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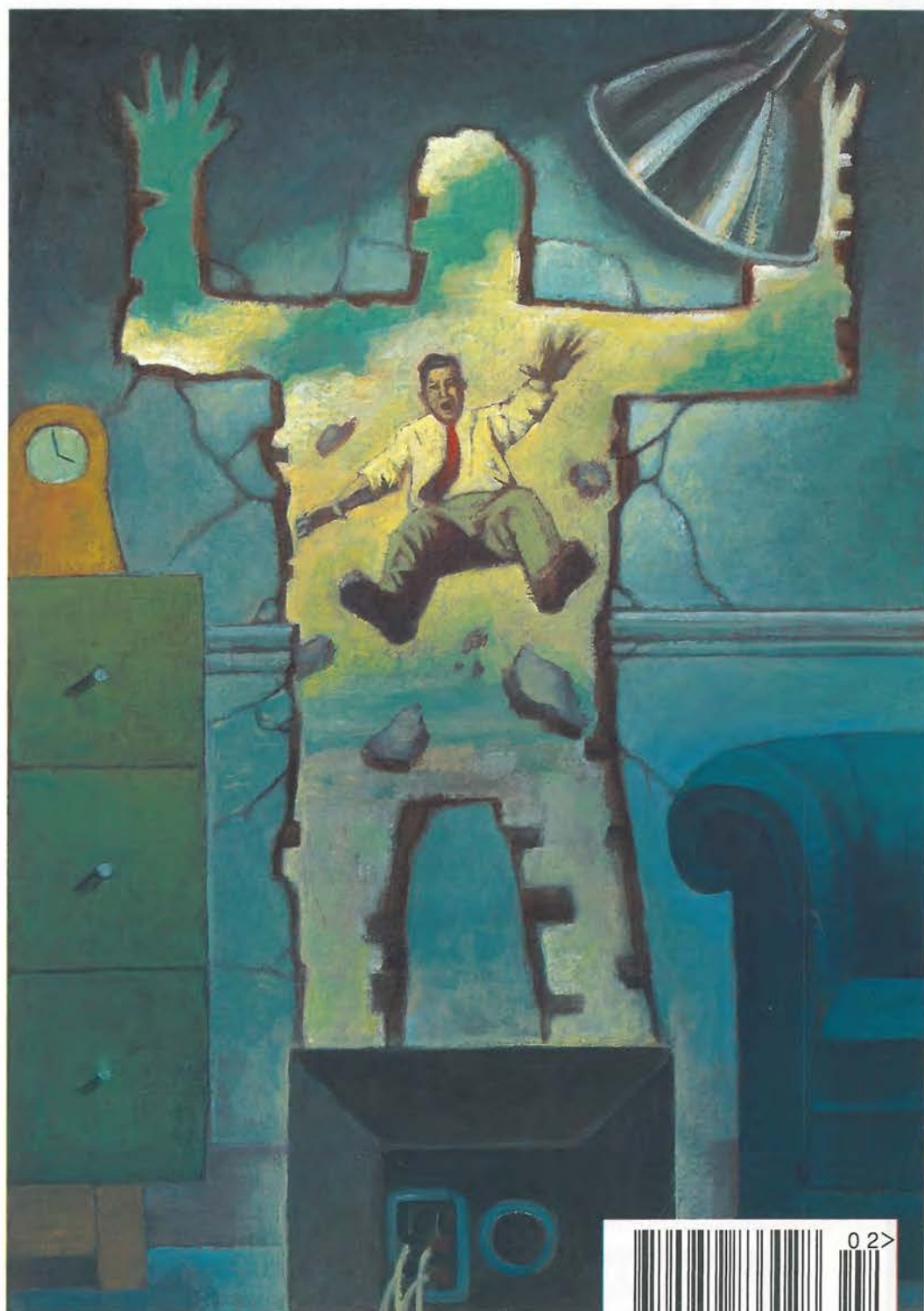
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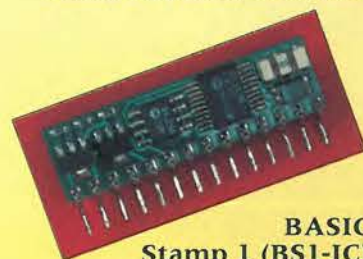


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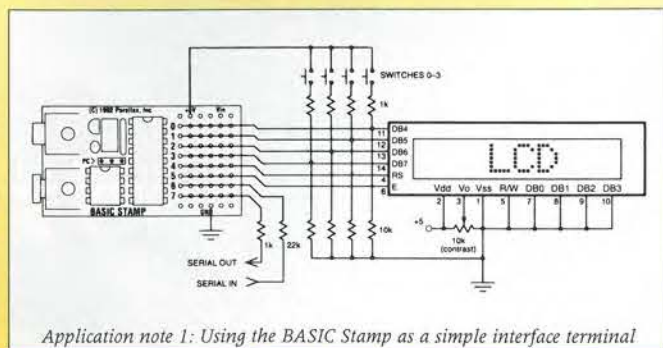
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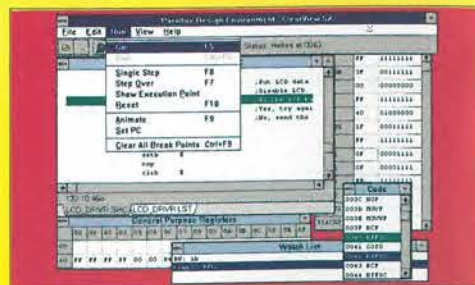


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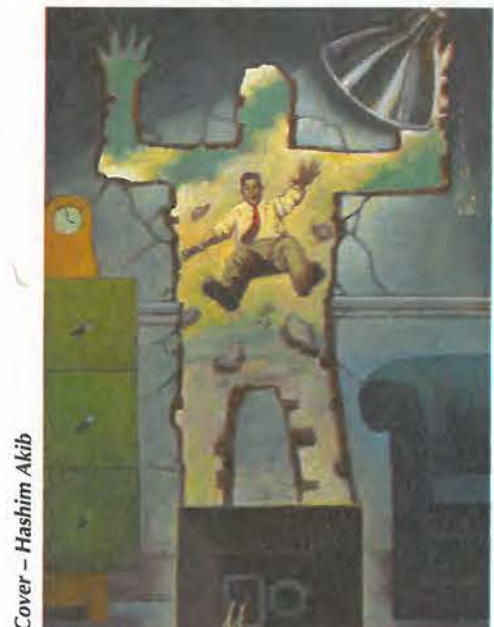


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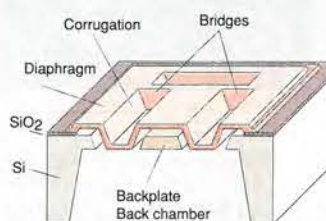
Analogue recording and playback systems suffer from speed fluctuations that usually worsen with age. **Christopher Kuni's** meter quantifies the problem.



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The Art Of In-Circuit Emulation

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REED BUSINESS INFORMATION

Europhile or Europhobe?

Europhile or Europhobe? It's the political question of the day. But there's one Euro-issue on which the electronics industry should be agreed - the value of pan-European research projects.

Look at the chip business. "In the early eighties everyone was saying that the European chip industry was dead", remembers Pasquale Pistorio president of SGS-Thomson Microelectronics. Then came collaboration.

First the Siemens-Philips 'Megaproject' to get Europe up to speed on first memory technology, then JESSI added a pan-European dimension to the German-Dutch formula, now the EU is backing the JESSI successor programme MEDEA which is open to all nationalities.

"Fifteen years ago people were saying 'microelectronics is not a European kind of thing - we'll buy that in from the Americans and the Japanese'", says Horst Nasko, the chairman of JESSI and MEDEA.

Now, in the myriad disciplines which constitute a microelectronics infrastructure - from materials through to production equipment - Europe has world-class performers.

And no one could deny that our three largest microelectronics companies - Philips, Siemens and SGS - are world-class in both technology and market clout. Philips top \$4bn in chip sales while Siemens and SGS have chip revenues of over \$3bn.

In a decade and a half, these companies have been transformed into serious players with which major worldwide companies want to collaborate, as witness the IBM/Toshiba/Motorola/Siemens alliance on memory technology.

So collaboration works - as the Japanese showed in the seventies, the US Sematech programme showed in the eighties, and the Europeans showed in the nineties.

Not that we had much to do with it. Shamefully the UK's contribution to JESSI was about a tenth of the contribution of Germany or France, about a quarter that of Holland, and about the same as that of Portugal.

Now, with MEDEA coming up for funding, it looks as if the same thing is going to happen again. The provisional commitment from governments is: Germany 32 per cent; France 29 per cent; Holland 19 per cent; Italy 10 per cent; Belgium 4 per cent and the rest of Europe 6 per cent. To our shame we are in the ROE group.

Moreover the UK DTI - an organisation supposed to be helping UK industry - has decided it will only pay 25 per cent of a project's funding whereas every other country is paying 38 per cent.

As if to rub in its contempt, the DTI has also put a ceiling of £250,000 on the UK government's contribution to any one project. This effectively debars British companies from involvement in projects costing over £1m - a pitiful sum in chip research terms. Under MEDEA accounting procedures, £1m buys an eight man project lasting one year. No big deal.

So there we have it - an attitude from the UK government that is mean-spirited, petty and chauvinistic. "The attitude is more in line with people who think Europe is something the other side of the Channel", acidly comments Dr Jurgen Knorr, group president of Siemens Semiconductor and chairman of the semiconductor committee of the European Electronics Components Manufacturing Association (EECA).

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...look at Eurofighter - a project costing \$40bn that has no apparent purpose...

Our government is something else. When it takes a shine to a scheme it will shovel out the money like a drunken sailor. Look at the grants being paid to Hyundai of Korea to set up a microelectronics factory in Fife - said to amount to several hundred million pounds. Lucky Goldstar and Siemens have also been given £100m+ financial inducements to set up microelectronics companies in the UK. Or look at Eurofighter - a project costing \$40bn that has no apparent purpose at all or at the really stupid 'Millennium Dome'.

Oh yes! We can dish out the lolly all right - no shortage at all when it comes to a pet government project - but when it comes to something not close to the government's heart it can be horribly mean.

Nowadays the cost of research is so great that not even regions, let alone countries, can afford to do it all alone. As Klaus Rupf from Germany's Ministry of Education, Science, Research and Technology told JESSI's last meeting, "If we really want to be world competitive we have to include partners from countries in other regions of the world."

The DTI has to wince up to the world. Otherwise the UK will be heading back to the days when we painted our backsides blue while continental Europe heads for a high-tech future.

David Manners

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Fears surround UK digital tv

Britain 18 SET to become a backwater in digital television thanks to the regulations announced by the Department of Trade and Industry late last year, writes Svetlana Josifovska.

According to the regulations, which although preliminary are close to being finalised, broadcasters wanting access to various digital technologies will need to ask the owners/controllers of the conditional access (C.A.) systems (needed for subscription) for a license. This, supposedly, will be granted "on a fair and non-discriminatory" basis, with Oftel acting as the watchdog.

Broadcasters such as the BBC, Channel 4 and others, fear that the regulations are not tough enough to guarantee open access and common set-top box standards. The broadcasters state that any digital tv operator should provide access to other service providers' technologies to ensure audiences have a choice. They also argue that CA-system owners should be made to license the technology to ensure common specifications.

The regulations threaten to delay the launch of digital satel-

Britain cool on Euro R&D

Fears for the UK's high technology future have escalated after the British displayed a minimal commitment to collaborative European R&D at last week's JESSI Day, which kicked off a four year microelectronics research programme.

"The UK needs to get its act together - we're so late getting involved," John Brothers, technical director of GEC-Plessey Semiconductors, told EW. "If we want to get a share of the action we've got to get involved in things that much earlier."

The programme, called MEDEA (Microelectronics Development for European Applications), proposes to send 2bn ECUs (\$2.5bn) over the next four years to maintain the renaissance in Europe's high-tech fortunes following the previous Megaproject JESSI European collaborative research programmes.

Brothers said that GPS would like to get involved in projects such as a SMIF (mini-environment) project. GPS is a leader in having a fully integrated mini-environment wafer fab.

lite television, originally scheduled for late 1997. Broadcasters, such as BSkyB, were waiting for the regulations before ordering the set-top boxes.

"At first glance, these regulations would seem to be stricter than all other countries with digital television," stated BSkyB.

"It is clearly in the interest of government and BSkyB that a workable solution is found to ensure that a digital launch is achieved during 1997," said Barry Rubbery, CEO at set-top box maker Pace.

David Manners
Electronics Weekly

DECT on one chip

VLSI Technology has released a single-chip baseband design for DECT, the micro-cellular in-building cordless phone system.

Called Vega, the chip has a dedicated DECT (digital enhanced cordless telecommunications) processor, a general purpose cpu and interfaces including those for microphone, speaker and keyboard. The only things not present are an lcd controller and the processor's rom.

Patrick Edmond, a spokesman for VLSI, said: "Once a firmware design is stable, we can add the customer's rom into a custom Vega chip."

Vega is designed for use in handsets and base stations. VLSI claims that the DECT processor is comprehensive enough to leave the cpu, which is an ARM Thumb, with nothing to do once a call has been established. Edmond said: "In a handset, the cpu can be shut down to conserve power, or it can be used for performance enhancement like echo cancellation in base stations."

Making a chip that serves base stations and handsets could leave room for a competitor to undercut VLSI in handsets which do not require the full power of the Thumb.

Edmond said: "The Thumb core takes only the same die area as a conventional eight-bit cpu core, and it is becoming a standard in handset applications, so there isn't really any scope to make a lower-cost handset chip. There is no such thing as a half Thumb."

Sample silicon for Vega (VWS23101) is available and production volumes are planned for the second quarter of 1997.

Steve Bush
Electronics Weekly



Raphael, a tool from Technology Modelling Associates, analyses the effects of parasitic interconnect in IC designs. When used with TMA Visual, the design can see the simulation results, in this case the electrical potential distribution in a static ram when the bit line is set to a higher voltage.

HART

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ASM 100 Module, complete with IEC mains lead, instructions and ASM - W20 cabinet drawings. Pt. No. V7000. £185.29
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Kit No. LK5963 Per Pair. £424.93

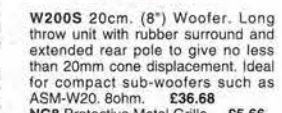
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Effect 80, Rear Speaker Kits, per pair. £39.00

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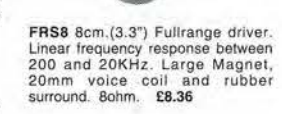
BG30NG 30cm. (12") Woofer. High efficiency, (95db) for sealed or vented cabinets. Peak power handling 250watts. 8 ohm. £69.61



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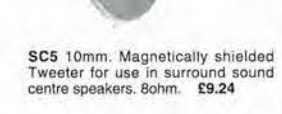
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Taiwan d-ram is not to blame

The great tidal wave of Taiwanese d-ram, blamed for this year's catastrophic drop in prices, is a myth. The Taiwanese are at least a year away from significant volume in d-ram manufacturing.

The three new d-ram manufacturing entrants in Taiwan are Nan-Ya Technology, Vanguard Semiconductor and PowerChip Semiconductor. The previously existing d-ram manufacturers are TI-Acer and Mosel-Vitellic.

TI-Acer is a captive supplier to TI and Acer. Mosel-Vitellic has always been a speciality memory house though it is about to enter the commodity business via a joint venture with Siemens called ProMOS Technologies.

"We'll start with the 16Mbit next year," said Mosel-Vitellic's vice-president for operations, Dr Nasa Tsai. "The first phase of the fab has been equipped for 20,000 wafer starts a month."

Most advanced of the new entrants is Vanguard. "Now we have Fab 1(a)



Mosel-Vitellic's Nasa Tsai: "16Mbit next year"

with 15,000 wafers a month and we're currently installing capacity in Fab 1(b) for another 10,000 wafers a month," said Dr F C Tseng, president of Vanguard.

Nan-Ya Technology, a subsidiary of Taiwan's largest company, Formosa Plastics, is running 16Mbit licensed from Oki Electric. "This month we will run 5000 wafers," said executive vice-president Charles Kao. "Our plan is to add another 22,000 wafers a month by the end of 1997."

The third new entrant, PowerChip Semiconductor, is a joint venture between Mitsubishi and the Umax-Elite group. "Currently, we're running 8000 wafers a month," said marketing manager K Y Tsai. "The plan is to be running 12,000 a month by March, representing three million pieces, and 15,000 a month by June."

That is the combined level of output from Taiwan this year, less than the monthly output of a Japanese or Korean fab - hardly a reason for the 1996 price collapse. "The last one in is always the scapegoat," was the wry comment from one of the island's executives.

David Manners
Electronics Weekly

US catching up with smartcard applications

Seven top US financial services companies have endorsed the Mondex smartcard technology in what is the USA's largest vote of confidence in the sector. So far the USA has lagged behind Europe in smartcard applications.

AT&T, Chase Manhattan, Dean Witter Discover, First Chicago NBD, MasterCard, Michigan National Bank and Wells-Fargo Bank have invested in Mondex USA Services, which will use the Mondex smartcard technology in a series of pilot programmes in the USA.

"The power of this group will propel Mondex as the pre-eminent electronic cash payment system in the USA," said Janet Hartung Crane, president and CEO of Mondex USA. In spite of its heavyweight backing, Mondex will still face stiff competition from Visa and American Express, which are pursuing their own smartcard projects.

The first commercial Mondex cards will be introduced in 1998, following the results of key trials. Mondex USA says it will license its technologies to other US companies to help further establish the technology.

In a separate move, Mondex announced its agreement with Sun Microsystems for the inclusion of its format in the Java Commerce Toolkit. This toolkit will allow the development of open, secure and integrated electronic commerce applications which will link Mondex to the Internet.

Computer learns user habits

Australian firm Formulab Neuronetics has launched a computer in the USA which it claims learns from its user and makes decisions.

The device, called the Richter Paradigm Computer, uses a parallel-processing architecture comprising 896 simple Risc processors, and costs \$3000. The company says that the low cost of the system will help establish a large market for a "reasoning" computer.

In a demonstration, Formulab said that it can run neural network applications 180 times faster than an Intel Pentium 166MHz system.

Formulab also revealed plans for a supercomputer based on its computer architecture, which has taken more than 14 years to develop. The supercomputer would combine as many as 6000 microprocessors and could be used for scientific applications. It would run a special operating system that could manage the difficult task of splitting a computational problem into separate tasks and assembling the results.

The company also said that it is working on an add-on card for pcs which could assist users by learning from their work habits. It also plans to shrink its technology so that it can be embedded into products such as cameras and consumer electronics devices. Formulab said it will license the technology to other companies.

Other applications include stock buying, with the system noticing differences in stock prices and trading patterns.

Tom Foremski
Electronics Weekly

Rockwell wins 56kbit modem support

The battle to establish the dominant 56kbit/s modem technology continues apace. Rockwell Semiconductor now says that it has won the support of Compaq, Hewlett-Packard, Toshiba, and AST Computer, while rival US Robotics has added Hitachi to its list of allies.

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MPEG-4: draft agreed

The main components of MPEG-4, the emerging audio visual coding standard, have achieved working draft status following a recent meeting in Brazil. The meeting also saw the Moving Pictures Expert Group kick off work on a follow-on standard, curiously called MPEG-7.

MPEG-4 aims to provide a universal mechanism for communicating audio and visual "object" data. Unlike the MPEG-1 and -2 standards, which encode

frames of video, MPEG-4 reflects the emergence of myriad media content, and is capable of manipulating such content whatever its shape.

"MPEG-4 offers lots of flexibility. It will give applications developers far greater scope," said Paul Fellows, a project manager at SGS-Thomson Microelectronics involved in MPEG-4 work. "The basic techniques have been designed - experiments have

identified the best techniques and these have been incorporated into the drafts."

MPEG-7, the follow-on standard, will add value to MPEG-4. According to Fellows, MPEG-7 will be concerned with "bits about the bits". It will offer ways to describe content whatever its guise, allowing identification in much the same way that search engines identify text on the World Wide Web.

40Gbit/s amp/filter breakthrough

A compact 35GHz amplifier suited to long-distance, high-speed optical communications has been co-developed by a group of UK and Portuguese academics. The device, which uniquely combines digital filtering and analogue amplification, is claimed to be the fastest of its type, capable of working with 40Gbit/s communication systems.

Dr Izzat Darwazeh of UMIST, and one of the academics involved in the design, said: "In today's fibre systems for long links, people use optical amplifiers which introduce some noise. Even 10ps of jitter could seriously degrade the system."

One solution is to use signal shaping filters at the receiver, traditionally an external filter after the front end. However, the filter itself introduces noise, and the physical link between the components on the board has inductance, further degrading the signal.

The circuit design, involving UMIST,

University College London and the University of Aveiro in Portugal, uses a distributed amplifier. "A distributed amplifier is functionally equivalent to a finite impulse response filter," said Dr Darwazeh. By forcing the gate and drain delays to be different, the amplifier gains the filter characteristics.

The circuit, implemented as a single monolithic microwave IC (MMIC), has been built by the Fraunhofer Institute in Germany using a 40GHz HEMT process. Different lengths of the transistors define the delays and coefficients of the filter.

Dr Darwazeh says that an integrated front end would have applications in very high speed optical fibre such as transatlantic or Pacific links.

The next stage is to make an adaptive system that adjusts the filter's coefficients and the amplifier's gain depending on the signal received.

Richard Ball, Electronics Weekly

Film promises 70% brighter liquid crystals

A British company is developing and marketing a plastic film which promises to increase the brightness of laptop lcds by 70% for no extra power.

The film, invented by Philips in Eindhoven and now being worked on by Merck, of Poole, Dorset, acts to pre-polarise light entering the rear polariser of a standard lcd.

The rear polariser of an lcd transmits 40% of light from the backlight and absorbs the rest. The new film alters the random polarisation of the light from the backlight so that most of it is accepted by the rear polariser.

Two new players in digital video disk arena



Hitachi and Akai are the latest companies to announce that they will sell DVD rom and DVD players. Hitachi will introduce its DVD rom player in early 1997, and has already begun shipping samples to key customers. However, a DVD player for the consumer market will not come until later in the year.

The entry of DVD has hit some snags in recent months with companies such as Toshiba and Fujitsu delaying the introduction of their products into overseas markets. A key reason has been the lack of DVD titles, since without content the DVD systems only have cd-rom based titles to play.

US media company Time-Warner has said that it will begin shipping DVD titles, which should help jump-start the market. Time-Warner said it will release four movie titles: *The Assassin*, *Blade Runner*, *Eraser* and *The Fugitive*.

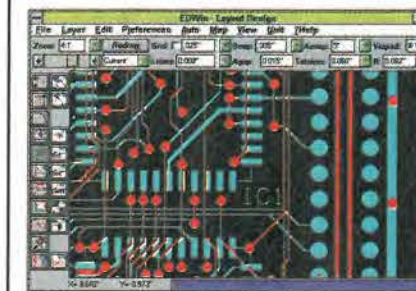
Toshiba, which was the first to introduce DVD players into the Japanese market, said that the players are selling well and that it has shipped about 30,000 units. Toshiba will introduce its DVD players into the USA in early 1997, followed by a European launch.

Philips set the pace in DVD roms. Now other companies are targeting the market.

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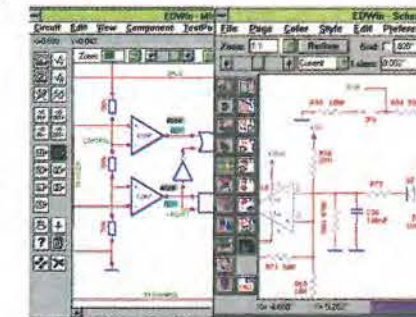
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RESEARCH NOTES

Jonathan Campbell

Single chip mic sounds the best yet

The thin, flexible diaphragm and rigid bookplate design of the condenser microphone can already be implemented on a single chip. But there have been problems in the past with residual stress in the diaphragms affecting the sensitivity of the device.

Now three Chinese researchers have manufactured a silicon condenser microphone with a a corrugated

diaphragm that shows a dramatic reduction of stress.

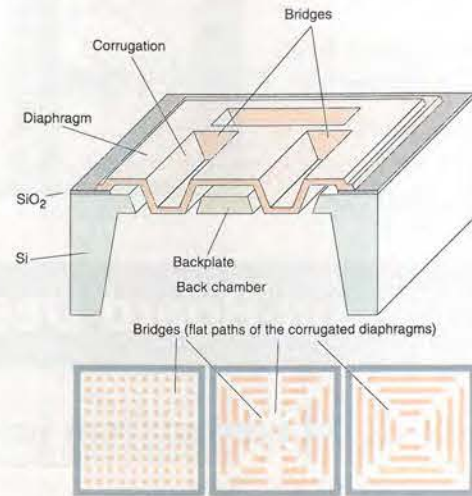
This is not the first microphone to use a corrugated diaphragm to reduce stress in this way. But the Chinese microphone has also demonstrated a flat frequency response and high sensitivity ('Design and fabrication of silicon condenser microphone using corrugated diaphragm technique,' Quanbo Zou *et al*, *J of Microelectromechanical Systems*, Vol 5, No 3, pp. 197-204).

The microphone capacitor consists of the corrugated diaphragm that acts as an active electrode, and a single crystal silicon bookplate with acoustic holes that acts as the stationary electrode. It is fabricated, using seven masks, on a single wafer by use of silicon anisotropic etching and sacrificial layer etching techniques. So no bonding techniques are required.

Up to now, the microphone has demonstrated a flat frequency response between 100Hz and 8-16kHz, and open circuit sensitivities as high as 14mV/Pa while using a low bias voltage of 10V

Further work will aim to improve the overall microphone performance, boosting sensitivity and flatness of the frequency response, by optimising the structure parameters and process conditions.

Contact Quanbo Zou, Institute of Microelectronics, Tsinghua University, Beijing, China.



Corrugated diaphragm results enhances single-chip microphone. Diagram shows three alternative bridge configurations.

Sun shines again for Pathfinder: The record-setting Pathfinder solar-powered research aircraft has resumed flight testing at Nasa's Dryden Flight Research Center, following its disastrous damage during a ground

accident 12 months ago. In its last flight, the Pathfinder set an altitude record of more than 16,800m on a flight from Dryden which lasted nearly 12h. However, its latest flight was just a low-altitude check-out flight over

the northern portion of Rogers Dry Lake at Edwards Air Force Base, California.

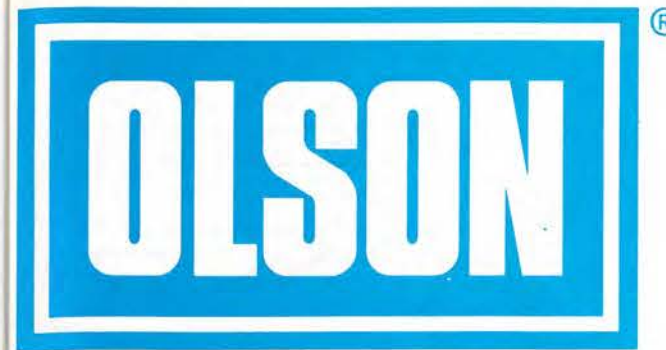
Low sun angle and limited hours of sunlight during the winter limit Pathfinder's altitude capability to about 6670m. This is one of the reasons why the project will be transferred to the Navy's Pacific Missile Range Facility on the island of Kauai, which lies at a lower latitude. Kauai's latitude and more favourable prevailing northerly winds will allow more opportunity for high-altitude solar-powered flying during a five-month flight test program.

Pathfinder is one of several remotely-piloted aircraft being evaluated under Nasa's Environmental Research Aircraft and Sensor Technology (Erast) program. The joint Nasa-industry alliance is seeking to develop technologies required to operate slow-flying unpiloted

aircraft at altitudes up to 34,000m on environmental-sampling missions lasting up to a week or longer. With a span of 33m, Pathfinder is basically a flying wing.

Only two small pods extend below the wing's centre section to carry a variety of scientific sensors and support the craft's landing gear. The solar arrays on the wing can provide as much as 7200 watts of power at high noon on a summer day to power the craft's six electric motors and electronic systems. A backup battery provides power for up to two hours to fly the craft after the sun is down. Built primarily of lightweight composite structure, plastic foam and a thin plastic covering, Pathfinder weighs about 230kg.

Contact: Fred Brown, Dryden Flight Research Center, Edwards, CA, USA



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Robot has an office at its heart

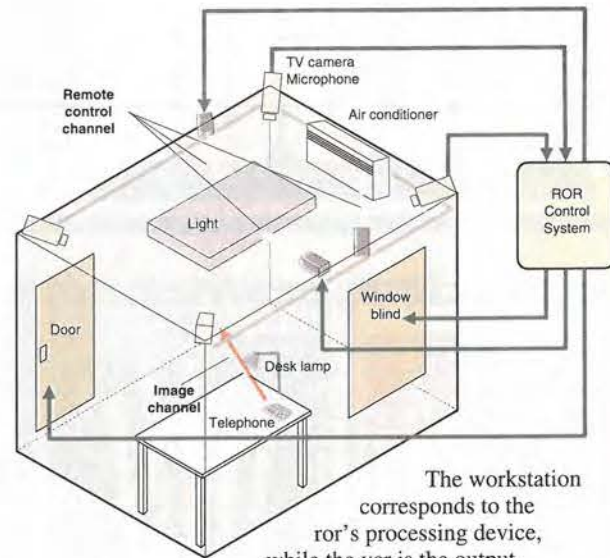
When we dream of using robots to carry out helpful tasks, we generally think about cyber-butlers precisely mixing our gin and tonics and waking us up with a newly ironed newspaper, rather than giving assistance at the office. (Though in some areas of the *EW* office, a gin-mixing, paper-ironing robot would fit in very well.)

