacitance-induced oscillation. This is possibly because the level of global feedback is fairly modest at 29dB at 20kHz.

The capacitances tried in parallel with 8Ω were 100nF, 470nF, 1µF, and 2µF, and a fast square wave was used to test for ringing. A 100nF capacitor caused close to continuous oscillation, but there was only well-damped ringing on the amplifier output A for 470nF, 1µF, and 2µF.

With the 6µH inductor replaced there was complete amplifier stability in all cases. The 6µH inductor was then cut in half, giving 2.3µH and 10.1mΩ dc resistance; this was also stable for all capacitor values, but has not been tested with real speakers.

An alternative method of stabilisation is a series resistor instead of an inductor. Even with 100nF, a 0.1Ω wire-wound output resistor completely removed ringing on the amplifier

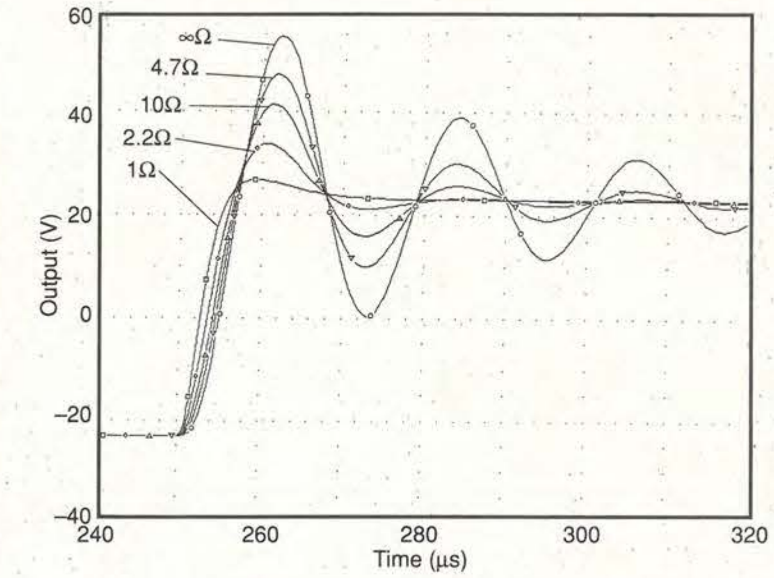


Fig. 8. Damping resistor value has a powerful effect on ringing; 2.2Ω reduces it greatly, and 1Ω prevents it completely.

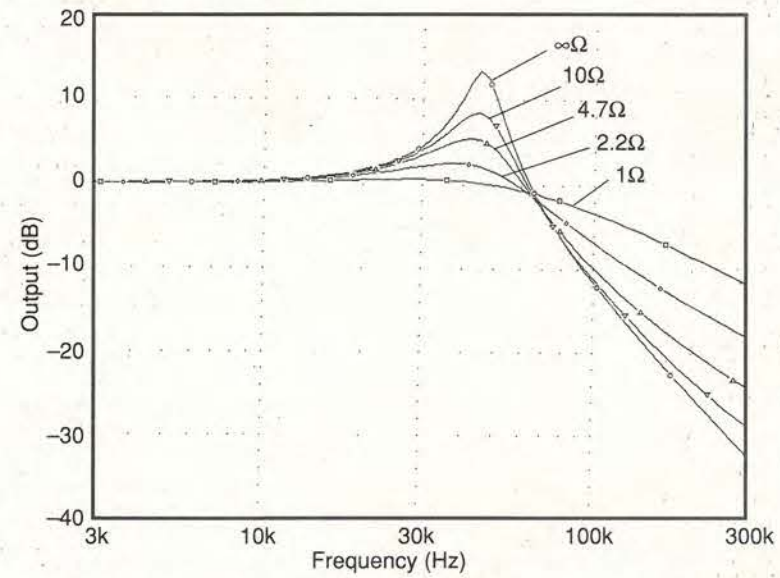


Fig. 9. The effect of damping resistor value on frequency response. Lower values damp the LC resonance better.

output. This is cheaper, but obviously less efficient than an inductor, as 100mΩ of extra resistance has been introduced instead of 10mΩ for our new 2.3µH inductor.

With 0.1Ω, the damping factor cannot exceed 80. A more important objection is that the 4Ω output power appears to be significantly reduced. An amplifier capable of driving 200W into 4Ω is reduced to a 190W unit. This doesn't look so good in the specifications – even though the reduction in perceived loudness is negligible.

The output inductor is only actually wanted for its isolating qualities, but it has at least two other effects. It adds to the cable inductance and the combination causes an high-frequency roll-off when driving an 8Ω load. Secondly, it introduces major confusion into testing with capacitive loads, as described next.

Effects of load characteristics

Amplifier transient response is usually tested with a squarewave. The output is loaded with 8Ω and 2µF in parallel to roughly simulate an electrostatic loudspeaker, as this is often

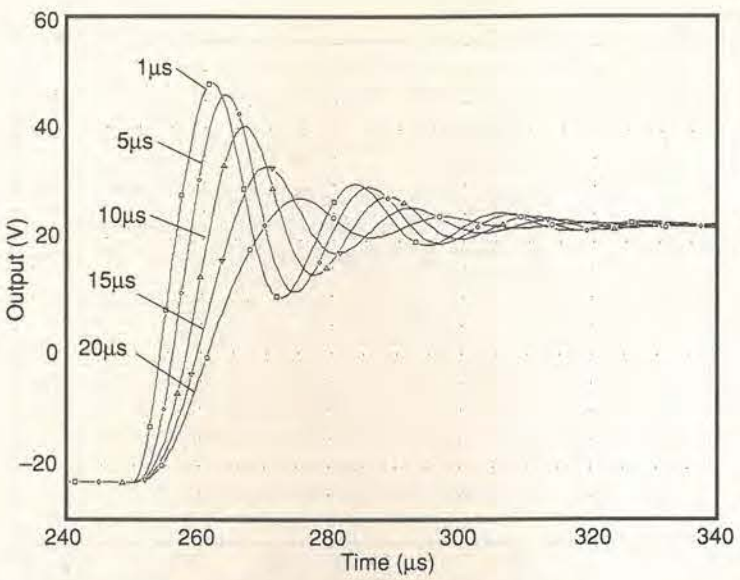


Fig. 10. Test-signal rise time has more effect on ringing than any other parameter. Varied here from 1 to 20µs.

regarded as the most demanding condition.

When there is an inductor in the amplifier output, and significant capacitance in the load, they resonate together. This produces a peak in the frequency response at hf, together with overshoot and ringing on fast edges. It is usual to put a low-value damping resistor across the inductor; this reduces the Q of the output *LC* combination on capacitive loading, and thus reduces overshoot and ringing.

The damped ringing that is seen during these capacitive loading tests is visible at point C but absent at A and B. It is due to the output inductor resonating with the test load capacitance, and has nothing whatever to do with amplifier stability; this test does not really examine amplifier response at all. The ringing is usually around 40kHz or so, and this is much too slow to be blamed on any normally compensated amplifier.

If a power amplifier is provoked by shorting out the output inductor and applying a capacitive load, the oscillation is normally around 100-500kHz, which is destructive of output transistors if it persists. It is nothing like the neat ringing seen in typical capacitive load tests. In the power amplifier itself there is no such thing as 'nicely-damped' ringing; damped oscillation at 500kHz means you are one shaved inch from disaster.

A 'blameless' amplifier is rather resistant to capacitance-induced oscillation. Paralleled with 100 nF directly across the output, a resistance of 8Ω induced damped ringing at 420kHz, while 470nF gave ringing at 300kHz, and 2µF at 125kHz.

The 8Ω/2µF test reveals nothing about amplifier transient response, but it is embedded in tradition, and it is too optimistic to expect its doubtful nature to be universally recognised. Several factors affect output ringing, and can be manipulated to tidy up overshoot and avoid deterring potential customers:

Output inductance. Increasing the inductance with all other components held constant reduces the overshoot, **Fig. 6**, and the amount of response peaking, but the peak moves downward in frequency so the rising response begins to invade the audio band, **Fig. 7**.

The damping resistor across the output coil. Reducing it to below 10Ω lowers the Q of the output *LC* circuit, reducing overshoot and ringing. It may be wire-wound without self-inductance problems.

Adding a 10Ω damping resistor reduces overshoot from 58% to 48%, and greatly reduces ringing, **Fig. 8**. Response peaking is reduced, with only a slight effect on its frequency, **Fig 9**. The damping resistor can be reduced to well below 10Ω, if stability remains dependable. This can cut transient overshoot further to 20%, and eliminate ringing altogether.

Load capacitance. Increasing this with the shunt load resistor held at 8Ω gives more overshoot and lower frequency ringing that decays more slowly. Response peaking is both sharper and lower in frequency, which is not a good combination. However, the capacitance is part of the standard test load and outside the designer's control, so it is not considered further.

Rise time. It is a commonplace that in audio technology, the real truth is quite different from the conventional wisdom. In fact, by far the most important factor affecting overshoot and ringing is the rise-time of the applied square wave. **Figure 10** demonstrates how the overshoot given by the circuit of Fig. 1 is 51% for a 1µs rise-time, but only 12% for a 20µs rise-time.

The 'transient response' apparently being measured depends critically on the details of the test gear and amplifier slew-rate, and can be cynically manipulated to give the result you want.

Minimising output impedance

Leaving aside specmanship, it is worthwhile minimising output impedance, not because it optimises speaker dynamics, but because it minimises frequency response variations due to varying speaker impedance.

The naked amplifier output impedance will be less than the inductor resistance. To minimise the latter, first determine how low the output inductance can be for stability with capacitive loads. Lower inductance means fewer turns and less wire. However, the inductance of the usual single-layer coil varies with the square of the number of turns. Halving inductance only reduces turns, and hence series resistance, by root-two.

It will also be desirable to minimise the resistance of the amplifier internal wiring, and carefully consider any extra resistance introduced by output relays and loudspeaker switching. ■

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Hands-on Internet

In addition to tips for helping you save time on the Internet, Cyril Bateman offers pointers to non-Spice electronics CAD software and details of a competent audio power amplifier design.

While searching Alta Vista for Smith chart software, following a link which turned out to be broken, to my surprise instead of the usual error message, I found myself at the 'Cern' site[1] in Switzerland, **Fig. 1**.

After a few moments I recalled that Tim Berners-Lee, the originator of the hypertext mark up language, was based at Cern in 1991 when the whole concept of the Internet Web page system was devised.

HTML is a method for linking to other relevant documents and forms the backbone of the present day World Wide Web. Originally it was a text based system. When I first used Internet in 1994, browsers for use on a pc were available. In their earlier days, for many users, Cern was an essential entrance gateway, using 'Telnet'.

Not previously having visited the Cern site, I spent a few minutes looking around. In 1991, to keep track of the Web system, Tim Berners-Lee started the WWW Virtual Library Project. It continued at Cern until handed over to the WWW Consortium[2] in January 1996, with Tim Berners-Lee as its director. With its continuing rapid growth, this library is now distributed between several caretaker organisations, according to discipline, **Fig. 2**.

My curiosity aroused, I decided to look into the background of Alta Vista[3], which is my favourite keyword search engine. Bearing in mind current disk and memory capabilities, the size, power and speed of the searcher seem quite impossible.

The main advantage of a keyword database is its ability to respond to a search query, regardless of the position of the keyword in any document. The disadvantage is the volume of results returned to the user and the storage needed.

Alta Vista's keyword index consists of more than 30 million pages and 10 billion words, taking up some 40Gbyte of disk space. Maintenance of this index requires the combined efforts of six Digital Alpha computers using 11.5Gbyte of ram and 463Gbyte of disk space.

This index is constantly rebuilt using a software robot called Scooter to crawl the Web, retrieving three-million pages a day. These are then indexed using Alta Vista's Web indexing computer, capable of

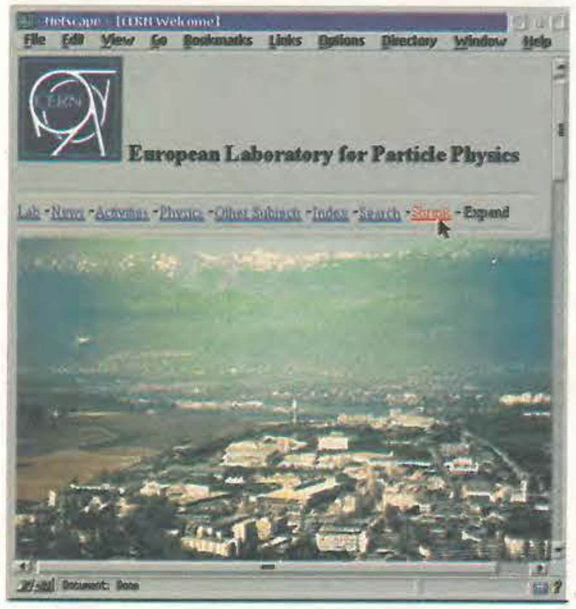


Fig. 1. Birthplace of the Web – Conseil Europeen pour la Recherche Nuclear (Cern). The World's largest scientific laboratory has two sites which bridge the French/Swiss border. It has the world's biggest large electron positron accelerator – 27.6km in circumference. Cern is a 19-member collaboration of international states, appropriately located near Geneva.

Where to look

1.	European Laboratory for Particle Physics.	http://www.cern.ch
2.	The Virtual Library.	http://www.w3.org/pub/DataSources
3.	Alta Vista search engine.	http://www.altavista.digital.com
4.	Cobb Group.	http://www.cobb.com/go/altavist
5.	The Internet Sleuth.	http://www.isleuth.com
6.	Industry Net.	http://www.industry.net
7.	Spread Spectrum Scene.	http://sss-mag.com/swindex.html
8.	Max Froding.	http://users.aol.com/maxfro/private/mstrip.html
9.	Rogers Corporation.	http://www.rogers-corp.com/mwu
10.	The Leach amplifier page.	http://www.ece.gatech.edu/users/207/lowtim

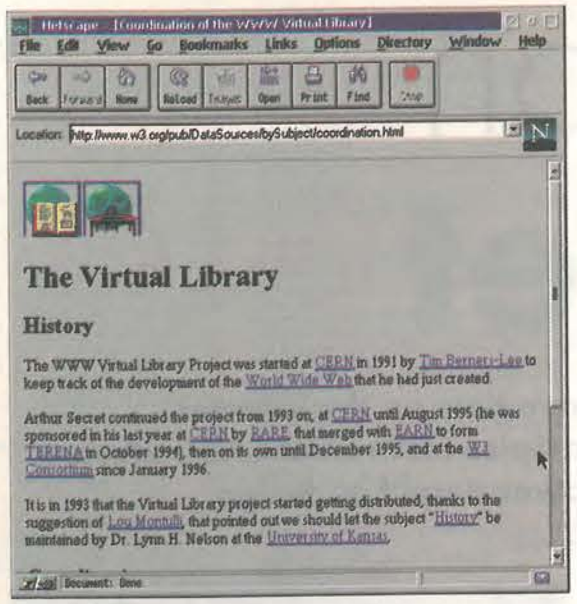


Fig. 2. The WWW Consortium manages the WWW Virtual Library. Engineering sections of the WWW Virtual Library are supervised by NASA.

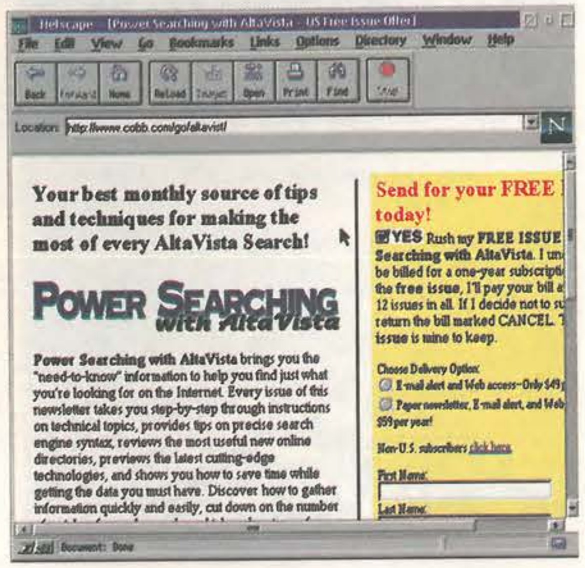


Fig. 3. Advanced techniques ensure the best search results. Find how to manage Alta Vista searches.

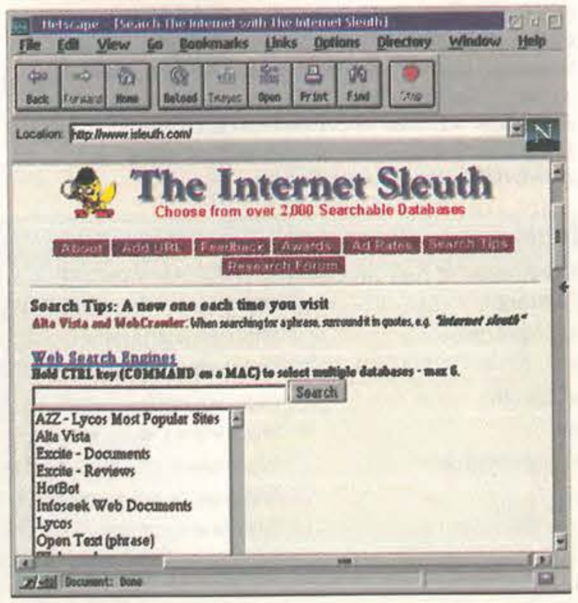


Fig. 4. Use a Meta-Search engine to maximise search information. Select your choice for multiple searches.

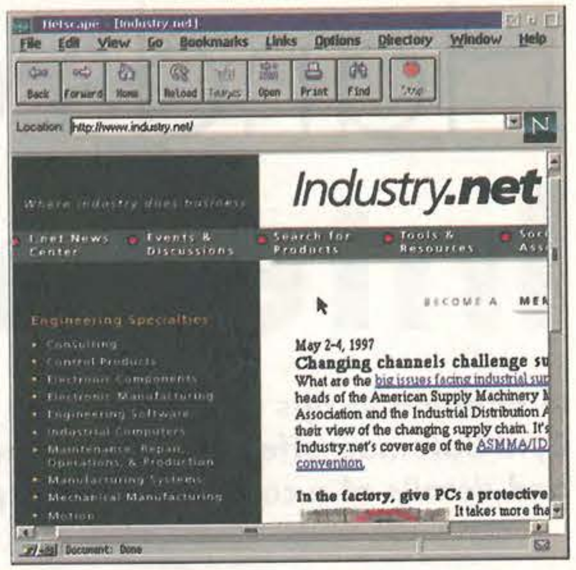


Fig. 5. Universal Engineering Manufacturing Forum. Excellent multi discipline engineering shareware repository.

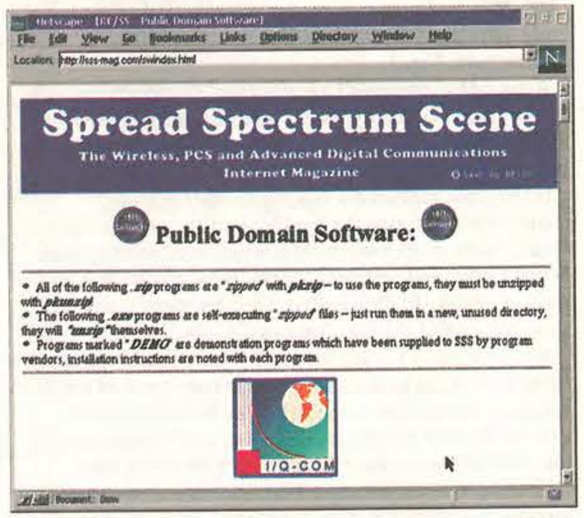


Fig. 6. Meeting place for all electronic designers – especially those involved in rf. Software library includes many hard to find utilities.

indexing 1Gbyte of text in an hour. However Scooter cannot update if the page is held on a computer behind a server gateway, or on one protected by a 'Firewall.' This explained my broken link.

The volume of 'hits' returned from searching a keyword index can be reduced using advanced search techniques. A newsletter publication 'Power Searching with Alta Vista'[4] from The Cobb Group claims to provide answers to search problems as well as using the Alta Vista Private Extensions software to access data within your own system, **Fig. 3.**

Even given the power of Alta Vista, it is not possible to ensure that any keyword search database is complete and up to date. Last month I covered use of the new limited area search engines. This can usefully target some engineering areas, but should both methods fail to locate the needed information, all is not lost.

The Internet Sleuth[5] provides the facility to search your choice of any six from over 2000 databases simultaneously. These include the popular Alta Vista, Excite and Infoseek indexes, which should guarantee a successful result. A possible downside, due to the volume of data searched, is that each search takes longer and can result in large amounts of returned data to be sifted, **Fig. 4.**

Simulation and design software

Sources of Spice simulators have been well covered in past issues, but electronics design encompasses many problems that cannot be solved using Spice. Two such involve metalwork of housings and radio frequencies circcuit design. An excellent source of shareware solutions targeted to design engineers in the widest sense can be found at Industry.Net.[6] This is a forum for all industrial manufacturing, which has an excellent thirteen page selection of essential engineering packages, both for mechanical and electronic design, **Fig. 5**.

This month's electronic topic centres on rf design needs. *Spread Spectrum Scene*[7] is an Internet magazine

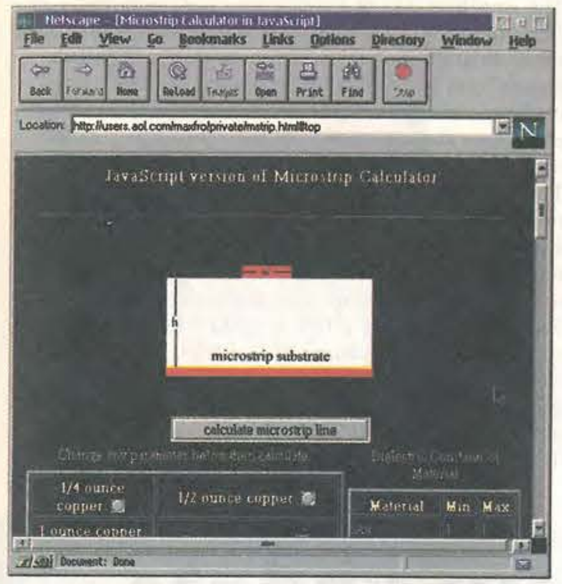


Fig. 7. Design microstripline widths/impedances on-the-page. Another unique Max Froding rf utility.

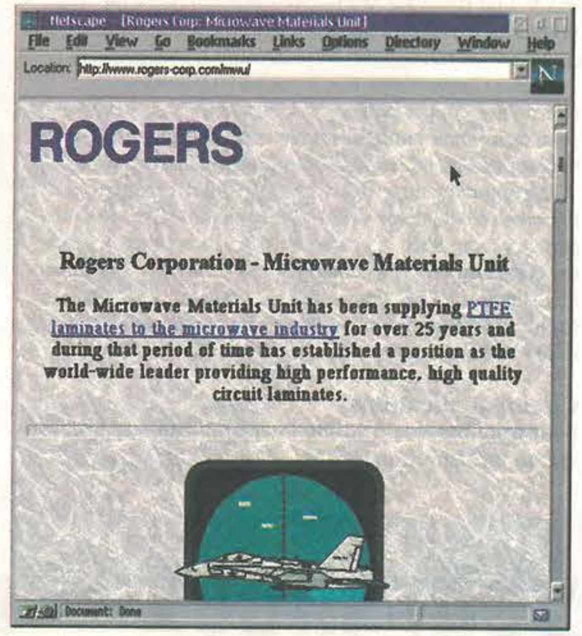


Fig. 8. Specialist supplier of low loss high-frequency laminates. Track widths designed using the laminate makers own utility, give maximum confidence.

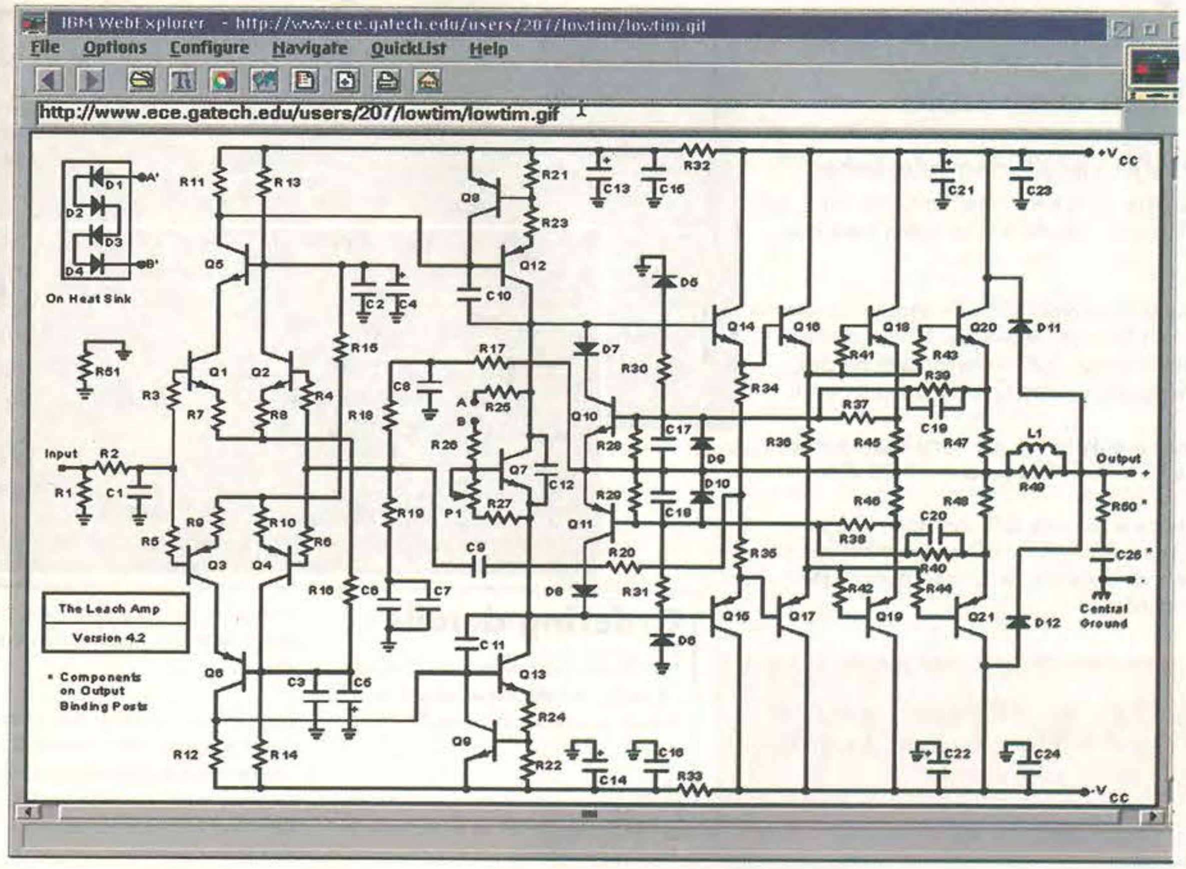


Fig. 9. The Marshall-Leach amplifier – an interesting well proven low-tim design. Fast slewing, low imd and full power into 4Ω, with this 120W version. Full constructional details for this design are published on this Web page.

embracing wireless, pcs and digital communication. Its public domain software page lists more than 150 useful packages that are easily downloaded.

Most packages are easy to find, but some are difficult. I spent some time trying to locate Motorola's excellent Smith Chart program on Motorola's page without success. I was able to download it immediately from SSS, **Fig. 6**.

In my November '96 Internet article, I outlined an add-on for generating net-lists from Max Froding.[8] It was produced for the ARRL rf circuit simulator. This month Max offers an on-the-page micro-stripline design tool in two versions. It uses JavaScript for the Netscape browser and Visual Basic for Microsoft's Internet Explorer. The Javascript version includes a useful table of dielectric constants for most common microstrip materials, **Fig. 7**.

Rogers Corporation[9] has supplied PTFE based microwave copper laminates for more than 25 years. In the past, using their laminates, I have relied on their excellent book of design tables when designing microstrip circuits.

The company now offers its MWI.ZIP microstrip impedance calculator which can be downloaded from the page, or from SSS. If you want to find out how successfully the company bonds its copper foils – as thin as $^1/_4$ ounce – to a slippery PTFE glass substrate material, visit the page for a description. Included are micro photographs of both rolled and electro-deposited copper foils, **Fig. 8**.

Circuit applications

Following the recent strong interest in audio topics, a search on 'amplifier design' led me to some pages written by Professor W. Marshall Leach, Jr[10]. He teaches both audio engineering and transistor circuit analysis at the Georgia Institute of Technology.

Professor Leach has contributed many published papers on audio, and on Spice models for loudspeakers and microphones. He is a Fellow of the AES, a senior member of IEEE and a member of the Acoustical Society.

His pages are particularly interesting since they include the circuit diagram together with full constructional detail for an interesting power amplifier. There is also a parts list with US sources.

The amplifier was originally designed some years ago following the vogue for designs achieving low transient intermodulation distortion. Over the years, countless versions have been built by his students, and all bugs have been worked out. The current version 4.2 is thus well proven and represents the results of many detail design refinements

This amplifier uses a cascode input stage together with feedback split into two separate high and low frequency paths. It also has feed-forward frequency compensation above 150kHz around the driver and paralleled transistor complementary output stages.

The design supplies 120W into 8Ω with SMPTE intermodulation distortion of less than 0.01% below clipping. It can drive a 4Ω load to full power without current limiting. Further features are an 8.5MHz gain-bandwidth product and a 70V/μs large-signal slew rate. It is also stable with capacitive loads, **Fig. 9**. ■

Camcorder dubber

As every camcorder owner knows, a car engine sound track does nothing to enhance scenery filmed from a moving vehicle. Ian Hickman's mixing circuit is designed not only to remove unwanted sound but also to replace it with whatever you want.

On a winter holiday recently, we took a camcorder with us, for the first time. Not wanting to tie up the expensive HI8 metal tapes for ever, and have to play holiday movies through the camcorder, the obvious step was to transfer the material to VHS tapes. But that raised the question of what to do about the sound.

In earlier times, with 8mm home movies, usually there was no sound, or at least it was added afterwards, by 'striping' the film. The camcorder, by contrast, gives one stereo sound, whether you want it or not. This is fine when shooting a street carnival, or for catching everything but the smell of a loco in steam. But often – as when shooting snow covered mountains through the windscreen of a car for exmple – the sound is more of a nuisance than a help.

A simple applique box

What was needed was a gadget to permit the sound on the VHS tape to originate from either the camcorder, or from a cassette recorder, at will.

Further consideration made it clear that it should be possible to 'cross fade' the two sound sources, to avoid any clicks due to switching transients. And a further useful facility would be a microphone input, so that comments, or at least a simple introduction, could be added to the soundtrack. To avoid further switching arrangements, the microphone input should operate a 'ducking circuit', to reduce the volume of the background music when speaking.

The outcome was an applique box, interconnecting the various items of kit as illustrated in **Fig. 1**.

The circuitry

Figure 2 shows the circuit of the applique box. Unity-gain buffers are provided for the camcorder sound and the input from the cassette recorder's DIN connector. These turned out to be at much the same level. A gain of 30dB is provided by the microphone input buffer.

The microphone and tape inputs are summed at the virtual earth, pin 13, of IC_1, and applied to one end of 1kΩ linear potentiometer, R_{18}. Buffered stereo sound from the camcorder connects to the other end, the wiper of R_{18} being connected to volume control R_{19} which is a 4.7kΩ logarithmic potentiometer.

Thus the sound output from buffer IC_1 pin 7 can be cross faded at will between the stereo output and tape/microphone inputs, and its level adjusted from normal down to zero. The

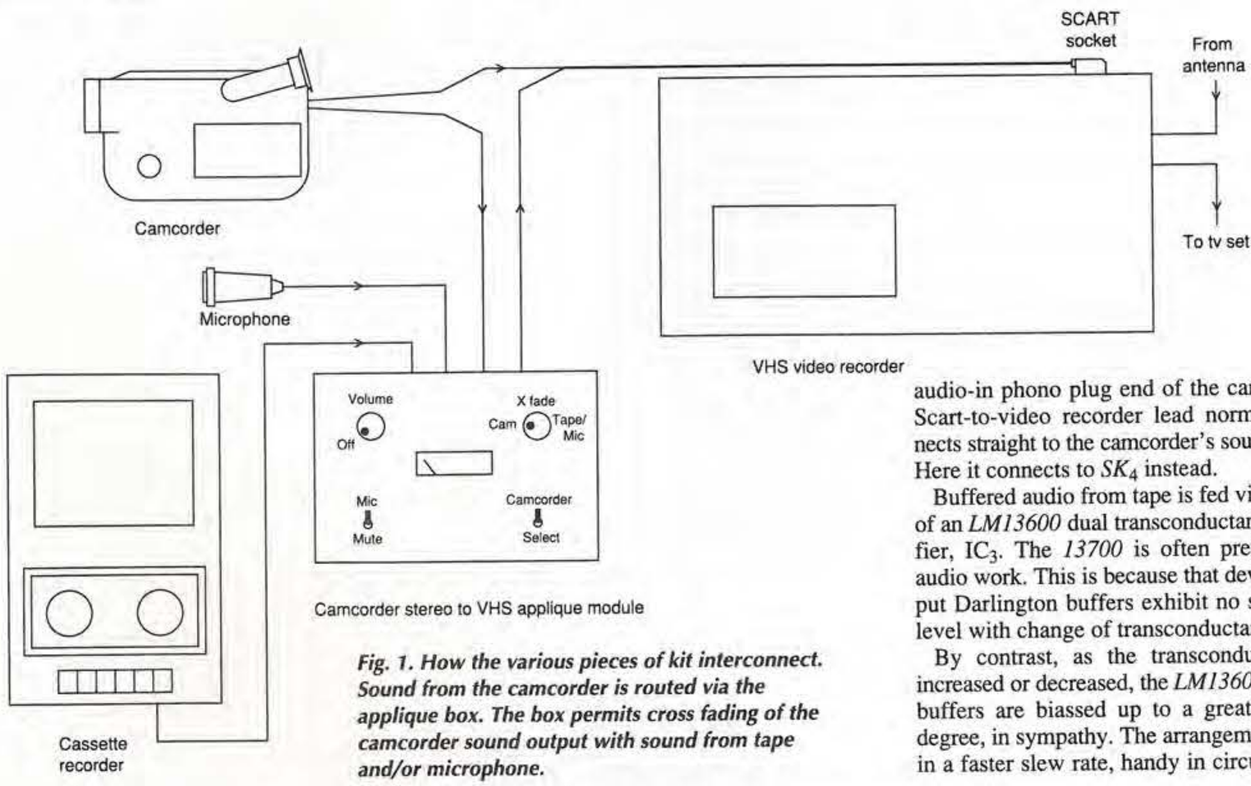


Fig. 1. How the various pieces of kit interconnect. Sound from the camcorder is routed via the applique box. The box permits cross fading of the camcorder sound output with sound from tape and/or microphone.

audio-in phono plug end of the camcorder's Scart-to-video recorder lead normally connects straight to the camcorder's sound output. Here it connects to SK_4 instead.

Buffered audio from tape is fed via one half of an *LM13600* dual transconductance amplifier, IC_3. The *13700* is often preferred for audio work. This is because that device's output Darlington buffers exhibit no shift in dc level with change of transconductance.