But concentration by Hiroshi Mizoguchi and colleagues on the office environment has now given birth to the robotic office room (ror) – a robot that supports human activity within the office.

For example, when a human worker points to an object and gestures to get it, the room understands the behaviour and takes the object to them, perhaps using a long reach manipulator centrally located on the ceiling. In effect the office worker, works inside the robot.

To collect information, the ror monitors behaviour through a tv camera. When it detects pre-defined functions of the motion of an object – moved by human behaviour – the ror starts to make a response. By choosing to monitor objects that are moved by human behaviour rather than humans themselves, computing power has been considerably reduced.

An experimental prototype consists of a telephone, a pen, a tv camera vcr, a workstation and an audio set. In this case, the telephone, the pen and the tv camera are the input devices of the ror since the system can infer human behaviour by monitoring them.



The workstation corresponds to the ror's processing device, while the vcr is the output device because human memory augmentation – through recorded images – is one of the supporting functions of the ror. This is also the point of the audio device.

So far, the Japanese ror has been able to recognise when the receiver in the office has been lifted up, and can start video recording the telephone call. The volume on the audio set is also reduced automatically when a telephone call is being made.

Contact: Hiroshi Mizoguchi, Research Center for Advanced Science and Technology, The University of Tokyo, Tokyo, 153, Japan email: hm@lssl.rcast.u-tokyo.ac.jp.

Working inside the robotic room as it collects information.

Raindrops reveal their charge

It is not often that experimental researchers of today have cause to cite 130year old lecture papers delivered to the Royal Institute as part of their introduction. But the reference to Lord Kelvin's 1860 paper on Atmospheric Electricity by José Fornés *et al* gives an indication of how long their particular area of study has been a mystery – and how little has been achieved recently to solve it.

The question Kelvin raised, and

which the Brazilian researchers have been trying to solve, is simply: "What is the pattern of charge distribution within falling or immobile rain drops?"

As the author's point out ("Evidence for multipolar charge distribution in falling water drops", *J. App. Phys.*, Vol 80, No 10, pp. 6021-6027), electrical charge determination in single water drops has been undertaken since the 1920s. But despite extensive investigation, results are scarce – partly because of the difficulties in constraining a water drop without charge exchange. A simple treatment might suggest that the result would be similar to a dielectric sphere in a uniform electric field, where a pure dipolar pattern would be observed. But this is not the case.

Fornés and his colleagues have devised an apparatus that can produce drops of near zero net charge by using an aluminium ring to enclose the drop at its formation, and force it to acquire a given charge. When the drop is freed, it retains that charge.

The measuring probe consists of a copper wire loop of 2.8mm radius

and a cross section of about 0.2mm, across which the drop falls. The ring is connected to a 10GΩ probe of a model 8900 Dagan patch-clamp amplifier, in the current-to-voltage configuration and set to voltage-clamp mode at voltage zero.

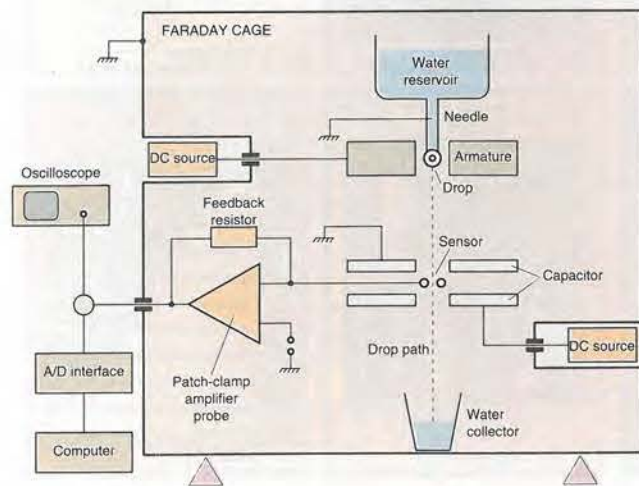
There is never any direct physical contact between the probe and the drop, the coupling being only capacitive. During measurement, the voltage output of the amplifier is acquired by an analogue-to-digital interface and fed to a computer.

To produce multipole charge induction, the drops (with their zero net charge) fall through a capacitor, typically maintained at 30V with 67mm plate separation. Each capacitor plate has a small hole to enable to the drop to pass through and the sensor probe is placed at mid distance between the plates.

Plotting the results show that the electrical charge on a water drop falling in the direction of an applied electric field is distributed in a multipolar pattern, possibly because of the contribution of the field-oriented water molecule multipoles.

Lord Kelvin would have been happy.

Measuring the charge on a raindrop – and not touching.



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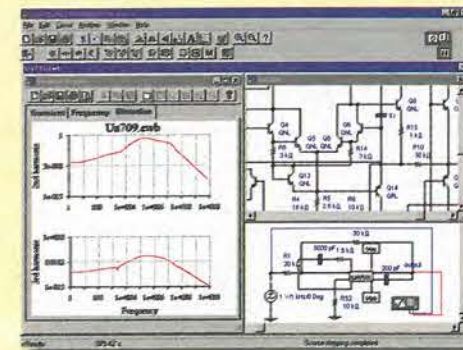
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4xxx ICs: 4000, 4001, 4002, 4008, 4011, 4012, 4013, 4015, 4023, 4025, 4028, 4030, 4040, 4049, 4066, 4068, 4069, 4070, 4071, 4073, 4074, 4085, 4086, 4107, 4019, 4024, 4027, 4028, 4030, 4040, 4047, 4075, 4077, 4078, 4093, 4101, 4106, 40103.

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CIRCLE NO. 111 ON REPLY CARD

Straightening out a pcb problem

Printed circuit boards, or pcbs, remain vulnerable to a simple, heat-induced threat: warpage. Unfortunately, warped pcb may cause a device to stop working, while boards that warp during manufacturing after expensive components are added can mean costly losses.

But a technique developed at the Georgia Institute of Technology and now licensed by Electronic Packaging Services (EPS) could provide a new weapon against warpage.

Charles Ume (right) examines the fringe pattern generated when a printed circuit board is heated in the new oven.

The experimental Thermoïre process provides real-time data about pcb warpage in simple and fast manner, so that manufacturers can avoid design problems.

"Electronic packaging companies can use the warpage information to make changes in their pcb design early," says Charles Ume, an associate professor in the School of Mechanical Engineering.

The heat that can warp pcbs is generated each time computers, camcorders or other pcb-run devices are turned on. Also, temperatures up to 230°C are an integral part of pcb processing.

In addition, if the pcb is small, thin and densely populated with components, as is the current industry trend, that is an invitation for warpage-related reliability problems.

For the new process, Ume developed a special oven with a glass grating top, through which the pcb placed inside is visible. A white light shines through the glass grating onto the pcb, and an inexpensive, compact, charge-coupled device camera captures warpage digitally as it occurs.

The flat glass grating is etched with equally spaced parallel lines. It is placed above and parallel to the pcb. A beam of white light is directed onto the glass at a specific angle, causing the etched lines to create a shadow on the surface of the pcb. When the surface of the pcb curves due to warpage, a moire pattern is produced by the geometric interference between the etched lines on the glass and the



shadow of those lines on the pcb's surface. The more the pcb warps, the greater number of moire fringes that appear.

Ume counts the number of fringes, puts them into an equation, and a computer determines how much warpage has occurred. The warpage process is displayed in real time on a television screen and recorded on video and on computer.

The Thermoïre technique can be used to simulate the three major kinds of soldering processes - infrared reflow, convective reflow and wave - and the automated oven system can reproduce any given soldering temperature history used in producing a board. In this way, the system can pinpoint which processes or designs may cause the most warping.

Ume says that companies can use the results to make design or process changes before production, such as changing soldering temperature profiles, reducing or extending processing times, relocating key components, and changing the materials used in constructing the pcb.

The ability to measure thermally induced warpage could also enable manufacturers to validate their numerical warpage predictions, created using finite element modelling techniques.

If a certain amount of warpage is allowable, the new technique also lets manufacturers measure initial warpage, rather than assuming the board is flat before transistors and other items are added.

Manufacturers can then determine how much additional warpage develops during further processing or attachment of components.

Contact: Charles Ume, Georgia Institute of Technology, Atlanta, Georgia 30332-0828, USA. email: charles.ume@me.gatech.edu

1997 will be the year of the blue laser

Operation of blue laser diode devices should now be possible, with commercialisation coming in 1997-98, according to one of the leading researchers in the field.

Shuji Nakamura, of Nichia Chemical Industrie, Tokushima, Japan, made his prediction in an interview appearing in *OE Reports*, published by Spie (The International Society for Optical Engineering).

Nakamura has already been behind some of the major breakthroughs in blue leds and lasers. He has also stood against the prevailing wisdom underlying much work in blue device research by concentrating on GaN technology, rather than using the more usual II-VI materials and then frequency doubling.

Now, when he is reported to be close to developing a commercial product, the significant degradation

suffered by II-VI blue lasers is convincing scientists that GaN is the way forward.

So far Nakamura has produced a pulsed blue laser diode operating at room temperature, with wavelength variable between 390-440nm by changing the In content of the InGaIn well layers.

Initially, the cost of the laser devices will be many times greater than the current \$1-2 price of blue leds, though volume will naturally force this down.

The attraction of blue lasers is the impact they could have on high density storage. Current limits using a red laser diode are around 5Gbytes per side. But the shorter wavelength of blue light should allow this capability to be increased substantially, perhaps by a factor of three.

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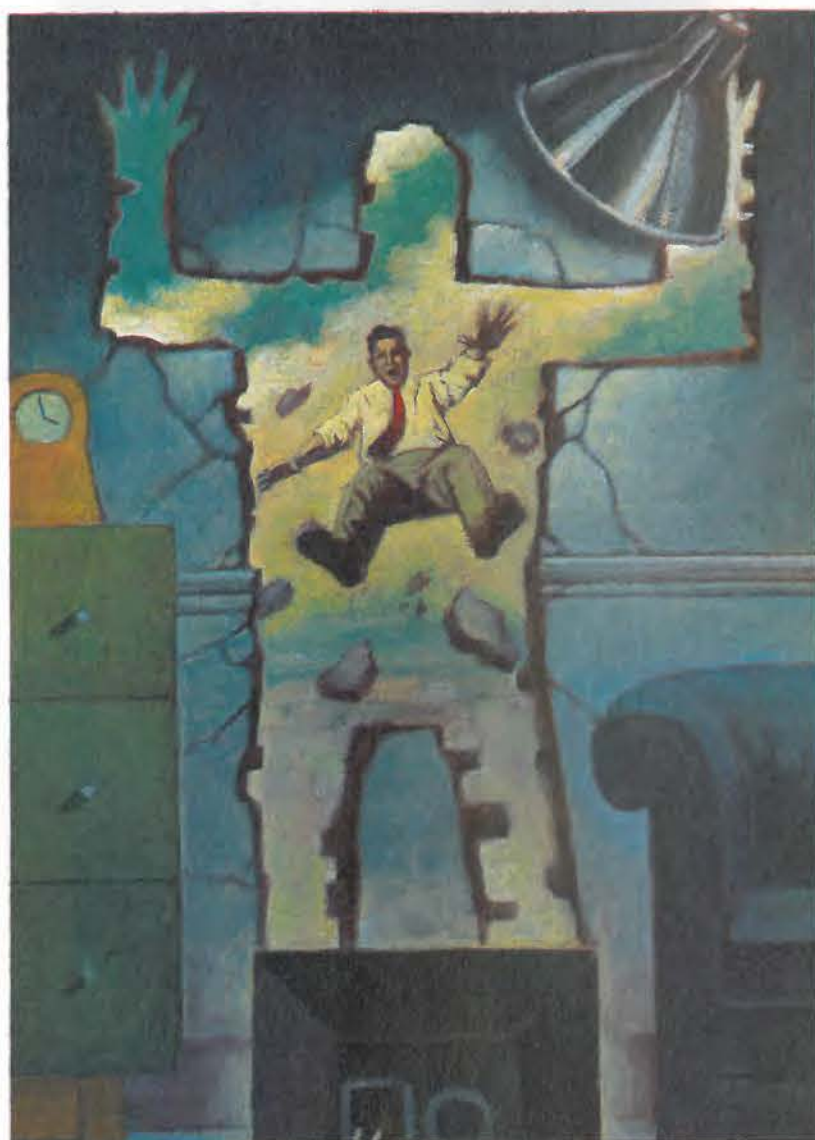
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CIRCLE NO. 158 ON REPLY CARD

Roaring subwoofer



Using motional feedback, **Russel Breden's** subwoofer produces flat response down to 15Hz – despite its relatively small enclosure. Feedback also makes feasible an infinite baffle rather than a reflex design, resulting in tighter bass.

Inspired by Peter Baxandall's 'Low-cost, high-quality loud-speaker' series of articles in *Wireless World*, I built my first sub-woofer back in 1978. It was a 2.4ft³ reflex using a KEF B139 driver and tuned to 30Hz.

Originally, this sub-woofer was passive, but it soon became apparent that adding a dedicated power amplifier and second-order low-pass filter produced the flexibility required to interface with my existing speakers. Built in the days when home computers were a distant dream, and Thiele/Small analysis was little known, it is surprising that it worked at all. As it was, it gave me a sense of what was missing from most of the other systems around at that time.

Nearly twenty years later, things haven't changed that much. Off-the-shelf speaker systems available today rarely produce an output below 60Hz.

Going lower

For a system to produce bass extending to at least 30Hz normally requires large boxes and expensive drivers. But by creative use of electronic circuitry, both size and expense can be cut to reasonable proportions. These techniques however require the sacrifice of hi fi's most sacred cow – flat amplifier response.

Today, we have all the tools necessary to design economical audio systems with a flat response, even though the system's component parts may be far from linear. The work of Thiele – extended by Small – provides comprehensive details about the response of a driver in an enclosure. All that is needed is to design electronic circuitry to compensate for the non-linear response of the speaker/driver combination.

Advantages of activity

A motional feedback system operates by sensing the speaker cone's motion and feeding this back into the power amp.

Providing negative feedback in this way forces the amplifier to produce a signal that corrects both for amplitude irregularities and the distortion generated by the speaker. The result is an acoustic output which is flat against frequency – even though both speaker and amplifier are operating in a decidedly non-linear fashion.

Motional feedback is not the only way of achieving this. You could use electrical equalisation, for example, but this would not reduce system distortion. In any event, motional feedback is an intellectually satisfying technique using well understood principles.

To produce a correcting signal, the speaker must be fitted with some form of transducer. In this design, I have used dual-coil drivers, one coil of which forms the pickup. As the cone moves, it generates a voltage signal in the coil which is proportional to cone velocity. This signal is then processed and used for correction purposes, see panel.

One major objection to motional feedback is that the feedback loop could try to force the driver beyond its limits. This

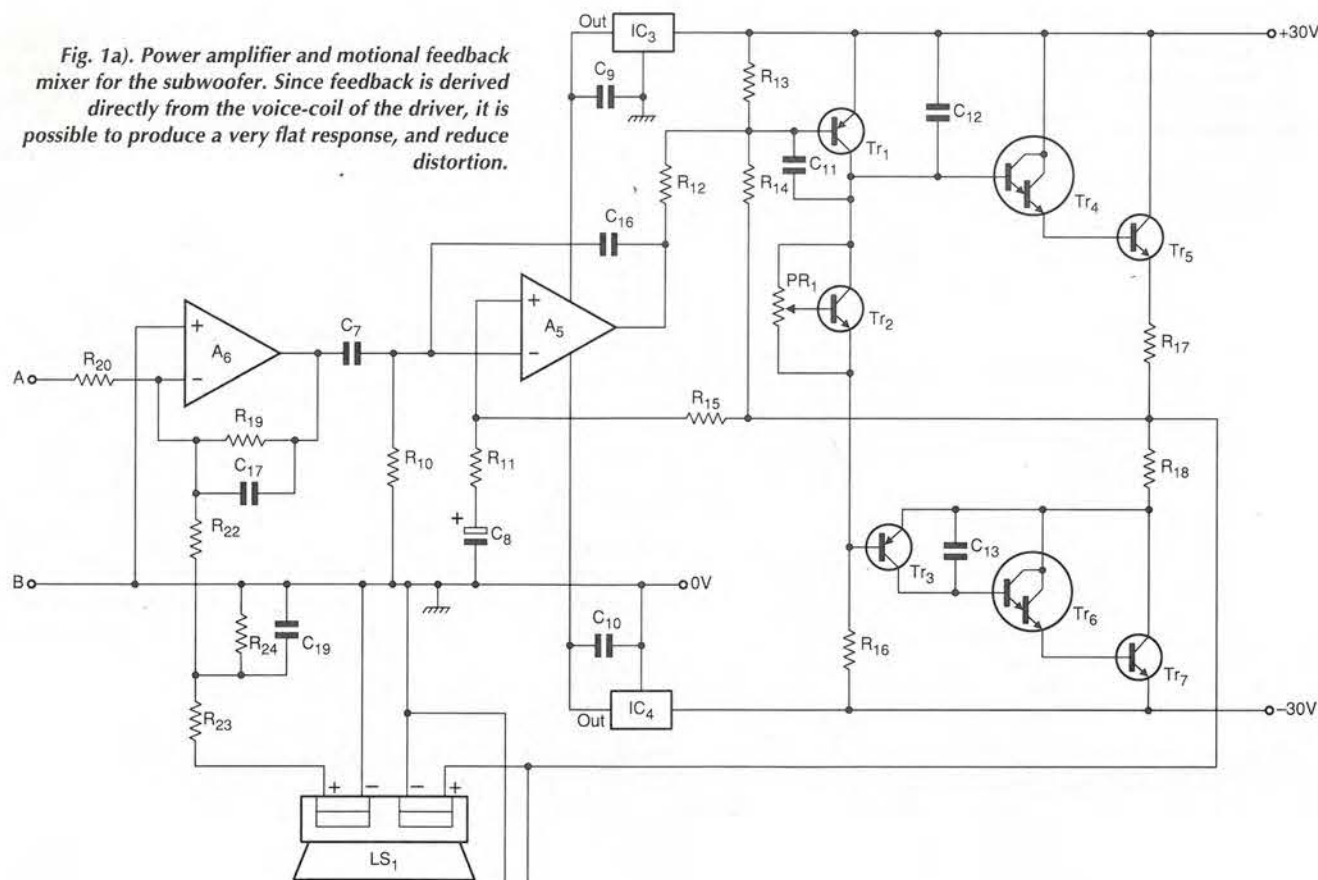


Fig. 1a). Power amplifier and motional feedback mixer for the subwoofer. Since feedback is derived directly from the voice-coil of the driver, it is possible to produce a very flat response, and reduce distortion.

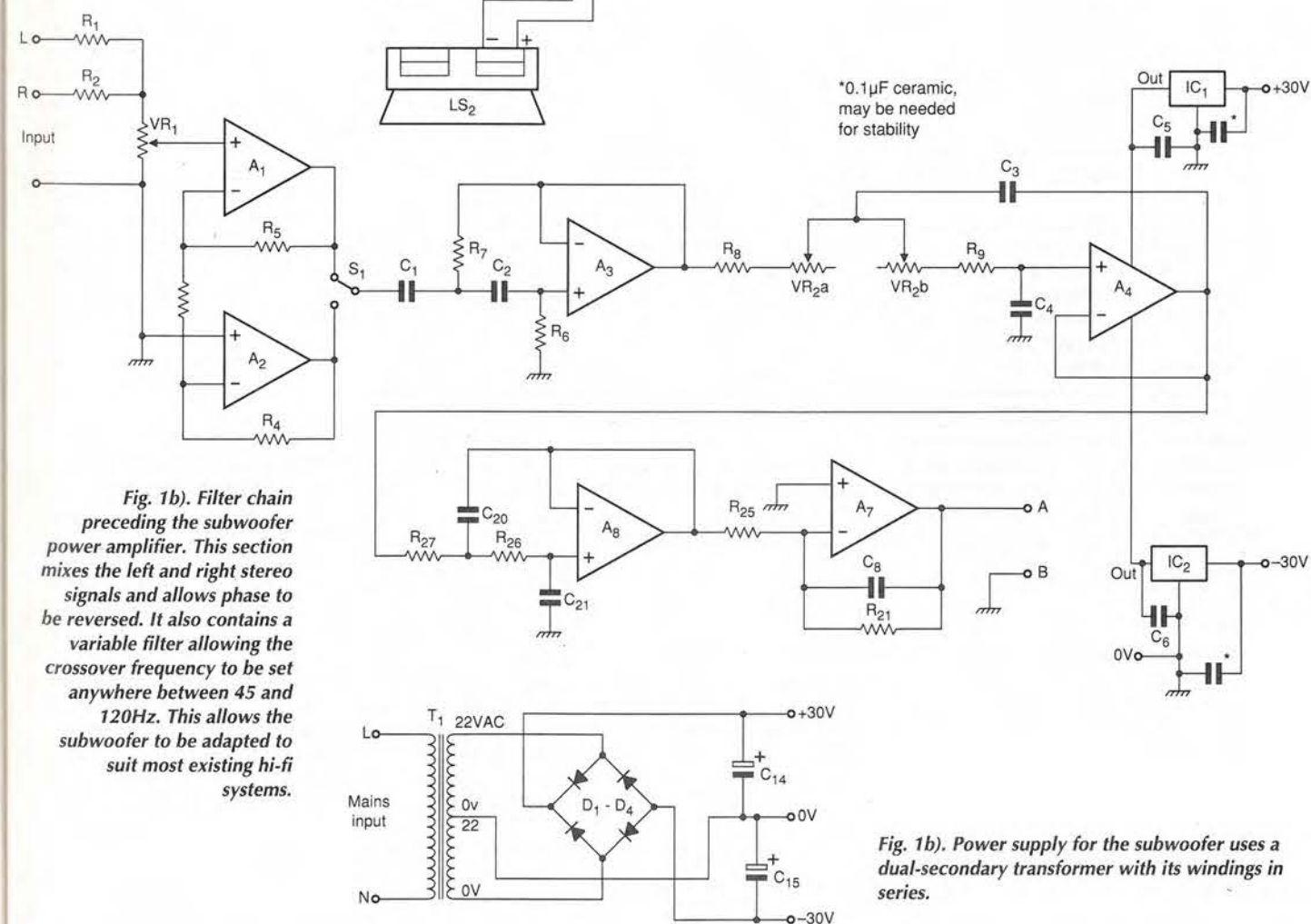
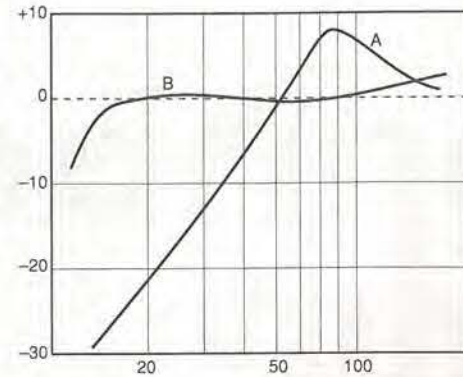


Fig. 1b). Filter chain preceding the subwoofer power amplifier. This section mixes the left and right stereo signals and allows phase to be reversed. It also contains a variable filter allowing the crossover frequency to be set anywhere between 45 and 120Hz. This allows the subwoofer to be adapted to suit most existing hi-fi systems.

Fig. 1b). Power supply for the subwoofer uses a dual-secondary transformer with its windings in series.

Fig. 2. Curve A is initial system response while curve B is final system response but excluding low-pass filtering. Both curves are measured near filed, i.e. 1cm from cone.



problem can be avoided by choosing a sealed box or infinite-baffle enclosure. Use can then be made of the natural roll-off of the driver to ensure that excursion limits cannot be exceeded (see panel).

Using an infinite-baffle enclosure means that the response is essentially that of a second-order filter. To produce a flat response down to dc – even if it were possible electrically – would require infinite cone excursion. To match such a driver, you would need an amplifier of infinite gain since the gain of a high-pass filter is zero at dc. For every octave of base extension, the power and cone excursion required increase fourfold. Obviously this cannot be carried too far. However the system described here has a flat resonance, with a -3dB point at 15Hz.

This frequency is at least an octave below the nominal cut-off of many subwoofer systems. Furthermore when I examined the output acoustically, sine wave distortion was below 2% at 15Hz.

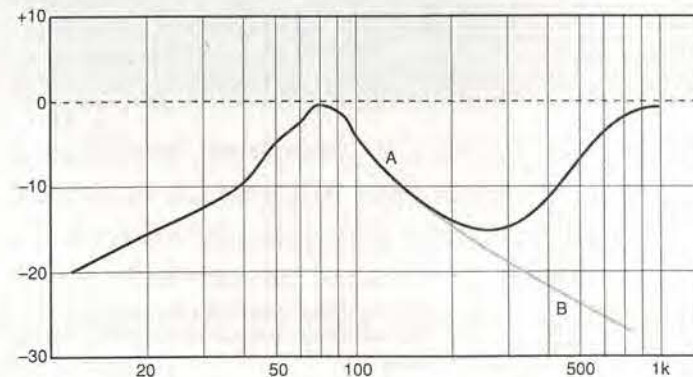
The reason for going so low is mainly to minimise phase shift. There is little musical information on most recordings below 30Hz. Too rapid a roll-off at this frequency produces phase errors between fundamentals and harmonics, leading to a muddy sound. One of the joys of listening to this system is the speed at which bass notes are delivered without the overhang associated with reflex systems.

One problem that has to be considered is how much sound-pressure level, spl, can be generated. Here, we are concerned with a domestic environment. Many more-than-adequate subwoofers use a 10in driver in a reflex cabinet.

Since reflexing is out in this design, I have chosen a pair of 10in drivers, operating in parallel. This has the advantage that both drivers contribute useful output over the entire range whereas a reflex system's vent only contributes output around its resonant frequency. Additionally, the price paid for vent output is a rapid bass roll-off that must worsen the transient response of the system.

For the purposes of analysis, the circuit can be split into three parts. First the power amplifier; the requirements from this element of the circuit include high power output. Furthermore, because the load impedance is around 3Ω, fair-

Fig. 3. Curve A is response from the pick-up coil of the loudspeaker and curve B response after high-frequency equalisation.



ly large current swings are required. An extended low frequency response also implies that the power supply will tend to sag under load. Precautions must be taken to make sure that this does not affect sound quality.

Taking into account these factors I make no apology for the use of rugged TO3 output devices, namely Tr_5 and Tr_7 , which are a pair of 2N3055s. These in turn are driven by a pair of Darlingtons, Tr_4 and Tr_6 . A quasi-complementary output stage is used with Tr_3 providing the necessary phase inversion for the lower output transistors.

Nested feedback loops

The entire circuit is based on TL074 quad op-amps, one of which is used in the power amplifier. However the low operating voltage and consequent low output voltage swings of this device provide insufficient power. For this reason the amplifier incorporates the idea of nested feedback loops, Fig. 1.

To explain further, the output stage operates from a split 60V power supply. The driver stage is built around Tr_1 . Usually, the output stage biasing voltage is provided by the V_{be} multiplier comprising Tr_2 and PR_1 . Resistors $R_{17,18}$ introduce emitter degeneration in the output stage, stabilising the operating point. Local shunt feedback around both driver and output stage is taken via $R_{12,14}$.

The value of R_{13} has been chosen to produce 0V output for a 0V input from A_5 . This local feedback loop reduces distortion from the output stage to well below 1% before global feedback is applied around the circuit.

Closed-loop gain from A_5 output to the load is approximately five. This allows the op-amp output stage to produce the required voltage swing at the output. An incidental advantage is that the op-amp output sees a relatively high impedance and therefore operates in push-pull, Class A.

Supply voltage for the op-amp is taken from the main power supply through a pair of 15V regulators, $IC_{3,4}$. Capacitors $C_{9,10}$ provide hf decoupling. Op-amp A_5 is the heart of the amplifier. Note that because of the inverting action of the driver/output stage, the inputs are used in the opposite sense.

Input signals are applied to the inverting input and overall feedback to the non inverting. The voltage gain of the amplifier is set by the ratio of R_{15} to R_{11} . Capacitor C_8 reduces the dc gain of the circuit to unity while appearing as a short circuit to ac signals. Resistor R_{10} defines the input impedance of the amplifier.

The closed-loop gain needs to be high since most of it is used to equalise the subwoofer and reduce speaker distortion. The other components to be mentioned are mainly concerned with keeping hf stability within the amp. This is the function of $C_{12,13}$ and C_{16} .

It could be argued that most of the parameters of the amplifier just described are well above those required for the circuit function. For example, the slew rate of the op-amps is thousands of times faster than required. Similarly, the distortion level of the amplifier is many times lower than that generated by the speakers. This is simply a reflection on the advancement of commonly available parts. This same level of performance allows supply line voltage rejection ratio of over 100dB, and this is of great importance, as mentioned earlier.

The second aspect of the design, whose overall performance is shown in Fig. 2, involves the manipulation of the pick-up coil voltage to produce motional feedback. Referring to Fig. 1, rather than feed the speaker coil voltage directly into the amplifier's feedback loop, it is fed via the mixer stage built around A_6 . The voice coil output feeds the network comprising $R_{22,24}$ and C_{19} . This is necessary because the voltage follows the impedance curve.

Below 200Hz the output is directly proportional to the velocity of the cone. Above this frequency output rises at

approximately 6dB/octave. To maintain the velocity curve – not to mention amplifier stability – the output must be suppressed at high frequency. Capacitor C_{19} flattens the curve to a straight line.

From here, the signal feeds the mixer amplifier A_6 , via R_{22} . This is configured as a virtual earth mixer. Feedback resistor R_{19} is shunted by C_{17} , which provides further high-frequency roll-off to the coil signal ensuring that the required response is obtained.

The net result of adding this signal to the amplifier input is that the acoustic response from about 10Hz to 150Hz rises at 6dB/octave. In other words, amplifier output voltage is proportional to cone velocity, and acts as a power differentiator.

To obtain a flat response, the amplifier output needs to become proportional to cone acceleration. Rather than differentiate the feedback signal exactly the same result is obtained by integrating the input signal. This function is carried out by A_7 , which, in conjunction with R_{21} and C_{18} forms the integrator.

At this stage we have produced a flat response speaker system – flat, at least, in the deep bass region. Plotting the response however reveals that the overall response is that of a high-Q low-pass filter. A glance at Fig. 3 reveals that further work needs to be done. The low-pass response is due to the voice coil inductance resonating with the reflected mov-

ing mass.

Rather than complicate the circuitry further, the solution used here is to tame the response by using a low-Q low-pass filter in series with the amplifier. When this has been done the final response is flat within 1.5dB from 15 to 150Hz. The mild penalty to be paid is that the response rolls off at 24dB/octave above 150Hz. Luckily, this is of little consequence in practice since this point occurs at least half an octave – and usually more than an octave – above the roll-off point required by normal speaker systems.

Filtering the input-stage

The main task of the input stage filtering is to extract the bass information from both incoming signals and present this to the power amplifier. In addition, the signal must be manipulated to allow 'seamless' integration of the subwoofer with the existing speakers.

For this design, I decided to drive the sub-woofer directly from the speaker outputs of the existing amplifier. This not only simplifies the design, it is the only rational place to take a signal feed. Once set up the sub-woofer will follow system volume adjustments. This is a particular advantage if, like me, you are always being told turn it down. You can also be assured plenty of drive signal.

Line outputs are rarely standard. The left and right signals

Infinitely baffling

In order to squeeze the maximum possible bass from an infinite-baffle enclosure, the volume has to be carefully calculated. Even motional feedback systems are not immune to the laws of physics. If the enclosure is made too large, the woofer will be driven beyond its excursion limit. If it is too small, maximum power input will not allow full excursion.

In order to calculate the required enclosure volume, Thiele-Small equations are required. When a circular piston is fed with a sine wave, it can be shown that the sound-pressure level generated at 1m into half space, A, is,

$$A(\text{dB}) = 40 \log_{10}(d) + 20 \log_{10}(app) + 40 \log_{10}(f) - 83$$

where d and app are the diameter and peak-to-peak cone excursion respectively, both expressed in mm and f is the frequency of interest.

From the term $40 \log_{10}(f)$, you will see that the available sound-pressure level falls with frequency at 12dB/octave. If the enclosure volume is chosen so that its response lies to the right of A, Fig. 4, then the driver will be protected from excessive excursions. If the response lies to the left of A then the speaker runs the risk of destruction from bass input. Ideal enclosure response coincides with A.

In order to calculate something useful it is essential to examine the efficiency of the driver and relate this to A. Maximum output that a driver can produce in the pass-band is independent of enclosure

size and can be calculated from the following equations, the first of which is for driver efficiency, η_o ,

$$\eta_o = \frac{k \cdot f_o^3 \cdot V_{as}}{Q_{es}}$$

where f_o is free-air resonant frequency, V_{as} is equivalent compliance air volume, Q_{es} is electrical Q and k is 9.64×10^{-10} when V_{as} is expressed in litres. Sound-pressure level in decibels at 1W and 1m distance into half space is,

$$112 + 10 \log_{10}(\eta_o)$$

Maximum sound-pressure level in

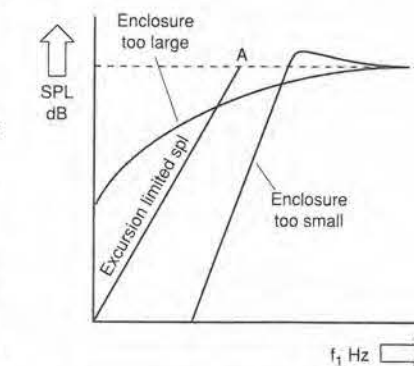


Fig. 4. Excursion sound-pressure level limit rises at 12dB/octave. Speaker responses that intersect the left of A are limited by cone excursion. Right of A are thermally limited but cannot produce maximum sound pressure level. Ideal speaker response coincides with A.

decibels at 1W and 1m distance into half space, B, is

$$112 + 10 \log_{10}(\eta_o) + 10 \log_{10}(p)$$

where p is the available amplifier output in watts continuous.

All drivers mounted in an infinite-baffle enclosure exhibit second-order high-pass filter response whose amplitude, C, is,

$$C = (\omega^4 + (d^2 - 2)\omega^2 + 1)^{-0.5}$$

where w is f_c/f and d is $1/Q_{tc}$, f_c being the resonant frequency of the driver mounted in the enclosure and Q_{tc} is the Q of the driver in the enclosure.

Unfortunately this does not help much because Q_{tc} and f_c are not known until the enclosure volume has been determined. But if you choose f at a low enough frequency, say 1Hz for convenience, then $w^4 \gg (d^2 - 2)d^2 + 1$ term. This makes it possible to simplify and rewrite the equation for an approximation of C as, w^{-2} .

To avoid the excursion limit, the 1Hz response must be $-A+B$ (in dB) down with respect to maximum pass-band sound-pressure level, B. The corresponding amplitude is $10^{(-A+B)/20}$, which is w^{-2} .

As f is 1Hz, ω must equal f_c so, $f_c = w = 10^{(-A+B)/40}$. Having obtained f_c , the enclosure volume can be simply calculated from,

$$V_b = V_{as} / ((f_c/f_o)^2 - 1)$$

Calculated volume is slightly conservative, but this is no bad thing considering the price of drivers.

Multichannel speaking monitor

Visual displays certainly have their uses but there are times when a more immediate message is needed. Heikki Kalliola's speaking monitor provides digitally-addressable and re-recordable spoken messages.

My design for a multi-channel speaking monitor is based on an Information Storage Devices speech memory chip. With no moving parts and needing few external components, the circuit is an economical means of making many kinds of announcements, including measurements and threshold warnings. One application of the speaking monitor is the reporting of exceptional conditions in vehicle environments. Using spoken messages, there is no need to divert your attention to the indicators on the instrument panel. This device can be mounted in an enclosure small enough

to hide behind the dashboard without any rebuilding. The ISD1016A* speech storage chip used can record and play back 16 seconds of voice in a number of individually-addressable segments — in this case four, of four seconds each in length. Each of the four messages can be triggered according to the situation. The chip family includes members with longer recording time, but 16 seconds is enough for the purpose described here. Inputs 1 to 4 are constantly monitored, Fig. 1. If one or more input is grounded, a message from the corresponding memory address is fed

to the loudspeaker. The message is repeated until the grounded contacts open. If all the inputs are used, there is time about four seconds for each message. If all inputs are not needed, the ones used can have more time allocated to them. If for example only line number 1 is used, the announcement can be a full 16 seconds. Recording and playing back Recording is performed by first selecting the message to be dictated on the sensor line. Selection is done with switch S₁. As drawn, S₁ points to line 1. After line selection, button S₃ is

memory location pointed to by S₁. Thus recordings are easily made. Once the messages are loaded, S₁ is turned to its base position, indicated as B on the diagram, and the device is ready for use. When one or more of the input lines is grounded, the message from that line's memory area is spoken from the loudspeaker and repeated until the line returns to its normal state.

Circuit logic

When an input line is grounded, diodes set the right address with bits A₅ and A₆. At recording and check playback this grounding is performed by S₁. The memory chip requires that during recording pins P/R and CE are grounded. Signal PD keeps power consumption extremely low when at +5V. To record and play back, this line must be grounded. The signal also acts as reset switch, in the event of a memory overflow. Resetting is carried out by pulling the line to +5V and back to ground again. When recording, S₃ pulls pins P/R and CE down to ground via diodes. At the same time transistor Tr₁ stops conducting for a moment due to grounding of the base via C₁. A positive reset-pulse is created from the collector to pin PD. Releasing of S₃ generates the end-of-message mark to the memory. To check play back, S₂ starts the message by pulling down the chip enable pin, CE, and

creating a reset pulse to PD with Tr₁. Playback starts at the address pointed to by S₁ and continues until the end-of-memory mark. Grounding of input line pulls also down the base of Tr₂. The transistor stops conducting, triggering the astable multivibrator timer chip IC₂. The chip puts out negative-going pulses equivalent to pushing the playback button S₂. If S₁ is at base position, B, these pulses pass through to playback triggering. Time interval between pulses, and therefore also the rate of message repetition, is dictated by the value of R₁₃. If, for example, only one input is used and the maximum length message recorded in one location, the interval time must be over 16 seconds to avoid rolling over. If on the other hand, all the inputs are used, the minimum interval is about five seconds, and the message length cannot exceed four seconds.

Anti down-out switch for vehicles

In vehicles, there are many possible stimuli for the monitor. Examples are oil pressure warning, over-high water temperature and low fuel indication. In many cases, easily accessible switch contacts for these are already installed in the car. When this design is used in vehicles, the usual power supply is the 12V battery, from which the voltage is dropped to 5V via regulator IC₃. In automotive applications, the car radio

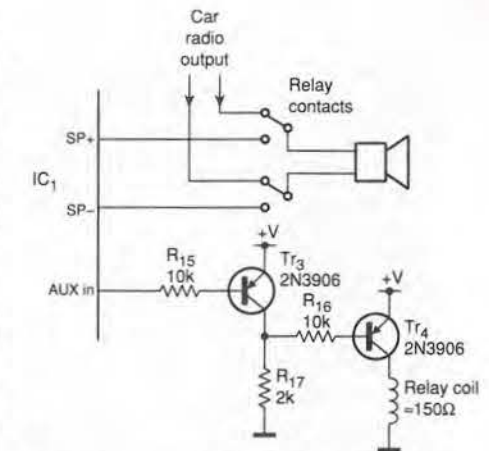


Fig. 2. In automotive applications, this switch circumvents the problem of a loud radio obscuring the announcement by automatically switching the radio's speaker over to the speaking monitor unit when a message is triggered.

might obscure messages. To avoid this, you can add the switching circuit of Fig. 2. With this modification, the radio loudspeaker replaces the monitor's loudspeaker and serves a dual purpose. Control for the output relay is taken from the speech memory's auxiliary input, which is held high during playback. Do not forget to add a 10Ω series resistor, if the loudspeaker has very low impedance. ■

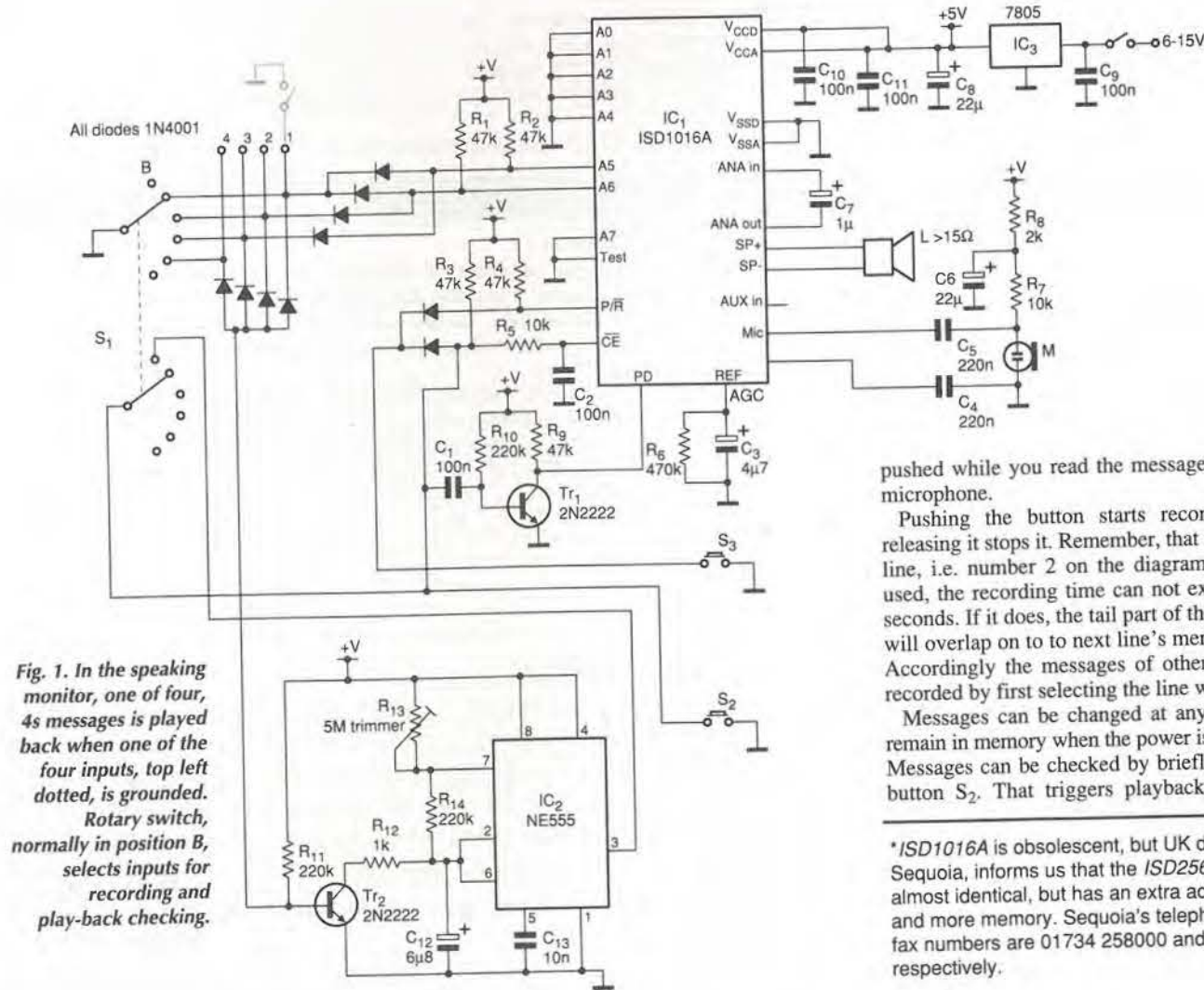


Fig. 1. In the speaking monitor, one of four, 4s messages is played back when one of the four inputs, top left dotted, is grounded. Rotary switch, normally in position B, selects inputs for recording and play-back checking.

pushed while you read the message in to the microphone. Pushing the button starts recording and releasing it stops it. Remember, that if the next line, i.e. number 2 on the diagram, is to be used, the recording time can not exceed four seconds. If it does, the tail part of the message will overlap on to next line's memory area. Accordingly the messages of other lines are recorded by first selecting the line with S₁. Messages can be changed at any time and remain in memory when the power is removed. Messages can be checked by briefly pressing button S₂. That triggers playback from the

*ISD1016A is obsolescent, but UK distributor Sequoia, informs us that the ISD2560 is almost identical, but has an extra address line and more memory. Sequoia's telephone and fax numbers are 01734 258000 and 258020 respectively.

OSCILLOSCOPE LIVE CHOP Page -0.25

MEASURE MODE
CH 1
CH 2
CHOP
ADD
COMPARE
X-Y PLOT

1.06 0.40 -0.26 -0.93 -1.60 -2.25 -2.93 -3.60 -4.26

16.5 13.8 11.2 8.53 5.8

0.30(sec) 0.30 0.90

FREQ TIME DIV TIME-MAG TIME-OUT HYSTERESIS SLOPE TRIGSOURCE PRINT COMMENT SETTINGS

SPECTRUM ANALYZER CH 1

HOLDING WINDOW
RECTANGLE
HANNING
MARKING
BLACKMAN
BARTLETT

3.70 3.20 2.70 2.20 1.70 1.20 0.70 0.20 -0.30

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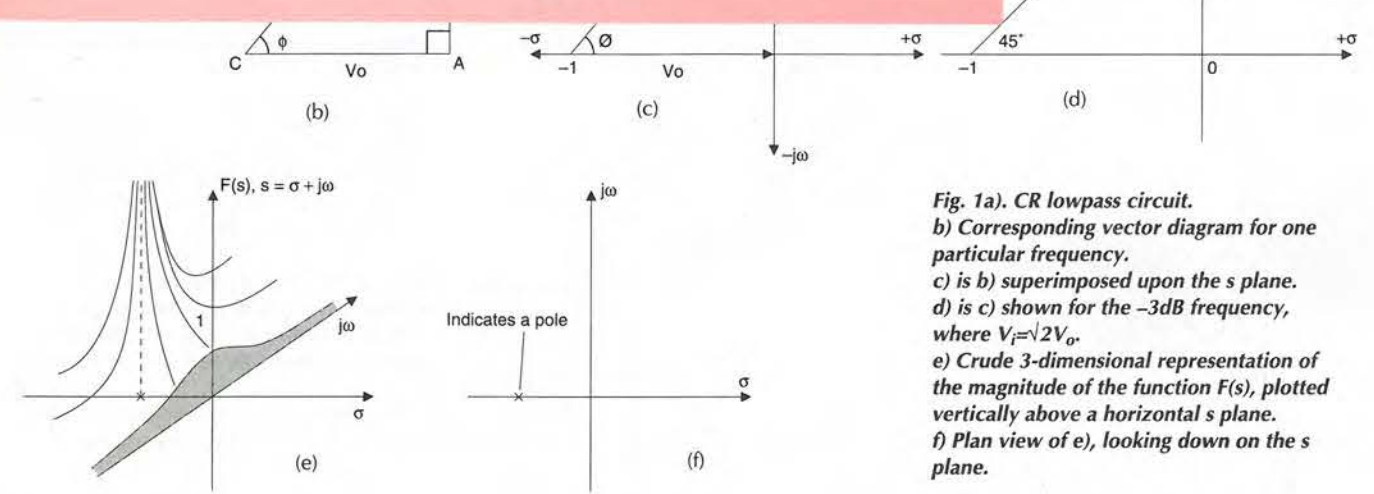


Fig. 1a). CR lowpass circuit.
b) Corresponding vector diagram for one particular frequency.
c) is b) superimposed upon the s plane.
d) is c) shown for the -3dB frequency, where $V_i = \sqrt{2}V_o$.
e) Crude 3-dimensional representation of the magnitude of the function $F(s)$, plotted vertically above a horizontal s plane.
f) Plan view of e), looking down on the s plane.

JRE

tor. Phase angle ϕ is given by:
 $\phi = \tan^{-1} I/R$
 R and I are the real and imaginary parts of the numerator. Clearly, $-\omega CR/1$ is zero at $\omega = 0$ and tends to $-\infty$ at infinite frequency, 0° , at 0Hz and tends to -90° at very high frequencies.

Two results give an important rule:
The amplitude of the response at any frequency due to a pole is inversely proportional to the length of a line from the pole corresponding point on the $j\omega$ axis. The phase is given by the angle that the line makes with the positive horizontal axis counting anticlockwise rotation from the pole as negative, indicating a lagging response.

Figure 1c) shows how the vector diagram with the pole/zero diagram. Figure 1d) is the same again but drawn for the -3dB frequency, the frequency where the reactance of capacitor equals the resistance R .

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Ian Hickman takes a further look at how circuit operation can be represented pictorially.

More in the PICTURE

An earlier article of mine¹ reviewed various ways of representing circuit action, with a view to showing how the different representations complemented each other. So the article covered vector diagrams, the circle diagram – a sort of generalised vector diagram showing what happens as the frequency varies – Bode plots – which also show what happens as the frequency varies – and pole-zero diagrams.

That was the intention, but my apologies to any readers who looked in vain for any zeros – they failed to materialise due to lack of space. This article rectifies the omission, and carries the story on another stage.

Poles

Well, just one pole to begin with, the one to be found in the lowpass CR circuit of the last article, the response of which was shown there as Fig. 6, and here as Fig. 1b). The equation giving the frequency response, as derived in the last article, is:

$$v_o/v_i = 1/(1+j\omega CR) \quad (1)$$

Note that, as last time, the base of the triangle (of length unity) is the vector v_o , then the two terms in the denominator take you from the pole at $\sigma=-1$ to the origin, and then a distance ω up the $j\omega$ axis (assuming as before that $CR=1$), where this distance represents the voltage drop across the resistor R .

Adding vectorially (the j indicating that these vectors are at right angles), this brings one to the tip of the sloping line, which represents the input voltage needed to give unity output voltage. Thus v_i/v_o at any frequency is proportional to the distance from the pole at $\sigma=-1$ to the corresponding point on the $j\omega$ axis. So the frequency response, v_o/v_i , is proportional to the reciprocal of this distance.

Equation 1 also indicates the phase, as follows. First, make the denominator real, by multiplying top and bottom by $(1-j\omega CR)$, the complex conjugate of the denominator. The equation then becomes:

$$v_o/v_i = (1-j\omega CR)/(1+\omega^2 C^2 R^2)$$

and, since the denominator is now just a number, the phase angle is given just by the

numerator. Phase angle ϕ is given by:

$$\phi = \tan^{-1} I/R$$

where R and I are the real and imaginary parts of the numerator. Clearly, $-\omega CR/1$ is zero at 0Hz and minus infinity at infinite frequency, so ϕ is 0° , at 0Hz and tends to -90° at very high frequencies.

These two results give an important rule:

The amplitude of the response at any frequency due to a pole is inversely proportional to the length of a line from the pole to the corresponding point on the $j\omega$ axis, and the phase is given by the angle that the line makes with the positive horizontal axis, counting anticlockwise rotation round the pole as negative, indicating a lagging response.

Figure 1c) shows how the vector diagram ties up with the pole-zero diagram. Figure 1d) is the same again but drawn for the -3dB frequency, the frequency where the reactance of the capacitor equals the resistance R .

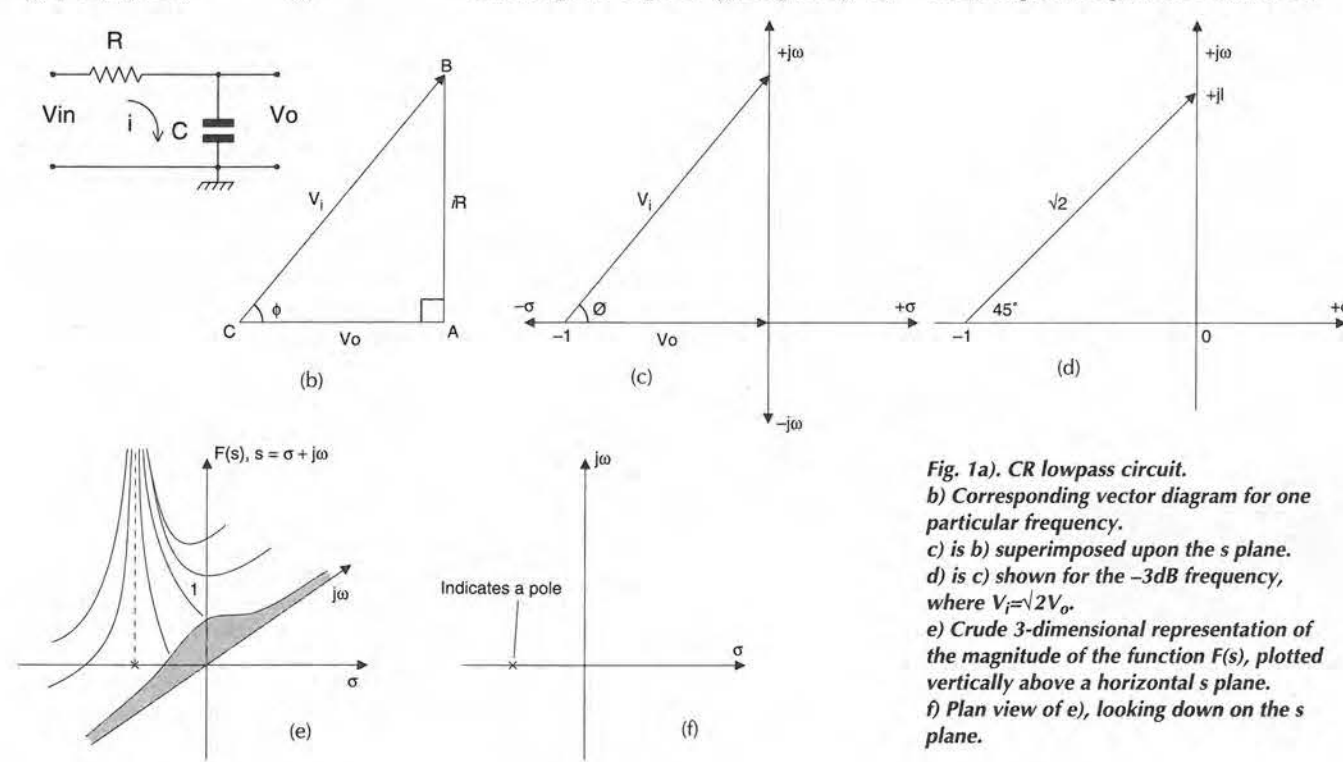


Fig. 1a). CR lowpass circuit.
 b) Corresponding vector diagram for one particular frequency.
 c) is b) superimposed upon the s plane.
 d) is c) shown for the -3dB frequency, where $V_i = \sqrt{2}V_o$.
 e) Crude 3-dimensional representation of the magnitude of the function $F(s)$, plotted vertically above a horizontal s plane.
 f) Plan view of e), looking down on the s plane.

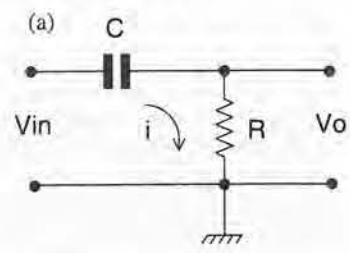


Figure 1e is a three-dimensional representation of the magnitude of $F(s)$, plotted vertically above a horizontal s plane. Only that part of the surface to the left of the $j\omega$ axis (where σ is negative) has been shown. The 'cut edge', above the positive $j\omega$ axis, gives the magnitude of the frequency response, as a linear plot against frequency (also linear). In plan view, the s plane looks like Figure 1f - there is a pole at the point where s has the coordinates $(-1, 0)$.

As noted in the earlier article, in terms of the complex frequency variable s , the transfer function $F(s)$ which gives v_o/v_{in} becomes $F(s)=1/(s+1)$, where $s=\sigma+j\omega$.

If the pole moves further and further toward the origin, the response will rise by 6dB every time the frequency is halved, and so on indefinitely. With a pole at the origin, you have an integrator - i.e., an ideal op-amp with capacitive feedback from the output to the inverting input, and with the input applied via a resistor.

...and zeros

Now for a 'finite zero', that is to say a zero at a finite frequency, which will appear on the pole/zero diagram. You get one with a high-pass (bass cut or passive lead) circuit, shown in Figure 2a, along with its circle diagram,

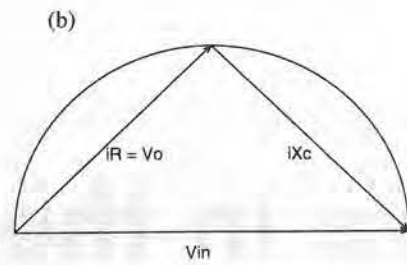


Fig. 2. a) CR highpass circuit. b) Circle diagram, with vector diagram for the -3dB frequency. Note: the input current i is in phase with V_o , i.e. leading V_{in} by 45° at this frequency. c) Vector diagram of b), superimposed on the pole/zero diagram. d) Crude 3-dimensional representation of the magnitude of the function $F(s)$, plotted vertically above a horizontal s plane.

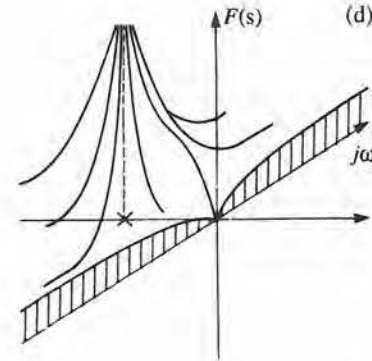
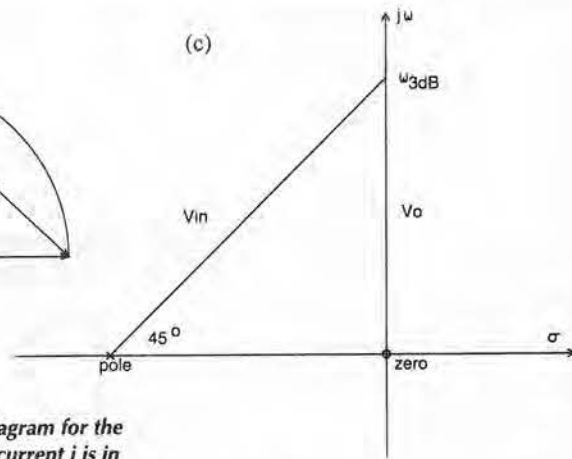


Figure 2b). Note that, by convention, a pole on the s plane (the $j\omega$ versus σ plot) is denoted by a cross, and a zero by a circle or nought. From the vector/circle diagram of Figure 2b), for a simple highpass CR circuit:

$$v_o/v_i = R/(R+1/j\omega C) = j\omega CR/(1+j\omega CR) \quad (2)$$

Thus the general expression for the transfer function $F(s)$ is $F(s)=s/(1+s)$, assuming that the frequency is normalised, or (effectively the same thing) that $CR=1$.

Clearly, as well as exploding to infinity when s is -1 (when $\omega = \text{zero}$ and $\sigma = -1$), $F(s)$ is 0 when s is 0 (when both ω and $\sigma = 0$), due to the s in the numerator. Figure 2c shows the

pole/zero diagram, with its zero at 0Hz, with the vector diagram superimposed, shown for the case where normalised $\omega=1$, the -3dB point.