By contrast, as the transconductance is increased or decreased, the *LM13600*'s output buffers are biassed up to a greater or less degree, in sympathy. The arrangement results in a faster slew rate, handy in circuits where

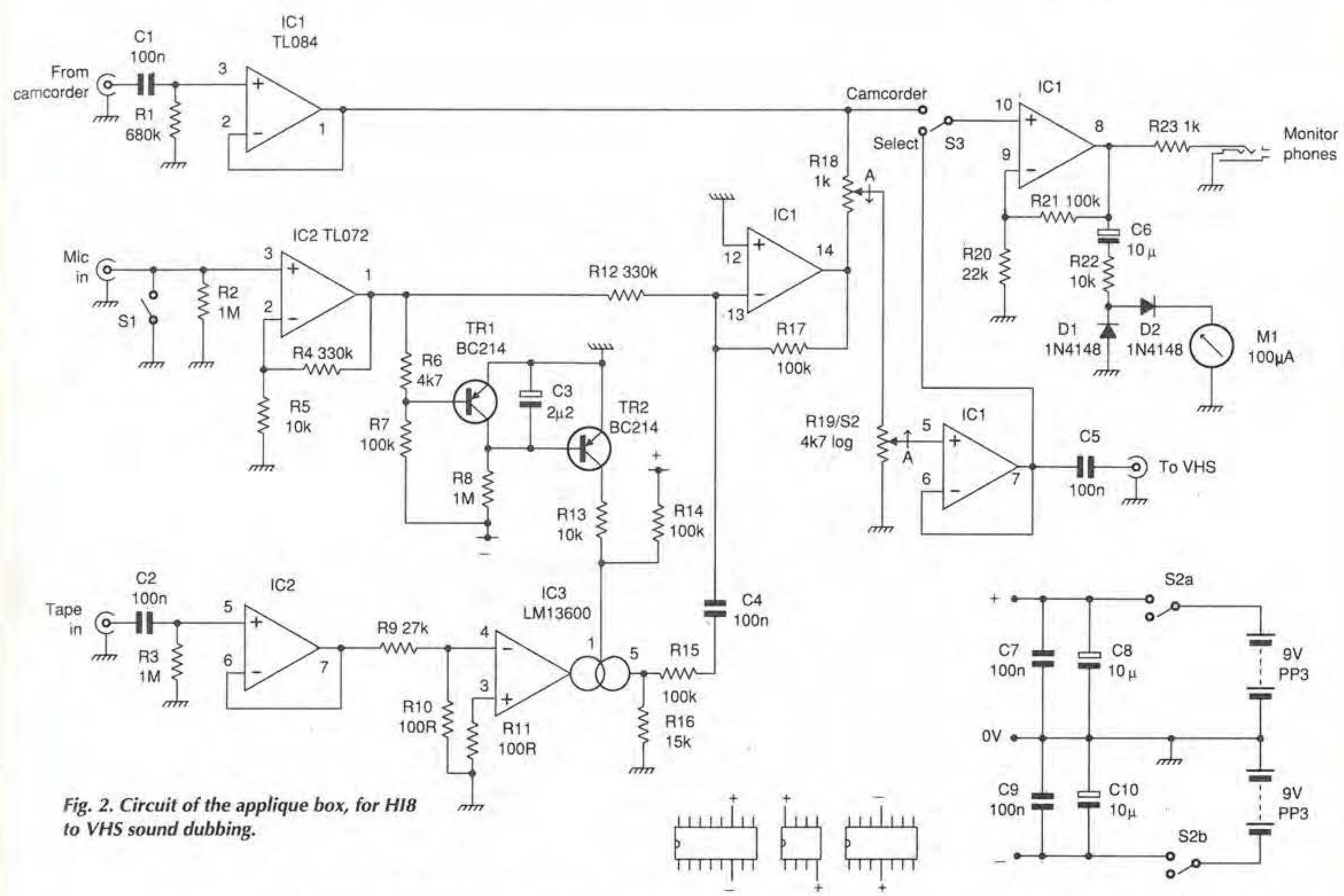


Fig. 2. Circuit of the applique box, for HI8 to VHS sound dubbing.

fast settling is needed. But it can result in 'pops' in an audio circuit, when there are rapid changes in gain. However, in this application the Darlington output buffers are not used, so either device will do.

The transconductance of IC_3, and hence its gain, is set by the bias current I_{ABC} injected into pin 1. This consists of two components, the larger proportion coming via $R_{13} - Tr_2$ is normally bottomed – with about another 25% or so coming from the positive rail, via R_{14}. When a voice-over output from the mike appears, the negative-going peaks at pin 1 of IC_2 bottom Tr_1. This discharges C_3 and removes the base current from Tr_2. The gain of IC_3 thus drops by some 12dB or more, this proving a suitable degree of ducking.

The microphone used was a small dynamic type with a 50kΩ output impedance. In fact, it needed only 20dB of gain to raise its output to the same level as that from the camcorder and cassette.

The extra gain, together with a little forward bias for Tr_1 via R_7, was incorporated to provide reliable operation of the ducking function. The extra microphone circuit gain was simply disposed of by making R_{12} 330kΩ, as opposed to 100kΩ at R_{15} and R_{17}.

Switch S_1 shorts the mike when not needed, preventing adventitious extraneous noises appearing on the soundtrack of the dubbed tape. Output from the fourth section of IC_1, at pin 8, acts as a buffer to drive monitor phones, which can be plugged into SK_5. It also drives a simple level monitor indicator, M_1. Switch S_3 draws the output monitor/meter buffer's input either from the camcorder stereo input or from the current 'select' input, be it camcorder stereo, cassette or microphone. In a stereo version of the dubber, potentiometers R_{18} and R_{19} could be ganged.

In my prototype, power is supplied by two PP3 9V batteries, the on/off switch being ganged with R_{19}. Capacitors C_7 and C_9 were duplicated adjacent to each IC, in accordance with good practice.

Using the dubber

When transferring video from a camcorder to VHS tape in the video recorder, the latter is set to use the Scart socket as the programme source. On our video recorder, this is achieved by setting its channel number to 0, which brings up the legend 'AV' in the channel number display – a fairly standard arrangement, I imagine.

When we view video tapes, our tv is usually supplied with baseband video via a Scart interconnection. This avoids the picture degradation caused by transferring the video via the rf modulator and demodulator.

While dubbing is taking place, the Scart socket is not available as it is required for the lead from the camcorder and applique box. But tuning the tv receiver to the recorder's rf channel enables visual monitoring of the Hi8 output as it is recorded. It also enables sound to be sourced via the dubbing box's 'select' setting. Thus the main use for the phone monitor is to keep an ear on the original camcorder sound track, ready to cross fade to it when appropriate.

Figure 2 is configured for my particular collection of kit. Depending on your particular microphone and cassette recorder, cd player, or even record deck, different gain settings of the input buffers may be needed. Here, the level meter M_1 is handy, as it indicates the typical level of stereo sound from the camcorder.

Stereo enthusiasts with a suitable video recorder can double up on IC_1 and use a quad op-amp in place of a dual at IC_2, to provide stereo working. A second transconductance amplifier is of course already available in IC_3. Stereo working for the voice-over channel seems a little over the top.

Instead of cross fading the two sound sources, R_{19} may alternatively be used in conjunction with R_{18} to fade one out and then the other in, if preferred. If voice-overs are going to be fairly infrequent, S_1 can usefully be a biassed toggle, so that the microphone input is permanently muted, except when required. ■

Sound from all angles

Home theatre and surround sound

Martin Colloms explains the benefits and pitfalls of home cinema sound options in this article, which is based on a section of his newly updated book High Performance Loudspeakers.

Domestic audio systems are becoming increasingly mixed-media installations – combining hi-fi loudspeakers allied to direct view or projection video. Described as home theatre, or home cinema, these systems include surround sound capability from video and non-video programmes. Stereo, two channel sources, range from traditional vinyl disc and analogue tape, predominately cassette, to RDAT digital tape, cassette-based data reduced digital tape known as DCC, compact disc, data reduced mini disc, and radio tuner. There is also videotape, laser disc, DVD, broadcast television and video disc, plus cable and satellite transmissions.

An increasing proportion of software – particularly video based – is multi-channel encoded. Where the installation gives surround sound, a full sound field is available all around the audience. In addition, audio processors are available which can extract and/or synthesise surround sound information from suitable sources.

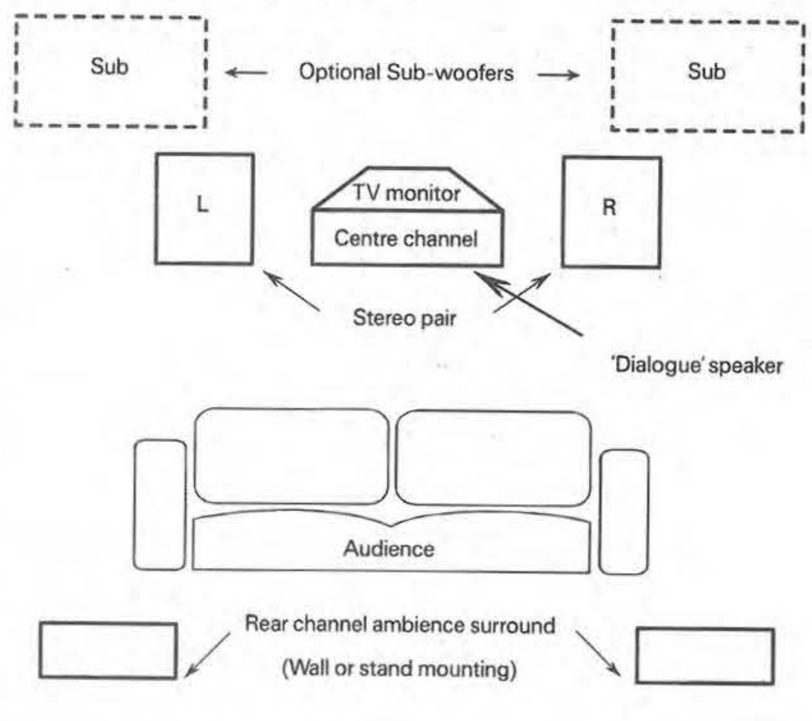
The most popular home theatre format at present is five channel, and typically includes a centre channel speaker; this is often called the dialogue channel because it generally carries the vital speech components of a movie.

The development of a sound stage, with characteristics of width and depth, is the responsibility of the front left and right audio channels. These are generally conventional high quality loudspeaker systems, often full-range types flanking, or placed relatively close to, the television monitor or projection screen. In some cases projection screens are acoustically transparent so that the centre-channel speaker may be hidden behind them. A superior blend of sound and visual image results, with better entertainment value, **Fig. 1**.

In basic installations, a pair of small ambience speakers located near the back of the room, preferably elevated to reduce localisation effects. Ambience channels should not draw too much attention to themselves; excessive level or proximity to the listeners can easily disturb the sound field balance.

The rear-channel speakers may reproduce a common, mono ambience channel, which may or may not be served by individual amplifier channels, as with Dolby Pro-Logic. More recent systems for digital discrete, multi-channel encoding, including DTS and the industry standard Dolby

Fig. 1. Basic home theatre surround sound system – a five-speaker layout using speakers with good low frequency performance for front left/right. Dotted sub-woofer(s) are optimal, only required if the 'stereo' pair have limited bass. Ambient speakers may be mounted on the side walls.



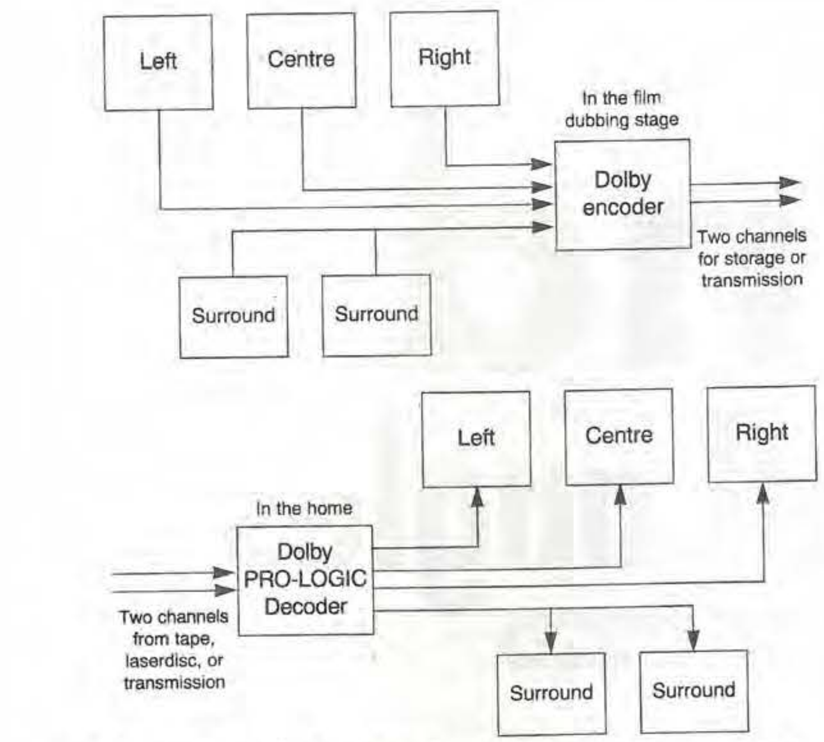


Fig. 2. Dolby surround Pro-Logic. Four channels are matrixed/encoded to two channels which reproduce as stereo. When decoded, especially with Pro-Logic steering circuity, improved separation is achieved for four output channels. Optional fifth channel for the sub-woofer is not shown (after Harley).

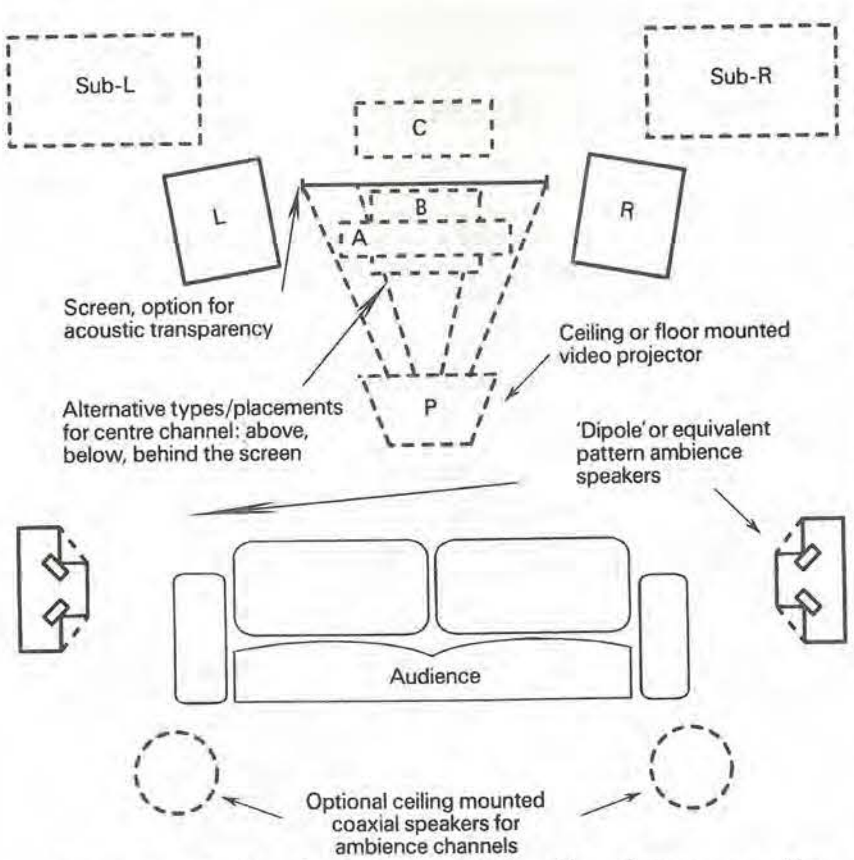


Fig. 3. THX based home theatre system. Projection video or large screen monitor, powerful sub-woofer(s) and THX style dipole or equivalent local ambience speakers. THX specification defines directivity; stereo, centre channel and ambience speakers. High sound levels typically used generally overcome the disadvantage of projector fan noise.

AC-3, can deliver discrete signals on five channels or more with the right processing – and up to eight in some instances.

Smaller, neater speaker arrangements can be used without compromising bass performance by adding a sub-woofer. In home theatres, bass is important in view of the powerful low-frequency effects commonly found on film sound tracks. Sub-woofers are generally recommended for all good home theatre installations.

Most popular surround systems incorporate decoders for Dolby Pro-Logic coded sound tracks. Decoders are available over a very wide price range and are often conveniently incorporated into multi-channel amplifiers. A specially filtered sub-woofer feed may be available, **Fig. 2**.

Stereo compatibility

It is possible to configure very good multi-channel systems, but there is always a degree of conflict between the requirements for best quality stereo reproduction. The best sound for a pair of high-performance speakers is obtained when they are the only significantly sized speakers in the listening room. Precision and power in the bass are diluted by the presence of additional speakers.

Furthermore, a direct view television frequently contains its own speakers and is placed in the centre stage position. It also constitutes a large acoustic obstacle, reflecting and redistributing the sound field formed between left and right-hand speakers. In addition, the speaker placement for optimal two-channel stereo is generally wider than that defined for good video-based multi-channel working. However, when a television monitor is replaced by a projection screen, this has rather less impact on the local acoustics and may be set farther back away from the stereo pair, **Fig. 3**.

Given the present state of the art, the fidelity of multi-channel electronics, processors and amplifiers in home theatre is not as high as that of discrete component stereo systems. Some compromise in respect of two-channel high-fidelity performance is inevitable.

Multi-channel advantage

It may be argued that multi-channel working conveys additional information and that the greater sensory experience resulting from surround sound working balances the loss in absolute fidelity when compared with pure stereo.

Greater versatility will be demanded from surround-sound systems as the new digital coding systems are introduced. These include the discrete channel Dolby AC-3. By comparison Pro-Logic is a two-channel system using matrix techniques for coding the additional directional information, **Fig. 4**.

THX and cinema-quality sound

The THX sound laboratory has set specifications, and provides design recommendations, relating to home theatre sound. These are intended to promote the production of a tonal balance and sound field in the home which more closely mirrors professional cinema practice.

In theory, the guidelines make for a closer sonic match with the film production intentions. Higher sound system cost and performance is the result, together with some divergence from normal Dolby surround practice.

THX-compatible decoders provide shaped audio signals, while speaker design also has some special features which will emerge in the following sections.

Speaker design

Specific requirements and acoustic environments for the different types of speaker system in a multi-channel installation need careful consideration. The objective should be a well-distributed sound field of uniform quality and tonality, such that no speaker or room region draws undue attention to itself and thus impairs the stability of the surround sound effect.

Equipment critics have found many instances of poor matching of the sound characters of the various speakers in a multi-channel system.

Design rules follow good speaker engineering practice. But in addition, they incorporate some specific factors, outlined below. Top quality systems conforming to the THX specification have their own set of standards and systems must qualify via direct evaluation at the THX laboratory.

Speakers for front left and right. Almost any good stereo speaker – floor or stand mounted – will serve for the front stereo pair. If the application does not include a sub-woofer, then full-range speakers with more powerful bass are advisable, probably of floor standing design.

Front speakers are likely to be close to each other, near to the monitor, and potentially not far from the rear wall. As a result, the bass alignment should be somewhat overdamped, for example with a Q of 0.5. This avoids excess energy in the 50-100 Hz range.

If the speakers are not properly damped, the reproduction of normal television speech, whether mono or two channel, may sound unnaturally heavy and boomy. This could cause early aural fatigue – even if the results are impressive on movies.

For THX working, a vertical driver array is favoured. This reduces the reverberant contribution from the floor and ceiling reflections. Mid-range tonal balance may also benefit from some subtle adjustment to take account of the proximity of the television monitor casing, and its likely supporting cabinet.

Rear channel. For simply reproducing ambient effects, an extended frequency response is unnecessary and compact slim-line enclosures are favoured. These are typically sealed-box types with a bass driver and treble unit with an alignment balanced for wall boundary location, possibly a downwards directed polar response. In some cases just a full-range single cone driver is used.

A maximum level of 100dB, an 85-90dB/W sensitivity and a 90Hz to 12kHz bandwidth is usually satisfactory, provided that the general quality, colouration standard, etc. match the primary system. Rear speaker bandwidth need only be a little wider than that imparted by the processor, **Fig. 5**.

In some installations ceiling-mounted speakers have been used successfully. These are usually coaxial types, which have the advantage of a symmetrical radiation pattern. Complications include the acoustics of the ceiling and the lack of a defined enclosure, which results in some variations in performance. The advantage is almost perfect concealment, while the additional height above the listeners aids dispersion of the ambient sound field, **Fig. 6**.

With digital multi-channel sound, the rear channel feed may be discrete, well localised, of higher quality and full bandwidth. For top quality systems superior 'ambience' speakers will be worthwhile, **Fig. 7**.

Centre-channel or dialogue speaker. For projection systems, acoustically transparent screens are available and any suitable high quality speaker may be concealed behind them.

Where a television monitor is involved, the dialogue speaker should be placed centrally to optimise the acoustic/optical alignment. Since the image cannot be obstructed it must be

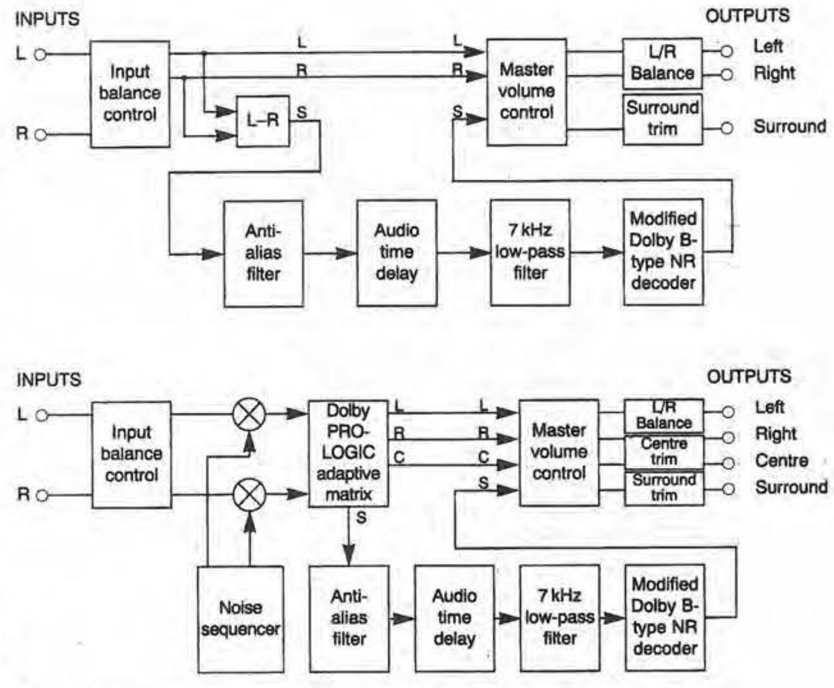


Fig. 3. Comparing the passive decoding of Dolby Surround with the active matrix system used for Pro-Logic for home theatre. Note that both systems include bandwidth limiting for the surround channel, specified at 100Hz to 7Hz –3dB. In practice, this reduces the demands made on rear channel speakers; a single full-range driver may be sufficient.

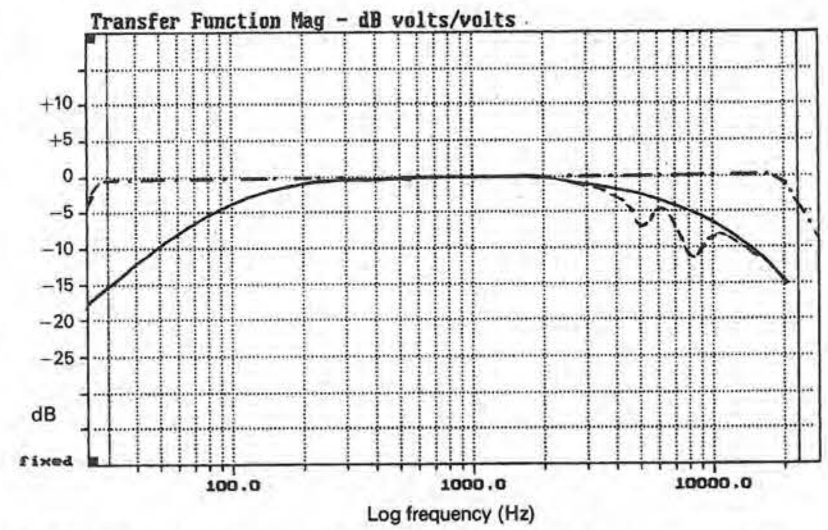


Fig. 5. Frequency responses for decoded rear channel signals. Pro-Logic (——) has a 100Hz to 7kHz bandwidth, THX augments with extra filtering (- - - -) while AC-3 digital decoding offers discrete rear channels with an additional user option – full bandwidth 20-20kHz (-·-·-·-).

located directly above or below the faceplate.

Vertically orientated drivers offer the best angle of horizontal directivity for the audience. Many THX speakers are vertically mounted. But aesthetic considerations generally dictate a low profile, slim-line, horizontally disposed enclosure, matching the monitor as closely as possible.

However, a horizontal driver disposition is the least favourable arrangement for audience coverage. Conventional two-driver designs have been tried, but inevitably the responses in one direction are unbalanced relative to the other due to asymmetry.

Symmetrical arrangements are favoured, and use a centre, preferably narrow chassis dome tweeter, slimmed by the use

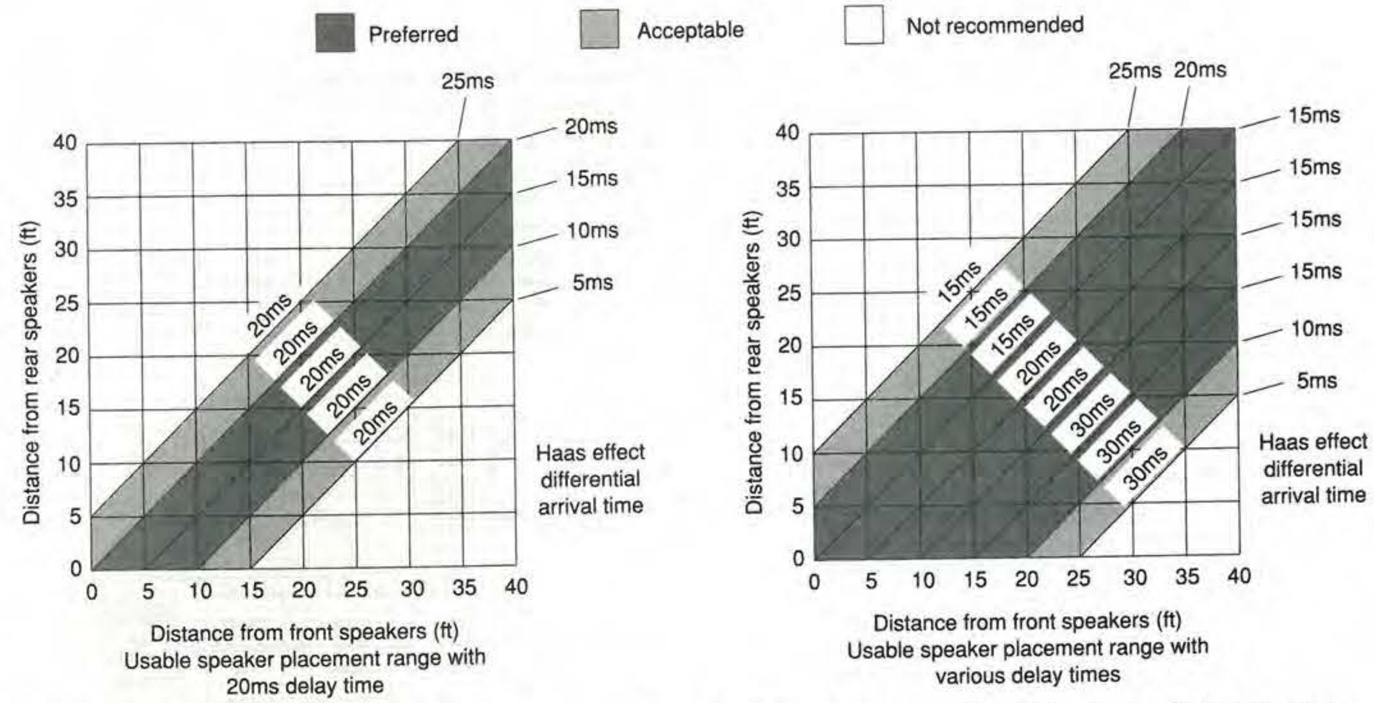


Fig. 6. Recommendations for speaker placement – rear channel relative to front – to allow good frontal localisation but avoid 'echo' effect from ambience channel feed. Some systems provide programmable delay for greater versatility but 20ms is a standard delay.

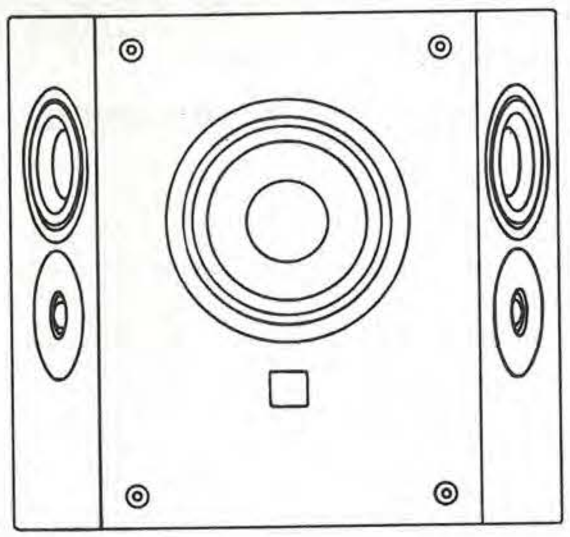


Fig. 7. An example of a dipole ambience field speaker system. Angled driver pairs operate in anti-phase above 300Hz. The centre driver gives omnidirectional radiation below 300Hz. Drivers are 170mm, 100mm and 25mm. THX application, after Aerial Acoustics Corporation, for side wall placement.

of a closed field, and thus magnetically screened, miniature magnet made from neodymium alloy. On each side is a bass-mid unit typically of 80-120 mm chassis size. The overall enclosure height is held in the region of 160mm.

Factors that help widen the radiation angle in the crossover range are high slope crossovers, and a lower than usual crossover frequency of around 2kHz instead of the usual 3kHz and 4kHz.

The use of a driver baffle structured to angle the bass drivers away from the central treble unit, for example by 10° or 20° helps to widen the acoustic lobe in the crossover range. This also needs to be well screened magnetically.

Note that these horizontal designs have another notch in their response, this time due to the relative delay, for the two mid sections appearing off-axis. This notch may be 12dB deep in the 1kHz region and adds a further complication.

A frequency range of 80Hz to 15kHz is worthwhile. Reflex loading, despite the small enclosure volume, is common to help improve the power handling. The centre channel may need to play loudly – at 103dB or more – and needs an 88-92dB/W sensitivity with up to 100W power handling.

Fine centre channel speakers have also been made with concentric drivers, such as the UNI-Q or the ICT. Such drivers have good directivity on all axes.

Specific voicing may be applied to a centre channel speaker. The proximity of the television screen also requires consideration. In addition, high clarity, intelligibility and articulation are paramount – even at high sound levels. The actor's words must be heard clearly, no matter how complex the mix of sound effects.

On balance there is a trend towards a taut, fast upper bass, occasionally some midrange and presence-range prominence of a few decibels, and an upper treble which does not draw undue attention to itself. For THX's narrow bandwidth setting, a range of only 60Hz to 10kHz is required, **Fig. 8.**

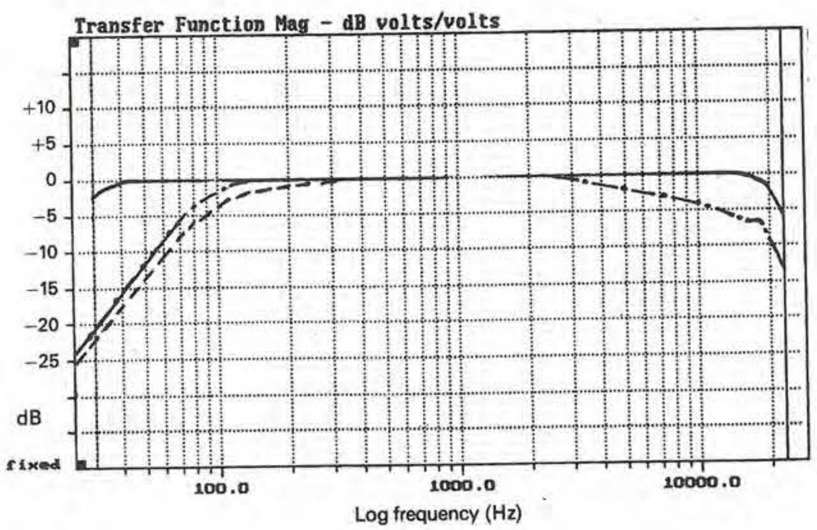


Fig. 8. Frequency response for decoded outputs of surround-sound processors, centre channel. Pro-Logic offers full bandwidth and high-pass –3dB at 100Hz options (——, -------). Processing via THX, aimed at producing a tonal quality close to cinema sound, also filters the bass and softens the treble (- • - • - • -).

Screening problems

Magnetic screening is now recommended for the front left and right speakers in addition to the centre channel units, in view of the magnetic sensitivity of the larger direct view television monitors.

Steel plates may be required to line the cabinet sides. These may be bonded with an appropriate visco-elastic adhesive to improve resonance damping. Even screened speakers may result in some colour shifts when placed on a monitor and this matter needs careful checking at the design stage.

Low frequency power

Considerable output is required in the bass for a suitably impressive film playback. It is rumoured that many of the 'foley' originated low frequency sound effects are simply shaped and/or gated bursts of third octave pink noise at 30Hz! These can reach peak level, which accounts for the THX requirement for 105dB, 30Hz at 1m for moderate listening rooms.

Two sub-woofers sound rather better than one – more than their arithmetic sum would suggest. Placement in the room corners generally gives the smoothest and most powerful bass. Good boundary matching is important because it reduces the demands made on the woofer, resulting in improved performance and a higher dynamic range.

Multi-channel discrete processors

Speaker systems for home theatre have evolved around matrix processed signals, often band limited to reduce the audibility of spurious processing artefacts. Such practice also makes good economic sense. There is no point in over-specifying the auxiliary speaker systems, centre and rear, if the result is to price the system out of the market.

However, with the introduction of discrete multi-channel processors, the European MPEG variants, the US designed DTS and Dolby AC-3 systems plus multi-channel (rather than synthesised) surround sound, all the audio channels may have a discrete, wide band, high quality identity.

Potentially, with the right replay system, the sound producer will be able to place a virtual acoustic image anywhere in the listening room, anywhere in the defined sound field. Even at this early stage, systems such as DTS can reproduce a full circle 'walkaround' with stable geometry, using a sound as complex and transient rich as closely miked orchestral chimes.