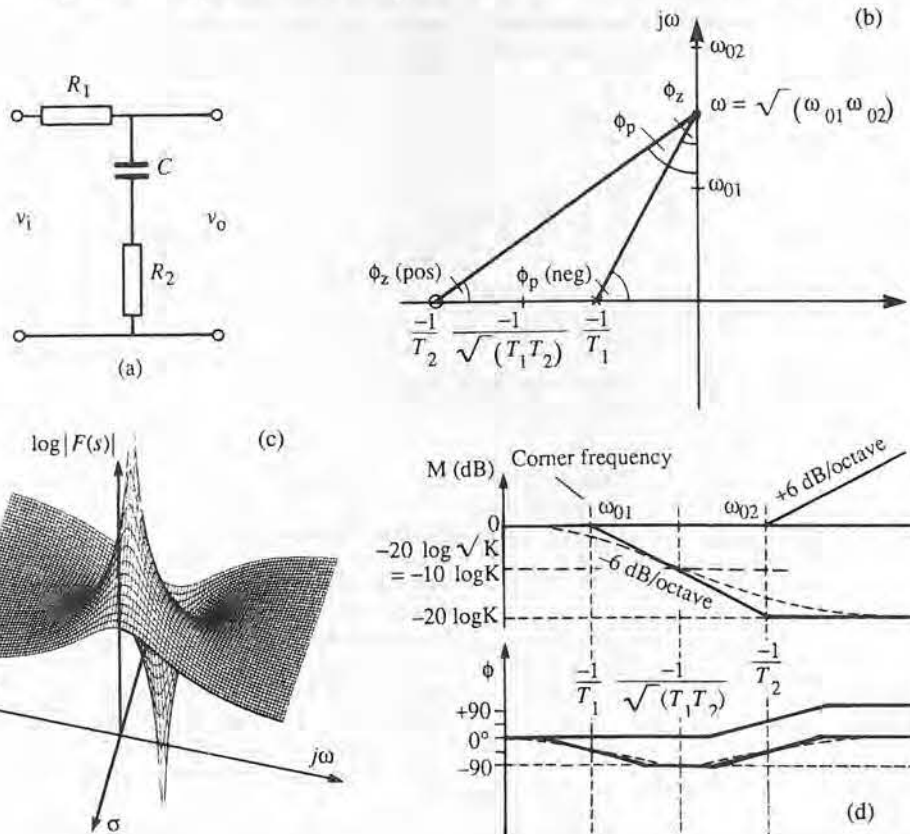
Note that in Figure 1c), v_o is represented by the line from the pole to the origin. This is because, in the vector diagram, this is the voltage iX_c across the capacitor. In Figure 2c), v_o is represented by the line from the origin to the point on the $j\omega$ axis representing the frequency of interest. This is because this line represents iR , across which the output voltage is developed, Fig. 2a).

You can see from Fig. 2c) and Eqn 2 that, for very low frequencies (where v_{in} is virtually horizontal and the denominator virtually equal to unity), v_o is directly proportional to the distance from the origin to the point on the $j\omega$ axis. Also, at very low frequencies, v_o leads v_{in} by 90°. This is made clear by the vector diagram, and can be checked by making the denominator of Eqn 2 real and finding $\tan \phi$, as was done above for the lowpass case.

These two results give another important rule:

the amplitude of the response at any frequency due to a zero is directly proportional to the length of a line from the zero to the corresponding point on the $j\omega$ axis, and the phase is given by the angle that

Fig. 3. a) Transitional lag circuit. b) The corresponding pole/zero diagram. c) Three-dimensional representation of the logarithm of the magnitude of the function $F(s)$. d) This circuit, has unity gain at 0Hz, falling to $R_2/(R_1+R_2)$ at very high frequencies



the line makes with the horizontal axis, counting anticlockwise rotation as positive, indicating a leading response, clockwise a lagging response. The angle is measured with respect to the $+\sigma$ axis if the zero is to the left of the $j\omega$ axis, but with respect to the $-\sigma$ axis if to the right of it.

In the case of Fig. 2a), however, the effect of the zero is not the whole story. To evaluate the frequency response (the magnitude of $F(s)$ along the $j\omega$ axis), you must take into account also the effect of the denominator of Eqn 2, representing the pole. So while initially, where v_{in} can be considered horizontal, v_o is simply directly proportional to the distance up the $j\omega$ axis, remember that it is also inversely proportional to the distance from the pole.

At very high frequencies, these distances become more and more nearly equal. The response is thus proportional to the distance times the reciprocal of an equal distance, result unity. Likewise, the phase is everywhere equal to the sum of the angles. So for the CR highpass circuit, the phase is $(90+\phi_{pole})^\circ$ where ϕ_{pole} is zero at 0Hz, lagging by 45° at the -3dB point and reaching -90° at infinite frequency. Thus overall, the phase of v_o starts off at 90° leading at 0Hz, dropping back to being in phase at very high frequencies.

Section (d) of Fig. 2b is a three-dimensional representation of the magnitude of $F(s)$, plotted vertically above a horizontal s plane. Only that part of the surface to the left of the $j\omega$ axis (where σ is negative) has been shown. The 'cut edge', above the positive $j\omega$ axis, gives the magnitude of the frequency response, as a linear plot against frequency (also linear). The zero at the origin can be seen to act as a thumbtack, pinning the surface $F(s)$ to the ground level at the origin.

If the pole is moved further and further to the left, the -3dB point, where the response is levelling out, will occur at a higher and higher frequency. If the pole moves out to infinity, the response will be rising at 6dB/octave indefinitely. With just a zero visible on the s plane, at the origin, you have a differentiator, e.g., an ideal op-amp with resistive feedback from the output to the inverting input, and with the input applied via a capacitor.

Zeros to the left of them

Zeros can occur anywhere in the s plane - not just on the $j\omega$ axis. Figure 3a shows a circuit known as the transitional lag, which starts off behaving like the top cut circuit of Fig. 1a but, instead of the response falling indefinitely as the frequency increases, it flattens out again, the response at very high frequencies being equal to $R_2/(R_1+R_2)$. This is illustrated by the Bode diagram of Fig. 3d), from which you can see that the response starts to fall at a 'corner frequency', but stops falling at some higher corner frequency. The actual response is shown dotted, and a -6dB/octave asymptote is shown joining the low-frequency and high-frequency response levels - unity and $R_2/(R_1+R_2)$ respectively.

The -6dB/octave asymptote is a straight line

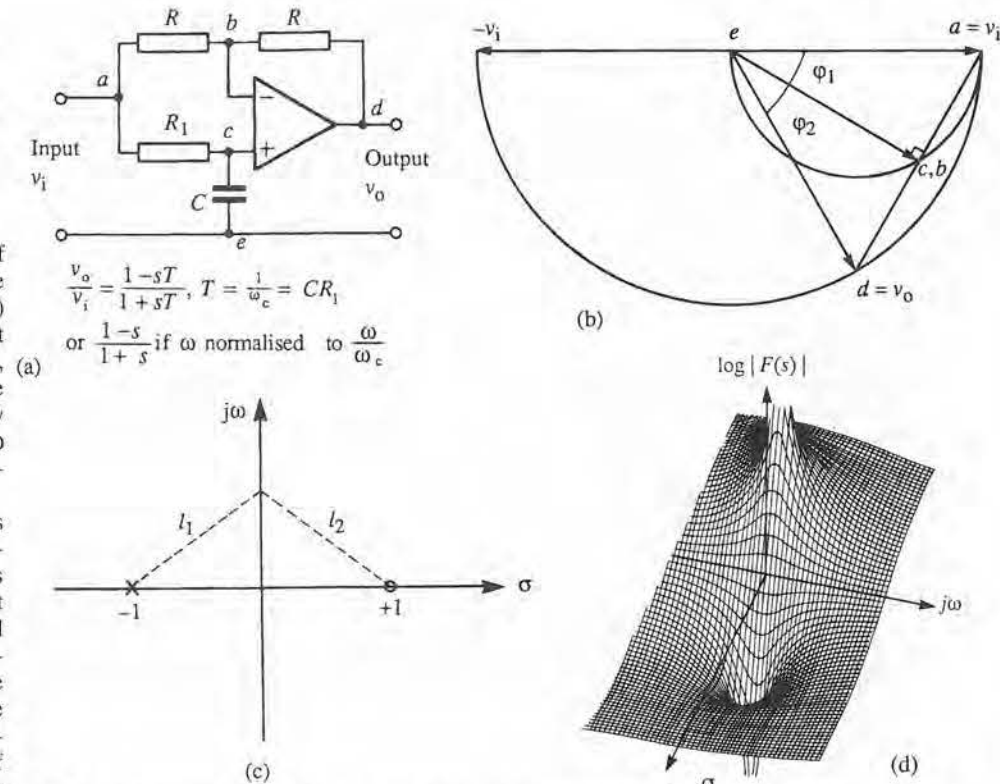


Fig. 4. a) First-order all-pass circuit with unity gain at all frequencies has a phase shift varies with frequency. b) Vector/circle diagram for the circuit a). c) Pole/zero plot for the circuit of a). d) 3-dimensional representation of the logarithm of the magnitude of the function $F(s)$, corresponding to c).

because the horizontal axis, representing frequency, is logarithmic, and the vertical axis, representing the magnitude of the response M , is in dB - i.e., also logarithmic. The actual response will never quite reach a 6dB/octave cut-off rate unless the corner frequencies ω_{01} and ω_{02} are infinitely far apart, and of course it is curved at the corner frequencies, unlike the notional asymptote.

Up till now, it has been convenient to work with normalised frequencies since, in both the low pass and high pass circuits of Figs 1 and 2, there is just the one corner frequency, where the response changes (gradually) from flat to -6dB per octave (or vice versa). But with the transitional lag, there are two different frequencies to consider, and these can conveniently be defined in terms of the CR time constants of the circuit.

Working out the response of the transitional lag of Fig. 3a gives:

$$\begin{aligned} \frac{v_o}{v_{in}} &= \frac{(1/j\omega C) + R_2}{(1/j\omega C) + R_2 + R_1} \\ &= \frac{1 + j\omega CR_2}{1 + j\omega C(R_1 + R_2)} \\ &= \frac{1 + j\omega T_2}{1 + j\omega T_1} = \frac{T_2(1/T_2 + j\omega)}{T_1(1/T_1 + j\omega)} \end{aligned}$$

where $T_2=CR_2$ and $T_1=C(R_1+R_2)$, or more generally: $F(s)=K(s+1/T_2)/(s+1/T_1)$ where

$K=T_2/T_1=(1/\omega_{02})/(1/\omega_{01})$. Term ω_{01} is the lower corner frequency, set by $1/[C(R_1+R_2)]$ and $\omega_{02}=1/CR_2$ is the upper, which means that $K=R_2/(R_1+R_2)$.

This time, the pole zero diagram Fig. 3b, instead of showing normalised frequencies, shows actual frequencies, defined in terms of the circuit time-constants. The point shown on the $j\omega$ axis represents the frequency which is the geometric mean of ω_{01} and ω_{02} , the point of inflection of the response in Fig. 3d). Up to this frequency, ϕ_p is increasing faster than ϕ_z , so that the output lags the input, but beyond this frequency, ϕ_z increases faster so the lag decreases again. Eventually, as ϕ_p and ϕ_z both tend to 90°, they cancel out, and the output is back in phase with the input, but reduced in amplitude to $R_2/(R_1+R_2)$.

Figure 3c shows the surface representing the magnitude of $F(s)$, drawn in three dimensions as a wire grid model. For clarity, it has been rotated so the the $+\sigma$ axis is pointing toward you and, in the vertical direction, it shows the logarithm of the magnitude of $F(s)$. This means that a zero, instead of appearing as a thumbtack pinning the surface to the zero level, now appears like an upside-down pole, extending indefinitely downward. Only that part of the surface over the negative σ axis is shown, and the edge above the $+j\omega$ axis (marked by a bold line) gives the amplitude response at any frequency ω , being unity at 0Hz and tending to K at very high frequencies.

If you imagine R_2 getting smaller and smaller, then the position of the zero at $-1/T_2$ in Fig. 3c migrates ever leftward, and K , the response at very high frequencies, gets ever smaller. Ultimately, as R_2 reaches 0Ω, the zero has disappeared leftward off the page to infinity, and you are back with the circuit and response of Fig. 1.

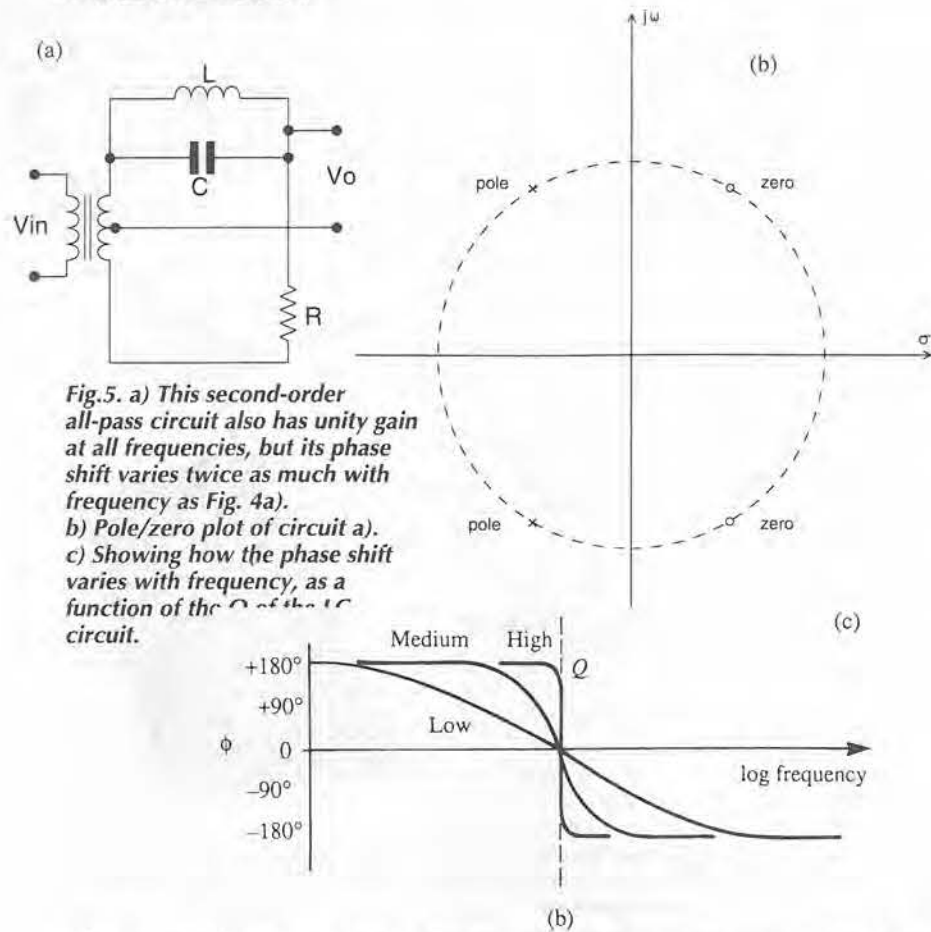


Fig. 5. a) This second-order all-pass circuit also has unity gain at all frequencies, but its phase shift varies twice as much with frequency as Fig. 4a). b) Pole/zero plot of circuit a). c) Showing how the phase shift varies with frequency, as a function of the Q of the LC circuit.

If, instead of being in series with R_2 , the capacitor in Fig. 3a had been in parallel with R_1 , the circuit would be a transitional lead, the response rising from K at 0Hz to unity at infinite frequency. In this case, the zero would be nearer the origin than the pole; the two have interchanged places. So as the frequency increases, starting from 0Hz, initially ϕ_z increases faster than ϕ_p and the phase of the output leads that of the input. Later, ϕ_p increases faster, and the lead disappears, as the gain gradually reaches unity.

And zeros to the right of them

An all-pass filter is one which changes the phase of a signal by different amounts, depending on the frequency, but without affecting its amplitude. In signal transmission systems, this enables distortion due to various causes to be cancelled, and the signal restored more or less to its original condition.

The simplest all-pass circuit, also called a phase-equaliser, is the single-pole variety, Fig. 4a). The signal at the op-amp non-inverting terminal is applied through a CR low-pass circuit, exactly as considered earlier. Consequently, relative to the input v_i (vector e-a in circle diagram, Fig 4b)), the voltage at the op-amp non-inverting input is the vector e-c. This coincides with v_i at 0Hz, falling towards 0 with a 90° phase lag at very high frequencies.

Negative feedback around the op-amp forces its output to do whatever is necessary to keep the voltage at its inverting input equal to that at the non-inverting input. So point c on the circle diagram is also point b. Since the two resistors connected to the inverting input are

equal, the output v_o (vector e-d) follows the larger semicircle shown, as though drawn by a pantograph set for two times magnification.

Thus the amplitude of v_o is independent of frequency, but its phase relative to v_i swings round from 0° at 0Hz, to -180° at infinite frequency. The all-pass transfer function $F(s)_{AP}$ is easily derived from that for the low-pass case, by observing that the output, in Fig 4b), is twice as big as the input, but shifted to the left by one unit - by the distance e-a. So $F(s)_{AP} = 2F(s)_L - 1$. But the $F(s)$ for the low-pass circuit was shown earlier to be $1/(s+1)$.

$$F(s)_{AP} = \frac{2}{s+1} - 1 = \frac{2}{s+1} - \frac{s+1}{s+1} = \frac{1-s}{1+s}$$

There is a zero at $\sigma = +1$, $\omega = 0$, since for $s = 1 + j0$, the numerator of $F(s)_{AP}$ is zero. This is shown on the s plane diagram of Fig. 4c). Following the rules noted earlier, the response at any frequency is directly proportional to l_2 and inversely proportional to l_1 . But since these are always equal, the magnitude of the response is always unity.

Also, as the frequency increases, ϕ_p increases from zero to 90°, in an anticlockwise direction indicating a lagging phase angle. And ϕ_z increases from 0° to 90° clockwise, this also according to the rule indicating a lagging response. Thus taking into account the phase contributions from both the pole and the zero, overall the phase drops steadily back from 0° at 0Hz to -180° at very high frequencies.

Figure 4d) shows a three dimensional wire grid model of the surface representing the magnitude of $F(s)$, or rather the log of the magnitude as previously. But this time the sur-

face along the + σ axis is shown as well, since this shows up the zero. Note that the surface is exactly skew symmetrical about the $j\omega$ axis, the magnitude of $\log F(s)$ along the axis being everywhere zero, or 0dB.

Like all the circuits that have been considered so far, this all-pass filter is a 'first-order' circuit, that is to say that the highest power of s in the denominator is 1. If the highest power of s in the denominator is 2, then you have a second order circuit, and if the numerator is a constant, the circuit is lowpass with a 12dB per octave roll-off.

If there is just a term in s in the numerator, the circuit shows a bandpass response with a 6dB/octave roll-off either side of the peak, while a term in s^2 in the numerator gives a highpass with a 12dB/octave low-frequency roll-off.

More zeros to the right of them

The first order all-pass filter described earlier causes the output phase to drop back from 0 to -180° in a fixed and fairly gentle manner, as the frequency rises from zero to infinity. If you want the phase change to take place over a smaller frequency interval, then a second order all-pass circuit can be used. This can be an entirely passive circuit, as in Fig. 5a), the corresponding pole/zero plot being as in 5b).

With this second-order all-pass circuit, the phase drops back from 0° at 0Hz, via -180° right round to -360° at infinite frequency.

If, at the resonant frequency of L and C , the reactance of each is large compared to R , then the phase change as the frequency changes will be very gradual. If on the other hand the reactance of each is small compared to R , then the phase will snap right around over a very small range of frequencies centred on the resonant frequency. This corresponds to the poles and zeros in Fig. 5b) being very close to the $j\omega$ axis. In this case, over most of the frequency range, output voltage will be determined by the signal coming via the inductor (below the resonant frequency) or via the capacitor. Only at the resonant frequency, and in its immediate vicinity, where the dynamic impedance of the tuned circuit is high, will the output be due to the signal coming via R .

This all assumes, of course, that L and C are ideal. A more practical arrangement uses active circuitry, in particular, the state variable filter, see page 143 of Ref. 2. If you add the filter's low and high-pass outputs, plus an appropriate proportion of the bandpass output, you again get a second-order all-pass response. The transfer function is,

$F(s) = (s^2 - Ds + 1) / (s^2 + Ds + 1)$, where D is the reciprocal of the filter Q. The s^2 term in the numerator is the highpass component, s the bandpass component, while 1 represents the lowpass component. You can see that when $s=0$ (so that $\omega=0$), the transfer function is 1/1, or unity.

When s equals infinity, you can forget the other terms, so the transfer function is just $\omega^2/\omega^2=1$ again. When $s=(\sigma+j\omega)$, with $\sigma=0$ and ω (the normalised frequency)=1, then $s^2=-1$, so the transfer function simply equals

$-Ds/+Ds$. The numerator indicates a 90° phase lag, and the denominator another, giving a total of 180° lag at the filter's normalised resonant frequency.

The pole/zero diagram makes it clear that the closer the poles and zeros are to the $j\omega$ axis (the higher the circuit's Q), the more more rapidly the phase changes as you pass between the upper pole/zero pair, travelling up the + $j\omega$ axis from 0Hz at the origin, towards infinite frequency.

The lower pole/zero pair has comparatively little effect on the phase in this region; it is there because both numerator and denominator are quadratic expressions, and their presence balances the upper pair, ensuring that the gain at 0Hz is unity and the phase shift 0°.

Lots of poles - and zeros

By now, I hope you have a feel for poles (s terms in the denominator of $F(s)$) causing phase lags and gain changes, and zeros (s terms in the numerator) causing gain changes and phase lags or leads, according to where they are. So here without further ado or any detailed explanation, are some more pole zero plots, and three dimensional wire grid models of $\log|F(s)|$.

Figure 6a), sections i-iii shows the pole zero plots for some fourth-order lowpass filters. Those of i and ii are 'all pole' filters, i being a Butterworth maximally-flat amplitude response, and ii a Chebychev response which gives a faster initial roll-off in the stop band at the expense of ripples in the passband.

In iii is a fourth-order elliptic response, which gives an even sharper cut-off in the stop band. This is thanks to the finite zeros situated on the $j\omega$ axis; but they do result in the stopband attenuation ultimately levelling out, rather than increasing for ever like the all pole filters. Figure 6b), sections i-iii shows the 3D wire grid representations of $\log|F(s)|$ for these three filter types, in each case cut to show only the part over the - σ region.

As before, the cut edge over the $j\omega$ axis -

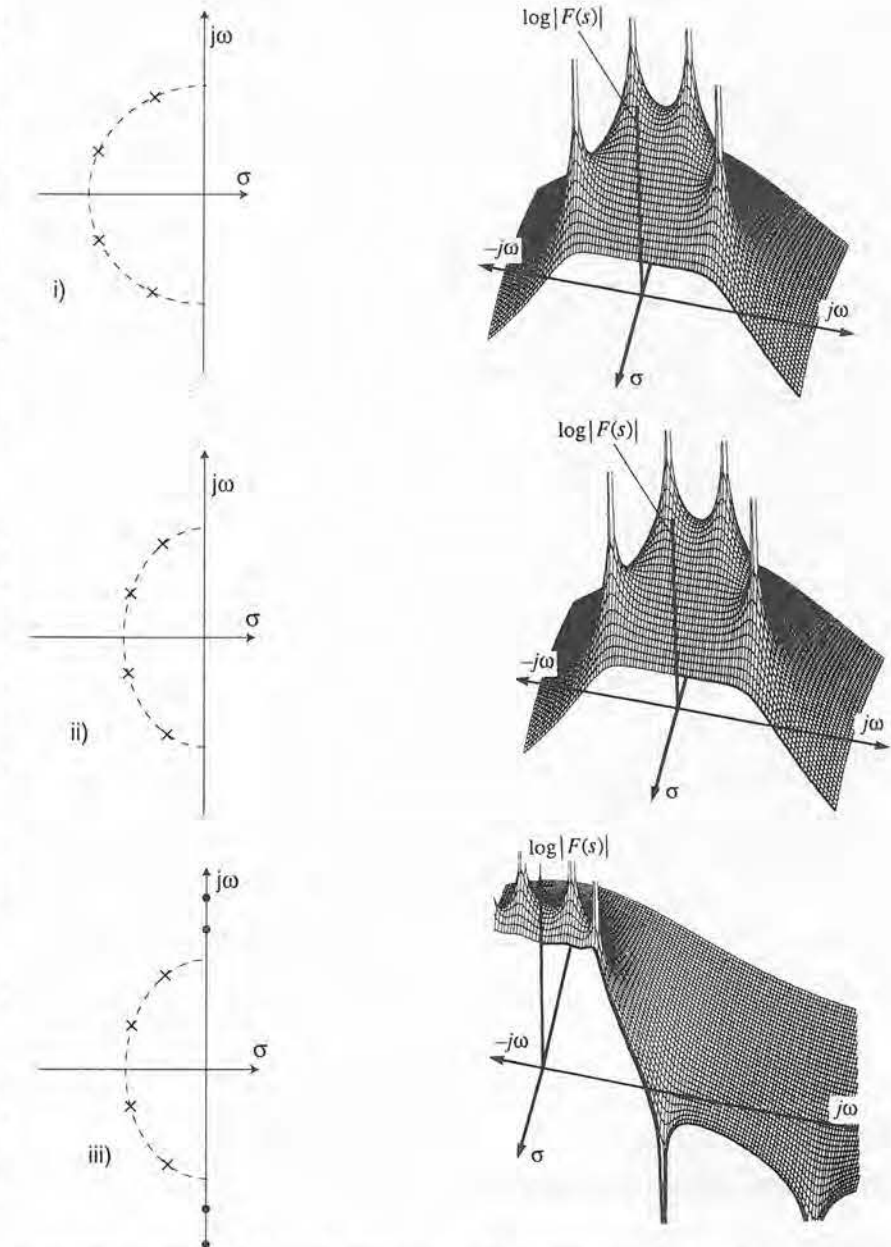


Fig. 6a) Pole/zero plots for three popular lowpass filter types; in each case, a four pole design is shown. i) The Butterworth design gives a maximally flat amplitude response. ii) The Chebychev design gives a faster cut-off in the stopband, at the expense of ripples in the passband response. iii) The elliptic or Cauer design gives an even faster stopband cut-off, but as well as passband ripples, has returns in the stopband b) Three dimensional wire grid representations of $\log|F(s)|$ for the three filter types. The frequencies of infinite attenuation in the elliptic case correspond to the positions on the j axis of the zeros in a) iii).

Transforms ahoy

It may seem unnecessarily complicated to bother about the whole s plane, when all you need to find a circuit's frequency response if the value of $F(s)$ for $\sigma=0$, i.e. how $F(s)$ varies along the $j\omega$ axis. But there is more to pole/zero diagrams than just a circuit's frequency response. The diagram can also represent an input signal to the circuit, and not just a steady-state sinewave either. A further mathematical trick - called the inverse Laplace transform - can then derive what the circuit's output waveform will look like, given its frequency response and the said non-sinusoidal input. But to cover that would take more space than is at my disposal. ■

References
 1. Hickman, I, 'In the picture', *Electronics World*, Jul/Aug 1996, pp 558-561.
 2. Hickman, I, *Analog Electronics*, ISBN 0 7506 1634 2, Butterworth-Heinemann 1993.

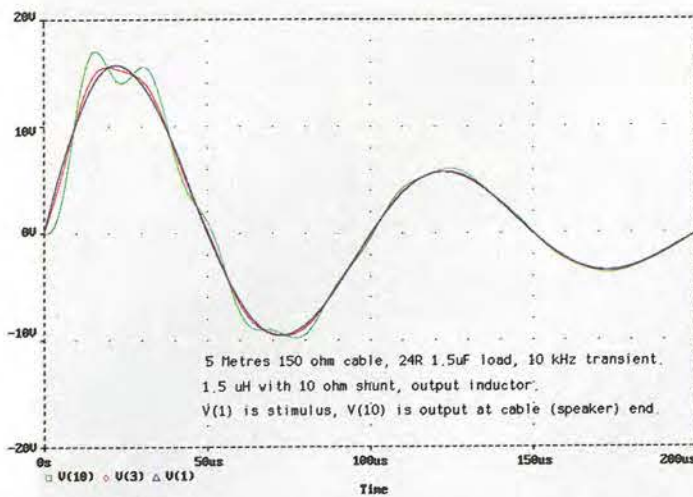


Fig. 1. Simulation of typical amplifier output stage when driving into a capacitive load with high-value shunt impedance, via a high impedance, typical figure-of-eight-style cable. This diagram was used for the first article.

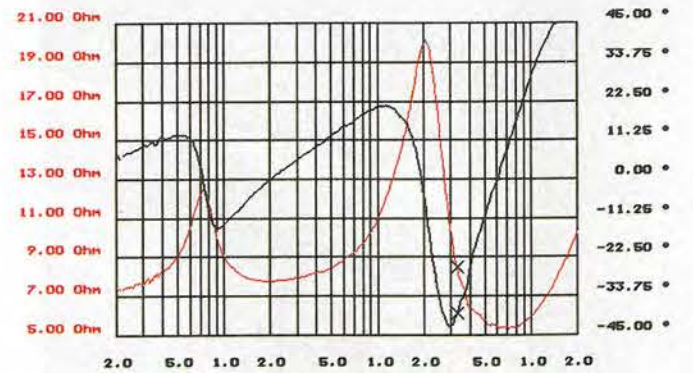


Fig. 3. Impedance and phase plot by frequency, of rebuilt two-way transmission line test speaker. Compare with published plot in reference 3.

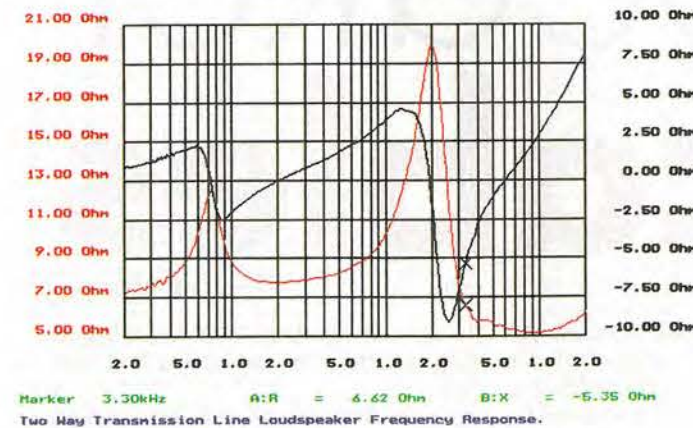


Fig. 5. The Fig. 3 plot again but this time displayed in 'R + jX' terms, clearly showing the inductive and capacitive parts of the frequency range, with negative X values being capacitive and positive X values representing inductive reactances.