With such a potential for high quality, it is obvious that the more costly home theatre and surround sound systems are going to need better speaker systems for all quadrants of the sound field. Revisions may be anticipated with regard to existing practice, performance specifications, response shaping and required bandwidth to meet these new demands. ■

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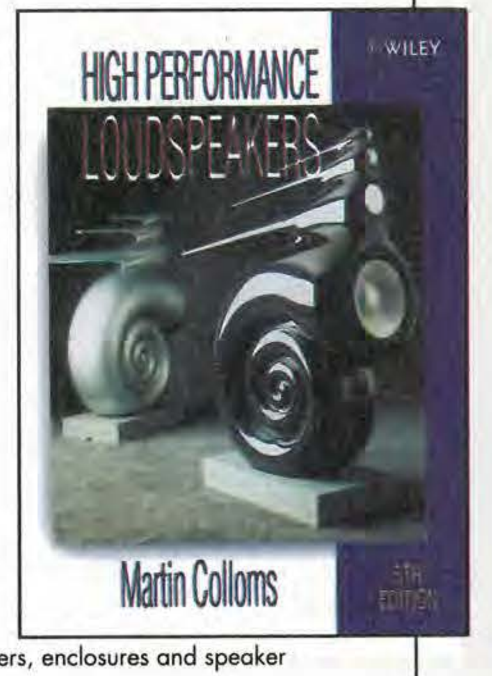


Genetically designed yagi

Richard Formato explains how a natural-selection like design process produces better antennas, and provides evidence in the form of a three-element Yagi example for 50MHz work.

Genetic algorithms, or GAs, are a class of optimisation techniques that mimic natural selection, i.e. 'survival of the fittest'. Such algorithms are applicable to many types of problems, and they are becoming increasingly useful in antenna design[1,2]. This note describes a genetically designed three-element Yagi that provides very good performance and illustrates how effective genetic algorithms can be.

Unlike deterministic optimisation schemes, GAs are based on random selection. A binary-coded genetic algorithm starts by creating a population of 'chromosomes' which are random one and zero bit sequences. Each chromosome contains a complete antenna design – in this example – a complete three-element Yagi antenna.

The chromosome is made up of 'genes' which are strung together one after another. Each gene corresponds to one of the antenna's design parameters.

The Yagi gene relationship appears in **Table 1**. A design is fully specified by eight genes: reflector length and radius *REF*, driven element length and radius *DE*, director length and radius *DIR*, and location along the boom *DE*/*DIR*. Gene length is its length in bits – for example, *REF* length is five bits.

The minimum and maximum values of each design parameter also appear in the table, and all dimensions are in wavelengths, 'waves'. The *DE* length, for example, cannot be longer than 0.6 wave or shorter than 0.4 wave.

Since each design parameter is a decimal number, not a bit sequence, the actual value of the parameter is computed by decoding its binary gene using the following transformation equation,

$$X = X_{\min} + \left(\frac{X_{\max} - X_{\min}}{2^L - 1}\right) \times D$$

where X is the decimal value of the parameter, D is the decimal value of the gene's binary sequence, and L is the gene's length.

To illustrate how this decoding scheme works, consider the 37-bit chromosome that contains the design for the Yagi discussed below:

0010111000011011111001011100010111100

The *DE* length is coded in gene No 3, which starts at bit No 10 and ends with bit No 14. The binary sequence for the *DE* length gene is 00110, and its decimal value is,

$$0(2^0)+0(2^1)+1(2^2)+1(2^3)+0(2^4)=12$$

Since gene No 3 is five bits long, the denom-

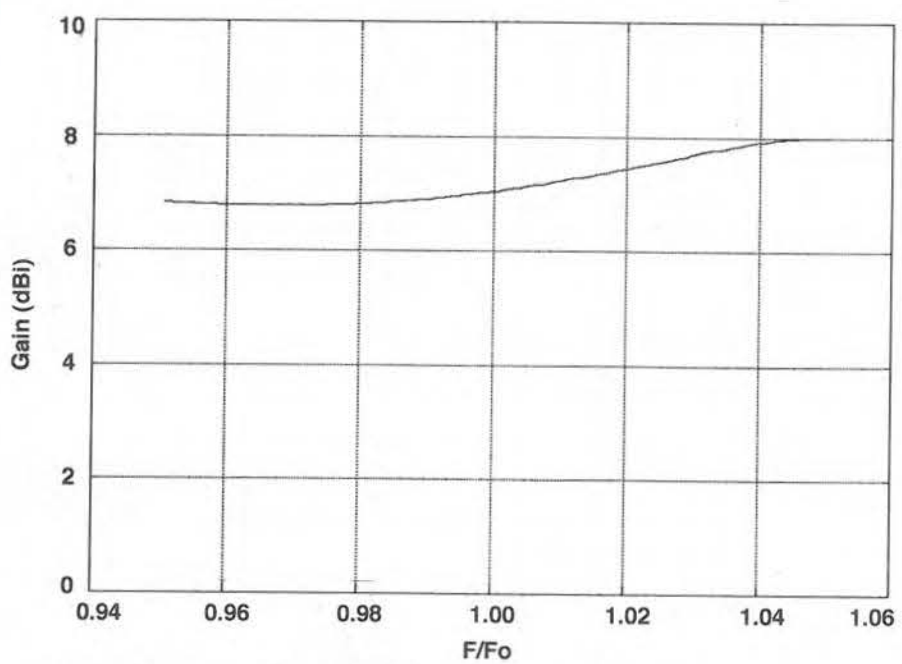


Fig. 1. Main lobe gain for the genetically-designed Yagi example.

Table 1. Gene table for three-element Yagi.

Gene #	Name	Length	Min	Max
1	REF Length	5	0.4	0.6
2	REF Radius	4	0.0005	0.002
3	DE Length	5	0.4	0.6
4	DE Radius	4	0.0005	0.004
5	DE Separation (from ref.)	5	0.05	0.3
6	DIR Length	5	0.4	0.6
7	DIR Radius	4	0.0005	0.002
8	DIR Separation (from DE)	5	0.05	0.3

* PO Box 747, Boylston, MA 01505-0747, tel. (508) 869-6077, fax (508) 869-2890

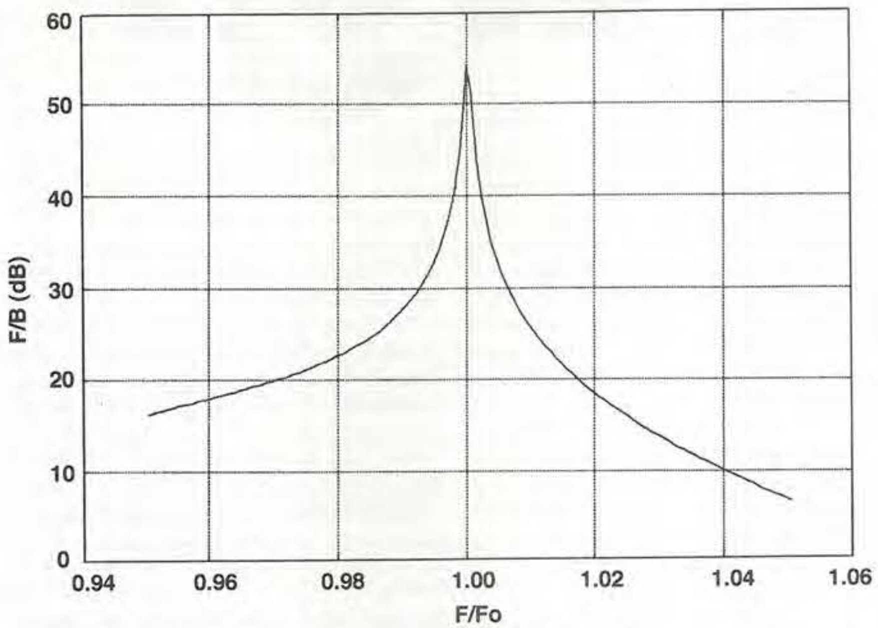


Fig. 2. Front-to-back ratio for the genetically-designed Yagi example.

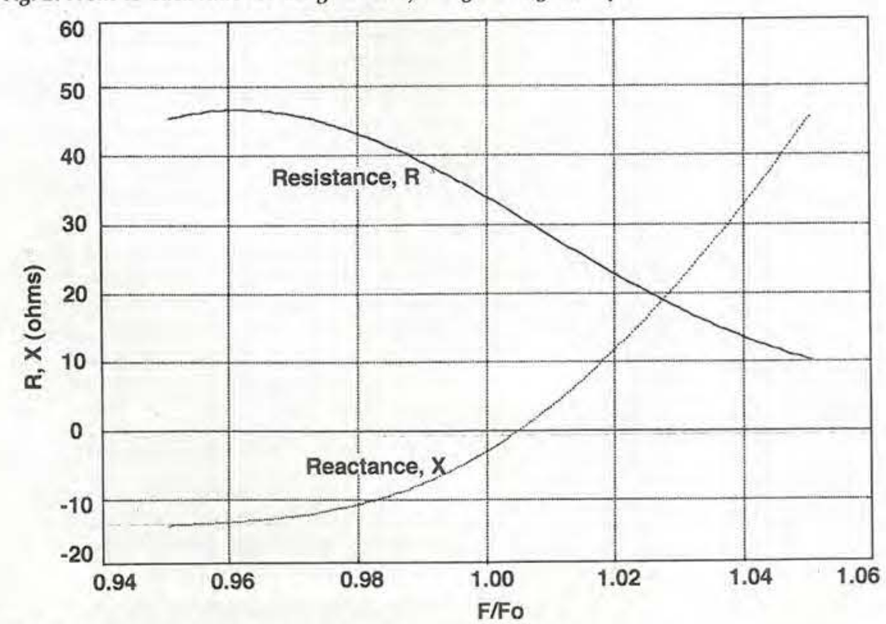


Fig. 3. Genetically-designed Yagi's input impedance.

inator in the transformation equation is $2^5-1=31$. This makes the *DE* length,

$$\frac{0.4+(0.6-0.4)12}{31}=0.477419355$$

wavelengths. Because the computer model used to calculate the Yagi's performance inputs the *half*-length of *DE* instead of its overall length, this value is divided by two and rounded to three places to give 0.239 wave. This decoding scheme is used to evaluate each of the Yagi's design parameters. The *DIR* radius, gene No 7, for example, evaluates to 0.0015 wave, and so on.

The genetic algorithm begins by creating an initial population of random 37-bit chromosomes. It then applies the operators of 'selection', 'crossover', and 'mutation' to filter out 'unfit' designs while retaining the better ones.

Successive applications of these operators create 'generations' of antenna designs, with each subsequent generation hopefully containing better designs than the previous one. But, because of the algorithm's inherently random nature, there is no guaranty of obtaining better designs. They may actually become worse from one generation to another.

Well-designed genetic algorithms, however, usually produce progressively better designs, at least on the average. Every new run holds the intriguing possibility of producing a previously unseen 'best' design.

The selection operator determines which chromosomes are fit enough to survive to the next generation. Some may be automatically discarded – for example, the worst 10% – while others are typically 'killed' at random, as they would be in nature. Others may be automatically retained – the best 5%, for example.

The algorithm designer is free to implement whatever selection process seems best. The crossover operator 'mates' two chromosomes, or 'parents', to produce two new chromosomes, or children', which become members of the next generation. Child chromosomes usually maintain a constant population from one generation to the next, although the population could grow if desired.

Each parent's chromosome is split at a gene boundary, usually randomly selected, and the pieces are swapped (concatenated together) to form two different chromosomes. This is the primary process by which genetic algorithm propagate 'good' genes from one generation to the next.

Finally, the mutation operator randomly flips a bit here and there with some small probability. This simulates the genetic mutation that occurs randomly in Nature.

Deciding which is best

In each generation, all of the designs, or chromosomes, are ranked from best to worst using a figure-of-merit. The figure-of-merit combines various antenna performance measures

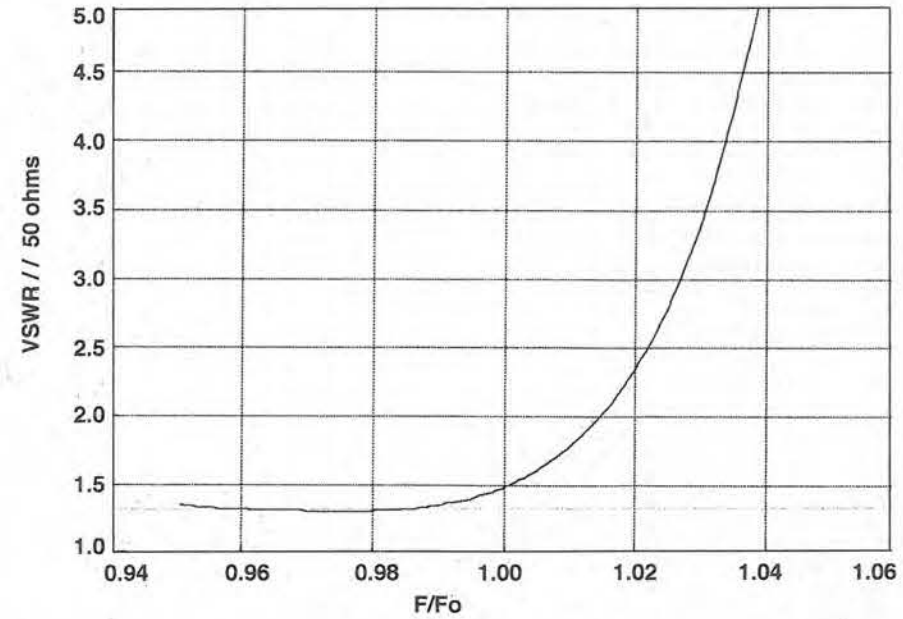


Fig. 4. Standing-wave ratio performance.

computed by a modelling engine, which is another computer program separate from the genetic algorithm.

Individual antenna performance parameters, for example, can be calculated with any suitable antenna modelling program. The figure-of-merit used for the Yagi described below is

$$\frac{5(G)+4(FB)-SWR}{10}.$$

This particular figure of merit gives slightly more weight to the main lobe gain *G* than to the front-to-back ratio *FB*, and relatively less weight to the input standing-wave ratio *SWR*.

The algorithm designer is free to define any figure of merit that reflects the relative importance of different performance measures, including even non-electrical parameters such as cost or time to build, or amount of material required, and so on. This feature is a major distinction between genetic algorithm and deterministic optimisations, which frequently cannot optimise arbitrary figures of merit.

Other significant differences are that genetic algorithms produce *groups* of designs with similar figures of merit, instead of the single 'best' design, and they usually require much less computer time than deterministic algorithms.

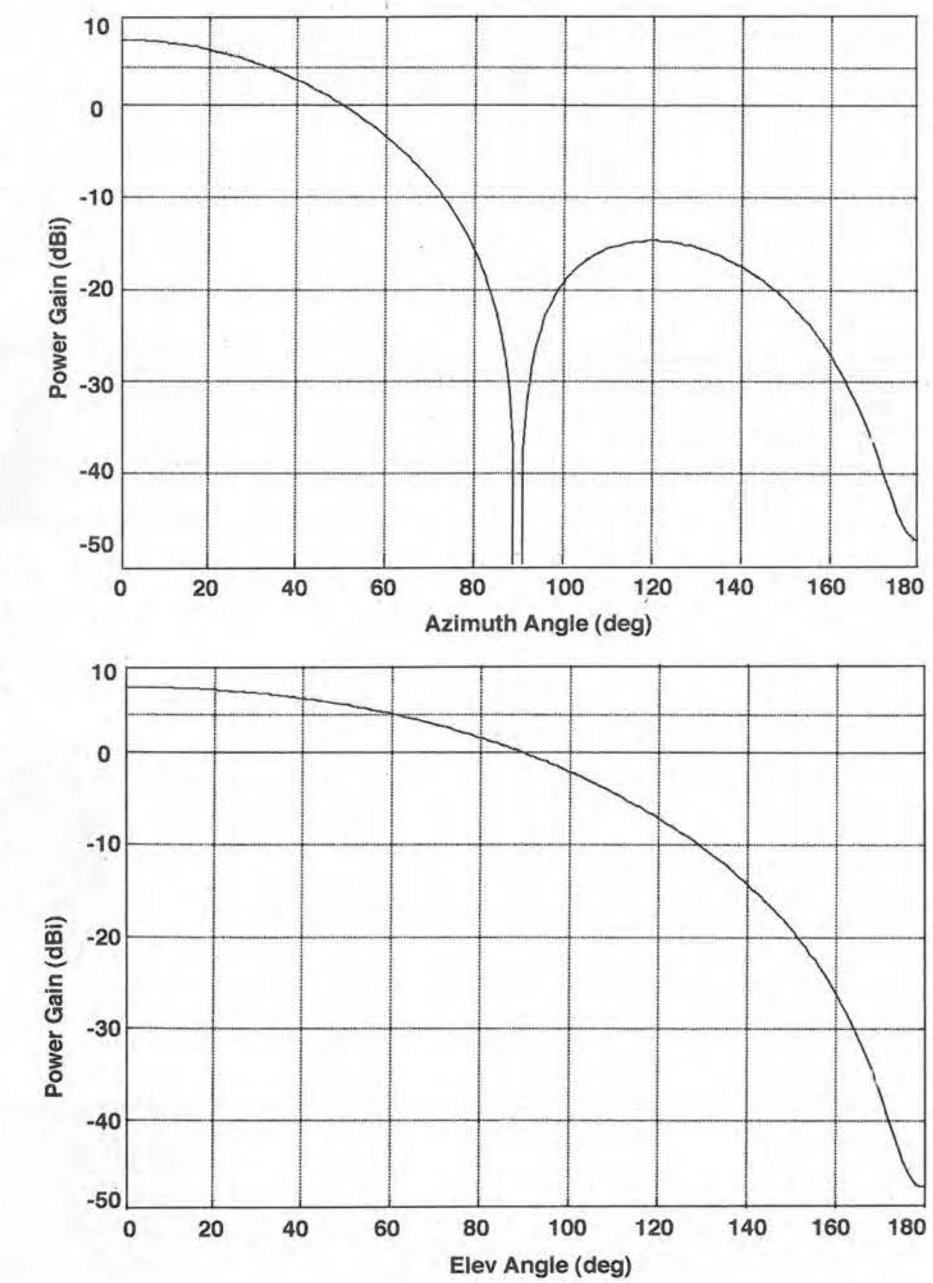
The genetically optimised three-element Yagi has the following dimensions, in wavelengths at the design frequency F_o:

Reflector length	0.530
Reflector radius	0.0008
Driven element length	0.478
Driven element radius	0.004
DE distance from REF	0.123
Director length	0.446
Director radius	0.0015
DIR distance from DE	0.106

The boom length – the sum of *DE*/*DIR* separations – is only 0.229. This is less than a quarter-wave, which is quite short. At the 6m amateur band frequency of 51MHz, for example, this Yagi is only 53in long. The *REF*, *DE* and *DIR* lengths are 122.66, 110.62, and 103.22in, respectively, with diameters of 0.37, 1.85, and 0.694in.

Gene *DE* is located 28.47in from *REF*, while *DIR* is located 24.53in from *DE*. It is interesting that the genetic algorithm converged to the maximum allowable value for the *DE* radius, because it is known from analytical considerations that increasing *DE* diam-

Fig. 5. Azimuth, a), and elevation pattern, b), of the antenna.



eter can improve Yagi performance substantially[3].

Free-space main lobe gain, front-to-back ratio, input impedance (resistance and reactance), and standing-wave ratio relative to 50Ω are plotted in **Figs 1-4**, respectively. These parameters were computed over a 10% band centred at the design frequency F_o.

The azimuth and elevation patterns at F_o appears in **Figs 5a)** and **b)**. Key performance measures are shown in **Table 2**.

The band-centre gain of 7dBi is typical of well-designed three-element Yagis, and the optimised antenna's *FB* of 54dB is exceptionally good. For comparison, this *FB* figure is more than 16dB better than the best *FBs* of typical quarter-wave designs described in W2PV's treatise on Yagi antennas[4] (see especially Fig. 2.9).

The optimised antenna also exhibits good *FB* bandwidth, with values exceeding 20dB from $0.97F_o$ to $1.017F_o$, which equates to 4.7%. The optimised Yagi is nearly resonant at F_o at an input reactance of 3Ω capacitive, which is less than 10% of the input resistance.

From $0.95F_o$ to $1.015F_o$, a difference of 6.5%, the standing wave ratio is less than two. If desired, this antenna can be fed directly with 50Ω coaxial cable, eliminating the insertion loss introduced by a matching network or antenna tuner.

Table 2. Performance of the genetically designed Yagi.

Gain	FB	Z_{in}	SWR	HPBW
7dBi	54.2dB	33.9–j3Ω	1.49	66°az, 122° el

Of course, a balun should be used to maintain feed system balance. But it would be interesting to build this antenna with and without a balun to see how much difference it makes.

For the 51MHz design, the standing-wave ratio is below two, and the *FB* is greater than 20dB, from 49.47 to 51.76MHz – a bandwidth of 4.5%. The lower band edge can be shifted up to 50MHz by increasing the design frequency to F_o=51.55MHz and recalculating the dimensions. Note that the wavelength is computed as $299.7956/F_{MHz}$, which is more accurate than the commonly used formula $300/F_{MHz}$.

The optimised Yagi's E-plane azimuth pattern has a characteristic two-lobe structure with a deep broadside null. The –3dB half-power beamwidth is 66°. The rear lobe is about 22dB down, which is quite low. The H-plane elevation pattern is plotted in Fig. **5b)**. It has a single, broad lobe with half-power beamwidth at 122°.

The genetically optimised, three-element Yagi is a very compact antenna that provides excellent performance. This example illustrates that genetic algorithms can produce very good antennas indeed. Such algorithms are easily implemented on a pc and can provide significant advantages over deterministic techniques.

Communications engineers will probably hear more and more about the genetic design approach. It certainly merits serious consideration by designers who are interested in antennas. ■

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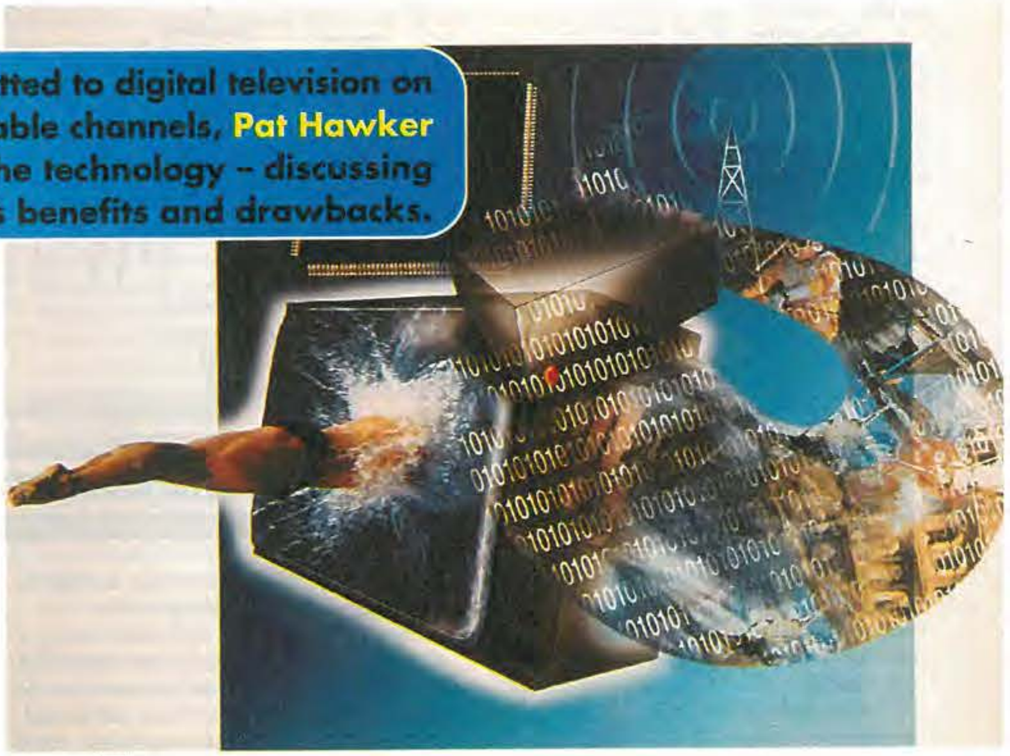
With the UK now committed to digital television on terrestrial, satellite and cable channels, Pat Hawker looks at the technology – discussing both its benefits and drawbacks.



Digital tv broadcasting

The BBC and ITC are now actively working towards an early launch of digital terrestrial television, or DTT for short.

Current viewers have a massive investment in analogue television receivers, many of which will last a good few years. Will they want to buy further equipment that will last them well into the 21st Century? Now is a good time to review the technology being held as opening a new era of multiple-choice television programming – and some of its inherent problems.

Unquestionably, the recent ETSI standards for digital video broadcasting, or dvb, represent an ingenious, if complex, means of squeezing many more programme channels into the available bandwidth. But you can take many of the claims made relating to the quality of digital tv broadcasting with a pinch of salt.

Picture quality will not be perfect, but subject to a new impairment caused when the data rate is reduced. As for wide-screen pictures, analogue PALplus has already been relegated to the back burner because British viewers are reluctant to accept 'letter-box' formats on 4:3 receivers.

Sound quality may be at compact disk levels, but then so is the current Nicam system. Picture resolution will be high too, but it is a pity that the millions spent on developing HD-MAC have largely been flushed down the drain along with the other MAC/packet systems.

Benefits of digital tv broadcasting

But we must put away the suspicion that dvb has been developed by a multitude of committees and may turn out to be a camel rather than a horse. The attraction of dvb to governments is the eventual freeing up of sellable vhf and uhf spectrum; to broadcasters dvb means extra channels available for pay-tv with a convenient conditional access system for terrestrial channels; for viewers, dvb will offer near unlimited 'choice' to those willing to pay subscriptions or pay-per-view. And for industry there will be the incentive of new set-top timers/decoder equipment that will be offered in advance of fully integrated terrestrial-satellite-cable receivers.

It is also worth noting that funding for the BBC will come from the Government ordered sale of its transmitters to the US firm Castle Television for £244-million.

Background to digital tv

As early as 1992, authorities in many countries were beginning to accept that digital television broadcasting systems were the way forward. This meant the eventual demise of enhanced hybrid analogue/digital systems such as MAC/packet or HD-MAC for satellites. In addition, only limited use is likely to be made in the UK of PALplus. This system can provide wide-screen displays with a 16:9 aspect ratio on compatible terrestrial receivers. It also offers less cross-colour impairment than conventional PAL, but involves the unpopular 'letter box' display on traditional receivers designed for 4:3 aspect ratio. But then digital 16:9 will also involve 'letter box' until receivers with wide-screen picture tubes become the norm.

In the early 1990s, research work funded by IBA showed that cofdm, short for coded orthogonal frequency division multiplex, was a multi-carrier modulation system that could overcome the long recognised problem that multipath reception seriously affects high-

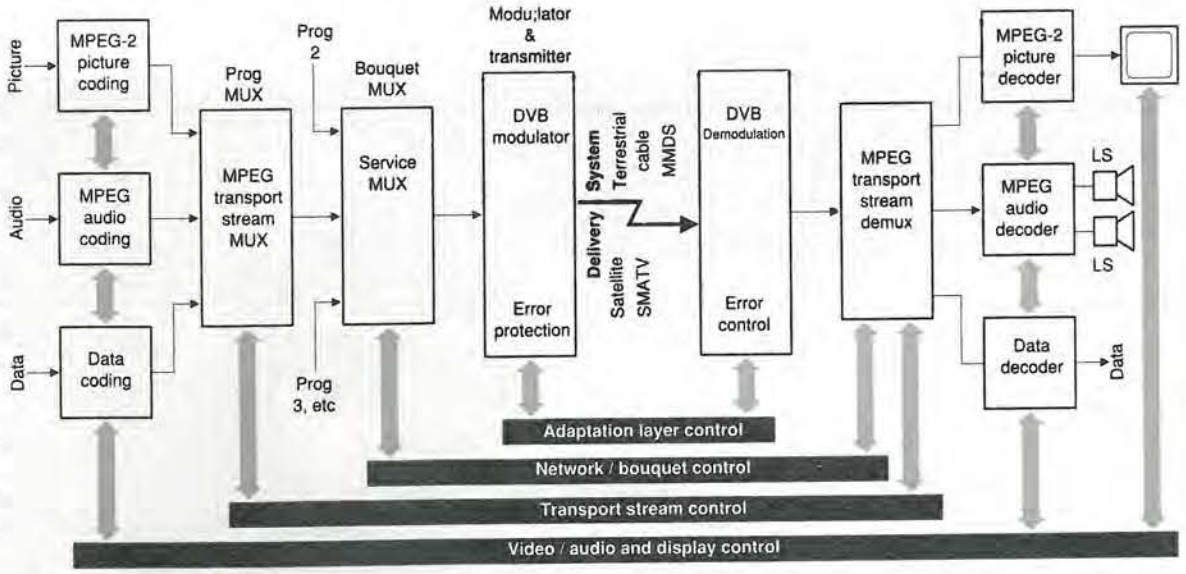


Fig. 1. Overall view of the European digital video broadcasting systems. Each delivery system provides a data container, best described as housekeeping (service information), and a multiplexed 'bouquet' comprising programmes and any other service offered by the provider.

speed digital data signals on terrestrial channels. When combined with bit-rate reduction, cofdm offered a realistic prospect of multiplexing a number of programmes within a single channel.

Earlier, in the 1980s, British and European engineers had shown that 216Mbit/s 625-Line component video, described in ITU-R Recommendation 601, could be reduced to about 34Mbit/s with virtually no degradation of picture quality. It could also be successfully applied to microwave and satellite contribution, as well as distribution links.

A turning point

Then, at IBC-1990 in Brighton, General Instruments demonstrated, using a short computer simulation, that its 'Digicipher system held the promise that much greater bit-rate reduction factors seemed feasible by using complex digital signal processing that took full advantage of the very large amount of redundancy in typical video material.

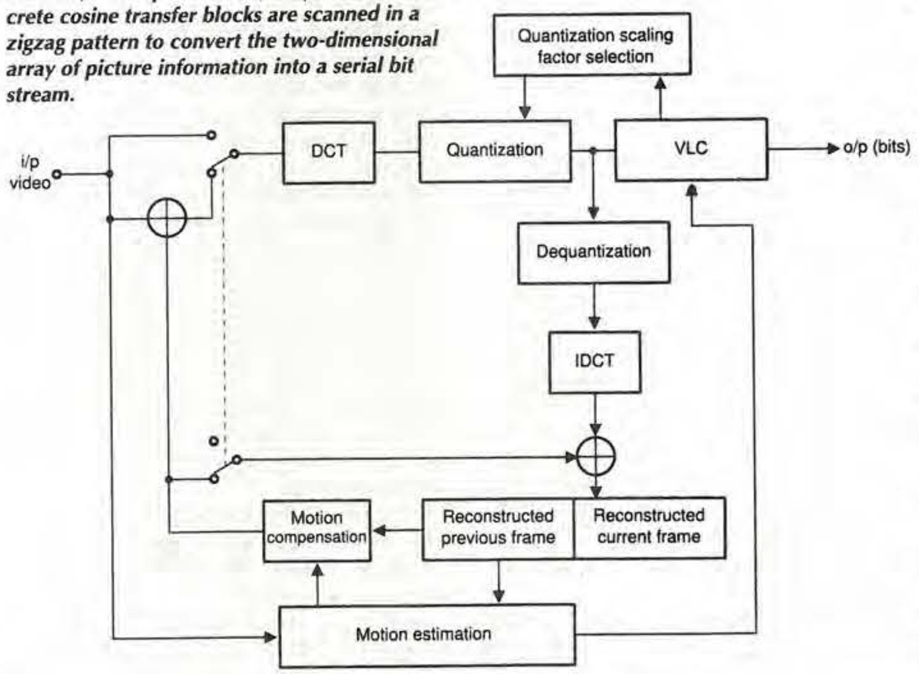
It was soon shown that such results were not confined to computer simulations. An ITU-R 601 216Mbit/s component bit stream can be reduced to 5Mbit/s or less, while still providing a picture quality subjectively no worse than analogue 625-line PAL.

A digital video broadcast project in Europe was formally set up in the autumn of 1993. It grew to contain over 200 members, representing broadcasters, programme production houses, transmitter companies, satellite operators, consumer electronics manufacturers, regulatory bodies and government representatives from over 25 countries.

The European Broadcasting Union in Geneva provided project management. In the USA, an advanced tv standard has been developed from a grand alliance based on the merger of several different proposals for advanced tv. This standard describes a system capable of providing high-definition pictures within the standard American 6MHz terrestrial channels. But it differs from the standards stemming from the European DVB Project in not accepting the use of cofdm.

The aim of Project DVB has been to formulate, with the maximum degree of commonality, practical digital transmission specifications for all forms of programme delivery. These include satellite using the DVB-S service, cable using DVB-C, terrestrial systems involving DVB-T and satellite master-antenna tv, or SMATV with DVB-CS. A multi-media system, namely MMDS will use DVB-MS and have an interactive channel.

Fig. 2. Simplified block diagram of an MPEG-2 encoder, After quantisation, 8 by 8 blocks of discrete cosine transfer blocks are scanned in a zigzag pattern to convert the two-dimensional array of picture information into a serial bit stream.



Dual subsystems

Two sub-systems common to all forms of delivery are service information, or SI, and fixed-format teletext, or DVBTXT. It is envisaged that a digital specification for an interactive channel will also be developed. All systems use the broadcast video and audio bit-rate compression systems of the Motion Picture Experts Group system MPEG.

It is claimed that much emphasis has been given to market requirements. This was not so with the earlier MAC/packet work, which reflected primarily the initiative of broadcast engineers.

For each class of delivery system, appropriate ETSI specifications have been developed, linked, initially, to MPEG-2 bit-rate-reduction coding. Each delivery system provides a 'data container' with a multiplexed 'bouquet' of programmes and any other data services offered by the service provider. An overview of the DVB system is shown in **Fig. 1**.

MPEG-2 bit-rate-reduction

Initially a Working Group of ISO/IECT JTCI, MPEG began life in 1988 with the aim of defining standards for digital compression of video and audio signals.

The group's first task was to define a video coding algorithm for digital storage media – in particular cd rom. In 1993 this system, MPEG-1, was published in three parts as ISO/IEC 11172. It is restricted to non-interlaced video formats, primarily used for pc applications. MPEG-1 supports video coding at bit-rates under 1.5Mbit/s.

In 1990, MPEG began work on an algorithm capable of coding interlaced pictures directly, and suitable for high-quality applications at bit rates up to about 10Mbit/s. It was later extended to cover high-definition formats in the range of about 15-30Mbit/s. This system, MPEG-2, is specified in ISO/IEC 137132. It was published in 1994.

Programme-dependent bit-rate adaption is provided for in MPEG-2. For standard-definition 625-line broadcast tv with MPEG-2 bit-rate-reduction, cinema film requires a data rate of 2.5 to 4Mbit/s. Film with only 24 frames a second can look better than electronically-generated interlaced tv pictures at equivalent bit rates. Educational programmes are allocated 2-4Mbit/s, general programmes 3-7Mbit/s, sport 5-11Mbit/s and audio 64-256kbit/s.

It has been claimed that MPEG-2 coding can provide high quality pictures and sound with data rates as low as 1.5 Mbit/s. For transmission however, additional forward error correction bits, etc, need to be added. Further research into sophisticated signal coding algorithms is leading to systems such as MPEG-4. These will permit hdtv pictures to be transmitted in the American 6MHz terrestrial channels.

How does MPEG work?

Video bit-rate reduction, i.e. digital compression, removes redundant and less important information from a video signal prior to transmission. Then, the receiver reconstructs an approximation of the original images from the information reaching the decoder.

In practice, there are three main forms of redundancy and all are exploited in MPEG-2. First are spatial and temporal redundancies. Pixel values are not independent but are correlated with neighbouring pixels both within the same frame and across frames. This implies that for most images, the value of any pixel is predictable.

Second comes entropy redundancy. In any non-random digitised signal, some code values occur more frequently than others. This enables shorter codes to be used for the more frequently occurring values. Compare this with Morse telegraphy, where frequently used letters have short symbols. The letter E for example is a single dot.