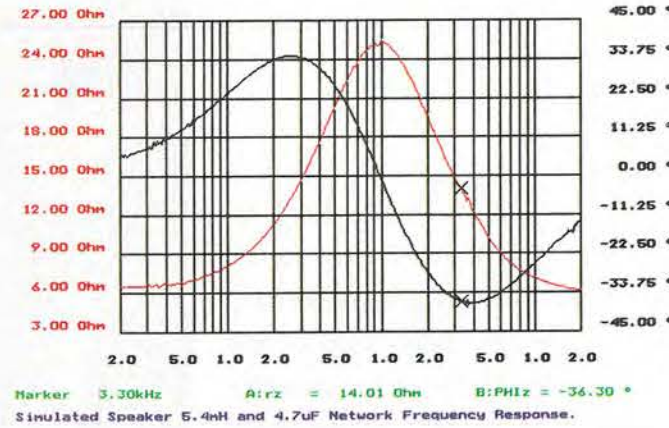


Fig. 2. Impedance and phase plot of the 'pseudo' IHF dummy reactive load, made using a 5.6Ω resistor in series with a parallel combination of 5.4mH, 4.7μF and an 18Ω damping resistor. This shows phase change with frequency for this high-impedance resonant circuit.

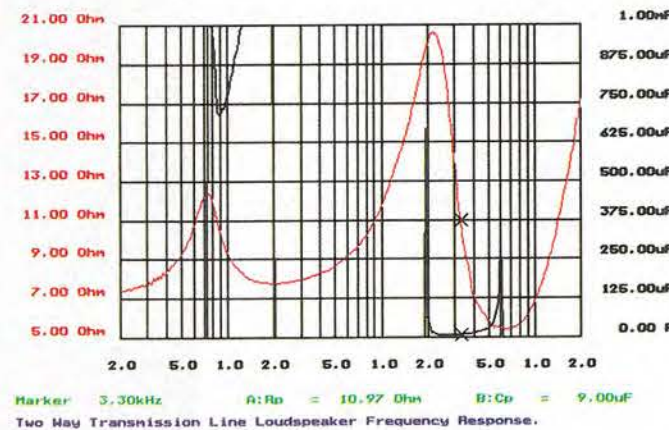


Fig. 4. Plot Fig. 3 displayed in terms of parallel resistance and equivalent parallel capacitance values. Note relatively large equivalent capacitances near bass resonance.

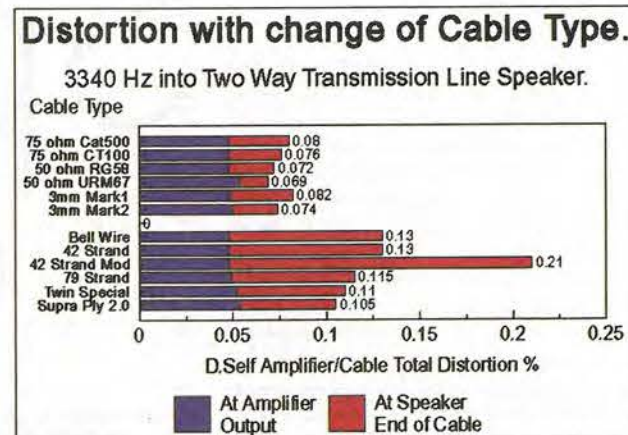


Fig. 6. This plot of harmonic distortion measured at the test speaker terminals, with the speaker acting as a capacitive load, shows clearly how distortion changes with cable. Great care was taken to maintain all other test conditions constant, as can be judged by the small changes in distortion measured at the amplifier terminals. This test was repeated on three separate days, all giving similar results, and only small levels of non-harmonic noise were noted.

Distortion with change of Cable Type.

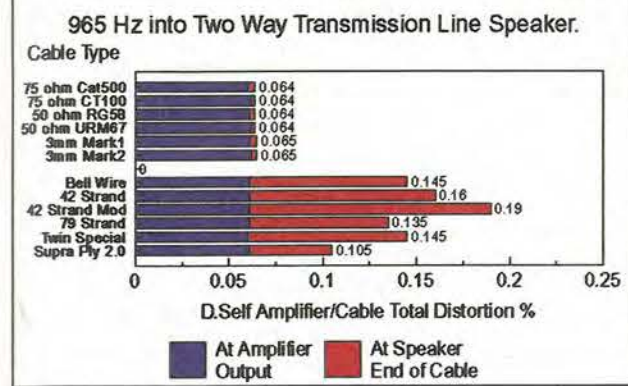


Fig. 7. Harmonic distortion measured at the test speaker terminals, with the speaker acting as an inductive load, shows smaller distortion changes with cable. During this test, non-harmonically related noise was observed for the figure-of-eight twin-line cables, clearly showing the coaxial cable's superior isolation from transmitted noise. This pick-up partly accounts for the poor performance of the Figure 8 cables with this inductive real speaker loading.

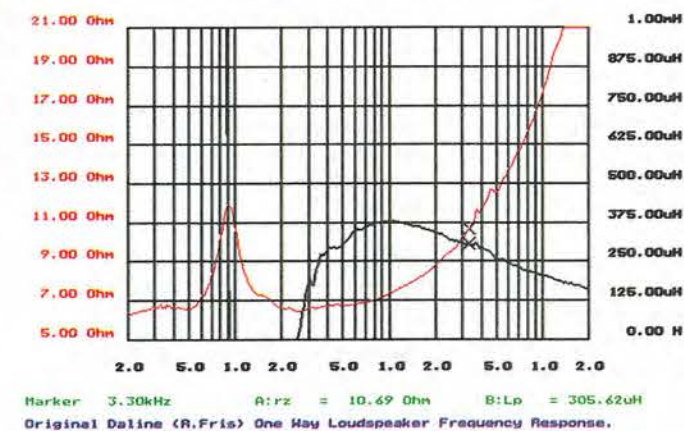


Fig. 8. Impedance and equivalent parallel inductance plot of the R. Fris original Daline cabinet, clearly showing high-frequency inductive behaviour. This cabinet was chosen since it has no crossover network.

way corner horns in my listening room.⁶ None of these is in an infinite baffle cabinet.

How can I measure the impedances and reactances of these speakers? While impedance plots can be made using nothing more than a signal generator, a known resistor and a suitable voltmeter, phase measurement is much more difficult.

Many personal computers have a 16-bit sound card. With suitable software, such a card can provide the heart of an extremely low-cost audio frequency measurement system, capable of measuring impedance and phase angle. While not state-of-the-art, this method can produce useful measurements. Two low-cost software systems for use with these sound cards are readily available (see panel on sound cards and software).

To ensure a resonant peak at crossover, I rebuilt my two-way transmission line as an 8Ω system, with a crossover based on the old Kef DNI3 design⁴, built using polycarbonate and polypropylene capacitors and used with a small cone tweeter. Using the *Elektor* software with my sound card, the resulting impedance plot was similar to that published for the Tannoy D700 system³, Fig. 3.

This conventional impedance phase plot can also be viewed in terms of equivalent inductance, capacitance or as 'R+jX' - whichever is preferred. Viewed as the parallel resistance and equivalent parallel capacitance, you can see that frequencies near 3kHz could provide loading similar to the simulations used for Fig. 1, with consequent distortion, Figs 4, 5.

Having established that the conditions needed for transient distortions

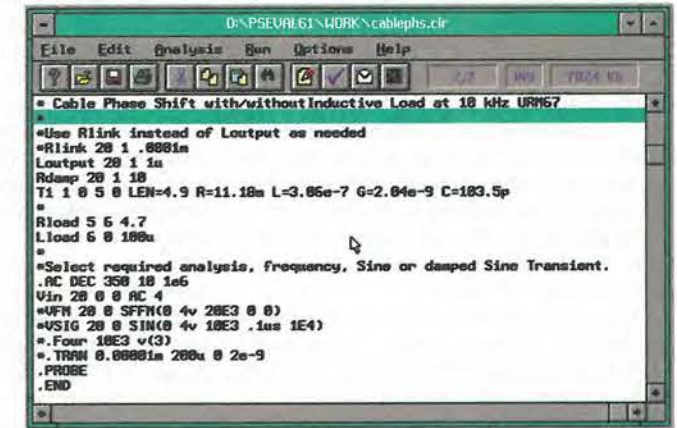


Fig. 9. PSpice net-list used to calculate the high frequency phase deviation results, with change of cable parameters, and various loading circuits, for Table 1. Note use of Spice 'SIN' waveform.

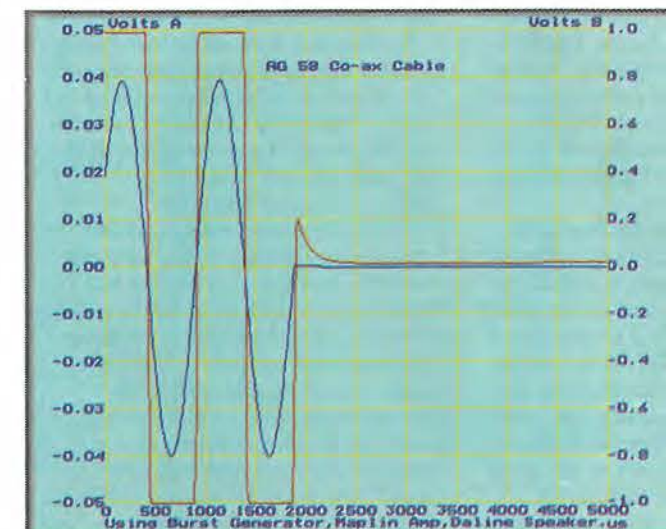


Fig. 10. Plot of loudspeaker damping, or lack of, using RG58 cable with the Daline speaker cabinet. This was made using a tone-burst generator and Maplin Amplifier set to 1.6V pk/pk output, with the Pico ADC100 virtual oscilloscope. Note the close similarity to the Duncan test plots.

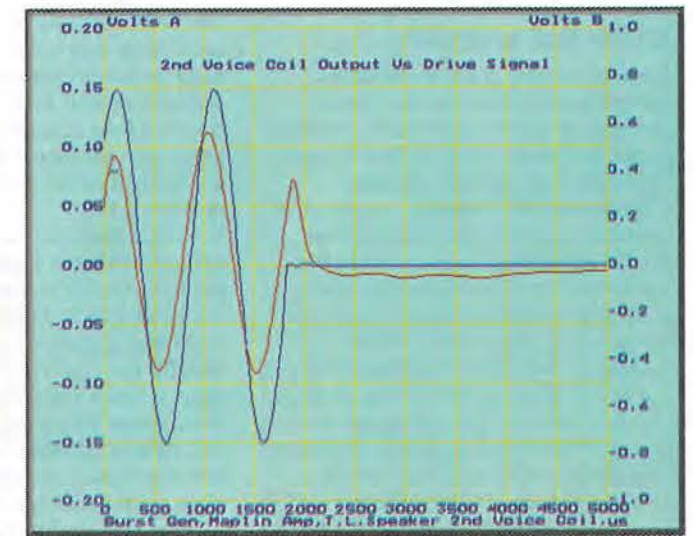


Fig. 11. Using the Fig. 10 set-up and RG58 cable to observe speaker cone overhang using the 'spare' voice coil of the transmission line speaker as a sensor. This clearly shows cone movement at reduced amplitude, continuing far longer than the initial overshoot spike.

Table 1. Simulated phase shift in degrees at 10kHz for the various cables tested. Shows low/high frequency deviations by cable, with amplifier and various dummy speaker loads. Ranked for least overall change with above permutations.

	No output inductor		With 1µH output inductor		Resistor only		Series 100µH		Ranking
	Resistor only	Series 100µH	Resistor only	Series 100µH	4.7Ω	8.2Ω	4.7Ω	8.2Ω	
Coaxial styles									
75Ω Cat.500	-1.485	-0.859	+0.034	-0.211	-2.234	-1.293	-0.232	-0.480	5
75Ω CT100	-1.414	-0.815	-0.114	-0.284	-2.159	-1.245	-0.380	-0.553	6
50Ω RG58C/U	-1.219	-0.708	+0.386	+0.033	-1.961	-1.139	+0.118	-0.237	3
50Ω URM67	-1.149	-0.660	-0.249	-0.320	-1.903	-1.094	-0.515	-0.590	4
3mm Mark 1	-0.685	-0.394	-0.148	-0.074	-1.446	-0.831	-0.342	-0.421	1
3mm Mark 2	-0.697	-0.401	-0.083	-0.155	-1.454	-0.836	-0.351	-0.427	2
Figure-of-8 styles									
2192Y 'Bell' Wire	-2.684	-1.566	-0.094	-0.326	-3.399	-1.984	-0.105	-0.591	9
42-strand	-2.904	-1.685	-0.303	-0.626	-3.647	-2.117	-0.563	-0.891	11
42-strand modified	-4.629	-2.687	-0.932	-1.257	-5.369	-3.118	-1.184	-1.518	12
79-strand	-2.545	-1.466	-0.621	-0.752	-3.301	-1.902	-0.882	-1.019	10
2mm twin special	-1.939	-1.117	-0.394	-0.527	-2.695	-1.553	-0.657	-0.796	8
Supra Ply 2.0	-1.394	-0.803	-0.212	-0.338	-2.152	-1.239	-0.478	-0.608	7

can exist, might these conditions also cause distortion with continuous sine waves? This region should be explored by practical measurements.

Conventional distortion tests

My variable frequency generator produces 0.05% distortion, so it can only be used for comparative, rather than absolute, tests. For the purpose of comparing cables, however, it should suffice. My Hewlett Packard 331A distortion analyser cannot measure distortions much below this level.

Using this generator with a two-stage series/parallel LC clean-up filter, I measured 0.045% distortion at 3340Hz at the terminals of my Self amplifier,⁷ loaded with 8.2Ω.

This amplifier, with the above instruments, was used with all twelve cables to test drive the rebuilt two-way transmission line speaker, making distortion measurements in sequence

Other test methods

Some of you may prefer to test cables using a low-cost tone burst test more in line with the Duncan test method. I include screen shots made using the Pico adapter with a low-cost, self-built generator.

Readers with a 50Ω output square wave generator and who wish to experiment using transient waveforms similar to that used in the Spice simulations can easily make a good 10kHz replica by loading the generator output with a parallel combination of 1mH with 0.22µF and 47Ω.

A variety of test signals can be found on the *Hi-Fi News and Record Review Test Disc III*. This includes a transient waveform very similar to the Spice exponentially damped sine wave, 'SIN', used for Fig. 1. However, the tone burst included on this disc equates to the IHF tone burst in having a change in level but no distinct off periods.

at both amplifier and loudspeaker terminals.

To help this rebuilt speaker survive prolonged testing, I used only half the drive voltage used by Duncan for his July/August article, i.e. 0.45V. Even at this reduced level, some ear protection is desirable.

Using a test frequency of 3340Hz, chosen to match my clean-up filter, notable changes in distortion at the loudspeaker terminals with change of cable were easily measured by my equipment, Fig. 6.

From the various simulations made investigating these cables, a similarly high but inductive impedance was expected to produce much less distortion. Change of test frequency to 965Hz, with appropriate changes to the clean-up filter, resulted in increased distortion at the amplifier terminals. However, as expected, incremental distortion was lower at the speaker end of the test cables, since the speaker was now behaving as an inductive load.

During this test, with the distortion residuals viewed by oscilloscope, non-harmonically related noise was observed for the figure-of-eight twin-line cables, providing a clear demonstration of the coaxial cable's superior isolation from transmitted noise. This pickup partly accounts for the performance of the figure-of-eight cables with this inductive speaker loading, Fig. 7.

At bass frequency, almost all loudspeaker designs exhibit a high-impedance resonant peak or peaks. It is reasonable to expect that similar distortion changes would also be measured with a change of cable. I have not tried this, for two reasons. Firstly, my test equipment is inadequate for measurements near 50Hz mains frequency. Secondly, the available speaker cabinets either have well damped low impedance resonance peaks or they peak close to the mains harmonic frequencies.

These 3340Hz distortions were much larger than expected, so I built a test circuit similar to the IHF (Institute of High Fidelity Inc, IHF publication A202) reactive load, using a 5.4mH inductor with a 4.7µF metallised poly-

carbonate capacitor to get a high impedance and similar phase to that of the speaker measured. With this test circuit, no realistic continuous sinewave incremental distortions were measured, Fig. 2.

A speaker is obviously a complex motor generator system giving results not easily represented by simple passive component models. Given time, perhaps this could be resolved – but not by me.

Relative phase

The final parameter influenced by cable

Sound cards and software

For these tests I used a Creative Labs *Soundblaster 16 Value Plug-and-Play* card with general purpose software (Elektor Electronics, part no. 966001-1). It is important to check that the dma channels and interrupts needed are not already in use in your system.

An alternative software for sound cards, targeted especially at speaker systems, is *AIRR* (Anechoic In-Room Response) by Dr J J Bunn, supplied by Old Colony Software. It is also available from Falcon Acoustics Ltd, Tabor House, Norwich Rd., Mulbarton, Norfolk NR14 8JT

Also available from Falcon and Old Colony is *Audiosuite* software, again for use with a sound card. It includes MLS functions but is much more expensive.

Those seriously interested in the design of loudspeakers rather than measurement might consider the Old Colony *IMP* (Impulse Measurement and Processing), which is a stand-alone system for a pc, available in kit form and reasonably priced. This also is available from Falcon.

Many other software and hardware systems are available, and the above list covers only some of those priced for the speaker amateur.

choice is change of high frequency phase relative to phase at lower frequencies. Every cable has a transit delay, influenced primarily by the cable's inductance and insulation material. This delay could change with frequency or test-circuit loading.

I am not well equipped to measure small phase angles but, knowing the cable's ac parameters, this could be simulated.

Most Spice-based simulators have transmission line models available but, to minimise transient simulation times, their default settings use a minimum of only two sampling points along the cable. This produces poor results with short cable lengths at audio frequencies. Also, the supplied small signal transmission line model is much simplified.

By overriding these defaults and forcing a slower simulation, acceptable results are obtainable. Many loudspeaker impedance plots show an inductively rising impedance above 10kHz. Thus, each of the twelve cables tested has been simulated using both 4.7Ω and 8.2Ω resistive loads and a series combination of these resistances with 100µH, Fig. 8.

Most – but not all – domestic amplifiers are

protected against reactive loads by a low-value inductor in series with their output. To cover all reasonable options, the above load combinations were simulated with and without a series inductance of 1µH and 10Ω damping resistor, Fig. 9.

These simulations were all made at 10kHz since, with these values, it represents a worst-case frequency. Regardless of the cable, only small changes in relative phase could be observed, Table 1.

In summary

The intention of these experiments was to quantify the effects of cable characteristics on amplifier speaker systems, rather than to choose a 'best buy'. This determined the cables chosen for test.

My use of coaxial cables as speaker cables is not new. Indeed, extremely low-impedance specialist *Mogami* speaker coaxial cable was reviewed by Nelson Pass⁸ in his 1980 article.

However, these experiments would be incomplete without forming at least a ranking of cable performance. This ranking was unweighted, taken from simple addition of place

numbers for each test. Although various test weightings could be used, regardless of this, the first four placings are unlikely to change, Table 2.

These rankings were supported by listening tests. However, for the reasons outlined in my first article, listening tests were performed only as the concluding test, long after completion of all published tables or figures, and with this text at the final editing stage.

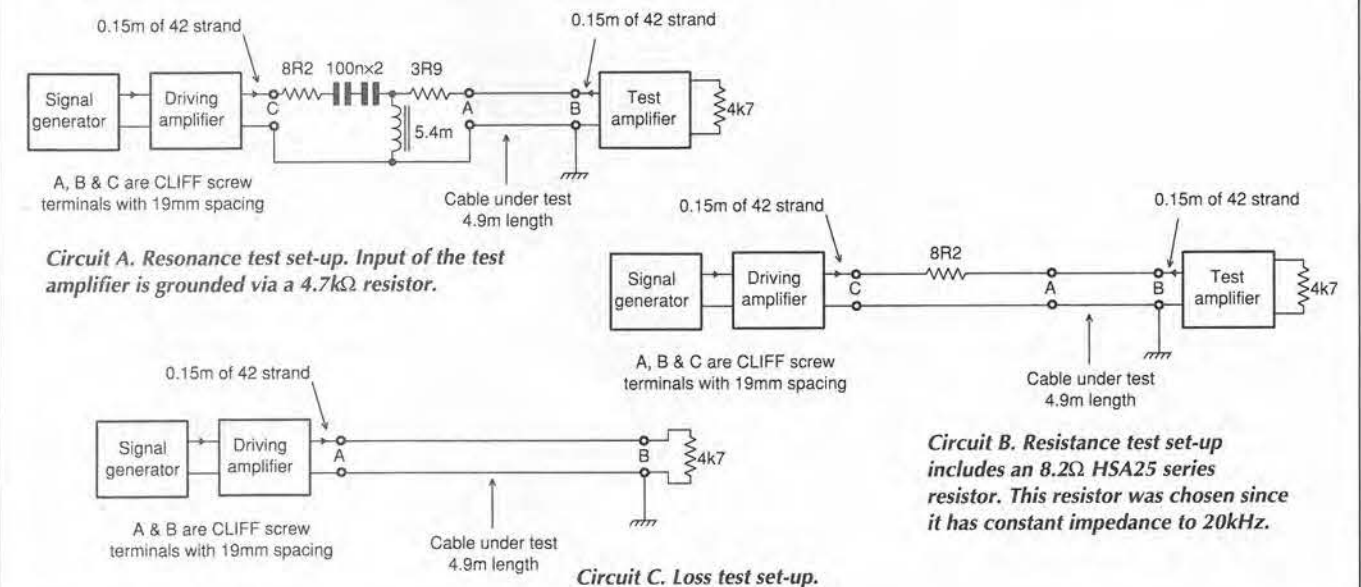
Regardless of these results, users of Naim and similar amplifiers not having an output inductor must use the maker's recommended cable and cable length, since the cable functions to replace the output inductor in protecting their amplifiers.

For all conventional inductor output amplifiers, the ideal speaker cable would have zero resistance, zero inductance and zero impedance. In other words, zero length – no cable at all⁸, since all cable degradations increase with cable length.

It is obvious from the results presented that a cable should have the lowest possible dc resistance, low characteristic impedance at audio frequencies, and minimal inductance.

Table 2. Summary of cable rankings by each test performed. Final ranking derived using equal weighting for all tests. All cables 4.98m long.

	Circuit A	Circuit B	Z ₀	Inductance	DCΩ	Loss	Phase change	THD	Overall
Coaxial styles									
75Ω Cat.500	8	8	9	7	8	7	5	5	8
75Ω CT100	7	=6	8	6	7	6	6	=2	=5
50Ω RG58C/U	10	10	6	4	11	2	3	=2	=5
50Ω URM67	3	3	4	3	1	1	4	1	3
3mm Mark 1	2	2	1	2	3	=3	1	5	2
3mm Mark 2	1	1	2	1	2	=3	2	=2	1
Figure-of-8 styles									
2192Y 'Bell' Wire	12	12	10	10	12	11	9	10	11
42-strand	9	9	11	11	=9	9	11	11	10
42-strand modified	11	11	12	12	=9	12	12	12	12
79-strand	6	=6	7	9	4	10	10	8	9
2mm twin special	5	5	5	8	6	8	8	9	7
Supra Ply 2.0	4	4	3	5	5	5	7	7	4



Voice-coil driver effects

Using my tone-burst generator with a Maplin amplifier and an oscilloscope, I tested my available speaker drivers, some of which were 15Ω, and a mixture of long- and short-travel cones.

In general, smaller, lighter cones produced less cable overhang voltage. Size for size, short-travel cones produced less than long-travel cones. Similarly, 15Ω drivers produced less overhang voltage than 8Ω drivers, which produced less than 4Ω drivers, **Fig. 10**.

I was able to compare voice coil impedances quite easily. The bass driver in my two-way transmission line system is an ancient Whitley Stentorian HF1016. This has duplicated voice coils, and is designed to easily configure and be used as 4, 8 or 15Ω impedances.

Configured as 8Ω I can use the spare voice coil as a cone-travel sensor, giving most interesting results when viewed on an oscilloscope, **Fig. 11**.

Is it possible that cables are now considered so important due to the change to lower impedance, long-travel speaker drivers over the years?

These characteristics are almost impossible to achieve using twin-line or figure-of-eight constructions without incurring considerable cable self-capacitance. They are much more readily achieved using the coaxial construction and with acceptable capacitance.

Other frequently discussed parameters, such as sections of individual wire strands and cable insulation material, may well have some relevance, but these tests indicate that they are of secondary importance.

One obvious disadvantage of coaxial speaker cables having dense outer braids is cable end preparation and termination. It is important to avoid damaging any braid wire strands when stripping back, since this would substantially increase distortion levels. However, a dense braid is extremely effective in reducing cable pickup into the amplifier's feedback loop.

As to the test results table, with one exception, the cables segregated automatically into distinct performance groups – coaxial and non-coaxial. Five of the six coaxial cables were outstanding compared to all but one of the non-coaxial constructions tested – including the hand-made, 2mm twin-line special, **Table 2**. Of the non-coaxial styles, the Jenving *Supra Ply 2.0* was the exception, performing well and coming fourth overall,⁹ **Table 2**.

By far the best cable of those tested is the custom made Mark II, followed closely by the Mark I. Neither of these is commercially available in cut lengths but they could be specially ordered in bulk, or further copies could be hand-made, as were my originals.

Both the Mark 1 and the Mark 2 are flexible, very low resistance and – being less than

6mm in diameter – easily installed. Mark 2 has 19 strands of 0.45mm inner wire insulated with polythene with an outside diameter of 3mm. Its outer braid is 240 strands of 0.127mm diameter. Heat shrink sleeve provided overall insulation.

Mk I is identical, except for its 37 strands of 0.32mm inner core. Designers interested in the materials used should send an sae to me via *Electronics World's* editorial offices.

When these tests commenced, I decided to use only cables easily purchased in cut lengths so that my results could be easily replicated. However, in order to understand some early results, I needed to measure a cable having lower impedance at audio frequencies than URM67, but I was unable to buy it. I certainly didn't expect that my crude hand-made cables would perform as they did (see panel on sources of cables used for test).

Of those cables commercially available in cut lengths, the URM67 performs extremely well, coming a close third overall. It is obviously by far the best commercial cable of those tested. However, being 10.3mm in diameter with a solid polythene core and a very dense braid, it is inflexible. It could almost be used as a car tow rope.

A not-so-close fourth, the Jenving *Supra Ply 2.0*, is a specialised, relatively low-cost, high-capacitance cable, available from Jenving's UK agent¹⁰ only in multiples of 10m. In direct contrast with URM67, it is much smaller and, being extremely flexible, drapes well. It is easily installed and hidden from sight.

Following closely on the *Supra Ply's* heels is a tie for fifth place between the 75Ω satellite tv cable CT100 and the 50Ω instrument cable RG58, both ranking equally well overall. However, depending on your needs, you may prefer to down-rank RG58 due to its poorer performance in the resonance and resistive circuit tests and for its high dc resistance.

CT100 and Cat 500 cables are both of larger diameter than RG58. They have a secondary inner copper-foil shield which could fatigue and crack with repetitive coiling and uncoiling, so they should only be used for permanent installations.

Once more, I ask that anyone wishing to shoot these findings down in flames should first repeat the experiments.

Cobbler's shoes

After all these time-consuming experiments, you may wonder which brand of expensive speaker cable I use. My listening room has the 'golden' dimensions of 6.7 by 4.6 by 2.9m, and needs cables somewhat longer than those tested.

My relatively low-cost system, based on Acoustic Research electronics with corner horn speakers⁶ crossing over at 250Hz, was cabled almost 30 years ago using the old-style 7 by 0.029in twin and earth ring mains cable buried in the plaster. I used this for no better reason than that I had plenty left over. I guess its only other merit is a cross-sectional area of more than 3mm, giving reasonably low resistance at the lengths needed. Since my corner horns are 15Ω impedance, it seemed acceptable.

Will I change this cable?

During these experiments I too, like Saul travelling to Damascus, have revised some views and intend to further develop Mark 2 for my own use. This design is now covered by Pat. No 9624876. ■

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Music used for listening tests

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Pictures at an Exhibition, Mussorgsky, Decca 417 299-2.
In The Region of The Summer Stars, The Enid, MNTLCD7.

Cables sources

Initially, I decided that all test cables should be obtainable either from: Electrovalue or Maplin. I bought URM67, CT100, 42-strand, 79-strand, and 300Ω feeder from Maplin, while the RG58 was purchased from Electrovalue. *Supra Ply 2.0* came from Future Film Developments¹⁰ however, and the Bell Wire and Cat 500 from local suppliers.

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One c-mos gate pack can form the heart of a frequency meter, a step-down switching regulator or a negative-supply generator, explains Rae Perälä.

Gates open up

There are many applications for the 4011 c-mos two-input-nand gate – other than just gating. This article presents three such circuits, namely a step-down regulator, a negative-voltage switching regulator, and a frequency measuring circuit.

Step-down regulator

Operating principle of a typical step-down regulator is shown in Fig. 1. Semiconductor switch S_1 is usually controlled by a pulse width modulator, or pwm.

While the switch is on, current flows through inductor L to the output. Simultaneously, magnetic energy in the inductor increases. Diode D stays in its reverse state. When the switch turns off, the inductor discharges its magnetic energy giving a current to the output. Diode D now becomes forward biased, providing path for current flow.

The on-to-off time ratio of the switch is controlled by the pwm circuit. This circuit compares output voltage, V_{out} , to reference voltage V_{ref} and changes the on-to-off ratio accordingly.

Figure 2 is a practical step-down circuit using 4011. It makes a new regulated output voltage V_{out} from a regulated +8 to +15V input voltage. Gates A and B work as an oscillator giving an asymmetrical 40kHz square

wave, Fig. 3. Gate C controls the switching transistor, which is a BS250 p-type enhancement mode mosfet. This same gate inverts the oscillator output voltage, presenting the transistor control voltage V_{GS} as in Fig. 3. The control voltage has long negative pulses interspersed with short intervals. The transistor conducts during the negative pulses.

Gate D regulates the output voltage. Its output is high when V_{out} is beneath the adjusted value. The high state activates gate C, which can now control the switching transistor. When the adjusted output voltage is reached, gate D output goes low, stopping gate C, which remains in the high state. The switching transistor has no control voltage and remains off.

The regulating circuit operates using the 'missing pulse' principle. It provides pulses when the output voltage is low and ceases pulsing when the output voltage is sufficiently high.

Output voltage is adjusted via the 470k Ω potentiometer. Usable output voltage range of the circuit is 0.5 to 0.9 of V_{in} . The circuit can be loaded to 100mA.

Negative voltage regulator

A negative voltage switching regulator is presented in Fig. 4. Gates A and B form an oscil-

lator, giving a symmetrical square wave voltage. Gate C controls the BS250 switching transistor.

During the transistor on time, current flows through the inductor, charging the magnetic energy within it. The magnetic energy discharges while the transistor is off through the diode, producing a negative voltage at the output.

Gate C regulates the output voltage. When the output voltage is too low, the regulating input voltage of gate C is high and the oscillator voltage controls the gate C output and therefore the transistor.

When the output voltage reaches the desired negative value, the regulating input goes low and it stops the output of gate C leaving it in its high state. The transistor stays off until the output voltage goes below its adjusted value.

This circuit requires an input voltage of +8V to +15V, which can be unregulated. A regulated +5V input acts as a reference voltage.

Frequency measurement on a dvm

A digital voltmeter can be used for frequency measurements to 1MHz using the circuit in Fig. 5. The frequency to be measured connects to the input of gate A. This input is biased with 330k Ω resistors to half the supply voltage. This allows low input voltages to change the output of gate A. Figure 6 shows the gate voltages. Output voltage of gate A changes each time the input voltage passes the half supply voltage threshold.

Gate A controls gates B and C. Gate B has a delay capacitor in its output. Figure 6 shows

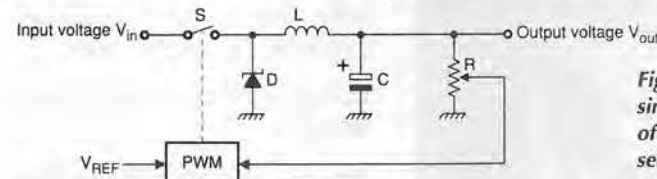


Fig. 1. Switching converter simplified. On and off times of switch S , invariably a semiconductor switch, are controlled by a pulse-width modulator.

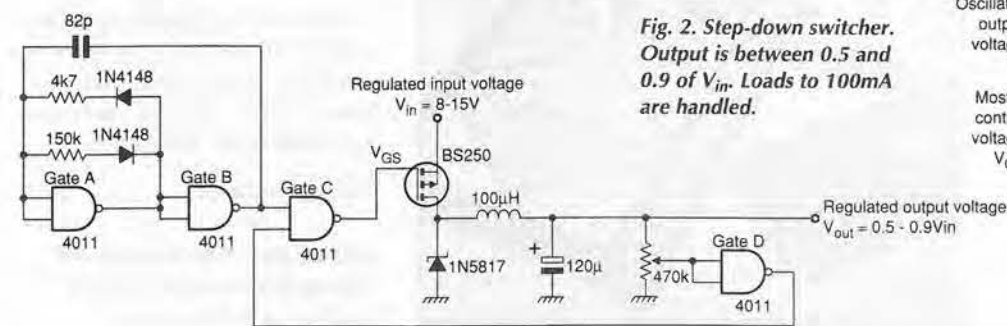


Fig. 2. Step-down switcher. Output is between 0.5 and 0.9 of V_{in} . Loads to 100mA are handled.

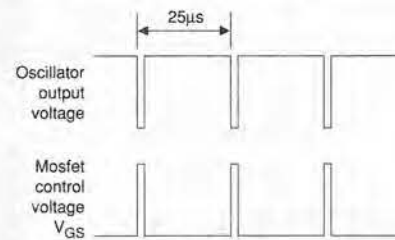


Fig. 3. Asymmetrical oscillator gates A and B of Fig. 2 runs at about 40kHz.

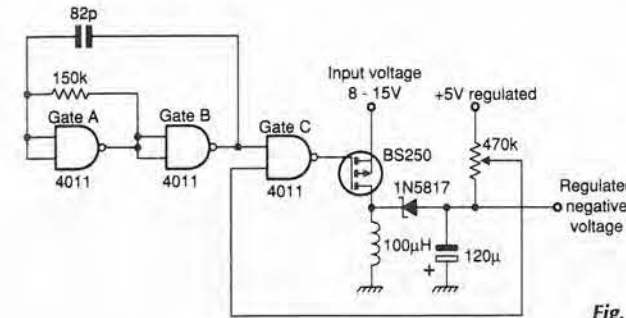


Fig. 4. Negative supply from a positive input. Again, gates A and B are an asymmetrical oscillator. Gate C determines whether drive pulses pass to the switching transistor.

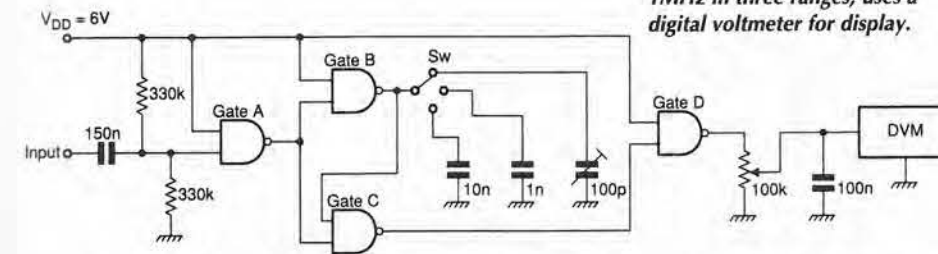


Fig. 5. Direct reading frequency meter, capable of measuring to 1MHz in three ranges, uses a digital voltmeter for display.

with its 200mV or 2V dc voltage range selected. The meter reading can be adjusted by the 100k Ω potentiometer so that the voltage reading directly represents the unknown capacitance.

Capacitance measuring ranges are selected by the switch, which connects a suitable capacitor to gate B's output. The 100pF capacitor should be a trimmer so that gate C's input capacitance can be compensated for. ■

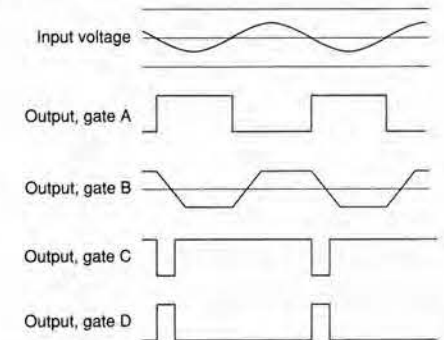


Fig. 6. Frequency meter waveforms. Top is the ac input, which is biased toward the switching point of the logic gate. In this way, the gate switches at a significantly lower input voltage than would otherwise be needed.

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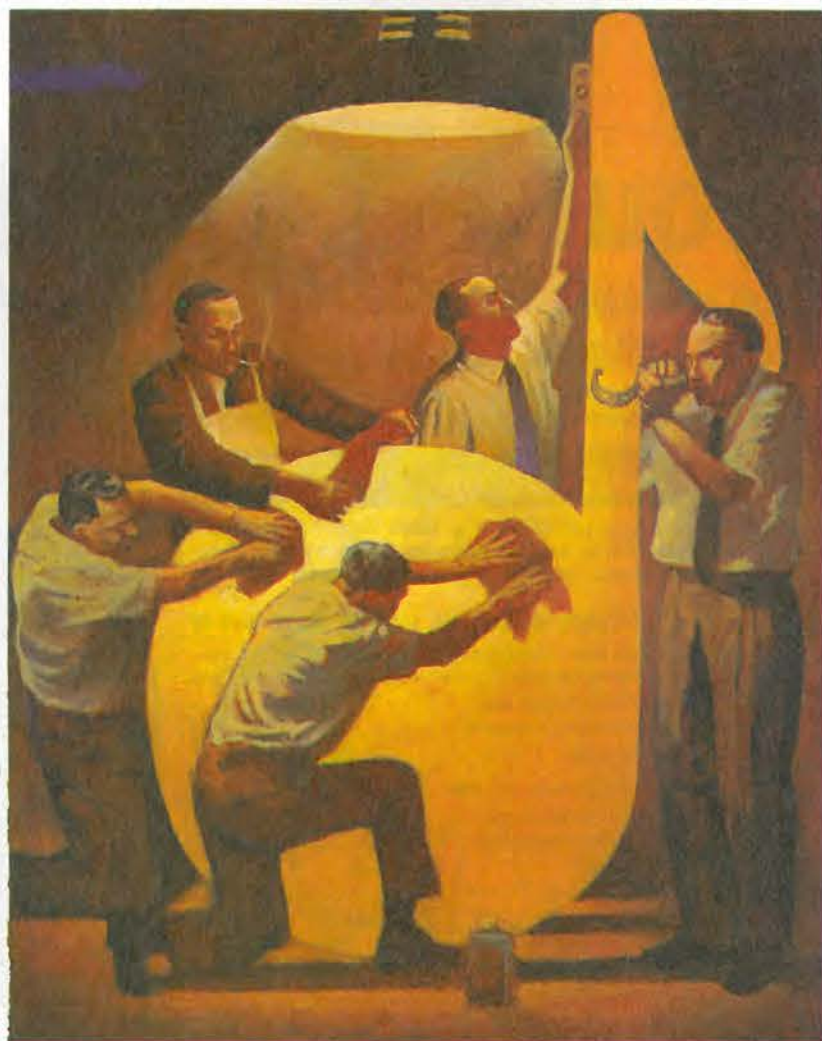
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OVERLOAD matters

Doug Self takes a look at the complex subject of preamplifier overload, and how to handle it.



There was no room in my Preamp '96 article for a proper discussion of the overload behaviour of RIAA preamp stages.¹ Like noise performance, the issue is considerably complicated by both cartridge characteristics and the RIAA equalisation.

There are some inflexible limits to the signal level possible on vinyl disc, and they impose maxima on the signal that a cartridge can reproduce. The absolute

value of these limits may not be precisely defined, but they set the way in which maximum levels vary with frequency, and this is perhaps of even greater importance.

Figure 1a shows the physical groove amplitudes that can be put onto a disc. From subsonic up to about 1kHz, groove amplitude is the constraint. If the sideways excursion is too great, the spacing will need to be increased to prevent one groove breaking into another, and playing time will be reduced. From about 1kHz to ultrasonic, the limit is groove velocity rather than amplitude. If the cutter head tries to move sideways too quickly compared with its forward motion, the back facets of the cutter destroy the groove that has just been cut by the forward edges.

At replay time, there is a third restriction – that of stylus acceleration or, to put it another way, groove curvature. This sets a limit on how well a stylus of a given size can track the groove. Allowing for this at cutting time puts an extra limit on signal level, shown by the dotted line in Fig. 1a.

The severity of this restriction depends on the stylus shape. An old-fashioned spherical type with a tip diameter of 0.0007in requires a roll-off of maximum levels from 2kHz, while a relatively modern elliptical type with 0.0002in effective diameter postpones the problem to about 8kHz.²

Thus there are at least three limits on the signal level. The distribution of amplitude with frequency for the original signal is unlikely to mimic this, because there is almost always more energy at lf than hf. Therefore the hf can be boosted to overcome surface noise without overload problems, and this is done by applying the inverse of the familiar RIAA replay equalisation.

Moving-magnet and moving-coil cartridges both operate by the relative motion of conductors and magnetic field, so the voltage produced is proportional to rate of change of flux. The cartridge is sensitive to velocity rather than to amplitude (and so sensitivity is always expressed in millivolts per cm/s) and this gives a frequency response rising steadily at 6dB/octave across the whole audio band. Therefore, a maximal signal from disc (Fig. 1a) would give a cartridge output like Fig. 1b – i.e., 1a tilted upwards.

Figure 1c shows the RIAA replay equalisation curve. The shelf in the middle corresponds with 1a, while an extra time constant at 50Hz limits the amount

of lf boost applied to warps and rumbles. The 'IEC amendment' is an extra roll-off at 20Hz, (shown dotted) to further reduce subsonics. When RIAA equalisation 1c is applied to cartridge output 1b, the result will look like Fig. 1d, with the maximum amplitudes occurring around 1-2kHz.

Clearly, the overload performance of an RIAA input can only be assessed by driving it with an inverse-RIAA equalised signal, rising at 6dB/octave except around the middle shelf. My Precision preamp '96 has an input overload margin referred to 5mV rms of 36dB across most of the audio band, i.e., 315mV

rms at 1kHz. The margin is still 36dB at 100Hz, but due to the RIAA low-frequency boost this is only 30mV rms in absolute terms.

The final complication is that preamplifier output capability almost always varies with frequency. In Preamp '96, the effects have been kept small. The output overload margin voltage – and hence input margin – falls to +33dB at 20kHz. This is due to the heavy capacitive loading of both the main RIAA feedback path and the pole-correcting RC network ($R_{24,25}$ and C_{20}). This could be eliminated by using an op-amp with greater load-driving capabilities, if you can find one with the low noise of a 5534.

The overload capability of Preamp 96 is also reduced to 31dB in the bottom octave 10-20Hz, because the IEC amendment is implemented in the second stage. The lf signal is fully amplified by the first stage, then attenuated by the deliberately slow initial roll-off of the subsonic filter.

Such audio impropriety always carries a penalty in headroom as the signal will clip before it is attenuated. This is the price paid for an accurate IEC amendment set by polyester caps in the second stage, as opposed to the usual method of putting a small electrolytic in the first-stage feedback path, rather than the 220 μ F used. Alternative input architectures that put flat amplification before an RIAA stage suffer much more severely from this kind of headroom restriction.³

These extra preamp limitations on output level are shown at Fig. 1e, and, comparing 1d, it appears they are almost irrelevant because of the falloff in possible input levels at each end of the audio band. ■

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- Self, D., 'Precision preamplifier '96', *Electronics World*, July/Aug and Sept 1996.
- Holman, T., 'Dynamic range requirements of phonographic preamplifiers', *Audio*, July 1977, p74.
- As [1], July/Aug 1996, p543.

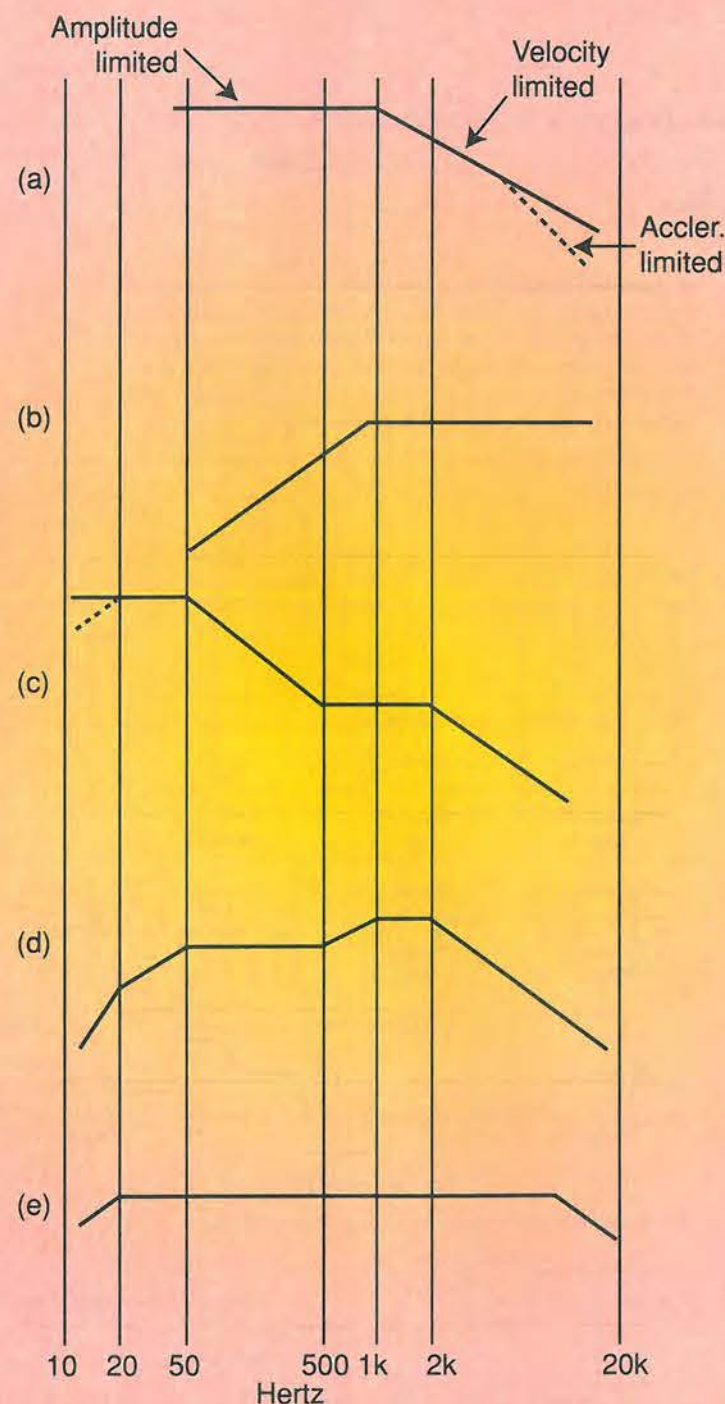


Fig. 1a). Restrictions on the level put onto a vinyl disc. The extra limit of groove curvature – stylus acceleration – is shown dotted. Fig. 1b). Response of a moving-coil or moving-magnet cartridge to a signal following the maximum contour in Fig. 1a). Fig. 1c). The RIAA replay curve. The IEC amendment is an extra roll-off at low frequency, shown dotted. Fig. 1d). The combination of b) and c). Fig. 1e). RIAA preamp output limitations. The high-frequency restriction is very common and is often much worse in discrete preamplifier stages with poor load-driving capabilities.

Programmable logic primer

In this first article based on his new designer's handbook - 'FPGAs and programmable logic', Geoff Bostock provides an introduction to standard logic families.

The primary building block of logic circuits is the logic gate. This is a device which operates on two or more logic signals to give an output which is defined by a logic operator. The standard logic operators are And, Or and Invert, which only acts on one signal.

There are two classical logic families - ttl, which uses a non-final 5V supply, and 4000-series c-mos, which can work from a supply of between 3V and 15V. In recent years, the 4000 series has been largely superseded by the HC high-speed c-mos family. A logic low signal is usually defined as being close to 0V, or ground, while a logic high signal normally sits close to the supply voltage, V_{cc} , or at least above half V_{cc} .

An And function is defined as only giving a high output when all the inputs are high; the Or function has a high output when any one of its inputs is high. The Invert function changes a logic signal from high to low, or vice versa. These functions may be written down as logic equations as follows:

And function: $Y=A*B*C$ or $Y=A\&B\&C$
 Or function: $Y=A+B+C$ or $Y=A\#B\#C$
 Invert function: $Y=!A$ or $Y=\bar{A}$

Alternative notations exist because different logic compilers have adopted different conventions. Using '*', '+' and '/' for logic clashes with their more familiar use as arithmetic operators in programming languages. For the remainder of this article, I will use '&', '#', and '!'.

Figure 1 shows an alternative way of showing logic relationships - the truth table. The symbol 'H' represents logic high, sometimes replaced by '1', while 'L' stands for logic low, with '0' as an alternative. The symbol 'X' represents don't care; the logic level can be high or low.

These three operators form the basis of all possible logic circuits. For example, the exclusive-Or gate has a low output if its two inputs are the same, but high if they are different. Its logic operator is written as $Y=A\$B$ or, sometimes $Y=A:+B$. It is logically equivalent to $Y=A\&!B\#!A\&B$.

These logic functions may also be represented diagrammatically. Standard gate symbols are shown in Fig. 2; as well as And, Or, and Invert, gates with the functions Nand and Nor are also depicted. The NAND function is an And followed by an Invert, Nor is Or followed by Invert; a small bubble on an output, or an input, signifies an inversion.

The exclusive-Or symbol, and its equivalent logic circuit, are drawn in Fig. 3.

Sequential logic

Output of a gate does not depend on the order in which the signals are applied. If both inputs of a two-input And gate are low, the output will also be low; if A goes high before B, or if B goes high before A, the result will be the same - two highs on the inputs yield a high on the output.

A	B	C	Y	A	B	C	Y	A	Y
1	1	1	H	1	X	X	H	1	H
0	X	X	L	X	1	X	H	0	L
X	0	X	L	X	X	1	H		
X	X	0	L	0	0	0	L		
AND				OR				INVERT	

Fig. 1. Logic function truth tables.

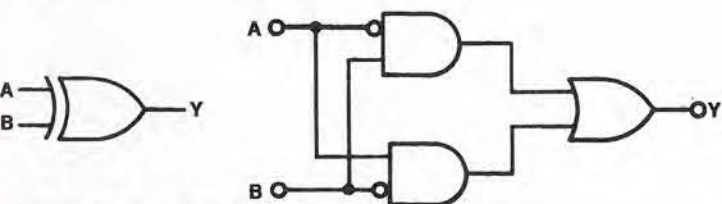
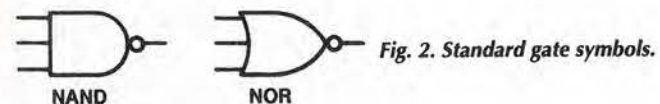
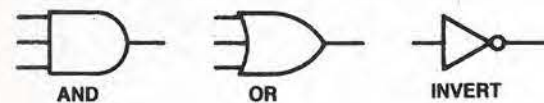


Fig. 3. Exclusive-Or symbol and its equivalent circuit.

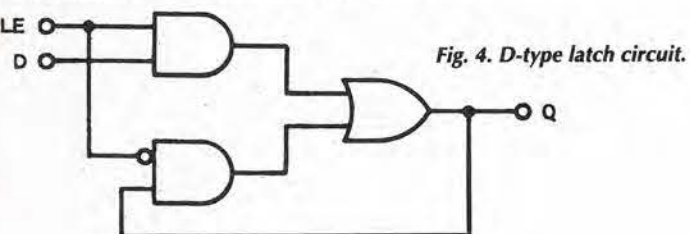


Fig. 4. D-type latch circuit.

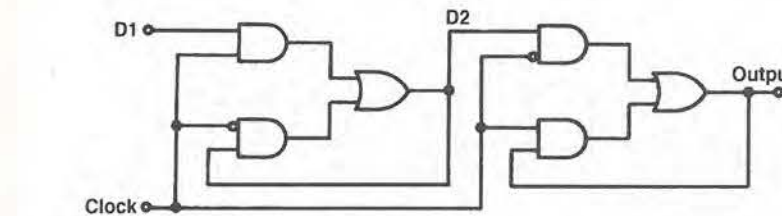


Fig. 5. Master-slave D-type flip-flop.

Fig. 6. One-to-eight line decoder and demultiplexer.

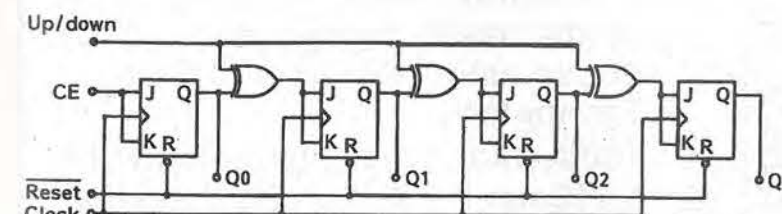
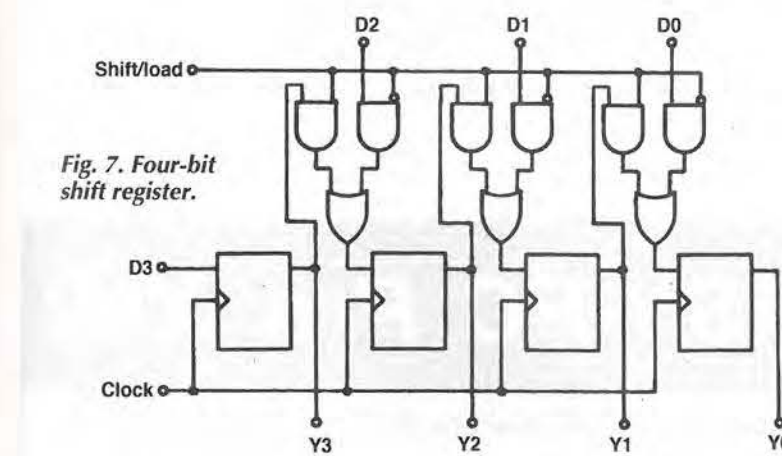
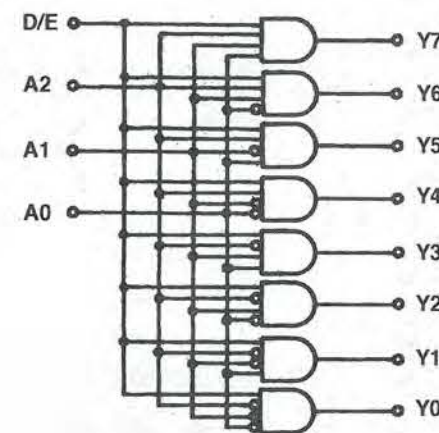


Fig. 8. Four-bit counter.

Table 1. Truth table for decoder/demultiplexer, Fig 6.

D/E	A ₂	A ₁	A ₀	O ₇	O ₆	O ₅	O ₄	O ₃	O ₂	O ₁	O ₀
0	X	X	X	L	L	L	L	L	L	L	L
1	0	0	0	L	L	L	L	L	L	L	H
1	0	0	1	L	L	L	L	L	L	H	L
1	0	1	0	L	L	L	L	L	H	L	L
1	0	1	1	L	L	L	L	H	L	L	L
1	1	0	0	L	L	L	H	L	L	L	L
1	1	0	1	L	L	H	L	L	L	L	L
1	1	1	0	L	H	L	L	L	L	L	L
1	1	1	1	H	L	L	L	L	L	L	L

Consider Fig. 4. If input LE is high, the output, Q, will be the same as input D. This is because the output of the lower And gate is low - irrespective of D. Suppose that LE is now taken low; the upper gate now has a low output so Q will follow the output of the lower gate. If D was high when LE went low, Q was also high, so the lower gate was high and Q will stay high. Conversely, if D was low, Q will stay low.

This function is known as a latch. While LE is high, the latch is transparent and Q follows D; when LE is low, the level of D when LE went low is latched into Q. Output of the circuit depends on the sequence in which the signals are applied, hence the term sequential circuit.

Figure 5 shows two latches in series, with the LE signal inverted between them. When the first latch is transparent, the second is latched, and vice versa. A signal applied to D₁ will appear at D₂ while LE₁ is high, but will not be transmitted to Q until CLK goes low, sending LE₂ high. At this time, LE₁ goes low and locks out any changes on D₁. It appears that the signal on D₁ is sent through to Q as CLK changes from high to low. This is the principle of the master-slave flip-flop or D-type flip-flop.

Practical logic circuits

Devices containing one or more gates, latches or flip-flops form the basis of the standard logic families. Circuit designers can use these integrated circuits to build up more complex functions by interconnecting these ssi, or small-scale integration, devices on a printed-circuit board. However, the device manufacturers anticipated these requirements by producing medium-scale integration, or msi, parts containing many of the standard circuit functions which can be built from gates and flip-flops.

A typical combinatorial msi function is a one-to-eight line decoder/demultiplexer. The circuit diagram for this function is shown in Fig. 6. The logic levels on the three address inputs represent a binary number in the range 000₂ (0₁₀) to 111₂ (7₁₀). The logic level on the data/enable input is transmitted to the output selected by the input address. The truth table for this function is shown in Table 1.

Similarly, msi functions can be built from sequential elements. Figure 7 illustrates this point with the circuit of a four-bit shift register. Data on the inputs is loaded into the flip-flops when the shift/load signal is low and there is a low-to-high clock transition. When shift/load is high, the data is moved one place to the right on a clock edge. This circuit can form the basis of some arithmetic functions, or it may be used in communications to change the data format from parallel to serial or vice versa.

A different type of flip-flop, the J-K flip-flop, is used in the counter circuit in Fig. 8. A J-K flip-flop behaves like a D-type when the J and K inputs are complementary, but if both are high, the output changes from high to low, or low to high, the so-called toggle function. If both are low, no change occurs on a clock edge; the output data is unchanged, the hold condition. As long as the count enable, CE, and reset inputs are held high, the counter function will increment by one on each clock transition.

Taking the reset low makes every output low, resetting the counter to binary 0000₂. Reset is an asynchronous function which is built into some flip-flops; it operates independently from the clock and allows the flip-flop to be put into a known condition. Some flip-flops also include an asynchronous set input which puts the output high.

Operation of the counter may be described by a state diagram. Each combination of output levels corresponds to a binary number from 0000₂ (0₁₀) to 1111₂ (15₁₀), and represents a distinct state of the counter. Figure 9 shows the counter state diagram. Each state is represented by a circle containing the output levels in that state. The arrows are labelled with the logic inputs which enable the jumps between states. A recursive arrow means that there is no jump and gives the hold

condition for that state.