Lastly there is psycho-visual redundancy. By exploiting the limitations of the human eye and brain, information beyond the limits of spatial resolution – fine detail which the eye cannot resolve – and temporal resolution which is the limited ability of the eye to track fast-moving images.

For MPEG-2 an assemblage of compression techniques are used in practice. However, unlike for MPEG-1, reduction of the horizontal sampling rate is not advisable for luminance or chrominance signals. Nor is temporal sub-sampling. Nevertheless, the frequency of the vertical chrominance can be halved with MPEG-2 suitable for 4:2:0 as well as 4:2:2 and 4:4:4 component video.

Removing redundancy

There are two main methods of exploiting redundancy. In discrete-cosine transform, or dct, an orthogonal transform similar to a discrete Fourier transform assists the processing that removes spatial redundancy. It does so by concentrating signal energy into relatively few coefficients. With the second method, motion compensated interframe prediction is used to remove temporal redundancy. This technique is similar to differential pulse-code modulation, dpcm. An outline of a MPEG-2 encoder is shown in **Fig. 2**.

After quantisation, the 8 by 8 blocks of dct coefficients are scanned in a zigzag pattern to convert the two-dimensional array into a serial string. They are then coded by counting the number of zero coefficients preceding a non-zero coefficient. This is run-length encoding. Finally they are combined with the non-zero coefficients and coded as a variable-length code, or vlc.

The discrete cosine transform coefficient quantisation, the run-length coding and the variable-length coding processes produce a varying bit-rate. The variation depends on the complexity of the picture information and the amount and type of motion in the picture.

To achieve the constant bit-rate needed for transmission, a buffer is required to smooth out the variations. To prevent overflow or underflow of this buffer, its occupancy is monitored with feedback applied to the coding processes to control the input.

An important feature is that the final bit rate at the output of the encoder can be freely varied; if this is reduced, the buffer will empty more slowly. The encoder compensates for this by making the dct coefficient quantisation coarser, reducing the quality of the decoded picture. Increasing the final bit rate improves the decoded picture quality. There is no requirement to lock input sampling rate to channel bit-rate or *vice versa*.

You should appreciate that this is only a bare outline of an extremely sophisticated bit-rate-reduction system which includes several methods for motion-compensation prediction. These involve both forward and backward prediction from other frames, or interpolation by averaging a forward and a backward prediction.

Where MPEG falls down

While MPEG-2 processing is well suited for the final distribution and broadcast transmission, it is far from ideal for contribution links. Here, further editorial processes, such as chroma keying, are likely to be required at the studio centre. For international satellite links etc an international 34Mbit/s standard has been agreed (ETS 300 174/ITU-T J81).

According to the ITC specification for UK television broadcasting, video encoding has to conform to ISO/IEC 13818-2 with the additional guidelines of ETSI-54. Mandatory video parameters include a frame rate of 25Hz and an aspect ratio of 4:3 or 16:9.

Mandatory audio parameters of ISO/IEC 13818-3 and guidelines ER 154 involve stereo at a sampling frequency of 48kHz with no pre-emphasis. Bit rates recommended for stereo are 256kbit/s and for joint stereo 192kbit/s.

The transport-stream multiplexing feature of MPEG allows a large number of video, audio and data services to be merged into a single

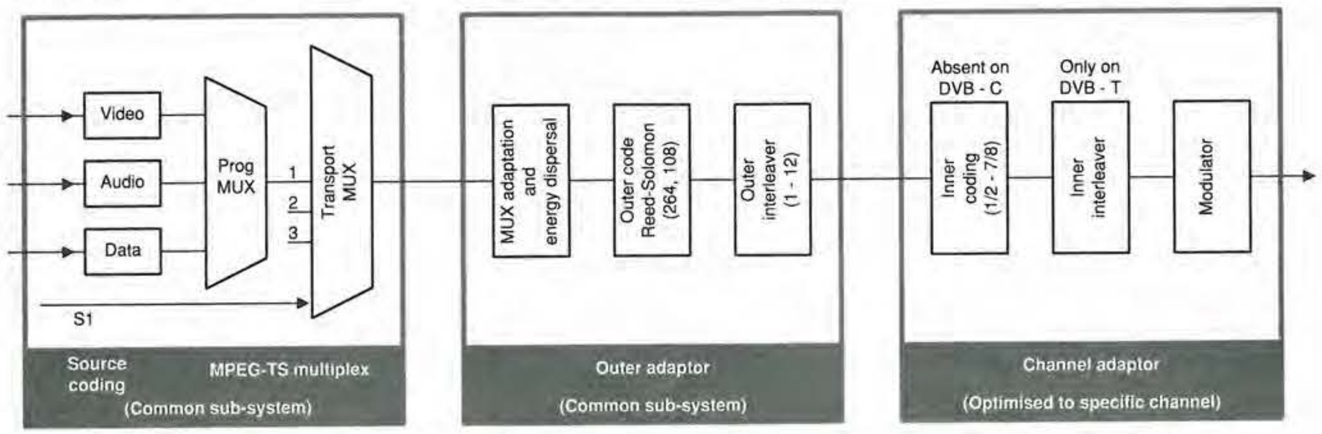


Fig. 3. Basic transmission chain for digital-video broadcast systems, showing how the error protection is added to output from the MPEG transport stream multiplexer. The first two blocks are common to all channels. The channel adapter is optimised depending on the type of information that the channel is carrying.

transmission stream. Packets are 188 bytes long and delimited by a sync byte.

Since no error protection is provided rugged forward error protection must be added to provide as near possible an error-free data stream to the receiver decoder. **Figure 3** shows how the MPEG-2 data stream is further encoded to provide the degree of error protection required by the various delivery systems.

In brief, all first-generation Project DVB systems use the common MPEG-2 transport-stream multiplex. They also use a common service information system providing the decoder with details of the transmission system, and the services being broadcast. All systems incorporate common Reed-Solomon forward error correction system. Where needed by the delivery system, they also have a common punctured convolutional code forward error correction system. Digital video broadcast service information – mandatory and optional – is specified in prETS 300 468, plus guidelines on its implementation and use.

MPEG housekeeping

Figure 3 shows how error protection is added to the MPEG transport stream multiplex with an outer adapter which is common to all systems and then a channel adapter optimised to the specific delivery channel. The programme-specific information within the service information includes the programme association table, programme map table and the conditional access table.

Service information also conveys the packet and transport packet layer guidelines, the time and date and three tables. The network information table represents the actual delivery system, the service description table is the actual delivery stream) and the event information table describes present/following events only.

Transmission parameter signalling, or tps, in effect tells the receiver which of the many options is in use. Since broadcasters are unlikely to change the modulation options frequently however, it is not essential to decode this information. Broadcasters are more likely to re-configure their multiplex to carry more or fewer programmes by changing the bit rates allotted to individual channels.

Pilot information is provided in two forms. A few carriers – called continual pilots – always have defined amplitudes and phases and are used primarily for synchronisation. Additional 'scattered pilots' are inserted in a regular pattern to facilitate coherent demodulation. ■

In a subsequent article, Pat will describe digital delivery systems.

Digital television glossary

ACATS – Advisory Committee on Advanced Television Systems (USA)
ATV – advanced television
BAT – bouquet association table (part of SI)
BER – bit error rate
B-pictures – bidirectionally predictive pictures (motion compensation)
BRR – bit-reduction rate.
CA – conditional access
CAT – conditional-access table (part of SI)
COFDM – coded orthogonal frequency-division multiplexing
CPE – common phase error
DAB – digital audio broadcasting
DBPSK – differential binary phase-shift keying.
DCT – discrete cosine transform
DPCM – differential pulse-code modulation
DTT – digital terrestrial television
DVB – digital video broadcasting, suffixed S for satellite, C for cable, T for terrestrial, CS for SMATV, TXT for fixed-format teletext and MS for MMDS.
EBU – European Broadcasting Union
EIT – event information table (part of SI)
EPG – electronic programme guide
ETS – European Telecommunication Standard
ETSI – European Telecommunication Standards Institute
FEC – forward error correction
FFT – fast Fourier transform
GOP – group of pictures (motion compensation)
IBC – International Broadcasting Conference
ICI – inter-carrier interference
IDCT – inverse discrete cosine transform
IEC – International Electrotechnical Commission
I-pictures – intra pictures (motion compensation)
IRD – integrated receiver decoder
ISO -International Standardisation Organisation
ITU – International Telecommunication Union
MMDS – multichannel, multipoint distribution system, alternatively, multipoint microwave distribution system
MPEG – video bit-rate reduction systems determined by the Moving Picture Experts Group
NIT – network information table (part of SI)
OFDM – orthogonal frequency division multiplexing
PAT – programme association table (part of SI)
PMT – programme map table (part of SI)
P-pictures – predictive pictures
PPV – pay-per-view
PRBS – pseudo random binary sequence
PSI – programme-specific information (part of SI)
PSP – programme service provider
QAM – quadrature amplitude modulation
QEF – quasi-error-free
QPSK – quadrature phase-shift keying
RLC – run-length coding
RS – Reed Solomon error protection
RST – running status table (part of SI)
SDH – synchronous digital hierarchy
SDT – service description table (part of SI)
SFN – single-frequency network
SI – service information, or housekeeping details added on to the video, audio and/or multi-media data stream
SMATV – satellite master antenna television
SMS – subscriber management system
ST – stuffing table (part of SI)
TDT – time and date table (part of SI)
TOT – time offset table (part of SI)
TPS – transmission-parameter signalling
TS – transport stream
VLC – variable-length coding

Changing times

For very little outlay, Ed Buckley's atomic clock interface takes the rather unfriendly serial data from an MSF receiver module and turns it into an RS232 time and data stream using a Basic Stamp. Output is a bcd data stream containing year, month, date, day, hour and minutes.

The *Basic Stamp* is a versatile, easy to programme microcontroller. As well as a stand alone unit the *Stamp* is also a very handy interface device- reading in non-standard data, performing some manipulation and then passing the data on in a more usable format.

The following design illustrates this function by combining a *Stamp 1* module with a readily available atomic clock module. To finish up, I will then describe how you can have your cake and eat it too – achieve low software development time by using the *Stamp1* and low production costs by using a standard *PIC*.

Design overview

We recently needed to fit a data logger to a series of truck-trailers. The data needed time stamping – no pun intended – but there was a serious risk of interruption of power to the real-time clock module.

The obvious solution was to incorporate an operator keypad to allow for clock resetting. As well as being expensive, this solution allowed the opportunity for operator error.

So we decided to look for a different solution. After looking at the atomic clock module now available, we decided to go down this route. Costs would be about the same as a keypad interface, but this alternative has the advantage that it is fully automatic.

Atomic clock summary

Atomic clock modules were featured in *Electronics World* March 1996. Here I will summarise the important features.

The off-air data stream is repeated every minute. High speed data is sent in the first 17 seconds, followed by null data that can be discarded. Next, low speed data is transmitted at 1bit/s. There is always a unique data format at the end of the minute period, which is 01111110_2. Data timing is shown in **Fig. 1** and the minute's data format in **Table 1**.

The circuit diagram with Stamp/Clock hook-up is shown in **Fig. 2.**

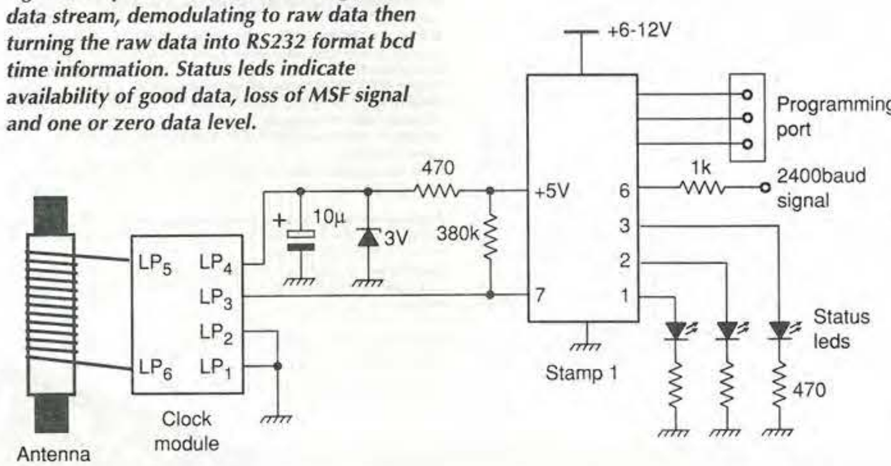


Fig. 1. Complete circuit for receiving the MSF data stream, demodulating to raw data then turning the raw data into RS232 format bcd time information. Status leds indicate availability of good data, loss of MSF signal and one or zero data level.

Writing the software

For our application, we require a burst of correct time data at 2400baud once a minute. The main system processor – a *Stamp 2* in this case – polls the signal line from the *Stamp 1* clock module. If valid data is present, it updates the system real-time clock.

The *Stamp 1* interfacing with the atomic clock is dedicated to reading and checking the validity of the incoming radio data. It only transmits the data if it proves valid.

We envisaged that the main problem would be loss of radio signal as the trailer moved around. For this reason the software always counts the number of data bits received from the last required data set until the unique format byte is received – in our case seven. If the counter is anything but seven, then a bit has been lost or found, the data is suspect and there is no transmission to the main system.

List 1 is the program listing. The input and output pins are allocated names to ease programme tracing and ram space allocated to the programme variables. The *Stamp 1* has up to 14 bytes of ram which may be addressed as bytes, for example b_3 or b_7, or words, as w_0 or w_3, where $w_3=b_6+b_7$. Further, two bytes may be addressed by their individual bits, that is bytes b_0 and b_1 provide bits bit_0 to bit_{15}.

The dirs statement sets the pins to either input if zero or output if logic one; the msb comes first.

We initially look for the 01111110_2 signature by reading the individual data bits using the Get_bit routine into the variable *signal*. The variable *counter* is incremented every time a new bit is read.

Once the signature has been received we now know where we are and proceed to 'continue' where the *good_data* flag is set to false. Here we check to see if we have a good data set by looking at *counter*. If *counter* is does not hold seven, then the data is bad. If the counter does hold seven then the data is good and the *good_data* flag is set to true. This condition allows transmission of the time data ten times during the fast data period at the beginning of the minute period.

Now begins the slow speed data section. The first 16 seconds are unused and discarded. Next we log the year, month, day of month, day of week – which we discard – hour of day and minute.

Now the program returns to the beginning. If all is well the next eight data bits should be the unique signature and counter should return seven to allow transmission.

Most of the grunt work is handled by two subroutines – Get_bit and Read_convert.

Get-Bit. Get_bit decodes the data from the atomic clock module.

The data falling edge is first detected. Once detected, the programme pauses for 150ms before reading the data – if logic one is returned the data is zero and if zero, a one is returned. Note that the clock gives inverted data, which we correct in the line bit0=clk+1. The program then waits a further 200ms to ensure the clock data is in the stable high portion of its cycle before continuing.

Read_convert. At the end of each chunk of received data we take the opportunity to convert it into a decimal and then a binary-coded decimal format for easy reading by the system processor. Our main system clock uses bcd format so using bcd minimises the calculations later on.

Rounding up

To aid fault finding and add value, we included several leds – the i/o pins were there unused so why not spend 15p? The three leds have the following functions.

- Sync_flag led is on when the *good_data* flag is set, ie the previous minute's data was good.
- Bit_read flashes every two seconds – a stable on or off here indicates loss of incoming signal.
- Data_bit indicates whether the data bit received was a one or zero data.

Check_sum is a byte variable. It is cleared every minute and all data is added to give a

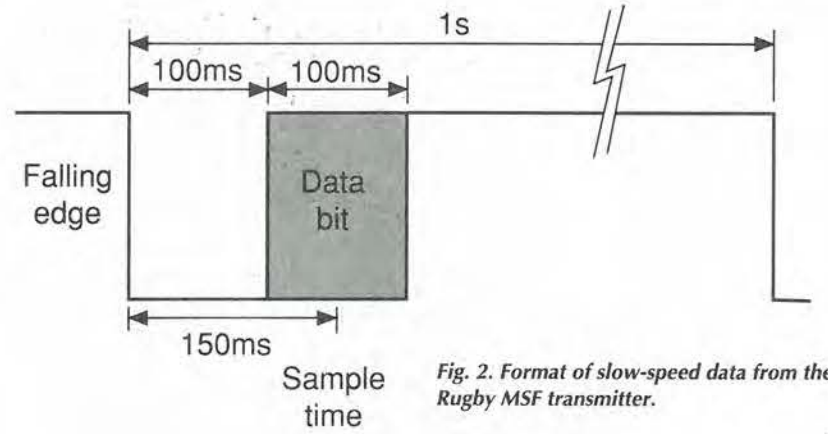


Fig. 2. Format of slow-speed data from the Rugby MSF transmitter.

Table.1. Time telegram of MSF transmitter – binary code for seconds 17 to 30 and data for March 1996. Note that each second value is one or zero. A one in second 17 for example represents 80 and a one in second 18 represents 40.

Second	Function	Value	Detail
0	Fast code	Not used in modules	
1	DUT1 code	Not used in modules	
2-16	–		
17	year (tens)	80	year 00-99, bcd
18	year (tens)	40	
19	year (tens)	20	
20	year (tens)	10	
21	year (units)	8	
22	year (units)	4	
23	year (units)	2	
24	year (units)	1	
25	month (tens)	10	month 01-12,bcd
26	month	8	
27	month	4	
28	month	2	
29	month	1	
30	day of month (tens)	20	day of month, bcd
31	day of month (tens)	10	
32	day of month	8	
33	day of month	4	
34	day of month	2	
35	day of month	1	
36	day of Week	4	day of week 1-7, bcd
37	day of Week	2	
38	day of Week	1	
39	hour (tens)	20	hour 00-23, bcd
40	hour (tens)	10	
41	hour (units)	8	
42	hour (units)	4	
43	hour (units)	2	
44	hour (units)	1	
45	minute (tens)	40	minute 00-59, bcd
46	minute (tens)	20	
47	minute (tens)	10	
48	minute (units)	8	
49	minute (units)	4	
50	minute (units)	2	
51	minute (units)	1	
52	always set to '0'	0	
53-58	always set to '1'	1	
59	always set to '0'	0	

resulting checksum which is used to detect errors between the *Stamp 1* and the main system processor – unsophisticated but works.

What about the cake?

We have demonstrated the versatility of the *Stamp 1* as an interface. The Stamp scores over raw *PIC*s in terms of speed of producing and debugging a simple programme.

If the application only requires a small number of units the *Stamp* is fine, either in its IC or discrete chip set format. For larger numbers however, the costs start to look a little heavy.

Realising this, Parallax, the manufacturers of the *Basic Stamp* has recently introduced a neat solution for *Stamp 1* users. By using the latest *Stamp 1* software and the Parallax *PIC* programmer, it is possible to programme a standard *PIC16C58* with both the *Stamp* interpreter and your programme. This gives low-cost development coupled with low cost production.

This system only works with the *Stamp 1*, *16C58* and Parallax's programmer, but it is a very useful tool for those quickie programmes that will fit into a *Stamp 1*. ■

Galleon, distributor of the clock modules, can be reached on 0121 359 0981.

List 1. Stamp 1 program to read and decode the Rugby MSF time clock.

```
' MSFCLCK.BAS
' Milford Instruments Original version dated March 1997

' Stamp decodes the one second pulses using Get_bit and
' keeps adding them to signal until the sync byte is received.
' First 17 seconds are ignored and remainder are clocked into
' relevant bytes. Results are converted to BCD format before
' storing in bytes. Time is only transmitted at the start of
' the minute - if data is judged to be OK. Program counter
' holding number of seconds from last valid time data
' until sync byte is received.. should be 7.

Symbol clk          =pin7   'clock signal on pin 7
symbol rs232        =6      'rs232 signal on pin 6
symbol signal       =b1     'byte to receive data
symbol counter      =b11    'general counter variable
symbol good_data    =bit1   'good time message flag, 0=false
symbol checksum     =b7     'data checksum variable
symbol sync_flag    =pin3   'in sync led
symbol bit_read     =2      'data bit received led
symbol data_bit     =pin1   'data bit rec. led, on for log. 1

dirs=%01001110    'set up the pin directions

'First look for the 01111110 signature
start:
 counter=0
 signal=0
 loop:
 gosub get_bit                          ' go and get a bit
 signal=signal*2+bit0                   ' add it to signal
 if signal=%01111110 then continue   'Sync byte received?
' continue if not and increase counter by one
 counter=counter+1
 goto loop

continue:                    ' sync byte now received
 good_data=0                 ' reset valid data flag
 if counter<>7 then skip1   'check that we've kept in step
'if good data captured then set flag
 good_data=1
 skip1:
 counter=0
 sync_flag=good_data         ' light the in_sync led

'Now send the good data transmission whilst waiting to get
'past the fast code section at the start of the minute
'Total loop period approximately 1.4s
 for counter=1 to 10          'send the good data at min.start
 if good_data<>1 then loop2   'check for good data flag
'Transmit if good data
 serout rs232,n2400,("T",b2,b3,b4,b5,b6,b7)
 loop2:                       'if not jump to here
 bit3=good_data+1
 b9=bit3*35+100
 pause b9                     'get past the fast data section
 next

'now discard the first 16 seconds of info
 for counter=1 to 16
 gosub get_bit
 next

'Now start to read wanted data
'Reset the checksum variable
 checksum=0
'Now read the year data
 b8=7
 gosub read_convert
 b2=b10             'result in b10 written to b2
 checksum=checksum+b2

'Now get the Month
 b8=4
 gosub read_convert
 b3=b10             'result in b10 written to b3
 checksum=checksum+b3

'Now get day of month
 b8=5
 gosub read_convert
 b4=b10             'result in b10 written to b4
 checksum=checksum+b4

'Now get the day of the week - and discard
 b8=2
 gosub read_convert

'Hour of Day
 b8=5
 gosub read_convert
 b5=b10             'result in b10 written to b5
 checksum=checksum+b5

'And the Minute reading
 b8=6
 gosub read_convert
 b6=b10             'result in b10 written to b6
 checksum=checksum+b6

goto start         'return for the next minute

'=====================================================
'                    Sub-Routine to retrieve a bit
'=====================================================

Get_bit:
Marker:
 if clk=0 then read_bit                          'detect falling
edge
 goto marker

read_bit:
 pause 150                                       'wait until the
middle of the A bit
 toggle bit_read
 bit0=clk+1                                      'read the value on
clk into bit0
data_bit=bit0
 pause 200
return

'==================================================
'                   Read and Convert Routine
'==================================================
Read_convert:
 b9=0                                            'Clear b9
 for counter= b8 to 0 step-1                     'one bit at a time
 gosub get_bit
 lookup counter,(1,2,4,8,10,20,40,80),b10  'Get the
appropriate scalar
 b9=b10*bit0+b9                            'Convert to true decimal
 next
 b10=b9/10*16                                    'now convert to
bcd- first the high byte
 b10=b9//10+b10                          'now add the low byte
 return
```

Multilayer air-cored coils

If you want to calculate the inductance of a coil and you know all the coil's other specifications, Wheeler's equation provides a simple and accurate solution. But the designer's task is usually the reverse. Robert Kesler describes the return path.

The most practical way to calculate the inductance of an air-cored multilayer coil is via Wheeler's formula.[1]

$$L=\frac{7.87N^2M^2}{3M+9B+10C} \quad (1)$$

This formula is claimed to be accurate to within 1%, if the numbers in the denominator are about equal, that is, if the shape of the coil is similar to that shown in the diagram. Of course, the accuracy in practice is also determined by the tolerances on the coil's dimensions.

A given piece of wire, wound in a coil, yields the highest inductance value, if its proportions are, $3M=9B=10C$. **It can be proved** that this shape yields the highest inductance for a coil, wound from a given piece of wire, or the highest inductance/resistance ratio for a given weight of copper.

The Wheeler formula is fine for calculating the inductance of a coil when everything else is known. But usually, the designer's task is the opposite: at the outset, only the inductance is known. In addition to the four other parameters in the formula, the wire diameter has to be found too.

Working back

A coil's dc resistance can be expressed using the diameter and the length of the copper wire, which can be expressed using the number of turns, and the mean diameter of the coil,

$$R=\frac{NM}{14250W^2} \quad (2)$$

The relationship between the number of turns that can be packed into the cross section of the coil and the wire diameter is,

$$N\left(\frac{W}{P}\right)^2=BC \quad (3)$$

Considering only the ideal shape, there are two more useful equations,

$$B_i=\frac{M_i}{3} \quad (4)$$

$$C_i=0.3M_i \quad (5)$$

In this case, formulas 1 and 3 become simpler,

$$L=0.875N_i^2M_i \quad (6)$$

$$N\left(\frac{W}{P}\right)^2=0.1M_i^2 \quad (7)$$

Now the parameters can be expressed, using the formulas 2, 6 and 7,

$$M_i=0.354\sqrt{\frac{L}{R}} \quad (8)$$

$$N_i=1.07\sqrt{\frac{L}{M_i}} \quad (9)$$

$$W_i=0.253\frac{M_i}{\sqrt{N_i}} \quad (10)$$

the inner and the outer diameter of the coil,

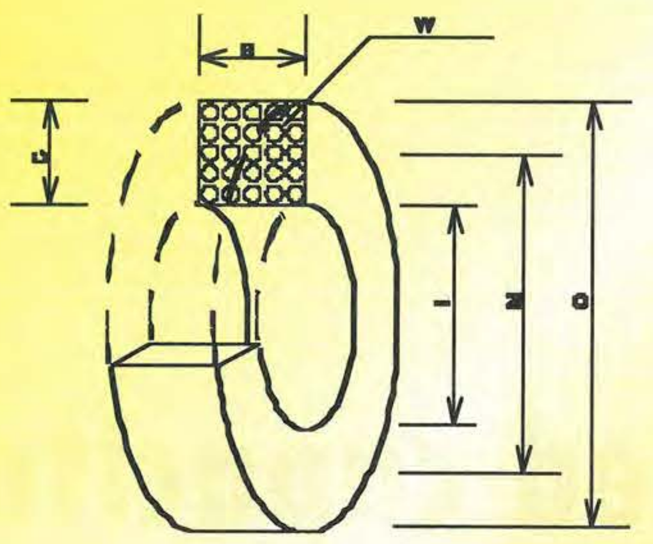
$$I=M-C=0.7M \quad (11)$$

$$O=M+C=1.3M \quad (12)$$

and two more useful relationships,

$$M_i=3.08L^{0.2}W^{0.8} \quad (13)$$

$$N_i=0.61\left(\frac{L}{W}\right)^{0.4} \quad (14)$$



L inductance in nanohenries
R resistance in ohms
N number of turns
M mean diameter
B width or length
C radial thickness
I inner diameter
O outer diameter
W diameter of the copper wire
P linear packing density*

(the wire diameter divided with the centre-to-centre wire spacing)
all dimensions in mm.

Air-cored coil features.

Note that formulas containing 'i' subscripts are valid only for ideally shaped coils.

Tips for designing multilayer coils

The above formulas are useful tools for designing multilayer air-cored coils. Two starting points are possible. With the first, the dc resistance is initially specified. For the second, dimensions of an available coil former need to be known.

If the coil is to be used in a high power circuit such as a loudspeaker crossover network, start by specifying the maximum allowable losses in terms of the dc resistance. Calculate the theoretical wire diameter using formulas 8, 9 and 10. Choose the nearest available standard wire size.

The calculated diameter need not be adhered to rigidly, bearing in mind that increasing the diameter will decrease the resistance, but increase the dimensions, and *vice versa*.

Use the chosen wire diameter to calculate the dimensions and the number of turns with formulas 1, 4, 5, 11, 12 and 14.

When working out from a given coil former, calculate the number of turns and the wire diameter, using formulas 1 and 3, with the dimensions of the former. Choose the nearest standard wire size. Using these new dimensions, recalculate the radial thickness and the mean diameter of the coil, then the number of the turns.

Calculate the dc resistance and decide whether or not it is acceptable. Note that in all the calculations, the linear packing density is assumed to be 0.8. Any errors that may arise from this assumption and other causes can be corrected at the final stage.

Measure the diameter of the ready wound coil. Before cutting the wire, calculate the inductance using formula 1 and correct the number of turns, as necessary.

Design examples

This first example shows how to proceed when the initial known quantity is the dc resistance of the coil. The inductance required is 200μH and the resistance assumed is 0.3Ω.

For mean diameter, number of turns and wire diameter,

$$M_i=0.354\sqrt{\frac{200.10^3}{0.3}}=28.9mm \quad (8)$$

$$N_i=1.07\sqrt{\frac{200.10^3}{28.9}}=89 \quad (9)$$

$$W_i=0.253\frac{28.9}{\sqrt{89}}=0.775mm \quad (10)$$

At this point, the calculated wire diameter can be modified to suit available sizes, provided that you understand the consequences.

Let the new wire diameter be 1mm. Use this diameter in the subsequent calculations,

$$M_i=3.08\times200000^{0.2}\times1^{0.8}=35.4 \quad (13)$$

$$B_i=\frac{35.4}{3}=11.8 \quad (4)$$

$$C_i=0.3\times35.4=10.6 \quad (5)$$

$$I=35.4-10.6=24.8mm \quad (11)$$

$$O=35.4+10.6=46mm \quad (12)$$

$$N_i=0.61\left(\frac{200000}{1}\right)^{0.4}=132 \quad (14)$$

Having finished the winding, remember to measure the outer diameter of the coil before cutting the wire. Calculate the inductance from formula 1, and correct the number of turns, if necessary.

The second design example is for when you are working from a specified coil former. Inductance required is 2mH. Assume that the former has an inner diameter, *I*, of 12mm, a width, B of 16mm and a rim, *C*, of 6mm.

$$2\times10^6=\frac{7.87\times N^2\times18}{3.18+9.16+10.6} \quad (1)$$

N=450

$$450\left(\frac{W}{0.8}\right)^2=18.6 \quad (3)$$

=0.392mm.

Choose the next available smaller wire diameter, *W*, of 0.35mm then calculate the radial thickness of the coil,

$$450\left(\frac{0.35}{0.8}\right)^2=16C \quad (3)$$

C=5.4 mm.

The new value for the radial thickness – and the mean diameter – may call for re-calculation of the inductance and correction for the number of turns. Alternatively, it may be left to the final correction, before cutting the wire.

Check the dc resistance,

$$R=\frac{17.6\times450}{14250\times0.35^2}=4.54\Omega \quad (2)$$

Decide whether this value is acceptable. If too high, choose a former of bigger diameter and/or cross section. ■

Further reading

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Can the switched-capacitor technique eliminate inductors in dc-to-dc converters? Ian Hegglun thinks it can, and explores how the technique can be applied.

Switched capacitor power supplies

Switching power supplies normally use inductors and/or transformers for voltage conversion. But this is not the only way to transform voltage efficiently. Voltage multipliers and voltage dividers are another option.

Voltage dividers are less well known. They are simply multipliers reconfigured for voltage step down. There are lots of similarities. Like multipliers, they need no inductors and like multipliers they are limited, mainly because they are difficult to regulate efficiently over a wide range of input or output voltages.

Pulse-width modulation can certainly vary the average voltage of inductorless circuits but the losses in the smoothing process are hardly any different to linear regulation. This is because the equivalent circuit of a switching capacitor arrangement is a low-pass *RC* filter which is inefficient when pulse modulated at low duty cycles. However, my earlier work showed efficient regulation is possible if a small value air-cored inductor is added to ac input of a full wave multiplier[1]. Regulation is then achieved by varying frequency.

This article looks at step down switched capacitor converters. In particular, converters that are capable of delivering tens of watts at 3V and converters delivering hundreds of watts at around 50V output. When cascades, these converters could form the basis for computer power supplies, but how practical is this?

240V Mains to 3V without inductors?

Breaking the task into two stages with an intermediate voltage at around 30 or 50V helps. This would work in well with the distributed supply technique where one main power supply delivers only one voltage which then supplies various boards, each board has their own converter for their particular needs.[2,3] This solves a number of problems and makes it relatively easy to add batteries to the intermediate voltage for an uninterruptible supply.

Each converter requires a step-down ratio of at least 8:1. Since there are so many stages, each stage must be very efficient. If you aim for, say, 70% overall efficiency then each of two converters need to be around 85% efficient. With two or three cascades within each converter, efficiency for each stage must be no less than 95%.

Although difficult, this is not impossible. The increasing use of synchronous rectifiers such as the *MA767* can give this efficiency even at the 3V level.[4] Recent developments with mosfets, namely Philips' *Trenchmos* and SGS-Thompson's mesh overlay, should reduce the cost of the silicon needed for low losses.

Other necessities for mains conversion are power factor correction and an earthable or non-floating output. Both of these can be provided without using an inductor or transformer.

Literature on voltage dividers

One reference to a voltage divider can be found in the *ICL7660* data sheet.[4,5] **Figure 1** shows the connections required and **Fig. 2** shows the internal switching with mosfets.

In the first half cycle when S_1 and S_3 are closed, the supply voltage divides across the capacitors in a conventional way proportional to their value. In the next half cycle when S_2 and S_4 are closed, the capacitors switch from a series connection to a parallel connection. This forces the capacitors to have the same voltage; the charge redistributes to maintain precisely $^1/_2$V+, across C_1 and C_2.

A Linear Technology application note describes using four discrete mosfets to convert 12V to 5V at 1A without using an inductor in a similar way[6]. Regulation over a small range is achieved by varying the charge duration while S_1 and S_4 are on.

Two of the switches, S_3 and S_4, can be replaced with diodes, **Fig. 3**. With diodes, efficiency is reduced for light loads. This is because mosfets acting as synchronous rectifiers in Fig. 2 can have a low forward voltage drop at low currents, unlike normal diodes.

In the diode version, the output capacitor can be removed and the circuit still operates. Ripple voltage increases, as expected, but the ripple frequency halves – something not expected.

Investigating further, it appears that a minimum output capacitance is needed to pull the ripple voltage average of the first phase to the same level as the second phase. Simulation on *Tina* showed this effect. The minimum capacitance appears to be around $^1/_3C_1$, although in tests the ripple current rating requires a minimum of $^1/_2C_1$.

I also tried simulation on *Electronics Workbench* V4. Although the simulation ran, it did not give the correct waveforms using the same models as *Tina*. In these simulations, the mosfet switches were simply voltage-controlled switches with a series resistance to simulate mosfet on resistance. Doing this speeds up simulation considerably and is obviously acceptable with *Tina*.

Simplifying the switches in this way may explain why *Workbench* gave incorrect results, but I am unable to repeat the analyses with proper mosfets since I no longer have access to the software. Anyone interested in seeing the Workbench file can do so by sending an stamped, self-addressed envelope to *Electronics World*'s editorial offices.

Using an equivalent circuit

The equivalent circuit is useful for predicting output voltage, ripple voltage and efficiency of a converter. **Figure 4** shows the equivalent circuit. The derivation is given in a separate panel. The voltage source is $^1/_2V_{in}$ and there is a V_D diode drop with a triangular ac ripple voltage (ΔV_{pp}) superimposed on the output.

Ripple voltage is determined mainly by C_1. Bear in mind that C_1 supplies output current for both phases in this type of converter. Therefore is best to place most of the available capacitance at C_1.

Finally, there is a series resistance of $R_{DS(on)}$ for the mosfet

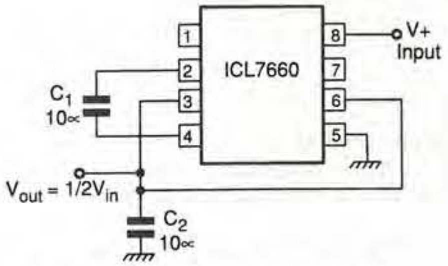

Fig. 1. Voltage divider using a popular IC.

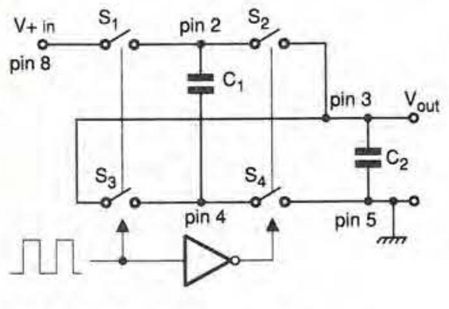

Fig. 2. Voltage divider internal circuit.

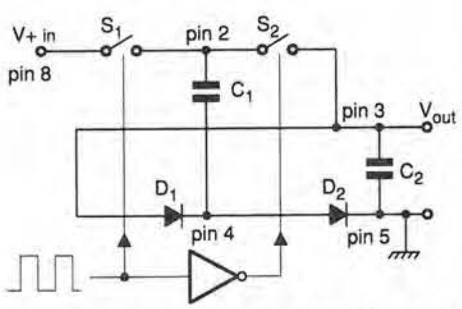

Fig. 3. Diodes being used in place of S_3 and S_4.

ESR of aluminium electrolytic capacitors

Losses in a capacitor are represented by an equivalent series resistance, or esr. Low esr is needed in switching power supplies. A low esr means low internal heating due to ripple current and low ripple voltage. It is common to parallel two or more small capacitors to give an esr lower than a single capacitor of the same total value.

Electrolytic losses vary with frequency and temperature. The graph shows the impedance of several aluminium electrolytic capacitors. At higher frequencies the capacitor becomes net inductive. Some capacitors have a flat region over one or two decades of frequency.

Minimum esr is also related to voltage and capacitance. In general, the same cans size fixes the esr and hence ripple current rating. This can be seen by scanning through the ripple current ratings for various capacitors of the same size within the same brand range.

For non-solid electrolytic capacitors reliability is related to power loss (average I^2esr) and the ambient temperature.

Philips' electrolytic capacitor data book 8 states that a 50% drop in reliability can be expected for each 10°C increase in temperature. Solid electrolytics are not degraded by temperature to the same extent and offer much longer life. Philips quotes a failure rate of 10^{-9}hours for solid electrolytics compared to 10^{-6}hours at a 60% confidence level for their non-solid electrolytics.

The company also gives an equation for calculating ripple current

$$I_R = \sqrt{\frac{P}{ESR}} = \sqrt{\frac{\alpha S(T_C - T_{amb})}{ESR}}$$

where I_R is the ripple current (A), P is heat dissipation (W), a is the heat-transfer coefficient (W/m^2/°C), S is surface area (m^2), T_C is the case temperature (°C) and T_{amb} is the ambient temperature (°C).

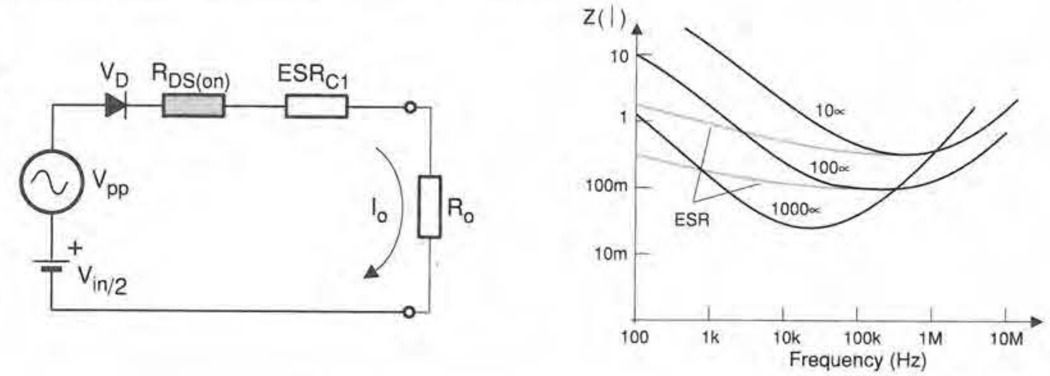

Fig. 4. Equivalent circuit for Fig. 3, and curves showing the variation of impedance with frequency for various aluminium electrolytic capacitors.

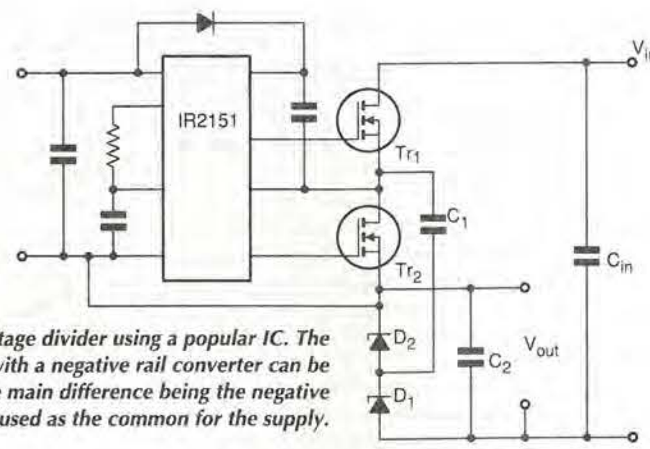

Fig. 5. Voltage divider using a popular IC. The similarity with a negative rail converter can be seen – the main difference being the negative rail is used as the common for the supply.

switches plus ESR_{C1}, as in the diagram in the panel. With significant dead time the losses must be increased according to the form factor.

A 30 to 15V converter

Figure 3 is a relatively easy circuit to implement. Test circuit **Fig. 5** uses an *IR2151* to drive two mosfets. Three 6.8µF, 20V tantalum capacitors measuring 7 by 5 by 4mm were paralleled for C_1. Each capacitor has an esr of 400mΩ at 100kHz. Two of these capacitors are used for the output C_2.

The input capacitor consists of three 22µF, 50V YXB aluminium electrolytics. These allowed the circuit to deliver 5A continuously. The input voltage was adjusted to give 15V output, requiring 34.9V.

Efficiency is easily calculated by 2×15/34.9, giving 87% efficiency. This calculation is possible because the ratio of output current to input current is precisely 2, due to charge conservation and current used by the driver is negligible at full load.

Analysing losses, 2W is lost in diodes, 2W in capacitors, mainly C_1, and 4W in the mosfets. This gives a total of 8W for 75W output.

Figure 6, top trace, shows output ripple voltage with a 5A load: the average of this ripple is offset 15V above ground. The lower trace shows the voltage at the output of the mosfets feeding C_1. Note that the ripple frequency is twice the oscillator frequency.

Dead time can also be seen. It totals about 2µs which means about 40% more current is required causing 40% more ripple and losses. Current through C_1 is 10A peak falling to 5A or 7.5A average. You may notice the similarity with a negative rail converter. The main difference is that the negative rail is used as the common for the supply.

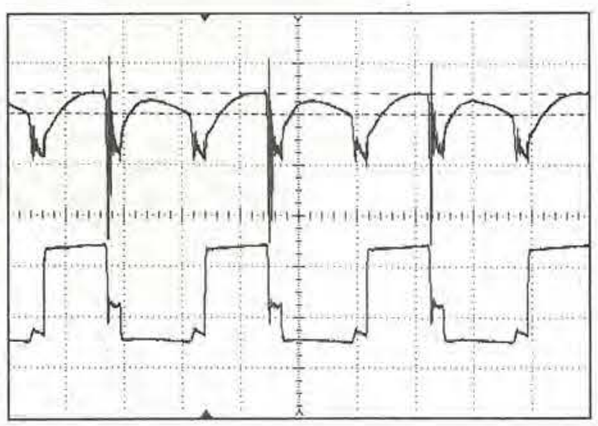

Fig. 6. Ripple voltage (upper trace 1V/div) and the ac drive voltage to C_1 (lower trace 10V/div) with a 5A load. Time base is 5µs/div. The ringing was traced to the driver. Adding gate damper resistors of 47Ω eliminates this ringing.

Due to the dead time, the ripple is increased by 50%. Increasing the frequency fails to reduce ripple because the dead time component remains constant. Adding a small 1µH inductor to form a low-pass filter between the load and output with a shunt capacitor across the load can reduce ripple to low levels if required.

The inductor used was removed from a 10A switching power supply output and measures 6mm dia. and 15mm long with five turns of 1.5mm wire. With a 220µF 200mΩ esr electrolytic the ripple was reduced by a factor of five with a 70kHz operating frequency. If the dead time could be reduced and the frequency increased to 150kHz the ripple would fall to around 10mV rms with this filter.

With 5A load, the tantalums ran at about 50°C, or about 35°C above ambient. Operating the converter at higher ambient temperatures, of say 50°C, the current would need to be halved. This means that ripple voltage would also be halved.

Another method for reducing ripple is by running two converters in parallel and clocking them 90° out of phase. This can reduce ripple by a factor of two – more with an *LC* filter – while also doubling the output current. This technique can also be applied when cascading two converters for higher step down ratios.

At about 1A per capacitor, this is a remarkably high power density. The chip inside measures only 3mm diameter and 3mm long, or $^1/_{25}$th the volume of an aluminium capacitor of similar rating. These capacitors were removed from a computer pcb some time ago and I don't know where they can be sourced. However, similarly rated special polypyrole *ECG* series electrolytics by Panasonic are readily available.

Later tests on an ECG 6.8µF/16V electrolytic revealed an esr of only 40mΩ – the data sheet quotes 0.4Ω maximum at 400kHz – and a minimum impedance at 2MHz. It appears some manufacturers are very conservative when it comes to esr values, possibly because there is a large spread in values. These ECG capacitors are rated at 1A at 105°C, but if they are sorted, most will run at higher currents. Although to make use of this capacity, the frequency needs to be closer to 1MHz.

I also tested a 4.7µF/16V multilayer ceramic in a 1206 package. Its capacitance measured on a dmm was only 3µF at 25°C and only 1µF at 1MHz. Minimum impedance occurred at 8MHz with an esr of 130mΩ. It uses Y5V dielectric and has a temperature coefficient of 1%/°C from 25°C which means the capacitance falls to $^1/_5$th of its room temperature value when heated to 80°C. These Y5V capacitors are not practicable for converters that operate above 40°C.

Comparison with buck

Comparison with a conventional buck converter is a useful exercise. The *L296* application note gives complete design details for a buck converter[7]. Equations for the value of the inductor and output ripple voltage are given below assuming the inductor ripple current is the usual 30% of the peak inductor current,

$$L = \frac{(V_i - V_o)V_o}{0.3V_i\, f\, I_o}$$

$$\Delta V_C = \frac{(V_i - V_o)V_o}{8V_i f^2 L C_O}$$

$$\Delta V_R = 0.3\, I_o\, ESR_{Co}$$

Note that the capacitor ripple voltage is made up of two parts. Since theses components are in quadrature they should not be added algebraically, although, if one component dominates by more than three times, the other can be neglected, due to the nature of quadrature addition. For a 2:1 buck configura-

tion and V_i=2V_o the inductance required and ripple voltages are,

$$L = \frac{V_o}{0.6 f I_o}$$

$$V_C = \frac{I_o}{24 f C_o}$$

$$\Delta V_R = 0.3\, I_o\, ESR_{Co}$$

Using the *L296* as described in the application, the frequency is 100kHz and two 100µF and 150mΩ esr output capacitors are paralleled. When loaded at 4A the ripple voltage is calculated to be 90mV pk-pk and esr ripple dominates.

For the switching capacitor converter of Fig. 3 the ripple voltage is approximately,

$$\Delta V \cong \frac{I_o}{4C_1 f} \cong \frac{I_o}{3\, C_O f}$$

where C_2 is $^2/_3\, C_O$ and C_1 is $^1/_3\, C_O$. The input capacitor is assumed to be the same in both cases. This equation comes from the equation $I_C = C^{\Delta V}/_{\Delta t}$, where the capacitor current pulse rate is 2*f* and current and the ripple voltage across C_1 is divided by two at the output and verified by measurements. It was also verified that there is no esr ripple component present – unlike the buck converter.

Using the same value of output capacitance as used in the *L296* example, the output ripple at 4A is 60mV pk-pk compared to 90mV. However, since the capacitors used in this example are only carrying 1.2A of ripple current (0.3×4A) the capacitance required for the same temperature and same life must be three times higher for a third the esr. A plus for the switching capacitor converter is lower electromagnetic radiation, lighter and more compact. The buck scores high on being able to vary the ratio over a wide range efficiently and requires less components.

Silicon utilisation

The buck converter requires one transistor, minimally rated at V_{in} and I_o (average) plus a diode for the same V_{in} and I_o. A switching capacitor converter such as Fig. 3 requires two transistors, each minimally rated at $^1/_2 V_{in}$ and I_o, plus two diodes each rated at $^1/_2 V_{in}$ and I_o. This is twice the number of devices although the total switching capacity (*V*×*I*) is the same.

With more parts there is a cost penalty but it also helps to distribute heat and improves reliability. Distributing heat makes natural cooling easier which improves reliability. In many cases a fan can be avoided which is often the most unreliable component.

Single converter divide-by-*n*

A circuit for a 4:1 step down, **Fig. 7**, has appeared in this magazine[9] and more recently in another magazine[10]. In this arrangement four capacitors, labelled $C_{1\text{-}4}$ on the circuit, are charged in series and later discharged in parallel. They operate as a Mosmarx[11] step down converter.

With this technique a 60V peak-to-peak input voltage can be stepped down to around 12V with an efficiency of around 75% using conventional diodes or around 90% with schottky diodes. Adding each extra stage in this circuit involves three additional diodes and one capacitor.

Figure 8 shows a modification that allows the step down ratio to be varied from 4:1 to 1:1 in 4 steps which is useful for regulation or power factor correction for off-line converters. A ratio of three for example is obtained when Tr_1 is held on for both phases, bypassing C_1 during the charge phase and therefore charging three capacitors in series rather than four.

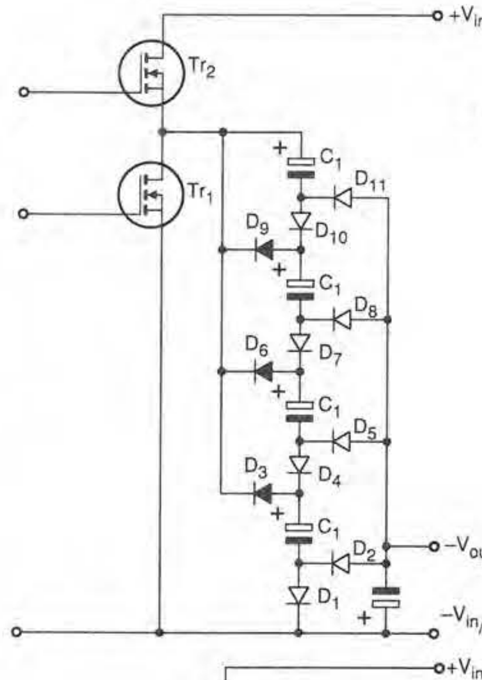

Fig. 7. This network converts V_{in} to negative $^1/_4\, V_{in}$.

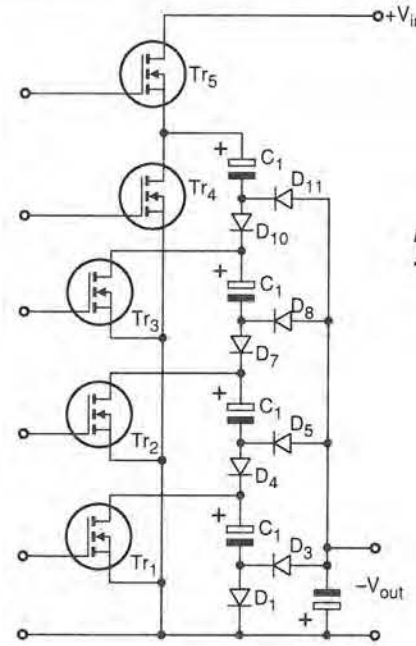

Fig. 8. Allows variable ratios of 4:1 to 1:1 in 4 steps.

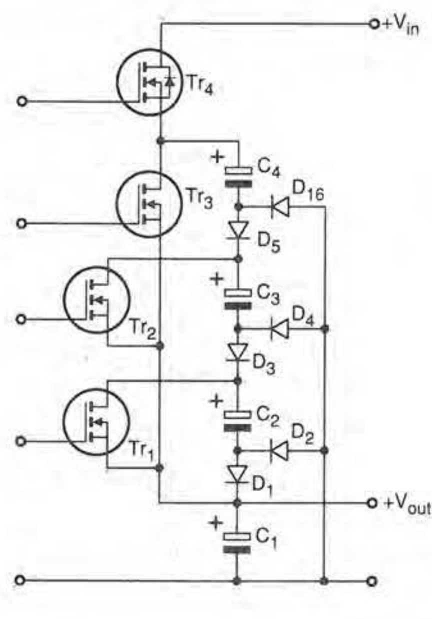

Fig. 9. Converts Vin to positive 1/4 Vin with 4 steps.

If a low drop series regulator is added after the converter, the ripple can be removed and regulation over a wide range of input voltages is possible with high efficiency.

Figure 9 shows a non-inverting version. It uses two fewer diodes, one fewer transistor and one fewer capacitor. This makes it slightly more efficient. However, the transistors in a selectable ratio converter requires individual high side drive making driving more difficult.

So can the Cockcroft Walton multiplier be reversed for step down to give ratios higher than 2:1? **Figure 10** shows how it can using mosfets on the improved version.[12] This allows the number of diodes to be reduced, improving efficiency.

Parallelling two dividers as in **Fig. 11** reduces the ripple in the input current and allows much smaller input and output capacitors. Minimal dead time is required for best results.

Equations for switching converters

Neglecting ac ripple component on C_1 and the effect of energy loss from capacitor charge redistribution, the equations for steady state output are as follows. In all cases the output voltage is slightly lower due to charge redistribution losses[4] but appears to contribute only a few percent of losses since ripple is usually less than 10% of the dc voltage.

Synchronous converter, Fig. 2. All switches are mosfets with equal $R_{DS(on)}$ since they carry the same peak current.
Charge phase,

$$V_C = V_{in} - I_O(ESR + 2R_{DS(on)}) - V_{out} \quad (1)$$

Discharge phase,

$$V_{out} = V_C - (ESR + 2R_{DS(on)})I_O \quad (2)$$

Substituting (2) into (1),

$$V_C = V_{in} - I_O(ESR + 2R_{DS(on)}) - V_C - (ESR + 2R_{DS(on)})I_O$$

So,

$$V_C = \frac{V_{in}}{2}$$

giving,

$$V_{out} = \frac{V_{in}}{2} - (ESR + 2R_{DS(on)})I_O$$

Single ended converter using diodes, Fig. 3. Each diode has a forward drop V_D and ohmic resistance R_D.
Charge phase,

$$V_C = V_{in} - I_O(ESR + R_{DS(on)} + R_D) - V_D - V_{out} \quad (3)$$

Discharge phase,

$$V_{out} = V_C - I_O(ESR + R_{DS(on)} + R_D) - V_D \quad (4)$$

Substituting (4) into (3) also gives,

$$V_C = \frac{V_{in}}{2}$$

hence,

$$V_{out} = \frac{V_{in}}{2} - V_D - (ESR + R_{DS(on)})I_O$$

From this, the equivalent circuit of Fig. 4 can be drawn.

Parallel converter, Fig. 11. This circuit shares the output current with $I_O/2$ through each stage giving,

$$V_{out} = \frac{V_{in}}{2} - (\tfrac{1}{2}ESR + R_{DS(on)})I_O$$

Converting from 30V down to 3V

This 30V to 15V converter gives valuable insights into the operation of a divide-by-two converter. At the time of writing, I had not tested the 30V to 3V converter, although a 24V to 12V synchronous rectifier version, as in Fig. 2, has been run, giving the target of 95% efficiency. To get this efficiency the output current is only a fraction of the mosfets current rating. The latest mosfets will improve this.

Philips now offers a 55V device in a TO220 package with an 8mΩ on resistance. Even lower resistance devices in smaller packages are likely soon.

Conversion from 30V to 3V requires a ratio of 10:1. This can be achieved a number of ways with dividers, either two divide-by-threes or three divide by twos. It appears that the $V \times I$ capacity is about the same in both cases.

A practical converter requires regulation of the output voltage. If the input varies over a wide range it is not practical to use pulse modulation as mentioned.

Recent tests using a small inductor within a divider demonstrated efficient regulation over a 2:1 input voltage range. This allows the 30V input on the low voltage converter to

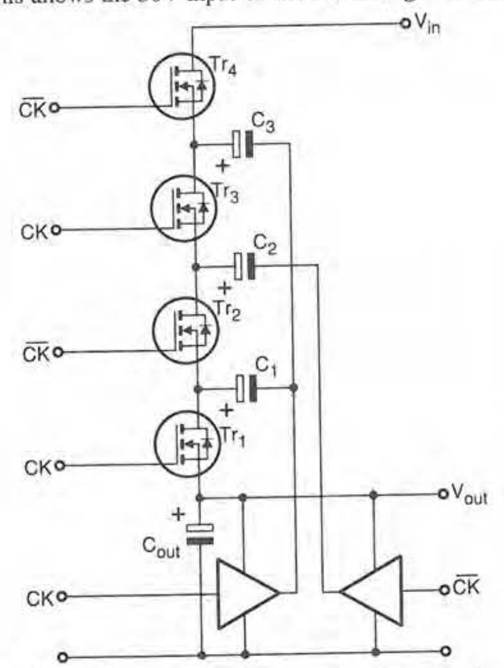

Fig. 10. Improved Cockcroft Walton multiplier can be reversed if diodes are swapped for mosfets.

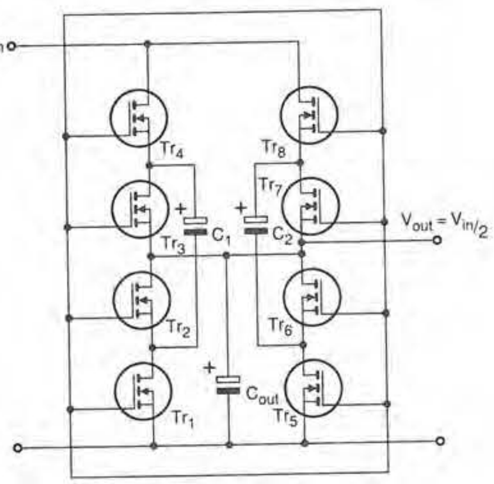

Fig. 11. Parallelling two dividers reduces the ripple in the input current and allows much smaller input and output capacitors. Minimal dead time is required for best results.

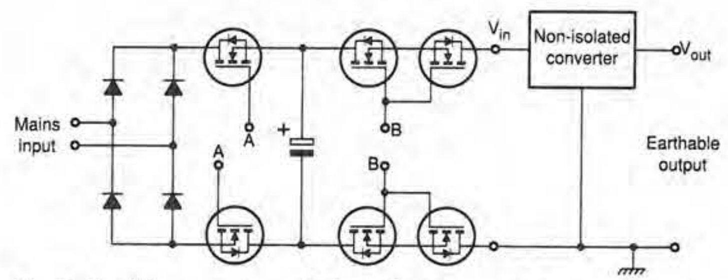

Fig. 12. The flying capacitor technique allows the normally floating output of a bridge rectifier to be connected to earth without damaging the circuit.

***Safety warning**

When Ian speaks of mains isolation, this does **not** mean safety isolation. Device failure could result in a direct path between the mains live side and the converter's output. As a result, all parts of any such converter should be considered to be lethal, and never touched. Wiring regulations in your country should also be consulted before such a circuit is connected – Ed.

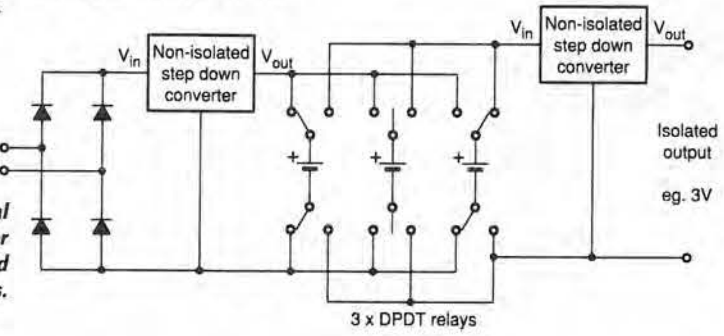

Fig. 13. Isolation to several kilovolts without a transformer is possible using three relays and switching batteries.

rise to 60V plus ripple filtering of the reservoir capacitor on this bus. I hope to present design details in a later article.

Mains to 50V conversion

The preferred arrangement for the mains converter is the circuit in Fig. 9, but with eight stages.

This arrangement uses fewer components and it allows the ratio to be varied in small increments for voltage pre-regulation and power factor improvement. It does this by varying the ratio over the mains half cycle. Current flows over most of the cycle in a sinewave like fashion.

At 50V output, this converter can use standard diodes giving about 2V in 50V loss and an efficiency above 95%. This is well above the target of 85% for the mains converter discussed earlier. It should be possible to get 80% overall for mains to 3V without using unrealistic amounts of silicon.

Isolating the output

The flying capacitor technique has been proposed as a way of 'isolating' the converter common from the mains[13]. Although this does not provide the same degree of isolation from the mains as a transformer it does allow the converter's low-voltage output common to be connected to an earth and it allows the mains to be rectified using a full-wave bridge rectifier.

Normally, the output of a bridge rectifier must be floating relative to the ac input terminals. The flying capacitor technique, **Fig. 12**, can be implemented with six mosfets or igbts switched at a high frequency.

Although this technique may be acceptable in some installations, it has safety limitations where people can touch the equipment being supplied. Even with a reliable earth leakage relay and the usual output over-voltage crowbar there are some people such as those with a pre-existing heart problem are at risk if exposed to a short duration mains shock while the earth leakage relay trips.

Using three batteries and three relays, **Fig. 13**, the output can be isolated to several kilovolts without a transformer,* with uninterruptible power supply as a bonus. With a 30V intermediate bus, three 30V batteries can be rotated in sequence, one being charged from the mains converter while the second supplies the load. The third ensures no-break changeover. A cycle life of around a second allows small batteries to be used, while giving an acceptable relay life.

This can match transformer isolation given suitable relays and sufficient tracking clearance of circuit board tracks. The batteries and their connections need to be suitably isolated from each other and the output side. Where the relays are exposed to humidity etc they must be protected or sealed.

In summary

Hopefully I have demonstrated that it is not impracticable to use switched capacitor dividers to convert mains to 3V without a transformer – and do it efficiently without huge amounts of silicon.

Certainly the number of components is greater than a conventional switch-mode power supplies, but when power factor correction and distributed converters are added to a switch-mode psu, the component difference reduces. ■

Anyone interested in communicating with the author regarding the concepts described here can do so by contacting Electronics World's editorial offices.

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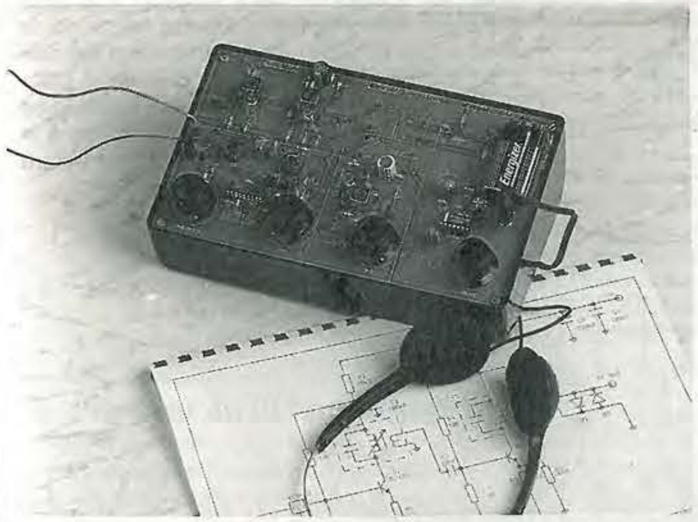

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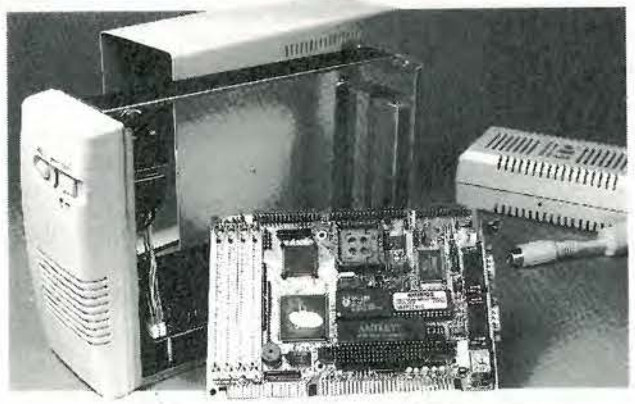

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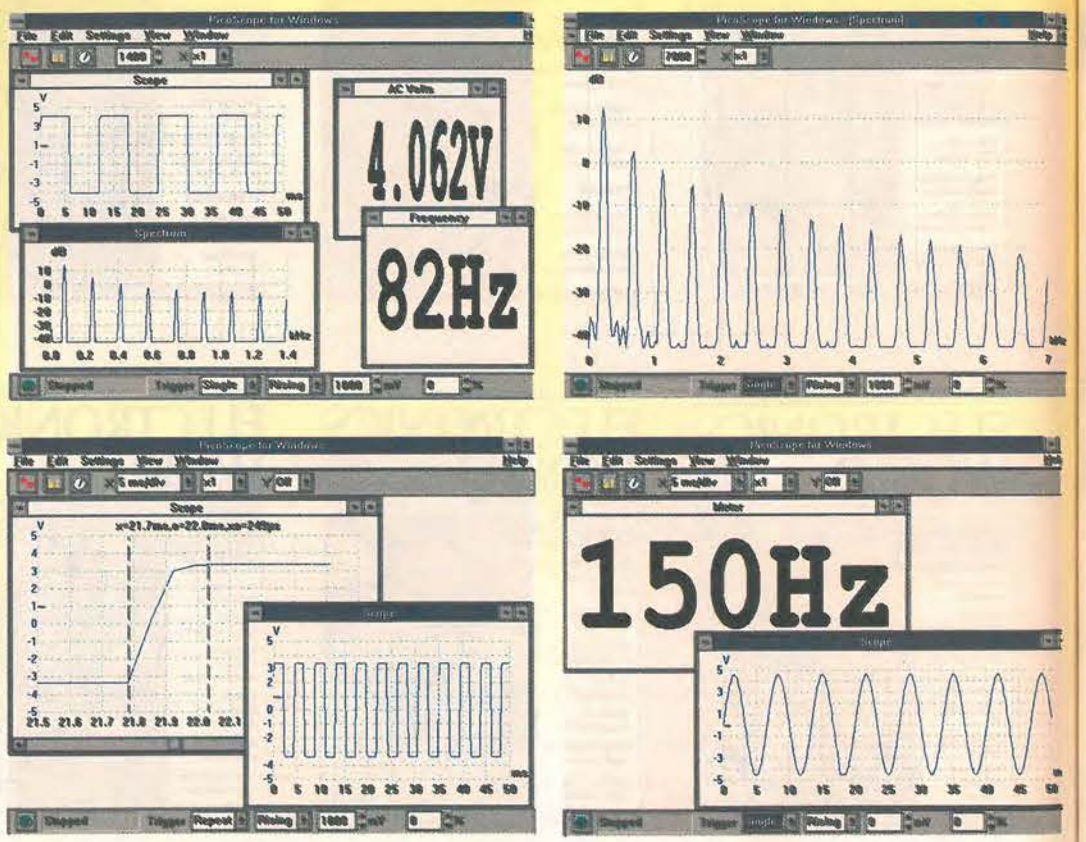
How to submit your ideas

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

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

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



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

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

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

Three-phase over/under voltage detector

When the voltage of each phase of a three-phase supply is within limits set by the user, the supply is connected to the output; otherwise the output remains off. No relays or contactors are used; zero-crossing, optoisolated triac drivers control each phase.

Each phase is stepped down by an ordinary 12V, 100mA transformer, whose output is rectified, filtered and passed to a *BB4115* window detector via the 5kΩ adjustment trimmer. Upper and lower limits V_{ut} and V_{lt} are set by the two trimmers on the output of the *7812* regulator. Depending on the outputs of the three window detectors and the following *MOC 3012* optocouplers, the *MOC3083* zero-crossing, optoisolated, triac drivers, the leds of which are in series, remain off or switch on. Switching logic is shown in the table.

The power circuit consists of three power triacs with optional snubbers, each driven by one of the *MOC3083*s.

At start-up, the two triac drivers that see zero voltage difference between two of the phases switch on, followed by the third when its voltage approaches the sum of that in the phases already on. This ensures zero-current turn-on of all three load branches, whether arranged as star, Y or delta.

When all leds in the *3083* switch off, all three phases switch off when current in two of them falls below triac holding current, creating zero current; the third phase then switches off. Switch S_1 controls the power circuit and an additional switch at the three-phase input might also be used. It might be useful to replace S_1 by an optocoupler, driven by an external circuit such as a temperature controller.

Triacs used in this circuit are *MAC223 10FP*, rated for an off-stae voltage of 800V, rms current 25A and peak surge 250A.

Fuse indicators shown illuminate red to show a blown fuse and green to indicate an intact one.

Porus M Mehta
Maharashatra
India

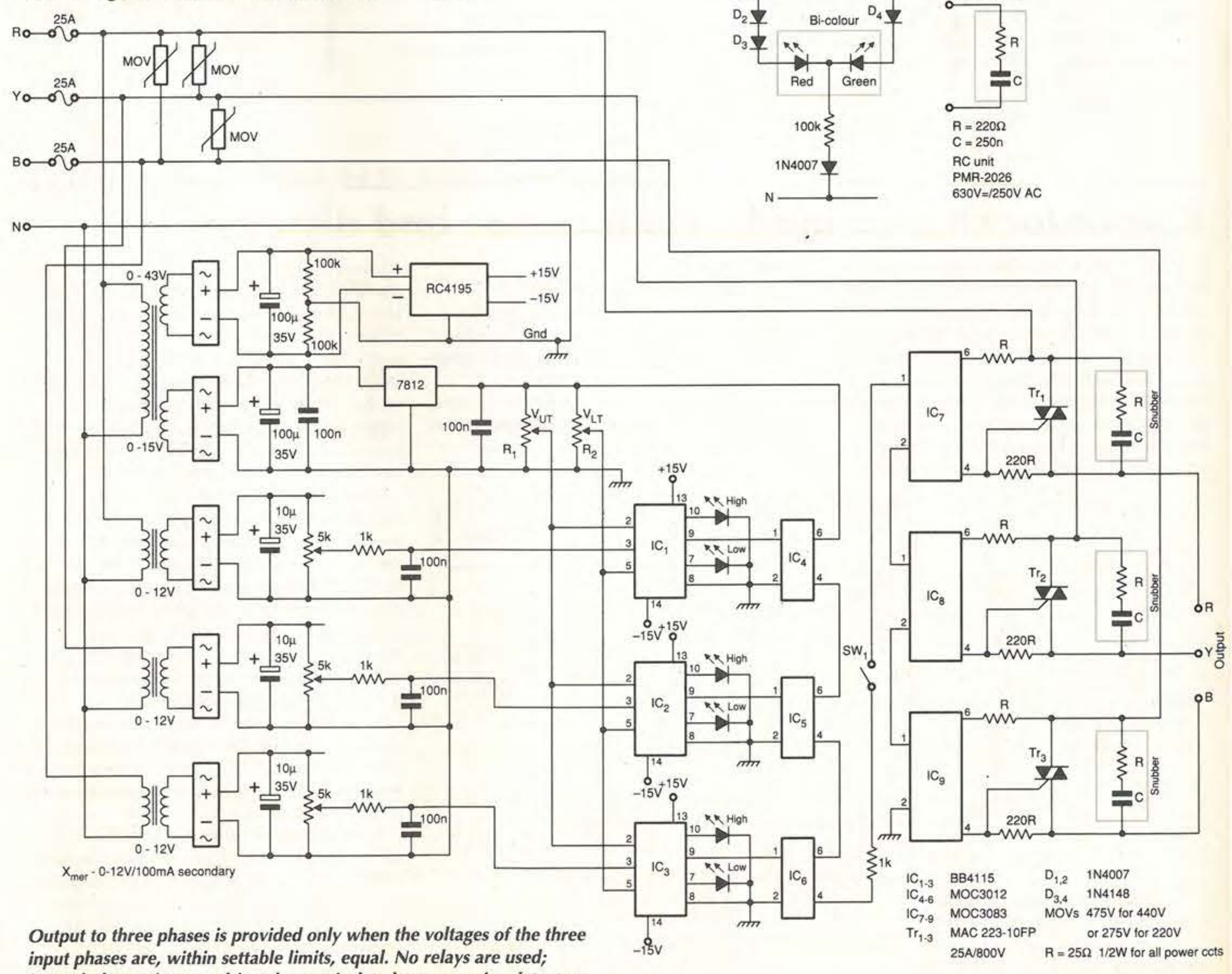
Logic table for the window comparator.