The CE must be true for counting to proceed; if CE is false, the counter holds its present state.

Many sequential functions are best described by a state diagram, just as combinatorial functions are often defined in a truth table.

In the next article, Geoff takes a more in-depth look at programmable logic elements.

This article is derived from Geoff Bostock's new book 'FPGAs and Programmable LSI - a designer's handbook'. This work covers designing FPGAs, large PAL structures, RAM and antifuse-based FPGAs and FPGA

selection. Comprising 215 pages, this book is available by sending a postal order or cheque with a request for the book to Electronics World, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. The fully-inclusive price is £27.50 UK, £30 Europe or £33 rest of world. Alternatively, fax your full credit card details and address on 0181 652 8956 or e-mail jackie.lowe@rbp.co.uk.

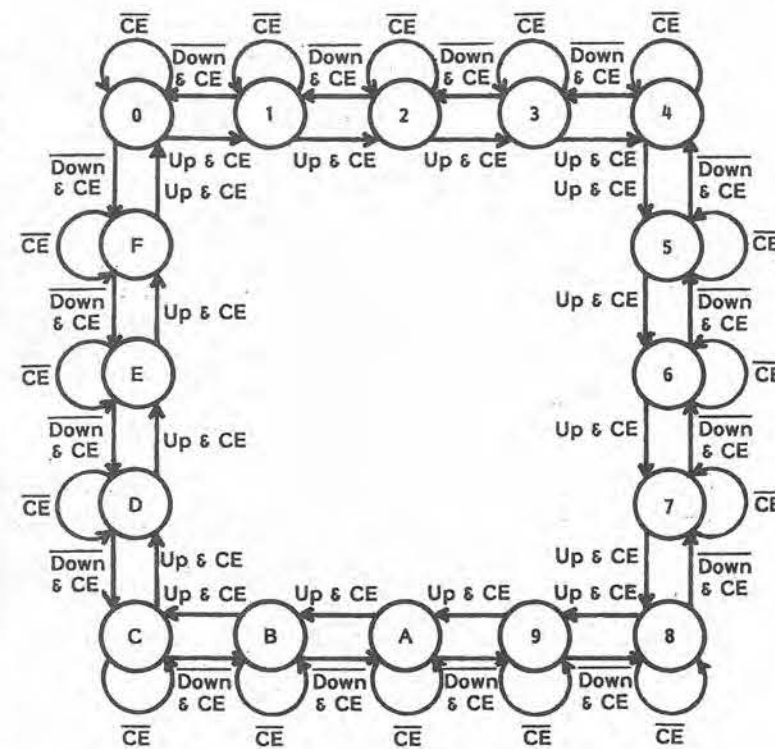


Fig. 9. Four-bit counter state diagram.

Designing power supplies

Ray Fautley describes an easy-to-use procedure for designing reliable full-wave rectifiers of the centre-tapped secondary variety.

In this version of the full-wave rectifier, the bridge rectifier is simplified to two diodes, but the transformer needs a centre tap, which becomes the ground connection.

Alternating voltage is applied to rectifier diodes $D_{1,2}$ where it is rectified and the output smoothed by reservoir capacitor C .

Fundamental frequency of the ripple voltage is twice that of the supply frequency. Resistance R_s represents the source of the supply and V_{sec} is the voltage across the whole of the secondary winding.

Design procedure

- 1) Specify required dc output voltage at full load $E_{dc(load)}$ (V).
- 2) Specify required maximum load current $I_{dc(load)}$ (A).
- 3) Specify maximum voltage ripple acceptable $V_{r(rms)}$ (V).
- 4) Specify the ac mains supply voltage $V_{pri(rms)}$ (V).
- 5) Specify the frequency of the mains supply f (Hz).
- 6) Determine the value of the equivalent load resistance R_L :

$$R_L = \frac{E_{dc}}{I_{dc(load)}}$$

where E_{dc} is the design value of the dc output voltage. It is the required voltage across the load, $E_{dc(load)}$, added to the voltage drop across one of the diodes. As the voltage drop across the diodes occurs only while they are conducting, and they conduct alternately, the effective drop is that of just one diode.

$$E_{dc} = E_{dc(load)} + V_{rec}$$

where V_{rec} (the drop across the rectifier diode) is 0.9V, so:

$$R_L = \frac{E_{dc(load)} + 0.9}{I_{dc(load)}}$$

- 7) Determine average current through each diode. Half the average current, I_0 , will flow through each diode.

$$I_0 = I_{dc(load)}/2$$



- 8) Determine a value for source resistance of the supply, R_s . If the mains transformer winding resistances are known - and it rarely is - refer to step 8 in the design procedure the full-wave rectifier (September 1996 issue) for the method of evaluating R_s . Otherwise, assume that R_s is about 5% of R_L . Then for low resistance loads:

$$R_s = \frac{R_L \times 5}{100} + \frac{0.9}{I_0}$$

For high resistance loads, where the transformer winding resistance predominates:

$$R_s = \frac{R_L \times 5}{100}$$

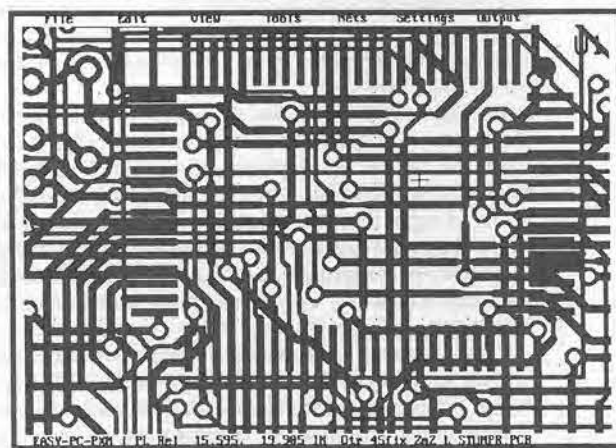
- 9) Calculate the ratio of R_s to R_L as a percentage:

$$\frac{R_s}{R_L} \times 100\%$$

- 10) Determine percentage ripple voltage from the specified maximum ripple and the dc output voltage:

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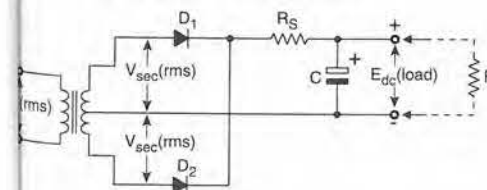
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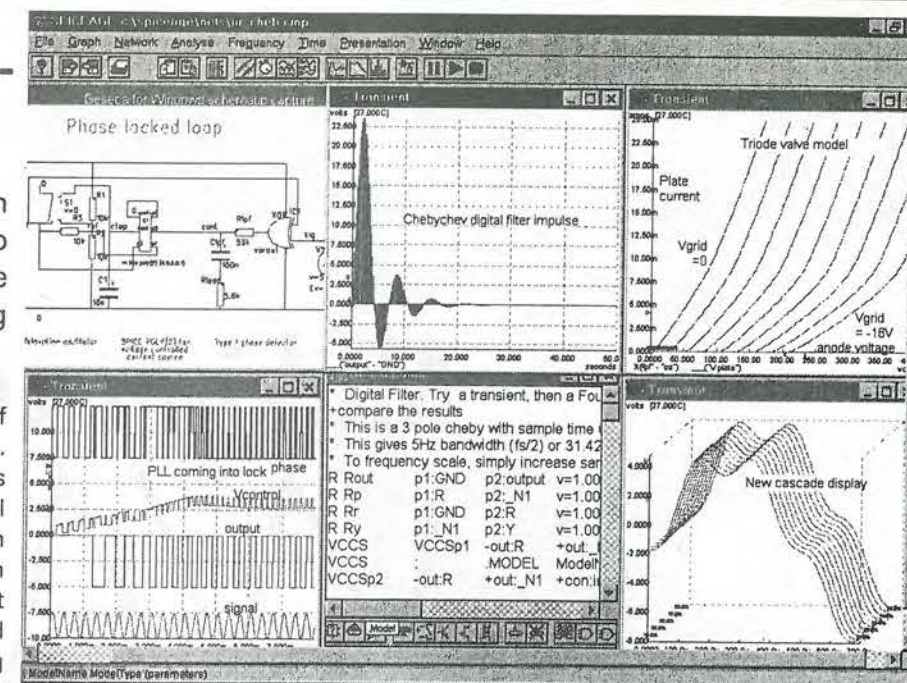
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Making continuous WAVES

With the competition for the first patent in wireless telegraphy won by Marconi, a new race began to emerge, to develop radio transmissions into a widely used medium of communication. Few guessed that it would result in the explosion of radio as we now know it. By Tom Ivall and Peter Willis

Marconi may have secured the world's first patent on wireless telegraphy (*Electronics World*, June 1996) but others were already working on similar systems.

Braun in Germany, for instance, inventor of the cathode-ray oscillograph, developed a spark transmitter in 1898 and had it in service for short-range marine communication by 1900. Later, he and Marconi were to share the 1909 Nobel Prize for Physics in recognition of their radio work.

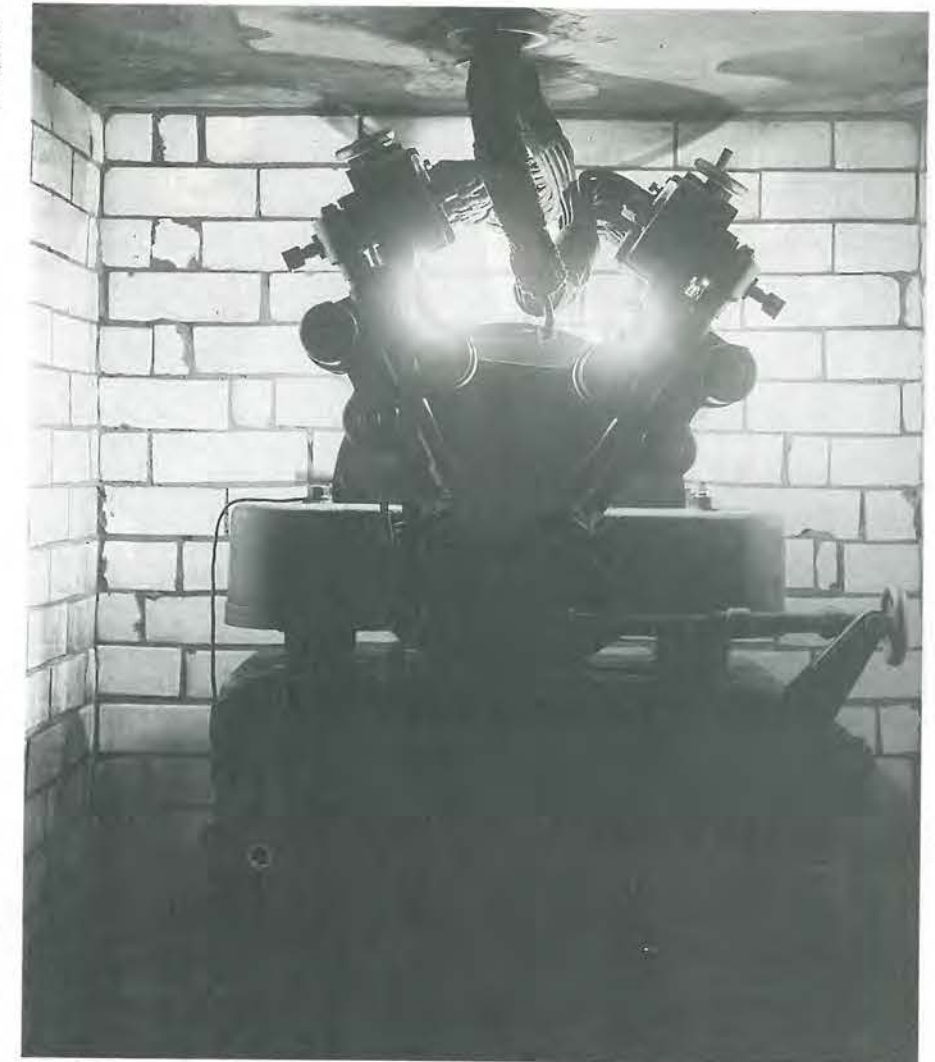
Braun and some business partners formed a wireless company called Telebraun, though this soon became a subsidiary of Siemens. At about the same time, two other German pioneers, Slaby and von Arco, supported by AEG, had set up a competing firm, Funkentelegraphie. In 1903, these two enterprises merged to form Gesellschaft für Drahtlose

Telegraphie and the result was the Telefunken system, a formidable rival to that of Marconi. By 1905, it had built more than 500 stations.

Meanwhile, in the USA, Fessenden and de Forest, separately, and the partnerships of Branly-Popp in France and Lodge-Muirhead in the UK, were also developing and operating spark transmitter systems of wireless telegraphy. All were striving for greater communication ranges through more highly powered transmissions and more sensitive receivers. They also wanted much better selectivity, to allow multi-station working without mutual interference.

One way to get more radiated power from a spark transmitter was to increase the oscillatory energy in the LCR circuit containing the spark gap. To do this, each high-voltage pulse had to get as much charge as possible into the capacitor. But the induction coil was restrict-

Part of a rotary spark transmitter at Poldhu, Cornwall, as seen, it is claimed, while actually transmitting news of Britain's declaration of war on Germany in 1914.



ed as a generator because it could inject only short-duration pulses of charge into a low-value capacitor.

So for higher-power transmission the induction coil was soon replaced by the engine-driven alternator, with an output of several kilovolts, plus a step-up transformer providing the necessary tens of kilovolts. This would produce as many as ten sparks, and hence damped wave-trains, per cycle of the alternator output.

As well as producing bigger charges, the alternator system could be designed to generate a relatively high spark frequency, thereby shortening the interval between wave-trains and increasing the average oscillatory energy. And this would give a musical note in receiver headphones which was more easily heard through static (atmospheric) interference than the rasping sounds produced by induction coils.

Radiated power was also increased by tighter inductive coupling between the oscillator and aerial circuits. But to achieve this, the spark had to be quenched rapidly, otherwise the interaction between the coupled circuits would cause beating and radiation on two component frequencies. This spark quenching also had the advantage of reducing arcing in fixed spark gaps. Rotary spark gaps or dischargers, driven from the alternator shafts, had the double purpose of preventing arcing and producing very regular, audible wave-train frequencies.

In the receiver, the coherer was too insensitive for long-range working and was difficult to operate. New detection techniques were invented and put into use, including anti-coherers, electrolytic and thermal principles, magnetic effects in an iron or steel medium, and rectification at contacts between crystals or between metals and crystals. Magnetic detectors proved robust in operation and were widely used.

In 1904, Fleming invented the thermionic diode, based of course on the Edison Effect,

which had been known since the late 19th century. The diode, however, did not immediately supersede the earlier devices but was used as a standby detector. Two years later, de Forest put in a third electrode, the grid, and produced the thermionic triode, or Audion as he called it. It was intended as a triggering form of detector but was not particularly successful as such. Only after years of development, on the device itself and on circuit applications, did the triode become really useful – but as an amplifier and oscillator. Subsequently, the non-linear parts of the amplifier characteristic were used for detection.

Even the earliest improvements in transmitters and receivers produced good operating results. The most dramatic was Marconi signalling across the Atlantic in 1901-02. Apart from the commercial implications, this supported early ideas of a reflecting layer and Heaviside's and Kennelly's 1902 prediction of the ionosphere, though the existence of the various layers was not fully confirmed till the 1920s. Transatlantic wireless telegraphy was also achieved by Fessenden, between Brant Rock, Massachusetts, and Macrihanish, Scotland, in 1906, and the following year Marconi started a full commercial service with stations at Clifden, Ireland, and Glace Bay, Canada.

By this time, radio was already well estab-

lished for marine communications. It had started in 1897 with fixed links such as those between the mainland and islands, lighthouses, lightships and moored vessels, and progressed to mobile communications with small ships such as ferries. The use of Morse code was universal.

During the first five years of the new century, large passenger vessels and warships were being equipped, and by 1911 wireless was compulsory on all ocean-going liners. In this first decade, several thousand lives were saved at sea through the use of radio.

Even before the wireless telegraph was fully established as a universal service, engineers were thinking about wireless telephony. They knew they needed continuous waves in place of damped wave-trains, and a means of modulating them with voice frequencies.

A short-lived proposal was to speed up the spark repetition rate to something above audible frequencies. But this idea was soon overtaken by two techniques which proved successful for nearly a decade, before the arrival of the triode valve oscillator made them obsolete. These were the multipole rf alternator, developed by Tesla and Alexanderson, and the Poulsen oscillatory arc, based on Duddell's 'singing arc' of 1900. The arc conduction process in hydrogen gas interacts with a series LC



Picture courtesy Barratts

David Sarnoff, who, in 1916, elaborated on Lodge's early suggestion for broadcasting and outlined the main technical features of the domestic receiver, or "radio music box" as he called it. Then a manager at the American Marconi Company, he rose to become chairman of its successor, RCA. He is seen here on a visit to London in or about 1925.



Picture courtesy Marconi

Music recital in 1922 from the 2LO broadcasting studio in London. Miss Olive Sturgess sings into a carbon-granule telephony hand microphone fixed in a laboratory-style clamp, raised to the required height on a decorative pot stand. Also singing is Mr John Huntingdon, while the accompanist in spats is Mr R Stanton Jefferies.

circuit shunted across the electrodes, so that the dc flowing through the arc varies in a periodic manner at rf.

The alternator, typically working at 30kHz, was the more stable and controllable generator, while the arc was smaller and cheaper. But both had the problem that they couldn't be voice-modulated at low level – the carbon granule microphone had to be inserted in the rf power circuits, either directly in series or by inductive coupling. This, of course, limited the rf power that could be generated and hence the transmitter range.

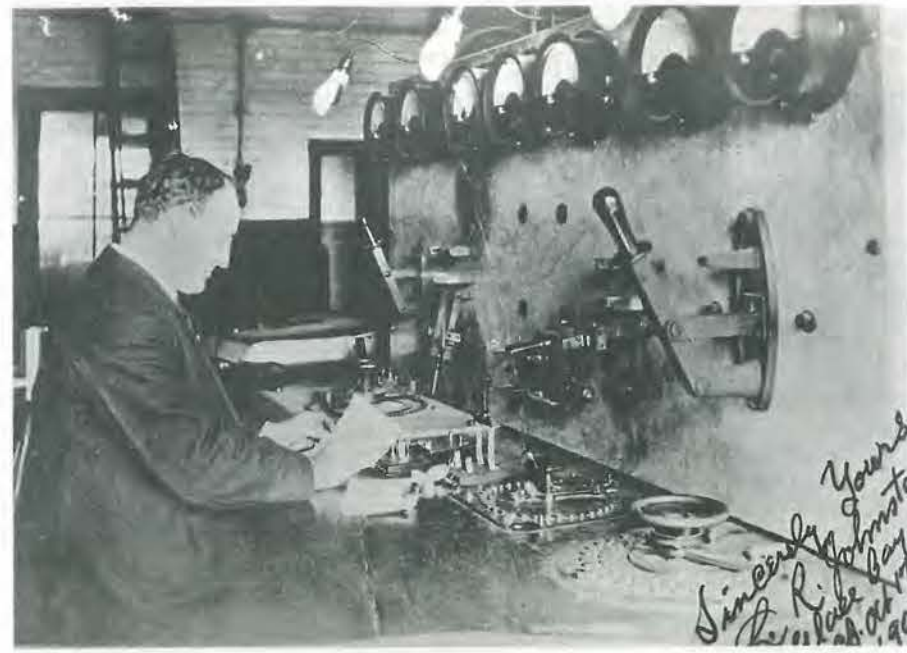
Nevertheless, in 1902 voice transmission was achieved, by Fessenden, and in 1906 demonstrations of radiotelephony were given almost simultaneously in the USA, Germany and Denmark. The advantages of voice communications over telegraphic signals were quickly recognised, particularly by the armed forces, and warships were the first to be equipped.

Civil applications were not far behind, and these were not limited to speech communication. After music had been transmitted on an rf alternator system in 1906 and Caruso had sung over a radio-telephone in 1910, thoughtful people began to realise that the new technology could do somewhat more than convey messages from point to point. After all, music concerts and lectures had already been distributed to subscribers over wired systems, the precursors of cable radio, in the 1890s.

Under the stimulus of the 1914-18 World War, the high-vacuum triode valve was developed into a robust and reliable device, capable of being manufactured in quantity. It now had a good amplification factor and, as a result of

Meissner's 1913 positive feedback circuit invention, had emerged as a new means of generating continuous oscillations, though at low frequencies.

Radio-telephone transmitters using valve oscillators began to appear. In 1913-14, speech and music were transmitted from Marconi House in London, and in 1915 the Atlantic was spanned by speech from a naval sta-



The North American end of Marconi's transatlantic radio-telegraphy link was at Glace Bay, Nova Scotia. This signed memento shows operator L R Johnstone transmitting the inaugural message in October 1907, when the commercial service started.

tion at Arlington, USA, to the Eiffel Tower in Paris.

Triode valves were also coming into receivers. The idea of cascading several valve stages to give higher amplification appeared about 1912. Then Armstrong and others showed how the positive feedback principle could be used to reinforce the signal and improve selectivity – the technique of regenerative amplification, later called reaction. Fessenden had demonstrated the heterodyne principle in 1902 (already discovered acoustically as beats or 'resultant tones' by the 18th-century musician Tartini) and this eventually resulted in Armstrong's famous invention of the super-sonic heterodyne or superhet receiver.

The wideband fm which Armstrong eventually developed in the 1930s was the outcome of his and others' early studies of the ever-present static interference on am radiotelephony.

So thermionic valves became the basis of radio-telephony. The extension of point-to-point radio-telephony into broadcasting, already foreshadowed by the isolated experiments mentioned above, came about almost accidentally. Oliver Lodge was the first to suggest the idea of messages being 'broadcast to receivers in all directions', in 1907. Then in 1916 an employee of the American Marconi Company, David Sarnoff, not only predicted broadcasting more firmly but suggested its main elements: transmitting stations and what he called a "radio music box" for the home.

"This device must be arranged to receive on several wavelengths with the throw of a switch or the pressing of a button," said Sarnoff. "The radio music box can be supplied with amplifying tubes and a loudspeaking telephone, all of which can be neatly mounted in a box."

The plan was still under consideration when

the USA entered the World War in 1917. After the war, American Marconi was merged with General Electric at the insistence of the US government, which required wholly US ownership. The resulting company, with a rich crop of wireless patents to exploit, was the Radio Corporation of America (RCA), and Sarnoff was eventually to become its chairman.

Regular broadcasting actually started from the US station KDKA in Pittsburg. In 1919, Westinghouse, while testing the range of a radio-telephone transmitter, decided to play some gramophone records of music as an alternative to speech. By this time, wireless enthusiasts were building crystal sets. Many of them liked what they heard and wrote to the company asking for more.

As a result, and after a well appreciated broadcast of election results, Westinghouse started regular programmes in 1920, with the idea that KDKA would create a market for radio components and receivers – which in fact it did. In the same year, regular broadcasts began in Europe from The Hague.

In the UK, the accidental birth of broadcasting came about in a similar way. At the Marconi company, the standard method of range-testing transmitters was to read out the names of railway stations from timetables. These were eagerly received by numerous amateur enthusiasts on home-built sets. To

mark the introduction of a new transmitter in 1920, something less mind-numbing was devised – a concert featuring staff and local artistes, with three instrumentalists and two singers.

Despite appreciative letters from hundreds of miles away, the company persisted in its view that the future of radio lay in speech transmission, and that the purpose of broadcasts was to demonstrate the comparative ease with which messages could be exchanged. Accordingly, it substituted news transmissions for the concerts.

In June 1920, however, the firm was persuaded to transmit from Chelmsford a recital by the operatic soprano Melba. One consequence of the success of this event – gramophone recordings were made of the reception at the Eiffel Tower – was that the Post Office withdrew Marconi's experimental licence on the grounds of "interference with legitimate services".

This happened just as commercial broadcasting and the mass production of receivers were beginning to take off in the USA. It was not until 1922 that the Post Office relented and permitted Marconi's to transmit at low power, within its existing half-hour-a-week, a 15-minute programme of speech and music.

The station set up to exploit this opportunity was housed in a wooden hut at Writtle, near Chelmsford. Its call sign was 2MT and

the cheerful informality of its amateur entertainments immediately endeared it to its listeners. Transmission, initially on 700m and later on 400m, was from a 250ft-long four-wire aerial at a height of 110ft.

A second permit quickly led to another station, the more famous 2LO, being established at Marconi House, London. Transmissions, still nominally demonstrations, were permitted first at 100W then later at 1.5kW for one hour daily on 360m.

Before 1922 was out, this had become the British Broadcasting Company, set up at the Post Office's instigation by a consortium of manufacturers including Marconi. Revenue came from a 10% levy on all receivers sold, plus half of the 10-shilling (50p) receiver licence.

Regional stations were set up quickly. The 2LO London station was reconstituted as purpose-built studios at 2 Savoy Hill and a 6kW transmitter on the roof of Selfridge's department store.

The BBC parted company from the set-makers in 1926 and in 1927 became a public-service organisation, established under Royal Charter as the British Broadcasting Corporation.

Throughout all this, spark transmissions continued to be allowed and were not internationally prohibited until 1940. ■

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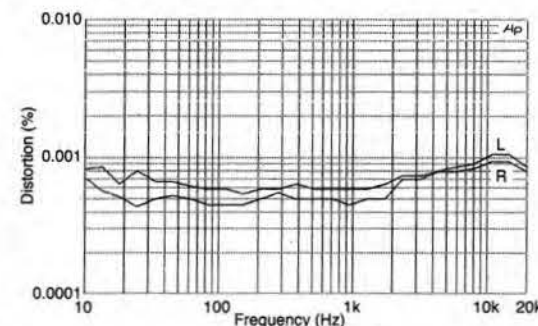
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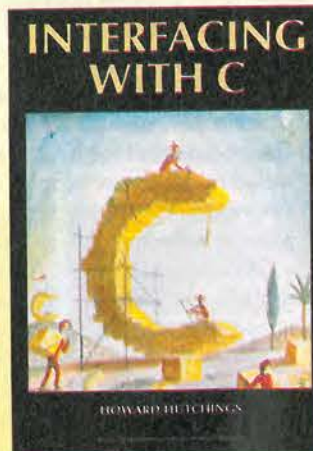
Howard Hutchings

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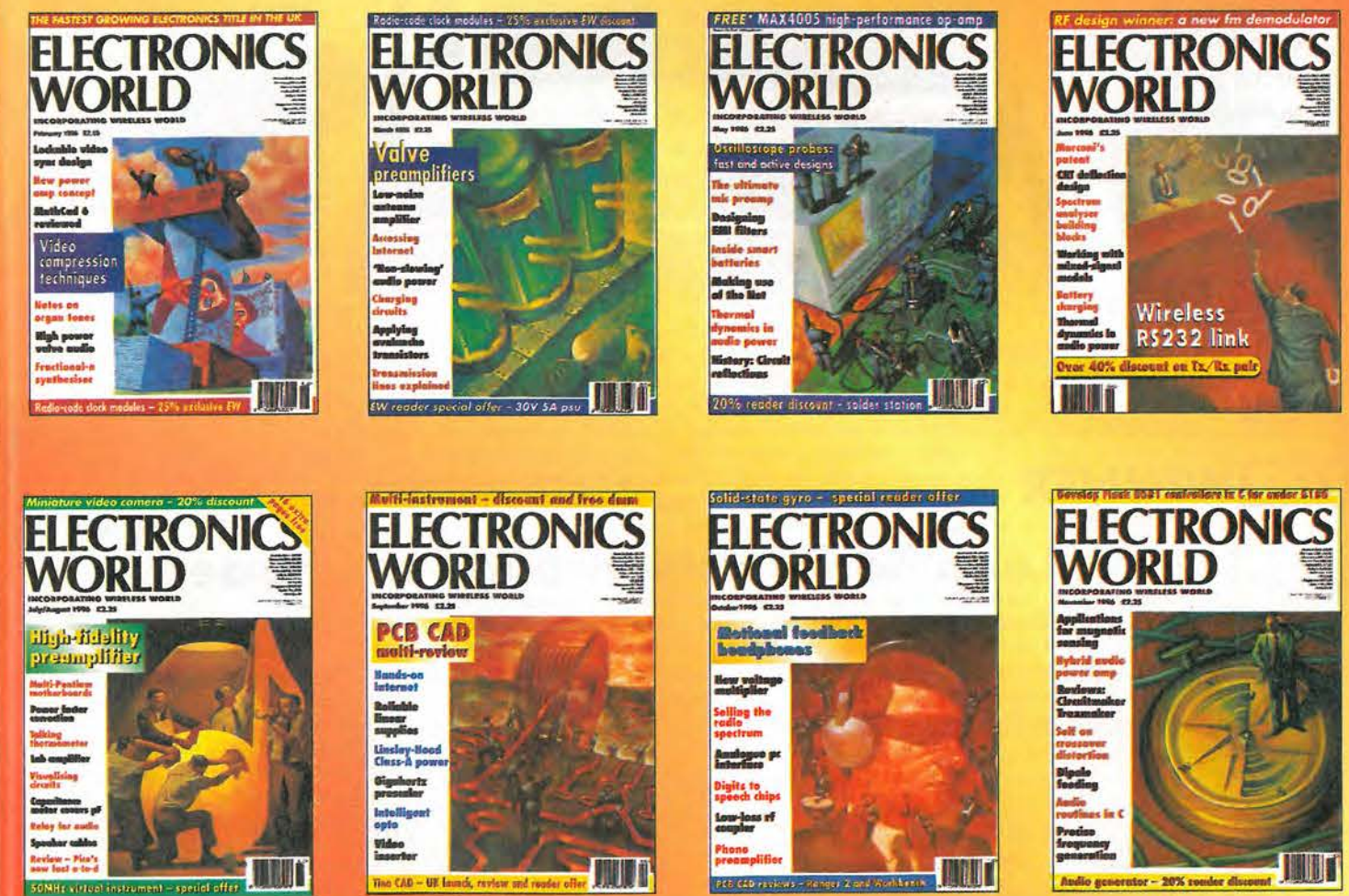
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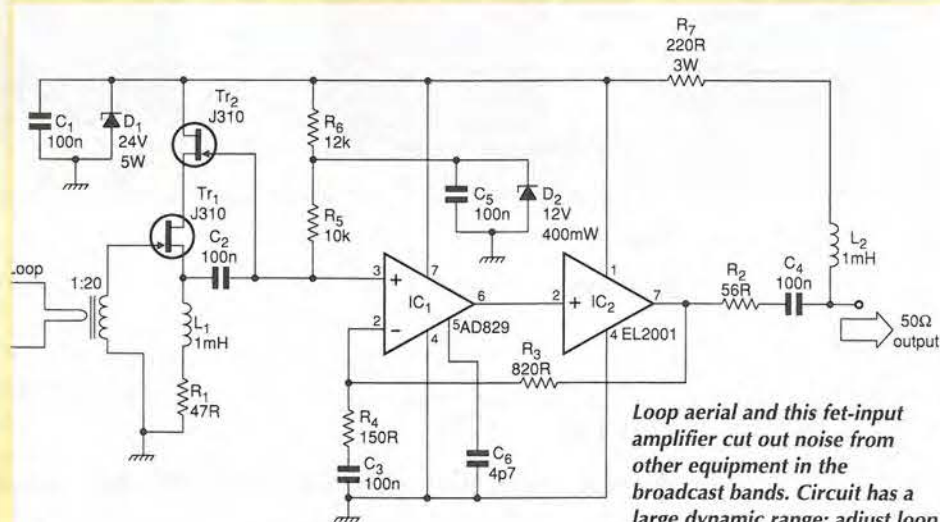
amplifier, working with a 1m square maximum loop, reduce noise to the background level of the bands. The business end of the circuit

is formed by the input transformer and fet source follower; at long- and medium-wave frequencies, fets show low noise figures at 10kΩ source impedance. Transistor Tr_2 bootstraps out the gate/drain capacitance of Tr_1 , the gate/source capacitance being low due to the follower configuration.