	$V_{in}>V_{ut}$	$V_{lt}<V_{in}<V_{ut}$	$V_{in}<V$
High (red led)	on	off	off
Opto drive	off	on	off
Low (green)	off	off	on

V_{ut}=upper threshold, V_{lt}=lower threshold.

BB4115 is obsolete

Please note that although the idea is express here is sound, the Burr Brown BB4115 is obsolete. I could find a direct equivalent, but the AD1317 window comparator might be suitable, with circuit modifications. Alternatively, it may b possible to duplicate the function using two AD790 comparators. If any of you know of a more direct replacement, would you mind letting us know please? Ed.



Output to three phases is provided only when the voltages of the three input phases are, within settable limits, equal. No relays are used; instead, three triacs are driven by optoisolated zero-crossing detectors

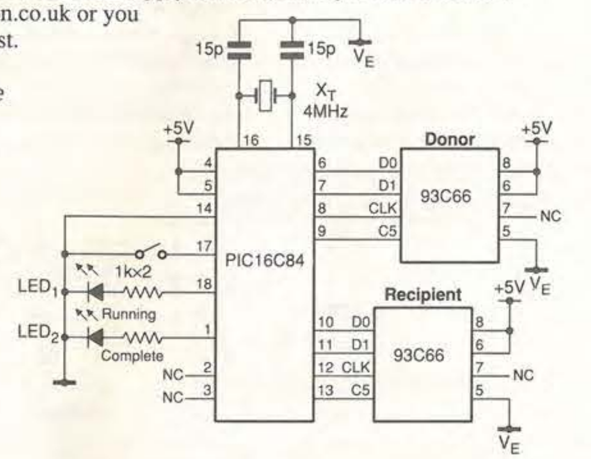
Eprom programmer/copier

Having previously used a pc-based method of programming eeproms, I produced this device to free the pc for other work.

A donor microwire memory device is read by the circuit one byte at a time, the information then being programmed into a recipient, which is automatically erased when the process starts; all locations are written. Furthermore, instead of the manual input needed by the pc programmer, this only needs one four-second switch operation. You do, of course, have to make one donor using the pc or some other method; donors can be kept to guard against hiccups of the hard disk. Pressing the switch starts the programming operation, which halts on completion; led 1 flashes while programming is in operation and led 2 glows on completion; opening the switch primes the *PIC16C84* for another run, but switch the power off when changing the recipient chip.

Opcode is shown here, but I can supply a 3.5in disk; my e-mail address is brian@briano.demon.co.uk or you can use ordinary post.

Brian Olliver
12, Fountain Drive
St Georges
Telford TF2 9DP



Instead of tying up a pc, use this programmer with donor and recipient devices to program eeproms.

```
:100000004A2800288610861106110000861586144DD
```

:100000004A28002886108611061100008615861 4DD

Calculator chip as maths coprocessor and display

Using a cheap microcontroller can make digital instruments cost-effective in small numbers, but the cost of displays and their demands on the processor pose a problem.

One solution to both problems is to use the ic and display from a cheap calculator as a combined coprocessor and display controller, all calculation and display functions then being carried out by this purpose-designed processor; the main microcontroller is then left to pursue its own affairs.

Calculator keypads are normally multiplexed and may be emulated from the microcontroller by monitoring column outputs and pulling the appropriate row input line high or low, depending on the calculator used.

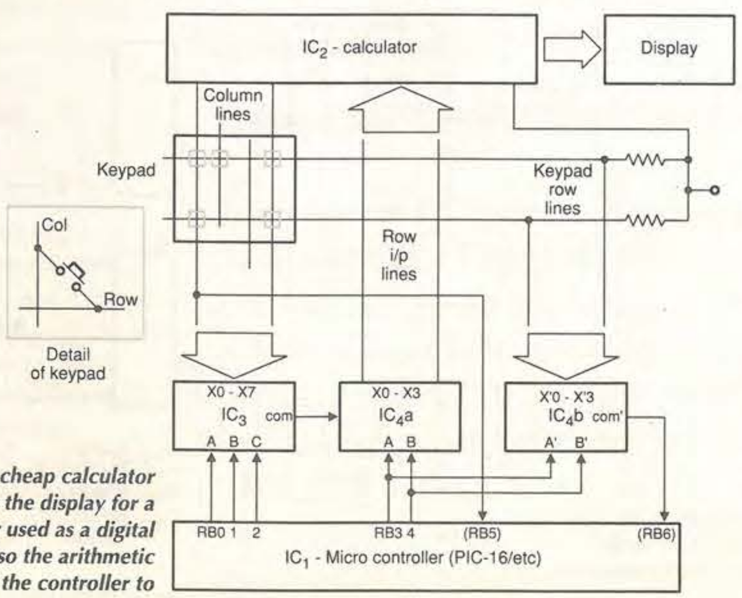
The diagram shows an alternative: two cmos analogue multiplexers, $IC_{3,4}$, will simulate the keypad directly, simplifying the software interface and reducing the number of port lines used. Only five lines, $RB_{0\text{-}4}$, are needed for calculators using the 6 by 4 or 7 by 4 matrix. Since these lines are only used when writing to the coprocessor, they can be multiplexed for other purposes – possibly using the enable pin on IC_3 to disable display writes. An unused column address in IC_3 is otherwise used as a No-op command and each digit or operand entry is achieved by writing one five-byte to the display controller.

As you can see from the diagram, the calculator keypad can be kept as a free control panel by using the spare 4:1 multiplexer in IC_4 and the extra port lines $RB_{5,6}$. Selecting No-op on $RB_{0\text{-}2}$ and the row address on $RB_{3,4}$ allows RB_6 to be monitored for a pulse. If columns rise in sequence, the active key column can be determined from the time delay from the first column at RB_5, or a priority encoder could be used to indicate a keystroke directly.

A variety of arithmetic functions is possible with eight-digit precision, even from cheap calculators; not only the basic functions, but with a little ingenuity, accumulation, average, delta and many others.

You can measure the mains frequency accurately without long gate times by using a *PIC16X* in timer mode and the calculator to find the reciprocal. Functions not normally found in commercial counter-timers are percentage or frequency change; these can be performed easily using this method.

Anthony New *Bristol*



Innards of a cheap calculator provide not only the display for a microcontroller used as a digital instrument, but also the arithmetic functions to free the controller to

High-current variable lipstick

If you remove all the metallic parts – and the red part – of a lipstick applicator and substitute a ferrite rod for the lipstick, the result is the core of a variable inductor capable of carrying an alternating current of several amps.

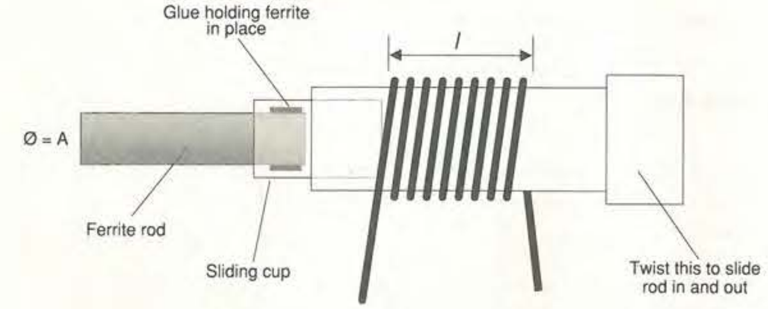
Maximum inductance is given by the standard formula:

$$L = \frac{n^2 A \mu_0 \mu_r}{l} \text{henries}$$

and the minimum inductance approaches the value given when the relative permeability μ_r=1.

Using the layout shown in the diagram with about 20 turns, inductance ranged from 4-20μH with my prototype.

Name not supplied
Edinburgh



Mechanics of a lipstick make a low-value variable inductor.

Monitor RS-232 activity

Using only a moving-coil voltmeter, you can observe the average activity on an RS-232 link and easily detect oddities.

A centre-zero meter in voltmeter configuration with a suitable series resistor is simply connected between common and the line to be watched.

Since RS-232 transmission is in the form of two voltages with opposite polarity to indicate binary 0 or 1. If you assume one start and one stop bit, maximum usage will show on the meter, scaled for ±100%, as zero. Idling shows up as –90% to –100%, fairly busy as –50% to –20% and flat out as –20% to +20%. The region +20% to +50% means that something is not quite right. Anything over +50% indicates a probable fault.

Trim the series resistor to obtain a –100% reading for zero traffic.

Keith Wootten
Reading

Transconductance square rooter

In the simple circuit of **Fig. 1**, output signal V_o is the square root of input voltage V_i. The circuit is based on the quadratic law connecting I_D and V_{GS} in an enhancement mode mosfet, $I_D=K(V_{GS}-V_T)^2$. This may also be written $V_{GS}=V_T+\sqrt{(I_D/K)}$, where V_T is the threshold voltage and K is a constant depending on the physical parameters of the mosfet.

Together with Tr_1, op-amp A_1 forms a voltage follower, so $V_{R2}=V_i$, $I_D=V_i/R_2$ and V_{GS}, which is buffered and referred to ground by a difference amplifier, is then $V_T+\sqrt{(V_i/KR_2)}$.

The second stage allows V_T to be subtracted, through RV_1, and it allows total gain to be adjusted through RV_2, in order to obtain $V_o=\sqrt{V_i}$.

To avoid thermal drift problems, the mosfet works close to the threshold voltage, with low values of I_D, where $I_D=f(V_{GS})$ is exponential rather than quadratic. However, a good adherence to the root law can be obtained by iteratively adjusting RV_1 and RV_2.

Figure 2 shows V_o versus V_i with an input range of 0 to 1V with a root function.

Giorgio Delfitto
University of Padova
Department of Physics
Istituto Nazionale per la Fisica della Materia
Italy

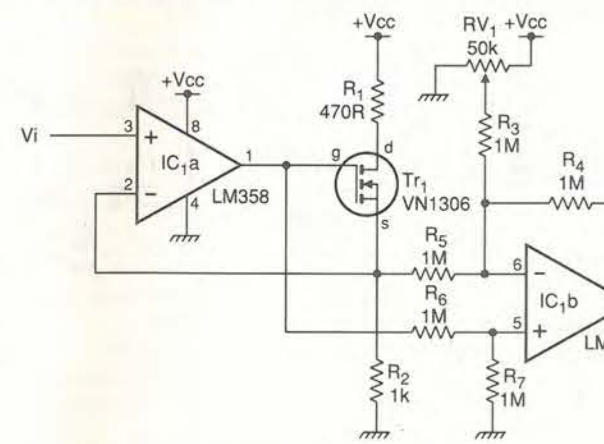
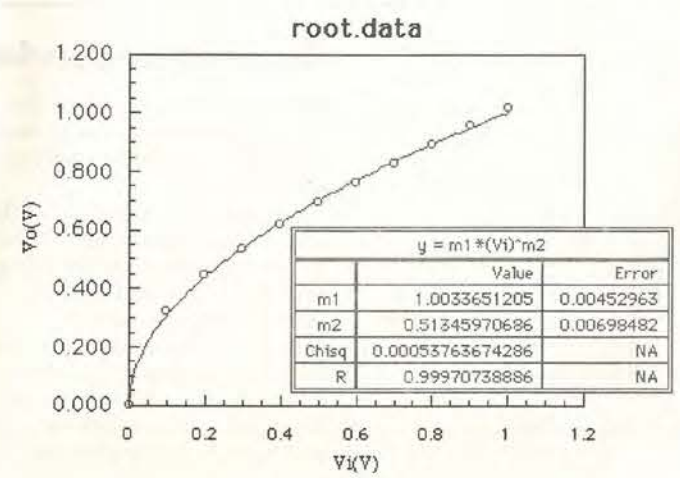


Fig. 1. Square rooting circuit based on transconductance.



	Value	Error
m1	1.0033651205	0.00452963
m2	0.51345970686	0.00698482
Chisq	0.00053763674286	NA
R	0.99970738886	NA

Fig. 2. Graph showing how voltage V_o closely approximates the square root of V_i.

Higher-order elliptic filter design

High-speed, current-feedback op-amps such as the *Burr-Brown OP603* make it possible to design active filters working in the megahertz range of frequencies.

Presented here is a quick method of calculating elliptic filters of any order, the example shown being a fifth-order type which is considered as being in two sections, as shown in **Fig. 1**.

Assume the poles $s=\alpha+j\beta$ and zeros Ω are given for a filter having the reflection factor of ρ and modulus α_0.

Design can proceed in three steps. Assume ρ=20% and $\alpha_0=20°$:

$\Omega_2=3.0653$ $s_{1,2}=0.382271\pm j0.67123$
$\Omega_4=4.8753$ $s_3=\gamma=0.49308$
$s_{4,5}=0.13647\pm j1.05129$.

1. If you calculate the constants A_1 and A_2 using the expressions,

$A_1=2\alpha A_2$ and $A_2=1/(\alpha^2+\beta^2)$, then for a third-order filter,

$A_2=1.6759379$ and $A_1=1.2813249$

and for a second-order filter,

$A_2=0.88918$ and $A_1=0.242865$.

2. Calculate normalised capacitances using the expressions in Fig. 1.

3rd-order

$c=1/\Omega_2=0.326232$
$c_1=2.40551$
$c_2=2.02807$
$k=0.134956$

and for a second-order filter,

$c'=1/\Omega_4=0.205116$
$c_1'=2.06648$
$k'=0.087087$.

3. For physical capacitance values, multiply the above c values by a factor of $1/2\pi f_0 R$, where f_0 is the cut-off frequency and R the reference resistance. In the case shown here, f_0 = 1MHz and R = 200Ω.

Kamil Kraus
Rokycany
Czech Republic

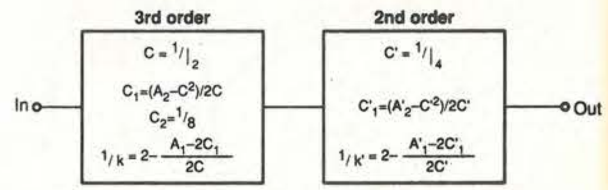


Fig. 1. Decomposition into third and second-order sections and formulas.

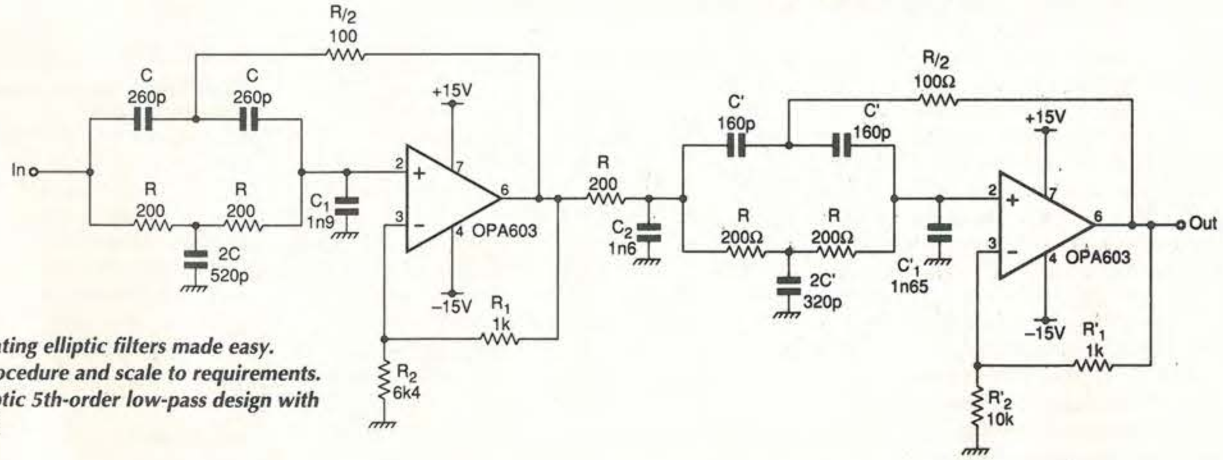
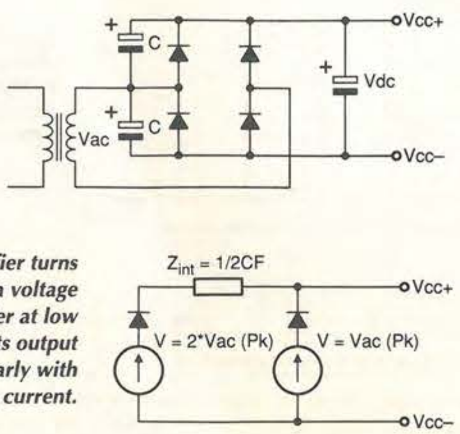


Fig. 2. Calculating elliptic filters made easy. Follow the procedure and scale to requirements. This is an elliptic 5th-order low-pass design with 1MHz cut-off.



Bridge rectifier turns into a voltage doubler at low currents, its output falling linearly with increasing current.

Current-dependent rectifier or doubler

This bridge rectifier becomes a voltage doubler at low currents, reverting to normality when current drain increases.

It is essentially a bridge rectifier with the extra capacitors *C*. When no current is taken, output voltage is twice that at the input and, as current increases, internal impedance causes the output to fall linearly according to $Z_{int}=1/(2Cf)$, assuming ideal components, until the output voltage equals the peak ac, when the circuit becomes an ordinary bridge rectifier.

An application is as a rectifier for lead-acid battery charging, in which a high voltage (25-30V) is needed to drive a small current into a 12V battery that has been discharged for a period and has an isolating layer on the electrodes. As current increases, a lower output from the rectifier reduces dissipation in the charger's output transistor.

Albert Pijuan
Girona
Catalonia

Frugal buzzer

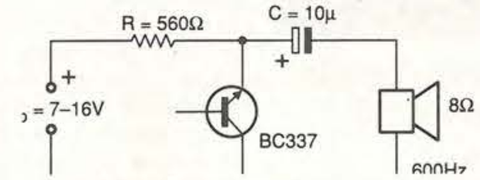
Four components make this very simple buzzer. Some n-p-n transistors oscillate at audio and low ultrasonic frequencies when polarity is reversed. Lowest working voltage depends on power rating, but it is about 7-8V for devices such as the *BC109, BC337* and *BC238*, 12V for the *BD139* and around 16V for power transistors *2N6543* and *BUX22*.

Oscillation frequency for the circuit shown is given by the empirical equation

$f=(V_b-5.5)/RC$

and is constant for devices of the same type. Power requirements are 5mA at 9V, 10mA at 12V; increasing *R* and decreasing *C* reduces power drain and also sound output. Higher voltage supplies can be used with increased values of *R*.

D Di Mario
Milan
Italy



Buzzer using reversed-polarity transistor as oscillator. Since the base is left open, it could be used to gate the buzzer.

10pF-10µF capacitance meter

A very simple circuit using a *555* timer and a few passive components forms a capacitance meter measuring full-scale ranges of 100pF-10µF.

Figure 1 shows the meter circuit, in which the *555* is connected as a free-running multivibrator, the frequency of the output rectangular wave being determined by

$f=1/T=1.44/(R_A+2R_B)C$,

and

$D=R_B/R_A+2R_B$,

T being the period and *D* the duty cycle, which is made assymetrical by setting $R_A=R_B$, since charging through the meter takes longer then discharging via the diode.

Unknown capacitors connect points A and B and C takes the form of six components selected by the switch in **Fig. 2**. If *V* is the amplitude of the *555* output, current through the meter is $I=VCf$, capaciatnce being therefore proportional to current for a constant frequency. The table shows the ranges and frequencies for a 10V supply. Without the damping components $R_{1,2}$,C_1, the meter bounces a little on the highest range, since the frequency is 1Hz; at higher frequencies it is steady.

The charge-pump power supply shown in **Fig. 3** also uses a *555* to double the input voltage and produce a 10V output, filtered sufficiently by C_3. A regulated supply would give better stability.

Calibration in the original was by the use of known capacitors, adjustment being by the trimmer for one range.

Raj K Gorkhali
Kathmandu
Nepal

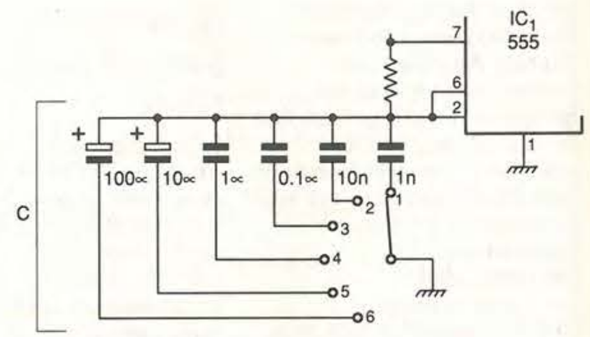


Fig. 2. Range-selection switch.

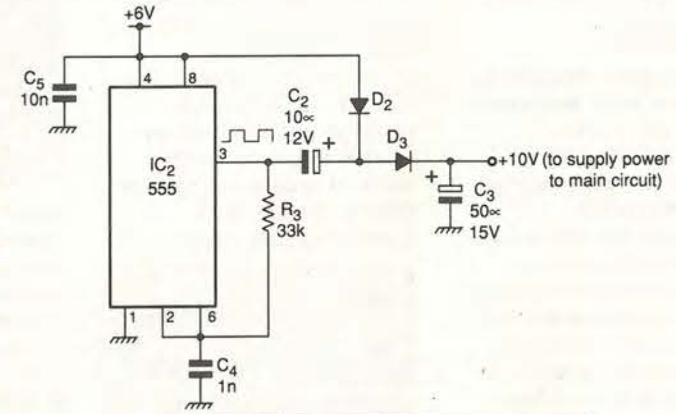


Fig. 3. Another 555 forms the power supply, a charge-pump voltage doubler to give 10V from the 6V supply.

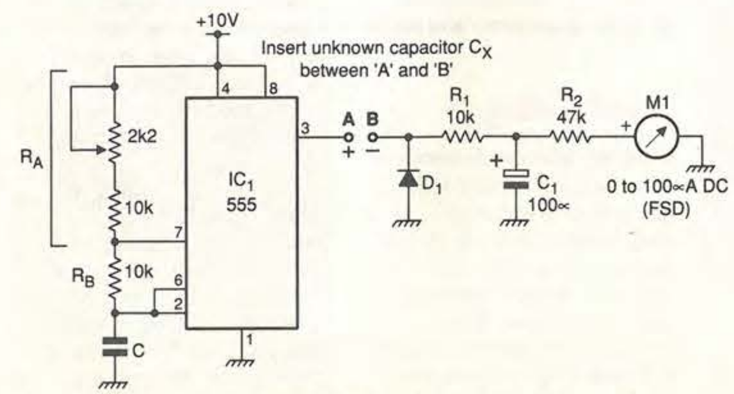
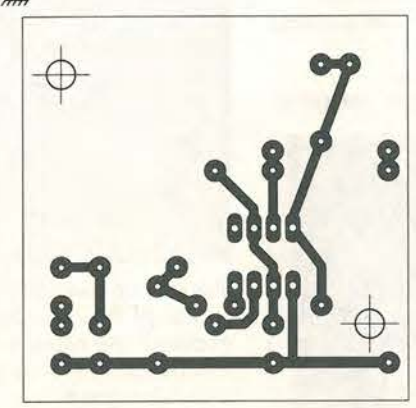


Fig. 1. Linear, direct-reading capacitance meter for values of 10pF-10µF.



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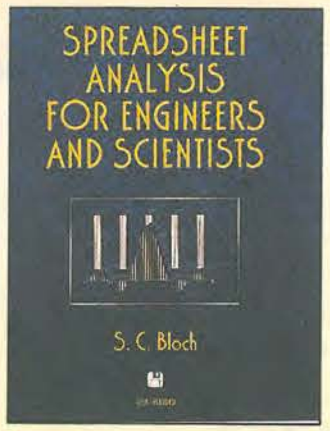

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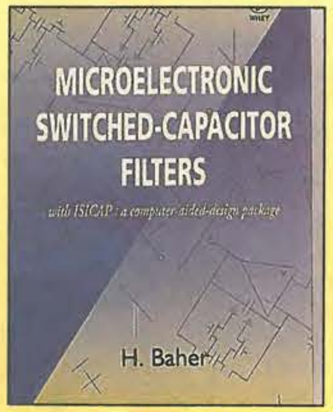

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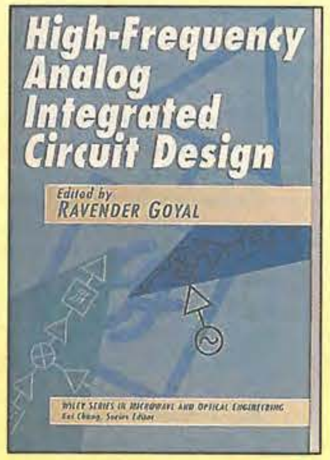

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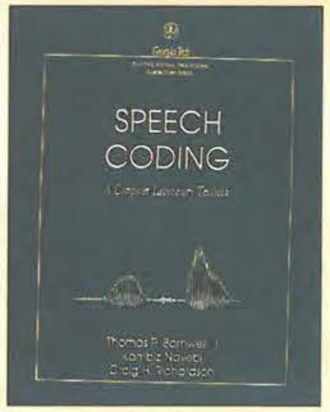

perform a wide variety of speech coding and speech processing experiments. The text presents and explains a set of basic speech coders analytically and in terms of the specific parameters controlling each coder. The manual leads the student through the experimental process of understanding how speech coders work and sound via over 70 exercises and projects. The class-tested menu-driven, Dos-based software can be operated by students with little or no training.

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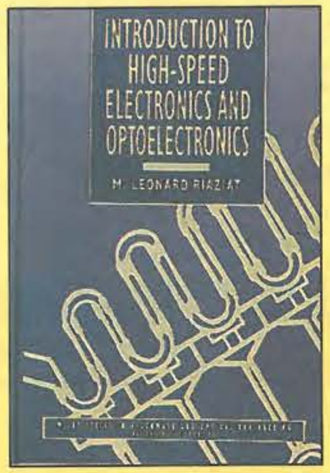

text/reference examines the complementary nature of electronics and optics and emphasizes high-speed technology in which the two fields are less differentiated. Beginning with an overview that develops a perspective and appreciation of analog high-speed technology in general, the book goes on to cover devices and circuits used at microwave and millimeter-wave frequencies, optical components, and opto-electronic integrated circuits and subsystems. Particular attention is paid to applications in the area of high levels of interest in this area and because many of the concepts are applicable in other fields. The book concludes with important coverage of the often-overlooked area of measurement and characterization of high-speed devices. Fully referenced and supplemented with hundreds of helpful illustrations, *Introduction to High-Speed Electronics and Optoelectronics* is equally useful as a professional reference or a textbook for senior undergraduate and first-year graduate courses.

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Measuring microamps

Not happy with the current measuring performance of today's digital multimeters, Robert Pearson has designed his own digital ammeter, which features a 20μA range and 0.1Ω shunt resistance.

There have been tremendous advances in the design and construction of portable digital multimeters. But although such instruments perform excellently on voltage ranges, most leave something to be desired when it comes to current measurement.

Before there were such things as digital instruments, it was easy to obtain a moving-coil instrument with a full-scale deflection as low as 50μA and with a full-scale volt-drop of as little as 50 or 75mV. Even lower figures were not uncommon.

In thermionic valve circuits, supply voltages were usually so large that milliammeter voltage drops could almost always be ignored. But modern semiconductor circuits are designed with very much lower supply line voltage requirements. A recent *Electronics World* news story concerned an op-amp that works with a supply of just one volt.

Many digital milliammeters produce a drop of 0.2V or more, at full-scale indication. This figure causes measurement errors that are often unacceptable. Another disappointment with modern digital multimeters is the almost universal common choice of 200μA as the most sensitive current range.

The meter's design

There are still many occasions when a separate voltmeter and an ammeter are desirable, which is why I designed this dc digital milliammeter. Its characteristics are at least as good as those available in common analogue meters 50 years ago.

The main trigger for designing the instrument was the offer, in *Electronics World*, of PM128 digital panel meters from Vann Draper Ltd. These compact and affordable basic meters have a 200mV range and very high input resistance. They also have a liquid crystal 3.5-digit display and require only about a milliamp from a 9 to 12V battery.

A factor that simplified the design was that integrated-circuit amplifiers with sufficiently low offset voltages and currents are now become cheap and readily available. There is no longer a need to resort to chopper configurations.

Circuit notes

Adding an amplifier like the *OPA177PG* to such a panel meter makes it easy to reduce the voltage drop of a digital milliammeter by at least one order, to only 20mV. Simultaneously, the op-amp makes it possible to reduce the lowest current range by the same factor of ten – i.e. to 20μA, **Fig. 1**.

For a 20mV drop on the 200mA range, a 0.1Ω shunt is required. With such a low resistance, the uncertainty of switch contact resistance becomes significant. An extra switch bank is added to ensure that the measured potential difference is solely that across the shunt resistor, and not across the contact resistance of the switch carrying the current to be measured, **Fig. 2**.

Including a higher current range or a lower voltage drop would require expensive four-terminal shunts and more complex switching. A wire fuse was not included due to its voltage drop. In any event, there is no need to protect the shunts since they are not expensive.

Some protection for the amplifier input is provided by a series resistor and a pair of back-to-back diodes. A similar arrangement following the amplifier offers protection to the panel meter. Negligible loss is caused by this extra circuitry since the meter has an input resistance exceeding 100MΩ.

Removing the unwanted

It is not easy to filter out low-frequency alternating current components that may exist simultaneously with the direct current being measured.

The main filtering is achieved by the a-to-d converter in the digital voltmeter section, by integrating over one or more complete cycles of the mains waveform. This technique provides rejection only near that one frequency, which needs to be 50Hz in Europe. It works adequately only if the a-to-d converter is not

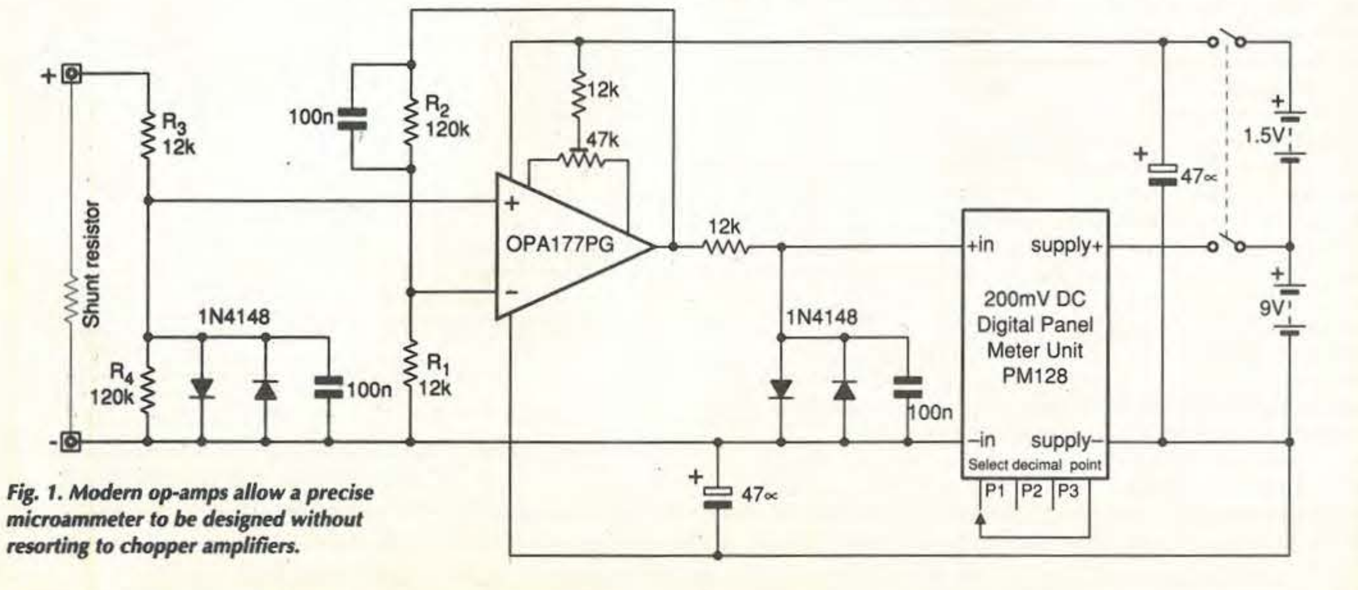

Fig. 1. Modern op-amps allow a precise microammeter to be designed without resorting to chopper amplifiers.

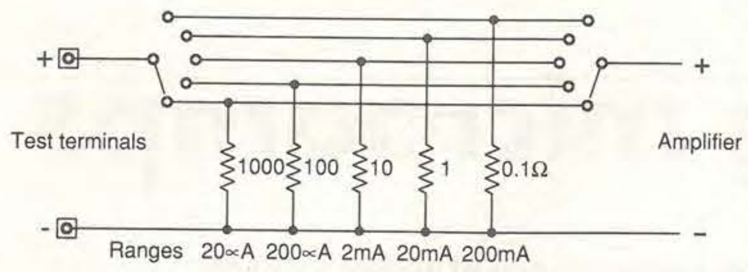

Fig. 2. Milliammeter shunt switching involves doubling up on the contacts to minimise effects of contact resistance.

pushed beyond its maximum input limits by the unwanted alternating signal. This is an important reason for including some signal amplitude limitation. The 100nF capacitors provide useful rejection of higher frequencies.

In Fig. 1, there appears to be no circuitry defining the vital 'common' rail potential for the amplifier input circuit. Since the panel meter section is completely inside the same box as the amplifier and uses the same battery supply, the meter must be allowed to set the level of its own input reference potential. This then becomes the common reference potential for the amplifier and the whole instrument. The reference potential is about 60% of the meter's positive supply voltage – a little too high to allow the very best performance from the op-amp. So although not absolutely necessary, the addition of one cheap AA dry cell, to raise the amplifier supply voltage above that for the meter, is a refinement worth the extra inconvenience.

Details and options

If the amplifier circuit is used to drive the 200mV range of a separate multimeter with its own battery, instead of the built in panel meter, then some changes should be made. The 1.5V cell is no longer necessary, but the remaining 9V supply for the amplifier requires splitting in half to provide a reference potential. Use two equal resistors, 47kΩ, one across each of the 47μF capacitors for example.

Also, some existing multimeters have an input resistance as low as 100kΩ on the 200mV range. If such an external instrument is used instead of the internal panel meter, the protection circuit between the amplifier output and the multimeter would cause an unacceptable loss in output voltage. As a result, the protection circuitry should be omitted.

Shunt switching for the five current ranges: 20μA, 200μA, 2mA, 20mA, 200mA, involves two switch wafers, each an 11-way make-before-break rotary switch. The six spare positions are wired with a short circuit in place of a shunt. This option of 11, rather than 5 switch positions is not shown in Fig. 2 but is strongly recommended. It conveniently enables the meter to be switched safely and quickly away from any sudden or unexpected large current, without interrupting the circuit under test.

A third switch wafer can be added to select the decimal point display of the internal panel meter. This merely completes a connection to one of three points, labelled P1, P2, P3. Note that in the samples I tried, these were in the opposite sequence to that suggested in the accompanying leaflet.

If the 11-way option is chosen then this third switch wafer can also be of the same kind as the others, but if only a 5-way switch is used, the decimal point wafer alone, should be break-before-make.

Useful additional protection would be a further pair of back-to-back diodes, wired across the feedback resistor R_2. I tried this, but rejected it because in a very warm room, the diode forward current was just beginning to upset indications near full scale. To make such a scheme satisfactory would need four diodes, two in series in each direction.

To keep the overall cost very low, all resistors used were metal film types with 1% tolerance. Thus the overall uncertainty limit could, at worst, reach just over 3% plus 0.5% for the meter. The shunts contribute only 1% to the overall uncertainty limit and are cheap enough to be expendable in an accident.

A further 2% depends upon the resistors R_1 to R_4. If these four resistors are upgraded to 0.1% tolerance, their overall contribution to uncertainty will fall to 0.2% and this is certainly worth considering. ■

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Linear integrated circuits

G-p op-amp. Micrel has introduced the *MIC6211*, which is said to be the first general-purpose op-amp in an SOT-23-5 package, whimsically known to some as an *IttyBitty*. It is similar to one of the devices in the quad *LM324*, the miniature package allowing four single op-amp *6211*s to occupy less space than in the *324*. Operating from either single 4 to 32V or ±2V to ±16V supplies, the *6211* provides high gain and is frequency compensated. Current drawn is 2mA, unity-gain bandwidth 2.5MHz and slew rate 7V/μs. Micrel Semiconductor Europe. Tel., 01635 524455; fax, 01635 524466.

Microprocessors and controllers

New PICs. New 8-bit microcontrollers from Microchip, the *PIC16C66/7* feature extensive communications and a 5Mips performance. They are eprom-based and the user operates at up to 5Mb/s, have 35 single-cycle 200ns instructions, 8K by 14 words of program memory and 368byte of data ram. Current taken from a 33kHz 3V supply is under 15μA and 1μA in sleep mode. Of the 40 pins, 33 are for communications, which include an 8-bit asynchronous slave port, a synchronous serial port for DPI or I^2C protocols, two pwm outputs at 80kHz and 8-bit resolution and a 16-bit capture/compare facility with 200ns resolution. The chips also drive leds and triacs directly. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Mixed-signal ICs

Digital video demodulator. Siemens' *TDA6180X* amplifier and demodulator ic handles qpsk modulation in digital video broadcast signals, converting signals at 300-600MHz to I and Q analogue levels. Functionally, the device is a qpsk-input amplifier, an agc oscillator which generates the 0° and 90°signals for demodulation and a qpsk demodulator. Filtered if goes to a variable-gain amplifier and is then demodulated to baseband. It is the split and passed to two low-level detectors, from which it is amplified and output as Q and I. Siemens plc. Tel., 0990 550500; fax, 01344 396721.

Optical devices

High-voltage isolators. Isocom has a 300V optoisolator, the *IS7000*, that withstands 300V across its output transistor; it is meant for use in positions such as fax modems and power supply interfaces. Input/output isolation is to 5kV and the darlington output stage produces a current output of up to 150mA on inputs down to 1mA, or much more with pulse inputs. Turn-on and turn-off times are 50μs and 15μs and performance is held over the –30°C to 100°C range. Isocom Components Ltd. Tel., 01429 863609; fax, 01429 863581.

Dil clocks. Dual-in-line clock oscillators from Total Frequency Control can be specified, made and delivered within seven working days. Frequency range is 4-30MHz to within ±5%, ageing rate ±1ppm/year and output is hcmos/ttl-compatible. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie @ tfc.co.uk.

PASSIVE

Cameras

Conference camera. *QuartzSight*, developed by Vision and Rockwell, is a digital colour camera, made to international standards, for use in pc-based videoconferencing. The camera itself is simply the lens and cmos single-chip colour sensor, while the *QuartzCapture* chip set contains many of the camera dsp functions. Digitised video signals, power and control are all carried by a 9-wire cable. Colour balance and exposure control are automatic, while the user controls focus, colour saturation, contrast, brightness, frame rate and resolution. An anti-flicker mode reduces the effects of fluorescent tubes. VLSI Vision Ltd. Tel., 0131-539 7111; fax, 0131-539 7141.

Communications equipment

Ringing generator. The *PowerDsine PCR-SIN01* is a 1W sine generator for telephony, putting out selectable 17, 20 and 25Hz frequencies for different countries. Current taken from 12V or 24V dc is 90mA and rf filtering is included. It is for use in single or dual line applications and an external inhibit control reduces idle power, the control also being used to generate the ringing signal intervals, so doing away with relays. Overload protection is provided. DIP International Ltd. Tel., 01223 462244; fax, 01223 467316.

Crystals

Surface-mount crystals. New crystals from Total Frequency Control are packaged in resistance-welded metal cases with preformed leads for automatic surface-mounting, special mounts inside the package conferring resistance to shock and vibration. Frequency tolerance over the –10°C to 60°C range is ±3ppm and ageing is 0.5ppm/year maximum; the company points out that this level of performance is not available in conventional s-m packages. Total Frequency Control Ltd. Tel., 01903 745513; fax, 01903 742208; e-mail, eddie @ tfc.co.uk.

Memory

Compact flash. New range of compact flash cards by *Kingmax* is intended to increase the number of pictures stored in digital cameras and the data available to mobile telephone users. Used with a camera, it avoids the need to download the data straight from the camera; many cards may be filled with data at high definition and interfaced to a pc. The range of capacities is 2-16Mbyte and a high-speed buffer allows transfer rates to and from memory at up to 8Mbyte/s, a speed compatible with the fastest digital cameras. Cards are compatible with PC Card services and Socket Services and, by the use of an adaptor, can be plugged into PCMCIA sockets. Premier Electronics Ltd. Tel., 01922 634652; fax, 01922 634616.

Displays

Small graphics panel. From Nan Ya, the *LMCDB078* liquid-crystal display panel provides 320 by 240 resolution and measures 134.5 by 117 by 14mm. It uses a transreflective technique and can be used with no backlight, so saving power. Neither does it need inverting circuitry and therefore also saves space. Mono displays in the range include types with an integral matrix touch screen in which, since no controller is needed, a simple buffer like those in membrane keypads can be used. These are transreflective or transmissive versions in blue and white or black and white. Anders Electronics plc. Tel., 0171-388 7171; fax, 0171-387 2951.

Filters

GSM saw filters. Surface acoustic wave filters in Fujitsu's *F4* range are meant for if applications in the 200-600MHz band, GSM filters being available at 246MHz and 400MHz. Inputs and outputs are balanced and attenuation in the stop band is better than 50dB. Surface-mounting packages measure 5 by 5 by 1.6mm for the 246MHz filter and slightly larger for the 200MHz device. Fujitsu Microelectronics Ltd. Tel., 01628 76100; fax, 01628 781484.

Hardware

Joysticks. Sakae joysticks, meant for plant control, robotics, wheelchairs and games, use conductive plastic tracks and have a life expectancy of up to five million operations. Models cover the range of sizes from minute to huge and can be had in 1, 2 or 3-dimensional form with optional switches, being built to ISO9001 standard. They withstand vibration to 10g and shock to 30g and may be supplied with fittings and a choice of knobs. Dust gaiters are available. Techni Measure. Tel., 01527 854103; fax, 01527 853267.

RCA jacks. Shiua Chyuan of Taiwan supply RCA jacks in a number of forms and they are now available in the UK. The jacks come in single units or with 2-8-way types, with optional emc screening. They are meant for high-volume use where fast installation is needed. Plating options are gold, half-gold and silver, resulting in a minimum contact resistance of 30mΩ and reliability sufficient for 5000 cycles. Colours can be specified and both horizontal and vertical barrels are available. TW Electronics Ltd. Tel., 01635 278585; fax, 01635 278122.

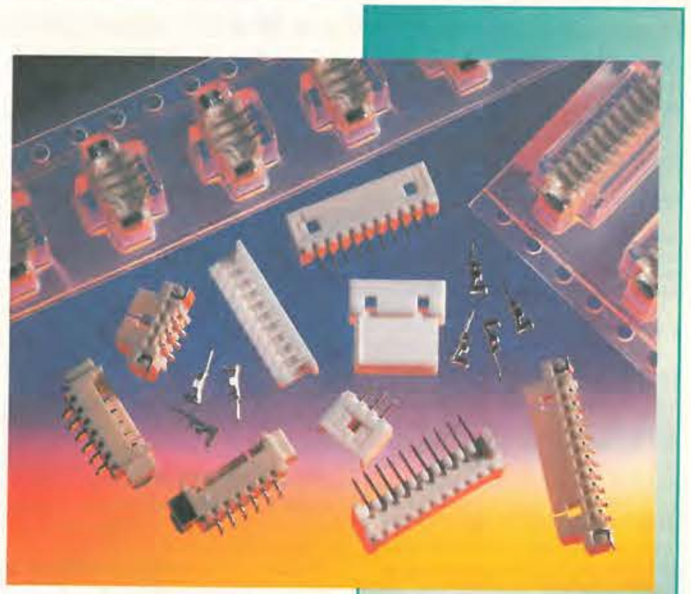
Ceramic adjusters. *Torayceram* tools, now available from Toolworld, are for the adjustment of trimmers and inductors, ceramic business ends avoiding any interference with component characteristics even at high frequencies. A manufacturing process confers resistance to bending forces and the chipping that can occur in zirconia ceramics. The tools come in a range of sizes, with interchangeable bits and with long thin handles for awkward positions. Also available are ceramic tweezers, which provide the same advantages of long life (ten times longer than metal and 100 times longer than plastic) and no magnetic materials. Toolworld Ltd. Tel., 01249 821234; fax, 01249 816723.

Cable tidy. Richco International's latest offering is a little plastic bit that clips into a 5mm hole on the edge of a pcb and holds any odd wires coming off the board in a 'saddle'. It is known as an edge holding wire saddle and is made in nylon 6/6, RMS-01 or flame-retardant nylon 6/6 and RMS-19. Richco International Co., Ltd. Tel., 01474 327527; fax, 01474 327455.

Chip coolers. Sanyo Denki *San Ace MC* coolers for Pentium and Alpha microprocessors are now available in the UK. They consist of miniature fans with integral heatsinks and come in four sizes from 45mm square by 18mm to 66 by 62 by 30mm, each being made in 5V or 12V versions. A silicone compound is used as a thermal interface. Noise level is low at about 28dB(A), there is locked-rotor protection and an alarm output is optional. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Test and measurement

Acoustics and vibration. Developed by Müller-BBM GmbH, the *PAK* multi-channel acoustic and vibration measurement and analysis system is based on a Hewlett-Packard VXI front end with an *HP9000 Series 700* workstation to run the *PAK* software under the HP-UX os. Analysis software handles time and frequency domains, octave analysis, order tracking and psychoacoustic analysis; additional parameters for engine analysis can also be captured. Both Yard and Oracle databases are supported and a report generator is provided. Strategic Test and Measurement Systems Ltd. Tel., 01734 795950; fax, 01734 795951.



Exotic multimeter. Capable of capturing data 3.3 times per second and of true-rms measurement, the *Escort EDM89S* handheld multimeter measures 500mV-1000V dc and true-rms alternating voltage, 5mA-10A direct and alternating current, 500Ω-50MΩ resistance, frequency, 5nF-50µF capacitance and dBm with selectable reference impedance. A range of features includes automatic or manual dynamic recording, 5000-count digital readout and an analogue bar graph. Bandwidth on ac measurements is 20kHz and there is dual-tone, dual-display logic detection, a relative mode for deviation, audible and visible input warnings and power-saving modes. Feedback Test and Measurement. Tel., 01892 653322; fax, 01892 663719.

2.4GHz counter/timer. Available from TTi, the *Tabor 6000* range of GPIB-interfaced counter/timers consists of three models: the *6010*, measuring up to 125MHz, extendable to 1.3GHz; the *6020* to measure to 225MHz and extendable to 1.3GHz; and the *6030* for 225MHz and 2.4GHz. The latter has two independent inputs for frequency measurements up to 300MHz, below which reciprocal measurement is automatically selected. Optionally, a third channel will measure up to 2.4GHz on an input of 15mV. Period and pulse measurements(5ns-2000s) and time interval A to B (0-2000s) can be made, with the provision for up to 500 programmed delay intervals to be inserted between start and stop of the time interval A to B. Stability with an optional oven-controlled time base is 0.15ppm/year. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Interfaces

GPIB interfaces. *ComputerBoards Inc.* announces a range of GPIB interfaces for ISA, PCI and PCMCIA, all significantly cheaper than competitors, the PCMCIA board being about half the common price. These GPIB IEEE-488.2 interfaces let any pc function as instrumentation control and data acquisition system. They use the standard CB7210.2 GPIB chip for full talker/listener/controller functions and allow up to 14 instruments to be connected to the pc. Supplied software supports programming under dos, Windows 3.1x and '95. Data transfer rates are up to 1Mbyte/s and the PCMCIA and PCI models are of the plug-and-play type. Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213.

Connectors and cabling

"Smallest" connectors. *Molex 1.25* connectors are said to be the smallest available. There are wire-to-board types and wires-to-other-wires versions, all with flat, twin-contact pins pitched 1.25mm, taking 1A maximum and in detachable through-hole form and for surface mounting. With 2-15 connections, the pins have a maximum contact resistance of 20mΩ and insulation resistance of 1GΩ. Connectors are polarised, the headers having friction locks. Hawnt Electronics Ltd. Tel., 0121 784 3355; fax, 0121 783 1657.

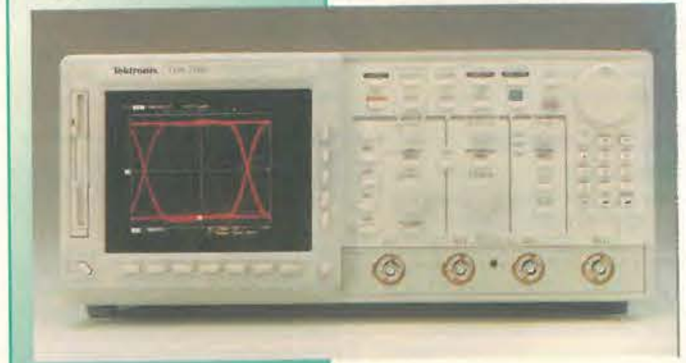
Literature

Optoelectronics. Onboard Electronics offers a 100-page catalogue of optoelectronic components from *Kingbright*, including super-bright led lamps and bright blue leds. Also described are surface-mounted, shaped and large-sized lamps, led clusters, seven-segment and dot-matrix displays. Onboard Electronics Ltd. Tel., 01256 818222; fax, 01256 840610.

Relays. Free from Ashwell Electronics, the 24-page *Teledyne Gentron* data book of solid-state relays, which covers devices from 10A board-mounted types to three-phase, 1000A devices. Ashwell Electronics Ltd. Tel., 01438 364194; fax, 01438 313461.

Servo amplifiers. Ninety brush and brushless servo amplifiers for motion control are described in Copley Controls' eight-page short catalogue. Power ratings from a few watts to 20kW, and single-axis and multi-axis types and sub-systems are all included, as are dc psus, transformers and amplifier mounting hardware. Most Copley pwm amplifiers can be made to match specific motors by plugging resistors and capacitors into a no-soldering header. Amplifier features include built-in power filtering or edge filters to reduce output rise rate to prevent rfi coupling to other circuits. Copley Controls. Tel., 001 617 329 8200; fax, 001 617 329 4055.

Scope for communications. Newly introduced by Tektronix, the *TDS 700C* and *TDS 500C* family of *InstaVu* acquisition digital storage oscilloscopes for use on communications equipment such as switches, routers and videoconferencing equipment, together with a set of accessories to enable connection of optical and electrical signals to the instruments. A feature of the oscilloscopes is the facility for mask testing of communications signals; there are 39 automatic telecomms and datacomms masks and custom masks can be built by the user. In compliance test, the mask for a particular standard is displayed, the signal being scaled and correctly offset for comparison. Statistical tools are included for analysis in the form of statistics and histograms. The two series are available in two and four channel models with sampling rates to 4Gsample/s, up to 8Mbyte of record length and 1GHz bandwidth. Tektronix UK Ltd. Tel., 01628 403300; fax, 01628 403301.



Emc gaskets

Holland Shielding Systems has a new engineering handbook on its range of emc gaskets for electronic enclosures. Most of the gaskets, which are in the 0.8-60mm range of sizes, can be used in existing equipment without design changes by virtue of the self-adhesive backing and low closure force required. The gaskets are in strip form or as ready-made gaskets, are in flame-retardant or high-temperature versions and are also available with an environmental seal. Holland Shielding Systems bv. Tel., 0031 78 6131366; fax, 0031 78 6149585.

Materials

Shielding tape kit. 3M supply an engineering kit for foil shielding tapes to allow engineers to obtain small quantities. The tape provides reliable point-to-point electrical contact for emi/rfi shielding and static draining and is provided with a range of backings and adhesive compounds. Tapes are 0.75in wide and 144in long, replacement rolls being available. Each of the nine tapes in a box is fully described on the box. 3M United Kingdom plc. Tel., 01344 858739; fax, 01344 858758.

Anti-emi coating. Chomerics *Cho-Shield 2052* is a conductive coating to give a high level of emi shielding on plastic materials such as polycarbonate, ABS, PC-ABS, Noryl and pvc. It consists of silver-plated copper particles dispersed in a one-component acrylic resin, which works with all available material application methods including high-volume, low-pressure sprays and propeller-agitated pressure pots. The material is abrasion-resistant, stable at high and low temperatures and withstands high humidity and moderately saline fog environments. Parker Hannifin plc, Chomerics Division. Tel., 01628 486030; fax, 01628 476303.

Thermal grease. *Easy-Ply* from Redpoint Thermalloy can improve heatsink performance by thirty times. It is applied to the heatsink at the factory and protected by a pull-off cover until needed, whereupon the whole heatsink with the grease in place is put into position with the size and shape of grease interface tailored to the application. Easy-Ply is non-toxic and stable and works over the −40°C to 204°C range, with an indefinite shelf life. Redpoint Thermalloy Ltd. Tel., 01793 537861; fax, 01793 615396.

Printers and controllers

Thermal print heads. Rohm's new series of thermal print heads come in printing widths of 2, 3, 4, 8 or 10in and with 8-dot/mm resolution. All models have Rohm's own drive ics. The *GC20* is a label printer and replaces the *C1S* series; it has superglass protection on the dot line, epoxy ic coating and a shrouded, polarised connector. *GC10* is similar, but smaller. Lowest cost model is the *GH1*, which is a small fax head, meant for epos printers or machines giving receipts. Kestronics Ltd. Tel., 01727 812222; fax, 01727 811920.

Production equipment

Metal wrist band. For static control, TBA ECP has a new metal wrist band that is more comfortable than elasticated types since it can be worn looser and is said to be safer than other metal ones on the market. It is fully adjustable and has a stainless steel linked inner layer with a plastic moulded outer cover, which is immune to scratching and cannot expose the metal, so preserving the wearer from possible contact with live parts. There is a 10mm stud to receive a standard coiled cable. TBA Industrial Products Ltd. Tel., 01706 47718; fax, 01706 46170.

Piezoelectric pump. Densitron's *Crystal pump* uses a piezoelectric stack element to give precise control of fluid flow. Metering of the fluid is obtained by varying crystal drive amplitude and frequency to give flows at rates between 0.05µl/min to 380ml/min; frequency can be at frequencies as low as 20Hz when noise might be a problem, although the noise at any frequency is small. Response is virtually instantaneous for the precise metering of small quantities of fluid. The pump body is available in Teflon, polyacetal, stainless steel or other materials to suit different fluids. Densitron Europe Ltd., 01959 700100; fax, 01959 700300.

Low-level switching. EAO-Highland's *84 Series* is a modular system of pushbuttons and indicators that can be assembled to make a wide range of switches by combining lenses, actuators, switches and mountings. The units are suited to low-voltage switching in industrial positions, for example in computer-based controls. Switch ratings are from 50mV to 42V and 10µA to 100mA ac or dc, maximum power rating being 2W and typical rebound time under 1ms. They can be had in aluminium or plastic and flush mount in a standard 22.5mm diameter cutout. Depth behind the panel is 18-31mm and the switches are sealed to IP67, resistance to vibration and shock is high and the metal fronts make them proof against electrostatic discharge. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Fluid dispenser. From Camelot Systems Inc., the *CAM/ALOT 680 Series* rotary positive displacement pump or glop dispenser accurately provides encapsulation and flip-chip underfill in precise quantities with no drips; it will pump 10-200mg of fluid to within ±2% by volume. It has a positive needle shut-off and the material is under constant low pressure to prevent cavitation. There

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is a micrometer built in to help adjust the gap between the dispensing auger and its seat and the interchangeable cartridge assembly can be cleaned in minutes. A heating cartridge brings the material to the required temperature. Dage Precision Industries Ltd. Tel., 01296 393200; fax, 01296 331585.

Power supplies

Micro-controlled 600W psu. Vicor's *FlatPAC* family of 50-600W supplies has a microprocessor-controlled front end and now has industry-standard input terminals and high-current output connectors. Input is autoranging. Thousands of configurations are possible from systems with one to three outputs at various powers, the *FlatPAC* family being modular in design; units contain up to three dc-to-dc converters, each providing up to 200mW, in a choice of chassis-mounted enclosures measuring 35mm high, 219mm long and 64, 125 or 186mm wide. Regulation is 0.05%, ripple 1% pk-pk, efficiency 80-90% and total remote sense compensation 0.5V. XP plc. Tel., 01734 845515; fax, 01734 843423.

Low-profile 150W supplies. First arrivals in a new series of low to medium-power supplies from Power-One single and multi-output switched-mode supplies are making an appearance. *MPU150 Series* units are in 1.5in 1U U-channel cases with optional covers and have power-factor correction to IEC61000-3-2 with a 85-264V ac input range. With air cooling given by the user's system or by an optional fan, each unit will produce 150W and other features include mains fail, current sharing, thermal shutdown, remote sensing and optional remote inhibit. The *MPU* series meets a truly impressive number of emi, rfi and safety standards. Power-One Europe. Tel., 01769 540744; fax, 01769 540756.

Compact 25W dc-to-dc. In a package measuring 2 by 2 by 0.375in, the *OB Series* of dc-to-dc converters by Amplicon give up to 25W, pin-out being to industry standard. Input is 12-48V dc, efficiency up to 86% and outputs 5/12/15V dc and ±5/±12 and ±15V dc, outputs being trimmable by ±10%. Regulation and stabilisation are both 0.2%, temperature coefficient ±0.02%/°C and ripple 150mV maximum. Cased or open models are available. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

Charger controllers. *ADP3810/1* by Analog Devices are battery charger controllers offering programmable current limiting and precise voltage limiting for the constant-current, constant-voltage charging of lithium ion, nickel cadmium and nickel metal hydride cells. The *3810* has its own precision resistors for 1% accuracy on Li-ion cells, while the *3811* uses external resistors for a wide range of of output voltages. Control voltage input, which is buffered, takes either an *RC*-filtered pwm microcontroller output or the divided reference voltage. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Charge-pump converters. National Semiconductor's *LM2660/1* are cmos charge-pump voltage converters for battery-powered equipment, PCMCIA cards and hard-disk drives. Conversion is 1.5V-5.5V at 90% efficiency, operating current being 120µA for most loads. Output impedance is 6Ω, giving a droop of 0.6V at 100mA. The *LM2661* also has a low-current shutdown mode of operation, in which it draws less than 1µA. National Semiconductor GmbH. Tel., 0049 1805 32 7832; fax, 0049 814103515.

3W dc-to-dc converter. Coutant's *PP3* low-profile converters have been reduced in price, due to new methods of manufacture and a move to another factory. The *PP3-48-12* is the first in the cheaper series and is suited to distributed and modular systems. Input is 32-63V dc, input current at 48V being 90mA. A single 12V, 120mA output is accurate to within ±5%, ripple is 150mV and efficiency is 75%, with stabilisation and regulation at 0.4% and 5% respectively. Coutant Lambda Ltd. Tel., 01271 865656; fax, 01271 864894.

Radio communications products

Receiver chip set. RF Monolithics' chip-set for its amplifier-sequenced receiver, which needs no frequency changing, for remote-control and security use is available from Acal. *ASH* chips include a low-loss surface acoustic wave filter for front-end selectivity, a saw delay line and rf circuitry to form the amplifier-sequenced gain and detector stages Typically, sensitivity is −100dB without the use of heterodyning or regenerative circuits; virtually no rf emissions are produced. Chip-sets for frequencies between 303.825 and 433.92MHz are available, all running on 3V lithium batteries and drawing 1.1mA. Acal Electronics Ltd. Tel., 01344 727272; fax, 01344 424262.

40GHz up/down converters. Transmission up-converters and receiver down-converters from Anglia Microwaves are for 40GHz video distribution. The *DUC-4042N810* and *DDC-4042N610* cost less than travelling-wave tubes and are more reliable, life-expectancy being ten years of full-power operation. Intermediate frequencies of 1.85-2.15GHz and rf at 40.5-42.5GHz are converted, local oscillator costs being reduced by the use of a multiplier before the mixer stage, which allows the use of X-band instead of K-band components. If/rf conversion gain in the *DUC* is 20dB to give an rf power output of 1mW, or 200mW with an optional amplifier. Oscillator leakage and sideband noise are a minimum of 20dB below signal. Down-conversion gain in the *DDC* is 30dB and noise 8dB in ssb operation. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Switches and relays

S-m fuses. Three new surface-mounted *PolySwitch* resettable fuses by Raychem are intended for use with Universal Serial Bus ports, protecting against shorts and overcurrent conditions when plug-and-play boards are used. In the presence of a fault, the fuses switch to a high-resistance state, reverting to normal when the fault is cleared. *SMD20* satisfies a demand for low resistance (50mΩ) and voltage drop on USB ports; *miniSMDC110* is for self-powered port protection, having a 1.1A hold current and 160mΩ resistance; and *miniSMDC075* is for bus-powered application, holding on 0.75A and showing 300mΩ resistance. Raychem Ltd. Tel., 01973 572692; fax, 01973 572209.

Transducers and sensors

Level and flow sensors. Crydom has ranges of liquid level and flow sensors with a large number of options to meet most applications; if a requirement in not available, there is a design service to produce it quickly. Level sensors use float switches operated by reed-switch and magnet in vertical and horizontal configuration and internal or external mounting; materials used for manufacture are specified for the liquid in use, from boiling water to acids. The flow detectors use double-isolated reed switches and are housed in Hoechst Hostaform, which is acetal copolymer. Models in the range work in hot or cold water low-pressure systems and in alcohol or dilute acids and there are variants with triac buffers for improved switching, particularly with inductive loads. Crydom. Tel., 01202 897 964; fax, 01202 891 918

"Lowest-cost" PIC Ice. Jointly developed by Neosoft and RF Solutions, *ICEPIC Junior-16* is a modular unit that reduces development time and cost by working with all *PIC 16C6/7/8/9X* processors. It runs in a Windows environment, using the same gui as others in the *ICEPIC* family, and is non-intrusive, being capable of source-level debugging in assembler or C at the maximum processor speed. It has a standard Microchip emulation chip, 8K of 16Cxx emulation memory and a fast RS-232 interface for communications with the host at 57.6kb/s. The basic emulator costs £299 and additional modules for other processors £99. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661

COMPUTER

Computer board-level products

Network controller for cards. *TDS2020* 16-bit and *TDS9092* 8-bit embedded computers by Triangle Digital Services can now be networked using the *TDS2020CAN* board, which fits under one of the *TDS* cards to connect it to a multi-drop RS485 CAN network. Users are then able to, for example, display messages on a remote lcd, toggle relays and lamps in remote equipment, acquire data and control remote equipment. CAN networking is over two twisted pairs over distances of up to 1km, with up to 110 nodes, each optically isolated. Triangle Digital Services Ltd. Tel., 0181-539 0285; fax, 0181-558 8110.

Computers

200MHz single-board computer. Eltech Solutions announces *Eurocom 128*, a VMEbus single-board computer using the 200MHz Pentium processor, with options of IDE or SCSI hard disk support by way of the P2 connector and a choice of operating software including *VxWorks*. Its 64-bit wide memory allows the use of 8Mb to 128Mb ram using standard 8Mb PS/2 simms, memory size being automatically detected; memory can be optionally boosted by 256Kb second-level cache and EDO rams with hyper-page mode. Bios is in updateable boot-block flash eprom and there is a 2K serial eprom for user's data. A variety of monitors, both crt and flat-panel displays, is on offer and the PCI-based graphic interface is compatible with the IBM VGA standard as well as backward to the EGA and CGA standards. The board has two asynchronous 16550 compatible serial channels with hardware handshaking and a bidirectional Centronics port supporting the IEEE1284 enhanced mode. Standard pc software is supported. Eltech Solutions UK Ltd. Tel., 01908 260044; fax, 01908 260012.

Workstation for clean rooms. Deeco, part of LucasVarity, has produced the *Sealtouch* clean room computer, which has virtually no effect whatever on its surroundings; it is completely sealed and has no need for a keyboard or separate mouse. No fans or filters are needed for cooling, since the computer is housed in a cast aluminium case that acts as a heat sink, and the infrared touch screen and flat-panel electroluminescent or active-matrix lcd colour display are sealed to prevent air entering or leaving. The touch mouse and on-screen keys can be used precisely even when wearing gloves. Lucas Control Systems Products. Tel., 0049 7024 971214; fax, 0049 7024 971240.

Data communications

Wireless RS-232. Radio Data Technology claims its *LPRS232* is the first reliable application of radio to data communication using RS-232 protocols. The transceiver is available in 428MHz and 433.92MHz with a range of 120m line-of sight or 30m in buildings, with a longer-range version, the -300 type, providing a range of over 1km. Transmission speed is selectable between 4.8kb/s and 19.2kb/s. Normally, the unit remains in receive mode until data arrives at its RS-232 input or an rf carrier is detected, whereupon the data is transmitted in packets with a header block containing a sync. word, the number of bytes in the packet and a checksum. The unit is compact, measuring 114 by 63 by 17mm, uses a single 8-12V supply and draws 45mA unloaded. Low Power Radio Solutions Ltd. Tel., 01993 709418; fax, 01993 708575.

Development and evaluation

Low-cost *68HC(7)05* development. Motorola has announced *JICS*, which is a £30 development kit for the 8-bit *MC68HC(7)05J1A* microcontroller. It contains everything needed for development and debugging, including assembly programming, editing, software simulation and in-circuit simulation. The kit provides a sample device, cabling, power supply and full documentation. Motorola. Fax, 01354 688248.

Software

Power analysis. Voltech has Windows software for visual power analysis to assist in the remote configuration, control and monitoring of single and three-phase power measurements. The software is designed for use with the company's low-cost *PM100/300* power analysers and eliminates the need to program the analyser's front panel, enabling remote set-up, running and recording of measurements including real and apparent power, crest factor, efficiency, harmonic distortion and inrush current. Results are shown graphically or numerically and both ASCII and CSV formats are used for transfer to word processors and spreadsheets. New capabilities conferred by the software include modification of the thd formula, adjustment of averaging depth and the setting of fixed ranges for all power measurements. Voltech Instruments Ltd. tel., 01235 861173; fax, 01235 861174. ■

Skin effect

Nick Wheeler outlines skin effect and describes why it is irrelevant at audio frequencies.

Everyone involved with electronics will have heard of skin effect. And most will know that it has the effect of increasing the effective resistance of a conductor as signal frequency increases. This is due to the current flow being increasingly concentrated on the outer surface of the conductor.

The effect can be so pronounced that in some cases a thin-walled tube is as effective as a solid conductor. Silver plating and the use of oxygen-free copper are commonly used to reduce the effects of skin effect.

Silver plating reduces resistance of the conductor skin while oxygen-free wire has a fractionally lower resistance than the standard-quality option. Oxygen-free cable can be simulated by simply increasing conductor cross section. The diameter increase necessary to do this is so small as to be negligible.

Note that effective resistance, as referred to above, is also known as 'rf resistance'. It is an ohmic effect, and nothing to do with reactance. It's operation can be detected thermally.

What causes skin effect?

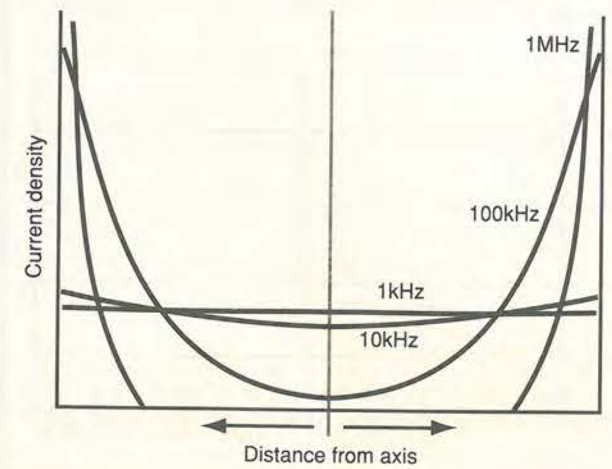
Consider a conductor made up of many strands of thin uninsulated wire, and carrying an alternating current. Current flowing through any single strand produces a field which cuts through the bulk of the surrounding conductors. In effect, these conductors form shorted turns as they are in intimate contact.

Currents flowing in the shorted turns oppose that flowing in the specimen strand, but those currents have to go somewhere. It is easy to show that they concentrate on the outer surface of the conductor bundle, or in the case of a single large conductor, on the surface.

Fields associated with the outside surface intersect less conductor material. As a result, the opposition to current flow is less on the surface, which is where it then flows. This effect depends strongly on increasing frequency and – because of the larger cross-sections involved – with increasing sizes of solid or uninsulated stranded conductors.

Table 1. Effect of frequency on rf resistance in two wires with different diameters.

Frequency (Hz)	Wire diameter 2mm	0.2mm
10	1	1
100	1	1
1k	1	1
10k	1.095	1
100k	2.626	1
1M	7.7	1.095
10M	23.8	2.626
100M	74.7	7.7



Current distribution across a 2mm conductor at various frequencies demonstrates skin effect.

Stranded versus solid conductors

Imagine a conductor comprising 55 uninsulated strands of 0.1mm diameter copper, such as RS356-167. This has a copper cross-section of 0.43mm^2 and an equivalent solid diameter of 0.74mm.

At high kilohertz and megahertz frequencies, this cable performs very much the same as 22SWG solid, with a diameter of 0.71mm which has, as predictable from theory, fractionally less rf resistance. A conductor made up of 15 strands of enamelled 36SWG, amounting to a cross-sectional area of 0.425mm^2, as opposed to the 0.43mm^2 of the 55 uninsulated-strand wire, has about half the rf resistance at frequencies in the 1 to 2MHz region. This is, I think, as high as you need to consider if you are interested in audio.

Note that these principles are not new. They have been known and applied since the earliest days of radio, in the form of Litz wire.

The facts

It is a matter of simple scientific fact that – to three decimal places – a conductor of 0.2mm outside diameter has negligibly more rf resistance at 100kHz than its dc resistance. What this means is that, up to this frequency, the current distribution is uniform across the conductor cross-section, thereby making oxygen-free copper and silver plating irrelevant for audio work.

At 1MHz the rf resistance is up by 9.5%, where silver plating and so on really do start to be useful. On the other hand a conductor of 2mm outside diameter has 2.625 times the rf resistance, compared with dc resistance, at the significantly low frequency of 10kHz.

Figure 1 from reference 2, illustrates this, and **Table 1**, from the same source, summarises some experimentally determined data. Of course, if a conductor of very large cross-section is used, the dc resistance, and consequently the rf resistance, will be low in comparison with the highly variable impedance of the loudspeaker load.

Domestic twin-and-earth cable of adequate cross-section may well be the most cost-effective route[1]. In terms of power loss, rf resistance is irrelevant in audio applications. What matters is that the low output impedance of a good amplifier may not be seen as effectively damping hf loudspeaker transients.

In summary

What this boils down to is that cables made up of many strands of thin uninsulated wire offer no advantages over the same cross-section of solid wire, except mechanical flexibility. Separately insulated multistrand cable can offer genuinely superior performance.

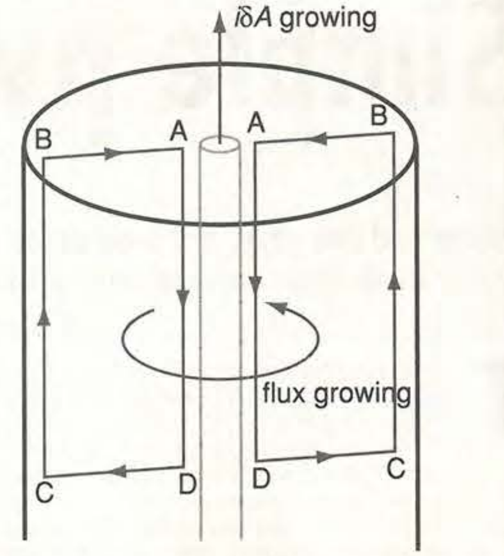
It would be prohibitively expensive to make up a coaxial cable of which the outer braid was composed of insulated wires, commoned at each end, and with an insulated multi-strand inner, but this must be the ultimate cable. ■

Reference
1. Speaker Cables, C. Bateman, EW Feb 1997
2. Services Textbook of Radio Volume 1 HMSO 1956.

Skin deeper

The following is an extract from reference 2.

Current density in this wire is iA/cm^2. Current $i\delta A$ flows through an area δA on the axis of the wire and grows, creating magnetic field *H*. Through any radial section of the wire, such as ABCD, flux due to H is growing, resulting in an emf between A and D along the axis, and from C to B on the outer side. This emf tends to reduce current flowing along the axis and increase that flowing along the outside of the wire.



Skin effect. Any change in current along the axis causes induced voltages which oppose this change along the axis and promote it in the outer regions.

Simple pwm power amp

Designed for efficient low-frequency and dc power driving, Richard Lines' pwm amplifier can deliver up to 5A at 25V.

This note describes a pulse-width modulated amplifier capable of producing a unipolar output of up to 25V at 5A.

Designed for simplicity, the circuit is based on an *LT1074CT* switch-mode power supply IC from Linear Technology. It is suitable for use in servo systems, or as a heater driver or thermoelectric cooler, where power conversion efficiency is a priority and the control signal is a few hertz or less.

The design is especially useful in low-voltage high-current applications, where a standard linear regulator would waste a lot of power.

Circuit operation

The chip has an internal oscillator running at 100kHz. The square wave generated is pulse width modulated with a duty cycle controlled by the voltage on pin 2 (out). The chip's main output appears at pin 4 (switch); this is the dc input chopped according to the duty cycle.

Components D_1, L_2, C_4 form the conventional flyback circuit and low pass filter. Inductor L_2 needs some consideration: the magnetic core needs to be large enough not to saturate at peak currents.

My prototype consisted of 32 turns of 1mm wire on an *RM400* pot core. If the full 5A output is needed then the inductance should be dropped to 50μH in accordance with the device data sheet. Capacitor C_4 needs to be rated at the power supply voltage since the output voltage will rise to this level if the load is accidentally connected.

The feedback pin connects internally to the error amplifier input. In the usual power supply circuit, it samples the output voltage via a potential divider. Also, the output pin connects internally to the error amplifier output. This makes it possible to build an inverting amplifier from the error amp with gain set by R_1/R_3.

I found a swing of something less than 1V at pin 2 enough to sweep the duty cycle from 0% to the chip's maximum of about 90%. In the prototype application, an offset current was required into the summing junction. This is provided by R_2 from a regulated 5V source.

Components R_4 and C_3 are recommended by the manufacturers for circuit stability. Capacitor C_2 was found necessary for the same reason.

It is important that D_1 is a high current Schottky barrier type. Resistor R_5 and C_5 were fitted to reduce ringing; they are not necessary and are not recommended by the chip manufacturer on the grounds of slightly reduced efficiency. However the output was less noisy with them fitted and this was an important consideration. An obvious alternative is to fit an extra low pass section in the output lead.

The chip contains an analogue multiplier. This is used to reduce the duty cycle with increasing supply voltage thus keeping the output voltage approximately constant – depending only on the control input. Without this circuit element, any power supply variations would be seen in the output in proportion to the duty cycle. The presence of the analogue multiplier gives a useful degree of ripple rejection.

Input filter L_1/C_1 is recommended to stop switching transients getting back into the general power supply. Capacitor C_1 should be close to the IC.

Setting up

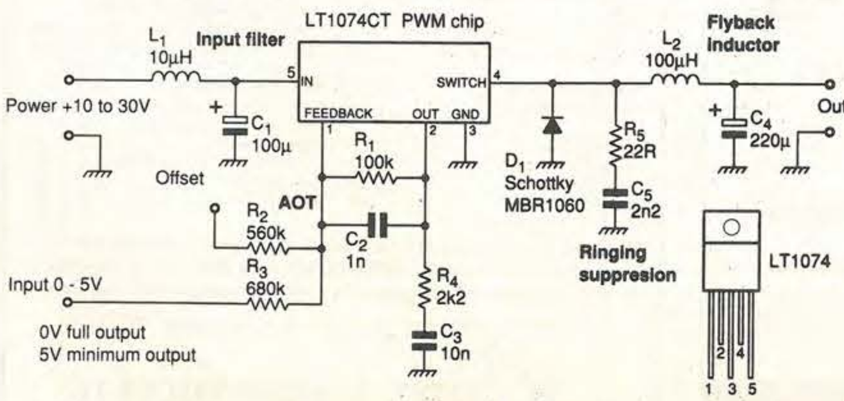
First consider the maximum duty cycle required. This will depend on the ratio of output voltage to power supply voltage and don't forget to allow about 1.5V lost in the switching transistors.

With the input open circuit resistor R_2 should be adjusted so as the output voltage is about half the expected maximum with a suitable dummy load connected. Check the waveform on pin 4 and see if the duty cycle is as expected.

Now adjust R_3 if necessary to ensure that the output goes from nearly 0V to the expected maximum as the input control voltage goes from 5V to 0V. Pin 1, feedback, will always be at 2.21V with respect to ground if the circuit is operating correctly.

I used this circuit as part of a temperature controller servo loop. With a power supply of 15V, the prototype produced 0 to 11V with the component values shown. This was used to run a small heater of 10Ω resistance.

Efficiency was measured at over 80%, except for very low outputs. The circuit is quite linear over the middle 80% of the range but deviates at both extremes. This was not of significant for me since the circuit was inside a temperature controller loop. ■



Designed as part of a temperature controller servo loop, this circuit produced 0 to 11V with the component values shown using a 15V supply. The controller was used to run a small 10Ω heater. Zero volts at the input gives full output while 5V gives minimum output.

LETTERS

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

Charge alkaline cells

I read with interest the article on recharging of alkaline cells in the March 1997 issue of *Electronics World*. There are two points that appear to be missing. They are of a supplier of suitable pcbs and a supplier of the special value resistors R_c, R_r and R_t.

With the aid of a standard 12-segment switch unit it would be possible to configure it as a 3-pole-4-way unit so as to be able to switch between the various battery types/ratings for each charger unit and a double switch unit for a pair of chargers?

Even better would be a 4-pole-4-way switch so that the respective socket could be switched as well so ensuring that the switch position matched the cell(s) being charged.

Philip T. Bellamy
Briitten
Switzerland

Rod replies

Some consideration was given to producing a pcb, but it was shelved as it was thought not to be an economic proposition at that time.

However, it is likely that I will be producing another article on recharging silver-zinc button cells using the same techniques, and I *will* be designing a pcb for this. If this article is published, the pcb will be designed to accommodate alkaline cell charging too.

It is possible to switch between cell types using a multi-pole mechanical switch, but the contacts needs to be good quality. I tried this with wafer switches, but the reliability of the contacts fell off rapidly with use so I did not consider it to be very safe. As you may have noticed from the article, the accent was on safety.

If this type of switching were to be attempted, I would suggest a switch with gold-plated contacts and a good wiping action, totally enclosed to exclude dust. I have yet to discover a reasonably priced switch of this specification.

You would also have to ensure that a high charge was never administered to a small cell by wrong switch selection. This is easy to avoid in practice – use individual cell holders for each cell type rather than the expendable 'universal' cell holders you often see in commercial battery chargers.

As for the 'special' resistors, no special values are needed. Need I explain the trick?

Thank you for your interest.

Noise in emc

Ann Baker's letter in the January issue, and also a recent editorial on the subject of emc prompts me to write to you.

My credentials include a major paper entitled 'Crosstalk (Noise) In Digital Systems' in IEEE Trans Comp. EC-16, Dec. 1967 and a major paper in IEE/IEPE Proceedings (IERE Pub. No 60) of the September 1984 International Conference on emc entitled 'The Fundamentals of Electromagnetic Energy Transfer'. In addition, I was employed as the 'emc expert' on sundry major British weapons projects including Stingray.

Working in the USA in 1965 on military equipment design, I knew that the emc requirements for military weapons were fraudulent – a slush-fund where US Navy admirals and the like presumably owned most of the stock in the accredited test houses.

Don White later became king of the pseudo-technical aspect of emc, and had his 'academy' suitably situated near Washington DC. Any 'engineer' appointed 'EMC Expert' on a mirror British weapons project would pay £400 for the Don White 'Bible' and copy slabs of it into his reports – duly to join the torrent of paper going into the MOD.

An attendant at a Don White seminar told me that Don was anxious to get hold of my books.

The key point is that the technical level of competence of the emc sub-culture is the lowest of any branch of electrical and electronic engineering. There will be no correlation between whether two modules have passed the emc tests, and whether they subsequently interfere with each other. This is because the theoretical basis for these tests is hopelessly flawed.

When I heard, some fifteen years ago, that Britain would join up with Europe on emc standards, I telephoned the government department responsible and could find no evidence of any technical element in our government's decision making on this matter. The whole thing was just a paper empire.

Europe EMC standards, which follow the US Don White style, will only result in products made in Europe costing some 10-20% more, and it will also drive out small companies and stop the manufacture of customised products. Companies in the one-off business are closing down.

In 1956, when I graduated in engineering, I took up electrical engineering, not mechanical. This is because civil and mechanical engineering had nonsensical, unscientific standards and regulations. These nonsenses – such as the fact that the architecture of a Roman Point high-rise was not culpable when it collapsed because he kept within regulations – has now taken over electronics.

The decadence of the IEE has meant that we could not work through it to protect our industry against the phoney emc nonsense.

After 30 years of emc talk and enforcement, you cannot work your lap-top in an aeroplane, and a car's brakes come on when you operate your mobile telephone. You cannot listen to your fm radio while your computer mouse is on.

At the same time, I cannot get any theatre in which to debate the theoretical base of these problems with the emc punkah-wallahs.

Ivor Catt
St Albans
Hertfordshire

Q & A

Q This aspect of digital signal processing which puzzles me. Consider a system sampling at 40.000kHz whose input is a sinewave just below the Nyquist frequency at 19.999kHz. Initially, assume that the sampling points coincide with the peaks of the input signal; the output from the a-to-d converter will alternate between maximum and minimum values.

A quarter of a second later, the system will have taken 10 000 samples but the input signal has only completed 4,999.75 cycles. The sampling points thus now coincide with the zero-crossings of the input signal, and the output from the converter will be zero.

After another quarter-second, the system has taken 20 000 samples and the input signal has completed 9 999.5 cycles. The sampling points once again coincide with the peaks of the input signal, only the other way up; the output from the converter alternates between minimum and maximum values.

After three-quarters of a second, the converter output will be zero again; and after a whole second, we are back to the starting point with maximum and minimum converter output.

When this signal is replayed, the output from the digital-to-analogue converter will not be a signal at 19.999kHz. It will be at 20.000kHz squarewave, whose sharp corners will be knocked off by the analogue filter following the d-to-a converter to give a 20kHz sinewave, whose amplitude falls to zero twice per second and whose phase reverses at the points of zero amplitude. This would be heard, by those golden-eared individuals whose hearing extends up that far, as a high-pitched whine going 'wuhwuhwuhwuh'.

More generally you can see that, for an input signal close to the Nyquist frequency, the output signal will be a sinewave at the Nyquist frequency, whose amplitude falls to zero at a rate of twice the difference between Nyquist and signal frequencies. So when attempting to record a 20kHz signal on a cd, you would expect an output from the cd player of a 22.05kHz whine – inaudible – and 4.1kHz whine – very audible.

Now I have heard many criticisms made of cd players, some more rational than others, but a tendency to transform the high frequency content of the input signal into strange whines and buzzes further down the scale was never one of them.

I have occasionally come across writings which purport to knock down this paradox. They generally start out by describing such a signal as I have done – perform some mathematical juggling which leaves me floundering – and triumphantly announce that in some manner averaged over many cycles we do get a 19.999kHz sinewave after all, But surely any filter able to perform such magic would have to delay the signal for many cycles to 'see what's going on'?

In the example above, unless the filter had a time delay of over a second, you would surely still get a high-pitched whine going 'wuhwuhwuhwuh'

Could someone please explain in as non-mathematical a way as possible why we apparently, do not?

Chris Bulman
Bedford

Crossover distortion and Miller capacitance

I read Prof. Cherry's letter with great interest concerning the position of the stabilisation capacitor in an audio amplifier[1]. I fully agree that the position recommended by Self[2] is inappropriate for reducing crossover distortion. I am sure, however, that most of your readers will realise that the position Self suggests, **Fig. 1**, is incapable of allowing crossover distortion to be fully suppressed.

Consider a simple output stage which has no bias applied as shown in **Fig. 2**. If the driver stage had to

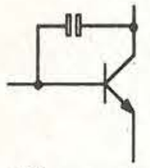


Fig. 1. Miller compensation.

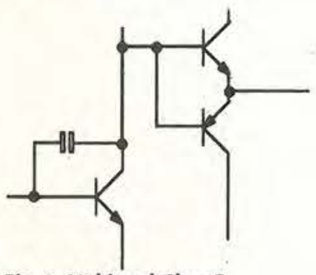


Fig. 2. Unbiased Class B amplifier.

provide a continuous signal for a given input, the coll.ector voltage will have to move in a very short time, tending towards being infinitessimal, by, about 1.2V – twice V_{be} – to ensure a continuous ouput current. The maximum speed at which a practical circuit could respond would be limited by the current which could be dumped into or extracted from the compensation capacitor, increasing the Miller effect, according to the equation,

$$\frac{dV}{dt} = \frac{i}{C}$$

Clearly, the larger the capacitor the higher this current would be. The burden for the charging current falls on the transistors ahead of the voltage amplifier stage. This causes

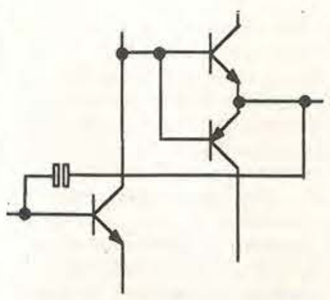


Fig. 3. Cherry's recommendation.

the concerns about the input stage overhead during fast signal peaks. In turn this endorses Prof. Cherry's view that fairly large resistors are needed in series with the input stage emitters to avoid possible transient intermodulation distortion. For example, a 1mA current source could only charge a 220pF capacitor at 4.5MV/s. Although high, this is comparable with a 60kHz signal at 20V amplitude.

In a circit which has a bias, normally intended to avoid the 'dead-band', some considerable reduction in crossover distortion is achieved, according to Self's results, although he admitted problems with residual distortion, which as I have indicated cannot be cured by his capacitor position.

Prof. Cherry's alternative position for the feedback stabilisation capacitor is to connect it to the amplifier output stage, **Fig. 3**, I have seen this once before, in a publication by RCA[3].

When I built a similar circuit, I did not experience any instabilities, but for other reasons the circuit was not particularly suited to high fidelity applications. Recently, I have built a fully complementary circuit using higher frequency transistors and I did observe some oscillation. However, I did not use damper capacitors across the drivers. as Cherry recommends, so there is a possibility that this will cure the problem.

My continuing concern with the capacitor is that even eliminating the effect it has on the driver transistor, it will still load the input stages. This of course is why Lohstroh and Otala developed an amplifier which explicitly did not use capacitors to roll the gain off in the Miller position[4].

Despite some criticism by Cherry, the Lohstroh and Otala approach has not to my knowledge been shown to be incorrect in any way as a methodology.

What Cherry and Self suggest is that lower distortions are possible using the conventional feedback method; except for the concerns that I have discussed in this letter.

I accept that transient intermodulation ought to be fully suppressed provided Cherry's observation – that the input resistors need to be high enough to prevent the input stages from ever cutting off – is met. This means that they have to drop a voltage at least as large as the input peak voltage. They will seriously reduce the open-loop gain if the input is to be as high as 1V rms – another point Cherry has made.

Another point to note is that often, designers involved with high-frequency amplifiers do not use Miller capacitors because of the degree to which they slow the design.

Many designers trying to minimise Miller capacitor effects use local feedback and low or medium impedance load resistors to define the gain of the voltage amplifier stage more precisely. Therefore, I do not accept that Self's amplifier can be given the label 'blameless' while the compensation capacitor could overload the input stages.

Perhaps if Self were to agree to increase his input resistors, and consider moving the compensation capacitor to the position Cherry advocates, Self's design would get rid of residual distortions he has been trying to eliminate recently.

If Self were to reoptimise his design, I suspect that there would be absolutely no grounds for any distortion components arising anywhere. It would then be very interesting to compare a Cherry nested differentiating feedback loop design[5] with a re-optimised Self amplifier and a Lohstroh and Otala.

I doubt that a critical listener could discern the differences. All three aim to avoid all known distortions; but could one really tell the difference between 0.03% and .003%? The Lohstroh and Otala design does peak at 0.1% at 20kHz, but then the nested differential feedback loop approach does not minimise hf distortion.

Finally, I still urge amplifier designers to listen to the type of music that their designs are intended to amplify. Try isolating one instrument from another while listening. Can you pick out a harpsichord cleanly in a Brandenburg? Or follow the 'cellos in a Beethoven symphony?

While this introduces a degree of subjectivism into the debate I will state for the record that poor designs absolutely fail this musical test – even though single instruments may sound reasonable. Good designs will not, whether or not the listener perceives any other subjective quality. Designs which pass this test include Bailey's 30W amplifier (from 1968) and Lohstroh and Otala's.

I would be happy to see these statements quantified by perhaps analysing the output from speakers digitally, using a microphone to record the output, and comparing the results to the original. This ought to be one way in which the amplifier designers could finally lay to rest this subjectivism – as well as being able to evaluate the effects of speaker performance on the overall result.

■

J.N. Ellis
Tavistock
Devon

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Towards a better World

In the July issue, some published circuits will disappoint without modifications.

In **Frank Ogden's fluxmeter**, on page 584, R_{scale} should be an unequivocal 10kΩ. This is because the meter and bridge rectifier are fed from a constant-current source in the feedback path and their resistance is immaterial.

However, this gives arithmetic-mean flux and for peak flux – assuming sinusoidal waveform – the resistor should be 6.32kΩ.

On page 586, **Phil Regalia's audio preamplifier** will spring to life once the lines crossing just left of the output terminal are joined together. And on page 590, **Sujit Liddle's phase -splitter** is useless until the right end of R_5, joined to IC_{2b}, is also taken to 0V; but to get the intended gain of 10.1 from IC_{2b}, R_3 needs a further 1kΩ in series, since the formula for its negative gain is different from that for IC_{2a}'s positive gain, being compared with $(1+R_6/R_5)$.

B Jones
London

Frank Ogden replies

Mr Jone's is quiet correct. The meter resistance is immaterial. This was a moment's mental aberration, doubtless caused by the excessive magnetic flux from the computer monitor on the word processor.

A difference resolved

In reply to Steve Winder's letter in the August issue, this is essentially a matter of definition; can you define a meaningful input impedance relative to ground when a floating input voltage source has no connection to that ground?

The input impedances for a floating voltage source are the same in the sense that a given voltage must cause the same current in each input. The voltages on them are far from equal, being indeed zero for the cold input, as described in detail in my article.

On mature reflection, it seems more sensible to say that a floating source can only have one meaningful input impedance, in this case between the hot and cold inputs, and we agree that this is 20kΩ.

Douglas Self
London

In defence of anarchy

Referring to the August issue Letters, I won't go into Jim Adam's reasoning because he shows so little of it. Besides, I am no expert. Having just discovered the magazine, I've not even read John Watkinson's loudspeaker article (June issue) so I cannot comment on it.

But when Jim came on like a tendentious boff he annoyed me. Jim Adams vilifies trial and error. No scientist would scorn observation and experiment. Even the best design involves trial, compromise and error. Lack of reference and design formulae in an article isn't bad practice. It is daring, maybe arrogant, but good, showing healthy scepticism of received wisdom, and real wisdom in not pasting human ideas over nature – as if that alone matters,

As to active versus passive, public address firms use active filters to avoid inaccuracy, power and phase problems – and expense. As amplifier costs fall faster than those for large, exacting, passive components, the music industry is buying more active monitors; and where they go, the home listeners often follow. Variable crossovers in amplifiers will preserve their choice.

As a final note, good listening, being subjective, is more about removing vice than adding virtue. Position speakers correctly, drive them moderately, and test them with simple waveforms, noise, and good well-known recordings. That's as objective as a varied world will allow.

Mr Crow
Bristol

Music speaker

John Watkinson's article 'Loudspeakers exposed' in the June issue makes much of the bad pulse response of standard loudspeakers. But this is surely not the point, given that the ear uses a series of tranducers responding to pure frequencies?

As regards polar diagrams, the JBL speakers with 'baby's-bottom' tweeter horns have admirable polar diagrams and frequency response. In addition, they are completely transparent for rock enthusiasts in relation to live concerts originally heard through JBL speakers. You could say these are integral parts of the electric guitars used, otherwise inaudible, just as huge speaker cabinets and Leslies had to be considered integral parts of electronic cinema organs.

Bernard Jones
London

The future of hi-fi loudspeakers?

It is entirely possible that the nature of hi-fi loudspeakers will change significantly over the coming years. I am not referring to the development of flat drive units, or any other new technology, so much as the likelihood of a large proportion of them becoming active units with active crossovers.

The market for these is currently approximately nil – despite the technical advantages of such an approach. There are strong reasons, however, for thinking this will change, and soon.

The impact of multimedia

Although the market for bi-amped and tri-amped speakers is tiny, the market for active speakers is now huge, as a result of the recent explosion in the cd rom and multimedia pc market. Multimedia loudspeakers are now sold by the million – and they are all active. They have to be, because computers do not include power amplifiers.

The multimedia sound source is the compact disc, which for domestic listening is well nigh perfect. To make the very best of it, extremely high quality loudspeakers are needed. Of course, one possibility is to plug the computer's audio line output into your stereo system, but this is not always convenient. There is undoubtedly a market for top quality speakers that can be driven directly from computers without the intervening stereo system,

The advent ofcd rom means that music cds will increasingly contain other data as well, a expanded 'sleeve notes'. They will also contain pictures of the artists concerned, and no doubt sales material, such as details of other titles on offer, interactive software to allow you to browse it, and something to click on if you want to place an order over the Internet. These developments seem natural and inevitable.

This means that cd-rom, multimedia, and stereo systems are moving closer together. Quality loudspeaker manufacturers will see that to gain as large a market share as possible they will need to offer products compatible with computer line outputs and existing stereo systems as well.

Computer manufacturers are not going to incorporate hefty high-quality power amplifiers into their products. But active loudspeakers can easily enough be made to work with existing stereos by bypassing the existing power amplifier – or at a pinch using the power amplifier output as the input to the active speakers.

Nasty this may be – but it works. You will no doubt be able to go down to your local hi-fi shop and buy a suitable adapter containing no more electronics than a couple of resistors to pad the level down a bit.

The upshot of all this is that there will be continued market pressure to make loudspeakers active, where in the past they have been passive.

Active versus passive crossovers

Having decided to relocate the power amplifiers into the loudspeaker boxes, the choice as to whether the crossover should be active or passive will be revisited.

A marketplace for active crossovers was always unlikely to arise when the amplifiers and loudspeakers were in separate boxes. All kinds of compatibility issues arise. In practice the most likely to emerge would have been to use a two-way active crossover between the bass and mid units, and a passive one between mid and top unit, if there is one. This has the advantage of avoiding a third amplifier section which would not always be required.

The lower frequency crossover is the more problematic one of the two and three-way passive networks, requiring bulkier and costlier reactive components which tend not to be very linear. Higher crossover frequencies are easier to implement. There are therefore also technical reasons why the bi-amp solution suggested above might have been appropriate.

But there still remain the issues of being able to adjust the crossover frequency, the balance between the low and high frequency outputs, and what happens if you want to switch between loudspeakers. By putting the power amplifiers in the loudspeaker these problems are removed.

Other problems disappear, too. The impedance of the loudspeaker drive units is no longer an external issue. The power amplifier comes with the right voltage and current capacity for the speakers. The top units will be harder to burn out because they will be driven by a small amplifier. Clipping in the bass amplifier will not harm them.

Active crossovers and multiple power amplifiers become a much more attractive proposition when the amplifiers and drive units are in the same box. This means that the arrival of cd rom is giving the market a strong push in a direction which favours active crossovers. That this will lead to higher sound quality is a welcome bonus.

There are no doubt some hi-fi obsessives who will disagree on this, but their impact on the market will be insignificant. Large numbers of bi-amped loudspeakers will be sold with monolithic power amplifiers like the *TDA2030* inside them. They will sound better than the current passive loudspeakers and passive crossover networks driven by separate power amplifiers with monolithic amplifiers like the *TDA2030* inside them.

Digital signal processing

DSP chips are already available which can do the job of an active crossover, and better, too. One thing dsp chips are able to do easily, unlike analogue filters, is introduce longer delays into the high frequency components of the signals, as well as the low frequency ones. This gives far more flexibility where design of the phase response is concerned. Another thing dsp chips are better at is precision and stability, whereas analogue filters are to varying degrees sensitive to component tolerances.

The main issue, however, is cost. The input signal needs to be converted to digital first, processed, and then the results converted back to two or three output channels to be amplified. Ultimately this is no barrier to their use, since the cost of the silicon required to do this is falling continually.

Once the digital threshold has been crossed, all sorts of new possibilities open up. A dsp-based loudspeaker can be made to produce higher quality output by compensating for some of the drive unit deficiencies. In theory, it is possible to completely compensate for the bumps in their frequency and phase responses. In practice of course it isn't anywhere near that easy.

This does not mean it is not worth doing. Such compensation can operate on three levels. Level one is simply to filter the-signal to compensate for the 'average' drive unit. Differences between units will mean the results are imperfect, but the improvement will be worth having.

Level two is to compensate for the failings of the units on an individual basis, The response of the loudspeaker is measured after it has been assembled, and the filter characteristics are adjusted for that particular loudspeaker. This would of course be automated. The results might be held in eeprom.

The third level is to adjust the filter characteristics to get the best from the loudspeaker in its environment. Clearly this is only possible at the top end of the market, where the customer has his new speakers installed for him by an engineer with the necessary equipment to set the system up. This is no wild speculation – equalisation of monitor loudspeakers in their environment has been common practice in the professional recording studio market for years.

DSP chips can do far more than just implement filter algorithms. Remote control is possible. The output signals to the power amplifiers can be limited to prevent burn-out or excessive mechanical stress in the drive units. The bass drive can be limited to prevent excessive excursion of the voice coil. The heatsink temperature can be monitored and the drive levels abated if it becomes too hot.

Once the low cost dsp capacity becomes available, as it inevitably will, these extra frills will be cheap to implement. These loudspeakers will tolerate gross abuse in a way you cannot possibly say of passive ones.

It is not difficult to predict that there will come a time when a dsp chip specifically targeted at active loudspeaker manufacturers will become available – including one a-to-d converter for the input – three d-to-a converters for the drive unit outputs, and on-board eeprom to hold the filter coefficients.

One of the three converters would be redundant in bi-amped speakers. Fallouts with one of the converters out of specification or not working properly would do for these.

Another likely development is a similar chip with pulse-width modulated outputs instead of d-to-a convteres. The PWM outputs would drive switch-mode Class-D amplifiers directly. They would be very easy to implement on silicon.

Chips like this can be expected to be targeted at the lower end of the market. Use of d-to-a converters and linear amplifiers is still likely to remain the norm where the highest quality is required.

Design freedoms

The establishment of a market in active loudspeakers will give designers more freedom. They will no longer be tied to making drive units with impedanees designed to match somebody else's drive capability, or to be compatible with another unit connected to the same passive crossover network.

One obvious possibility is to design a loudspeaker so that the bass unit is driven by a bridge mode amplifier, whereas the mid and top drivers are single ended, since the power output required is lower. These would all run off the same power supply.

The top unit might be made with a higher impedance, so its amplifier operates with a sensible output voltage swing, and is more efficient.

All kinds of other possibilities can be imagined. The connections to the drive units are all separately available to the drive circuits giving more design freedom than is available today.

Incorporating more exotic drive units becomes easier. If an unusually low or high impedance needs to be driven, the traditional constraint of having to couple it to a traditional amplifier with the wrong drive capability is removed.

Beyond the horizon

In the long run, it seems almost inevitable that digital interfaces between hi-fi units will supplant the current analogue ones. Such interfaces already exist in the professional audio equipment market. When this happens the input to the loudspeaker will be a digital one, and the a-to-d converters input to the dsp chip crossover will be eliminated.

Playing digital recordings will then involve an entirely digital path from the a-to-d converter on the microphone input at the recording studio – or wherever – to the dsp crossover outputs in the loudspeakers.

These things, however, are not imminent – or at least, not as far as I am, aware. Fairly soon, though, we can expect to see more and more active loudspeakers with active crossovers. And we can expect to see more and more dsp crossovers, starting at the top end of the market and working downwards. I, for one, think this will be a welcome development. The days of the separate power amplifier and passive crossover are deservedly numbered.

Alan Robinson
Holgate York ■

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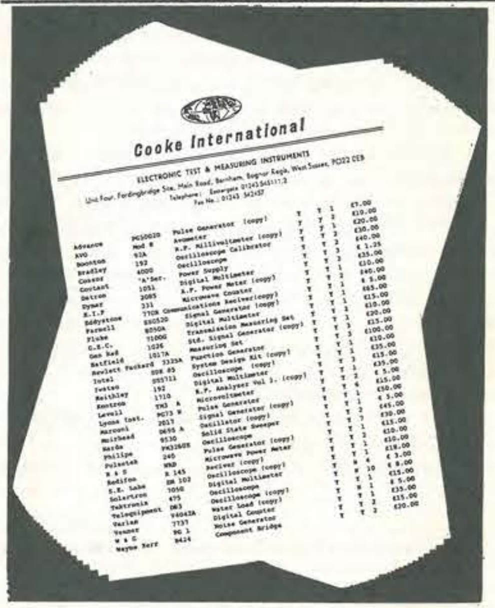

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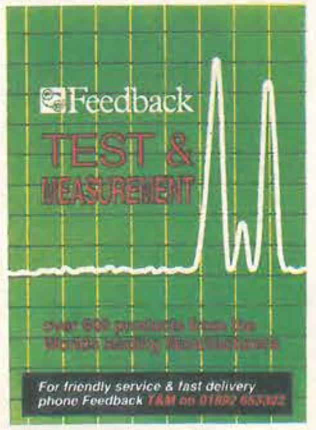

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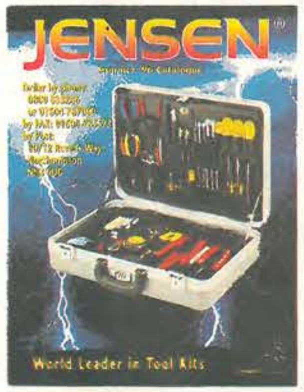

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Atmel Part Code	89C51	89C52	89C55	89S8252	89S53	89C1051	89C2051
Flash Code ROM (bytes)	4K	8K	20K	8K	12K	1K	2K
RAM (bytes)	128	256	256	256	256	64	128
EEPROM	-	-	-	2K	-	-	-
In-system re-programmable	-	-	-	YES	YES	-	-
I/O Pins	32	32	32	32	32	15	15
16-bit Timer/Counters	2	3	3	3	3	1	2
Watchdog timer	-	-	-	YES	YES	-	-
Interrupt sources	6	8	8	9	9	3	6
Serial UART (full duplex)	YES	YES	YES	YES	YES	-	YES
SPI Interface	-	-	-	YES	YES	-	-
Analogue comparator	-	-	-	-	-	YES	YES
Data pointers	1	1	1	2	2	1	1
Package Pins (DIL)	40	40	40	40	40	20	20

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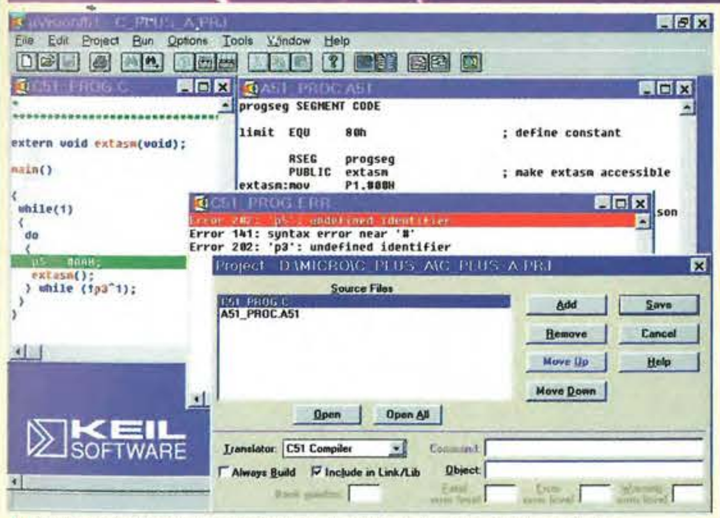

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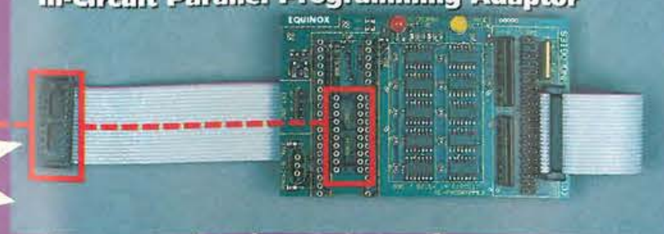

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