Maximising input transformer ratio while keeping shunt capacitance low results from the use of a toroid (Cirkit 55-40001 or Fair-Rite 26-43540001) with two primary turns of audio screened cable with the screen grounded at one end, and 40 on the secondary.

The op-amps form a low-noise amplifier driving a 50Ω cable and the other components form a phantom power supply, although a local supply could be used, in the 25-40V range.

J A Burnill
Camberley, Surrey



Loop aerial and this fet-input amplifier cut out noise from other equipment in the broadcast bands. Circuit has a large dynamic range; adjust loop size to give required sensitivity.

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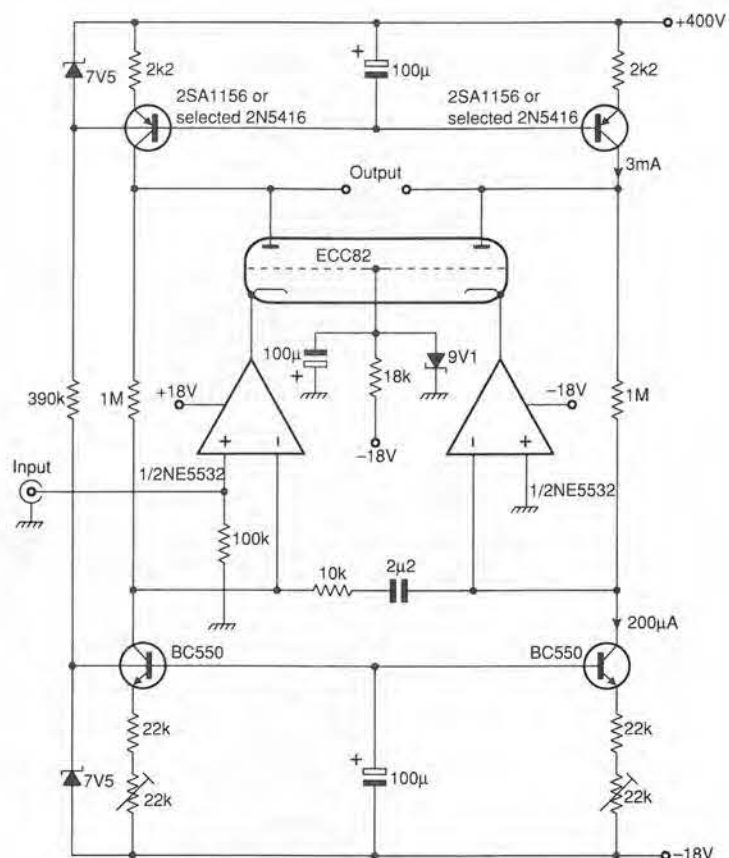
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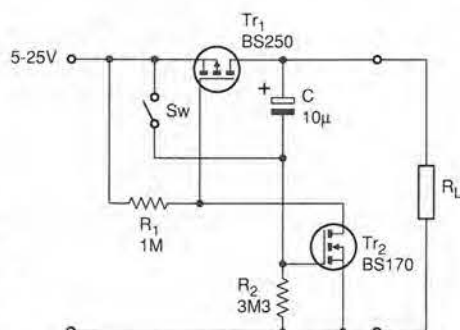
A high-voltage audio output to drive an electrostatic headphone comes from a double-triode stage, itself fed by op-amps. The whole thereby combines the robustness of valves and the high gain of op-amps. Common-grid drive to the triodes is the chief peculiarity, chosen to allow the output from the op-amps to be summed for the output and to exploit the greater stability of the configuration over the more usual common-cathode drive – all without loss of bandwidth. Current sources supply triode loading and carry all the current, the output therefore being protected against short-circuits. Further current sources for bias avoid the need for a split supply; trim for half the 400V on each output. Output is 200V rms into 200kΩ – or greater from 1V rms input, although gain can be altered by varying the 10kΩ feedback resistor. **Paolo Palazzi** Cervignano Italy

High-voltage audio for headphone drive. Variations include a differential input using the non-inverting input of the right-hand op-amp and the use of bigger triodes, with an adjustment in bias voltage.

Simple time-out saves batteries

The circuit described here can help you avoid the problem of drained cells in a battery-powered device by breaking the current off after a certain time, determined by an RC-circuit. The circuit is very simple, with only a few components. Transistor Tr_1 , which breaks the battery current, is a BS250 p-type enhancement-mode mosfet. When power is turned on, its gate is connected to a negative potential through Tr_2 – a BS170 n-type enhancement mosfet. Transistor Tr_2 turns on when voltage across its gate is positive. This voltage comes from an RC circuit formed by capacitor C and the resistor R_2 . At turn on, the capacitor has no charge. During operation, it is charged through R_2 . Gate voltage of Tr_2 goes down as the capacitor charge

increases. When the gate voltage reaches the enhancement value of the transistor, it switches off and can no longer supply gate voltage to Tr_1 . Pull-up resistor R_1 connects its gate to source potential and Tr_1 breaks the current. The circuit is released by momentarily closing the switch. The capacitor is discharged and Tr_2 has its gate voltage again. Tr_2 switches to on-state and gives gate Tr_1 a voltage, making it conductive again. Operating voltage of the circuit ranges from 5V to 25V. The device is well suited for use with common 9V batteries. Operating time of the circuit is approximately twice the time constant R_2C of the circuit, component values shown giving an on time of around a minute. Different values of the enhancement voltage of Tr_2 also influence on the operating time. Component values need to be selected if accurate timing is needed. Transistor BS250 has an $R_{DS(on)}$ -value of approximately 4Ω, causing a voltage drop in the circuit when loaded. Loading current should not exceed 50mA. Larger loads can be handled by using a more robust transistor instead of BS250. For instance, IRF9530 can easily reach loading currents up to 2A, if desired. Operating time can be lengthened at any time by simply closing the switch. The capacitor loses its charge and the operating time is renewed to its starting value. **Rae Perälä** Helsinki Finland



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The modifications include four additional circuit boards providing *Rechargeable memory and clock back-up *Balanced Audio line output *Reduced AM distortion *Buffered IF output for monitoring transmitted modulation envelope on an oscilloscope *Mains safety improvements.

The receiver is available in free standing or rack mounting form and all the original microprocessor features are retained. The new AM system achieves exceptionally low distortion: THD, 200Hz-6kHz at 90% modulation –44dB, 0.6% (originally –20dB, 10%).

*Advanced Active Aerial 4kHz-30MHz *PPM10 in-vision PPM and chart recorder *Twin Twin PPM Rack and Box Units *Stabilizer frequency shifters for howl reduction *10 Outlet Distribution Amplifier 4 *Stereo Variable Emphasis Limiter 3 *Stereo Disc Amplifiers *PPM5 hybrid, PPM9 microprocessor and PPM8 IEC/DIN –50/+6dB drives and movements *Broadcast Stereo Coders.

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Hands-on Internet

Following an update on searching the net, Cyril Bateman describes his latest Spice discoveries – among them fully working free evaluation packages.

Internet search engines described in my past columns can be used to locate many different information sources. They are also frequently used for transferring computer software programs, either by an FTP client or your Web browser.

If you know the exact file name you need, FTPSearch in Norway can almost completely automate transfer of files for you. However, if you do not know the file name, the main task becomes one of file name identification.

While Archie, discussed in the April 1996 issue, can perform searches using wildcard characters, and Wais can also help, many users who now rely almost totally on Web browser access do not have these packages. Search engines such as Alta Vista, using appropriate keywords, can be successful, but correctly identifying unknown software file names can prove difficult even for experienced Web surfers.

David Agbamu's home page on Demon¹ is dedicated to helping to solve exactly this problem. By identifying suitable search methods and providing dedicated links, the task is simplified. His page simply and effectively combines the essential information covered in the relevant FAQ documents with direct access to Internet sources – all by simply using your Web browser. David's section on using e-mail for file transfers is particularly helpful, Fig. 1.

If you need to identify a UK or European Business and you prefer to use the conventional Yellow Pages telephone direc-

Fig. 1. A well planned route to Ftp file name identification.

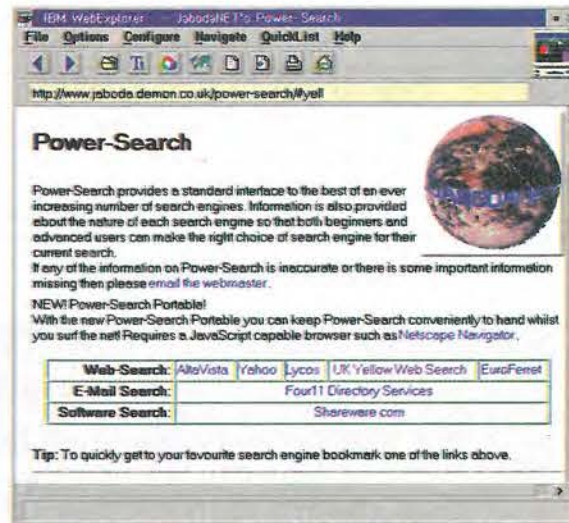
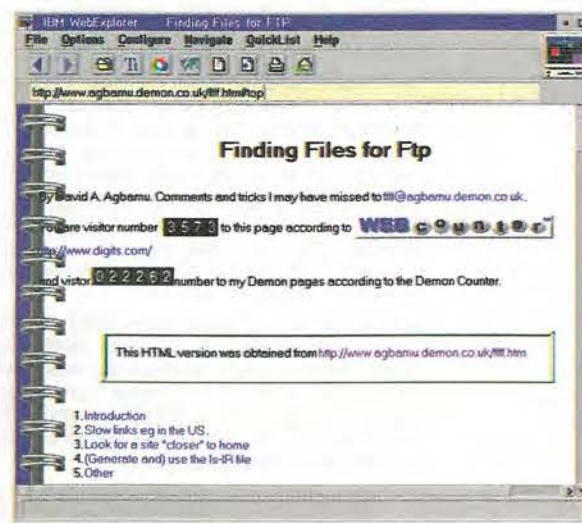


Fig. 2. Choose the most appropriate search method – use Power-Search – a uniquely English approach to Internet searches.

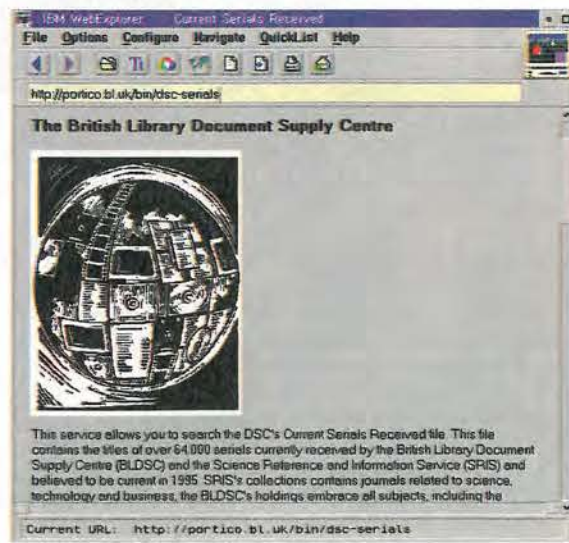


Fig. 3. The British Library Periodicals Holding on-line service. Deeper level searches are chargeable.

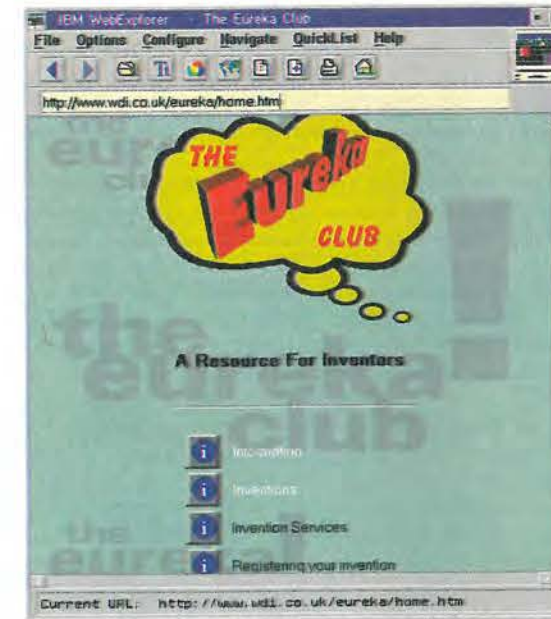


Fig. 4. On-Line Inventors Club. Mix and Match for inventors and manufacturers.

tory approach, Freepages.co.uk² or UK.Yellow Web³ may provide the answer. If not, a good alternative starting point for all Web searches is Power-Search⁴, also on Demon. This page gives a background description for each of the search engines listed, permitting more logical selection as well as direct access, Fig. 2.

Should you need to search for previously published information, the British Library Document Supply Centre⁵ and the Science Reference and Information Service hold more than 64,000 serial publications. These comprise journals on science, technology, business and general interest topics, with on-line searching from the British Library Page. Having established that your required document is in these holdings, it can be obtained using the Lexicon easy-order service or, within the UK, rather more economically by request to your local branch library, Fig. 3.

Perhaps you have a new design or invention, but lack the resources to market your idea. The Eureka Club⁶, while acting as an essential support for inventors needing assistance, also forms an on-line meeting place for designers with a new product but no production resources and manufacturers looking for their next marketable product ideas, Fig. 4.

Simulation software

Symptomatic of the explosion in the use of computer simulations for electronic circuit design, the numbers and variety of simulation software packages constantly increases. While this market remains dominated by derivatives of the Berkeley Spice 2G6 system, newer packages have emerged based on the latest Berkeley Spice 3F5 software core.

An interesting document by Filip Gieszczykiewicz, called 'Where to get Free Spice', can be found on his page at Paranoia.Com⁷. This up-to-date but rather lengthy paper, sub-divided by operating system, gives a good overview of the low-cost Spice-based systems available.

The basic Berkeley Spice software kernel is in the public domain, and therefore inexpensive. Many commercial packages however, being enhanced versions of the basic Spice core, and having improved input and output systems, can become quite expensive.

Spice simulations are only as accurate as the the models used in the simulator. While most analogue integrated circuit

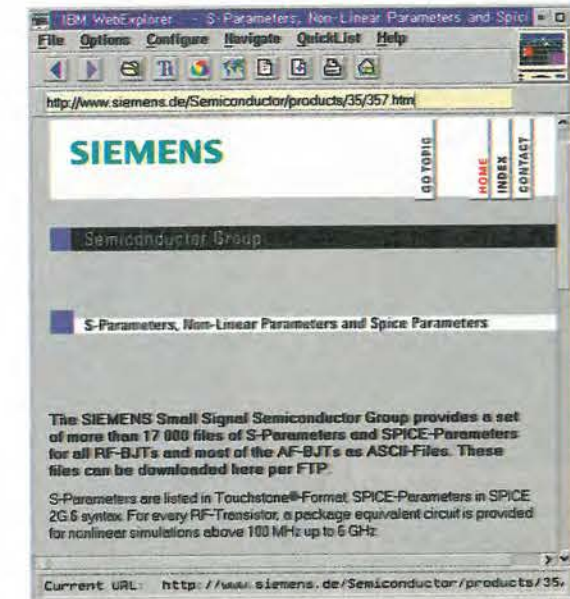


Fig. 5. Siemens Semiconductor Group's extensive model library. Modelling from dc to 6GHz and both Spice and 'S' parameters.

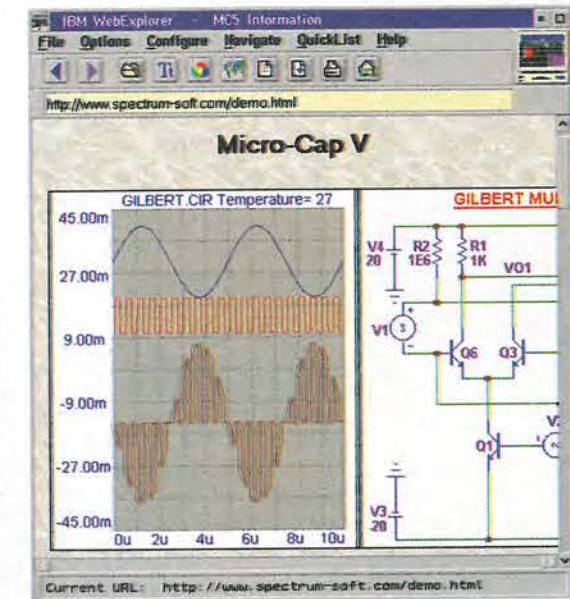


Fig. 6. Download your free evaluation demonstration version of Micro-CapV, a well established alternative pc simulator.

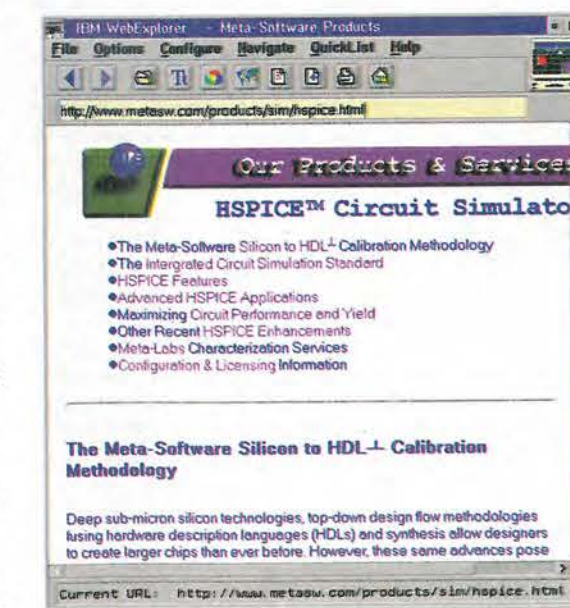


Fig. 7. Serious simulation at an appropriate price.

Fig. 8. One of the new-generation simulators. It costs nothing to try so why not download this software?



Fig. 9. If you are curious about the different variations of the basic Spice kernel, you should read the 'Free SPICE' document and the 'SPICE 3F2' document both by Filipg@paranoia.com.

makers provide free models for their products, models for discrete devices are generally only provided by the simulator software package. Contributing significantly to their cost is the extended model library now required, since older discrete devices must still be supported while new discrete devices continue to be added.

Spice models

One major commercial provider of Spice models is Symmetry⁸, part of Interface Technologies. Its *SymLib* library contains more than 7500 analogue and mixed-signal devices of proven accuracy.

Siemens Semiconductor Group⁹ manufactures both radio and audio-frequency discrete semiconductors. For these, the company offers more than 17,000 files of S-Parameters for downloading in Touchstone format, plus Spice-parameters in Spice 2G.6 format.

When I visited International Rectifier's site¹⁰ in April last year, I only found fifth-generation MosFet models, but a recent return visit provided models also for the company's older generation products.

Comlinear, now part of National Semiconductor, offers Macromodels for its signal-conditioning product line, which can be downloaded from National's pages¹¹. These Spice Macromodels are available either by individual device number or as a complete, self-extracting archive library for all types.

Simulator engines

Spectrum Software's¹² fifth-generation *Micro-Cap* simulator, *Micro-Cap V*, unlike earlier versions, runs under Windows

and is compatible with standard Spice model libraries. Note, however, that older S3 video card drivers and early Hewlett Packard printer drivers may need to be updated before running *Micro-Cap V*. A free 1.4Mbyte student/demo version can be downloaded from the company's Web page. It shows the result of simulating a Gilbert Multiplier to illustrate the use of *Micro-Cap V*, Fig. 6.

The Meta-Software¹³ version of Spice - *HSpice* - claims to be the most accurate commercial circuit simulator available, with the MetaMOS1 (level 28) transistor model giving faster, more accurate simulations. *HSpice*, which is available integrated into major simulation frameworks from Cadence, Viewlogic, Zuken and Mentor, is targeted to the design of silicon integrated circuits including ASICs, as well as more conventional circuit design needs, Fig. 7.

SIMetrix v1.1 is a Spice-based simulator with schematic editor for Windows 3.x available from Newbury Technical¹⁴. A free, no-time-limit version, fully working except for user definable menus, etc., called *SIMetrix intro*, can be downloaded from the company's page¹⁴. While some support for this free version is available by e-mail, those users downloading *SIMetrix intro* are advised to print out the Known Bugs page, Fig. 8.

This month's final Internet simulation offering is *AIM-Spice V2.0*¹⁵, also available for download in a free student version. This package was developed to provide a more user-friendly interface and take advantage of newer device modelling. It is based on the Berkeley Spice 3.E1 kernel and fully described in the book 'Semiconductor Device Modelling for VLSI', published by Prentice Hall (ISBN 0-13-805656-0), Fig. 9.

Readers curious about the different variations of the basic Spice kernel should read the Free Spice document and the Spice 3F2 document⁷, both by Filipg@paranoia.com. ■

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13. Meta-Software Incorporated http://www.metaw.com/products/sim/hspice.html
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Continued use of the phase/frequency comparator's charge pump system looks questionable in view of the increasing demands for spectral purity, argues Edward Forster.

Phase comparator purity

The phase/frequency comparator is one of the most widely used components in phase-locked loop technology. It is applied in countless applications, increasing by the day as more radio-oriented products appear.

Although the basic logic within the phase/frequency comparator is simple and well understood, the output interface to the analogue world has several variations.

The charge pump

The most popular output circuit is the charge pump system comprising transistors Tr_1 and Tr_2 , Fig. 1. Disregarding the logic for the moment, it is only necessary to know that when the phases are synchronised, the output on the up/down lines consists purely of short duration pulses, normally coincident, which occur as the comparator resets in every cycle.

The duration of the reset pulses only depends on propagation delays in the logic. These can be very short compared to the reference clock period. Resulting output for these

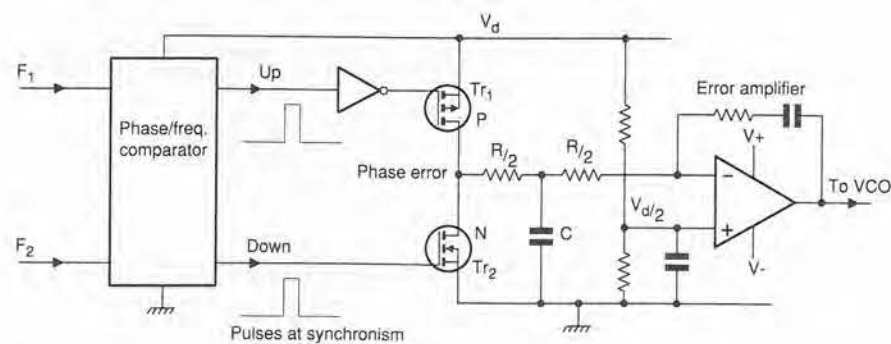


Fig. 1. Conventional phase/frequency comparator with charge-pump output.

short pulses is highly dependent on the matching of the transistors.

A perfectly complementary combination would probably avoid some of the variations in comparator gain which occur near zero error. This type of output circuitry is tri-state with the third state being a high impedance state. Presumably, this fits in with the fact that the logic also has three stable states. The fourth state, in which both up and down lines are high, is inhibited by reset.

The output logic circuitry sounds more like a digital engineer's idea of analogue design. However, the main outcome is that at phase synchronism the output is essentially in the high impedance state almost continuously except for a momentary clamping of the output ideally to $V_d/2$.

Figure 1 shows a typical error amplifier and active filter for this approach which has to be protected against the fast pulses, usually by splitting the input resistance and adding a capacitance to ground. But the extra delay due to the filter $R/2, C$ must not be made so large as to affect loop stability.

The capacitor associated with the charge pump is the integrating capacitor in the active filter and not this C . The dc reference for the amplifier is $V_d/2$ so that the loop will settle at zero phase error. Reference frequency suppression is then at a maximum and the gain of the comparator is $V_d/4\pi$ volts/radian.

The source impedance seen by the op-amp is sometimes a critical factor in determining the intrinsic noise of the error amplifier and through the loop, noise on the voltage-controlled oscillator, or vco. That is, the vco may have higher close-in noise sidebands than expected or desired.

The differential output, Fig. 2, shows another

er arrangement in which a differential error amplifier is driven by the up/down logic outputs directly so as to subtract them. The fast coincident pulses then become a common mode problem for the amplifier which only additional RC pre-filtering will satisfactorily resolve. Note that the nominal output/common mode voltage at phase synchronism is either very close to V_d or ground, depending on the logic polarity. This is often inconvenient in single supply systems.

Overall noise performance tends to be improved by the higher gain, $V_d/2\pi$ V/rad, and because there is no high-impedance state.

The resistive combiner

A better approach, which does not rely on the op-amp as a subtractor, is shown in Fig. 3. It is simply to take the up line in Fig. 1 and add it to the inverted down line in a 1:1 resistive network.

At phase synchronism the pulses disappear in the output which is a dc voltage of nominally $V_d/2$; the exact value depends on the high/low saturation voltages of the logic. As before, this is only true when a zero error control loop is used and adjusted correctly. The comparator gain is $V_d/4\pi$ volts/radian. Figure 3 shows that – in principle at least – infinite reference suppression is possible without filtering and that the interface is inherently suitable for wide band applications. In practice however, additional RC filtering is still necessary in front of the error amplifier.

Noise performance of the op-amp can be adjusted by setting the resistors R and the total input resistance to optimum values for the device. At phase synchronism, the output effectively shunts the supply line with a constant resistance of $2R$. Since this resistance may at times be quite low, the extra current drain must be considered. It may be seen as a price worth paying.

See it work

Typical discrete logic would use standard D-type bistable devices, namely positive edge-triggered 7474s with 'clear on low' inputs.

While it is possible to illustrate the waveforms, there is no better way to appreciate the circuit than to make one and test it. The simplest method is to take a signal generator, feed one input directly, and the other via a 100m of RG58, giving a delay of 500ns. By varying the frequency, all possible phase errors can be generated and the response seen.

While the comparator has a nearly -360 to $+360^\circ$ linear range you will find that it also has a 360° phase ambiguity, which depends on the initial conditions. As a result, it is not useful as an absolute phase comparator but it does excel in phase-locked loop applications.

Differential resistive combiner

Figure 4 shows a differential comparator which allows the op-amp to reject common-mode noise arising from the supply line, V_d . This comparator also has twice the gain ($V_d/2\pi$ volts/radian) of Fig. 3, which means that the effective op-amp noise is halved when

considering its relative effect on the noise sidebands of the voltage-controlled oscillator.

Lock detection is also shown and a complete comparator can be made with just two standard ics. Extended frequency range comparators for special applications are thus very simple indeed.

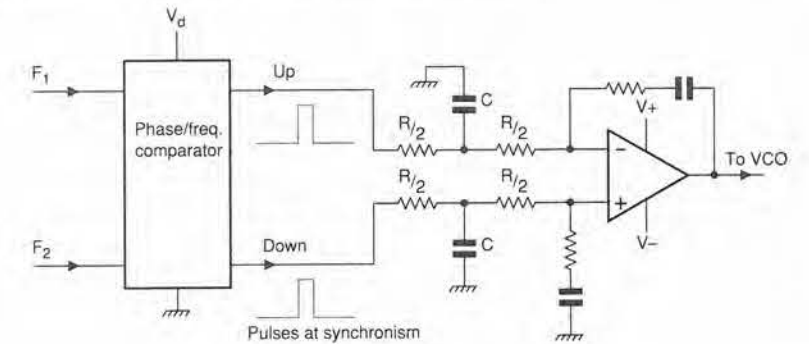


Fig. 2. Using a differential output with the conventional phase/frequency comparator improves noise performance.

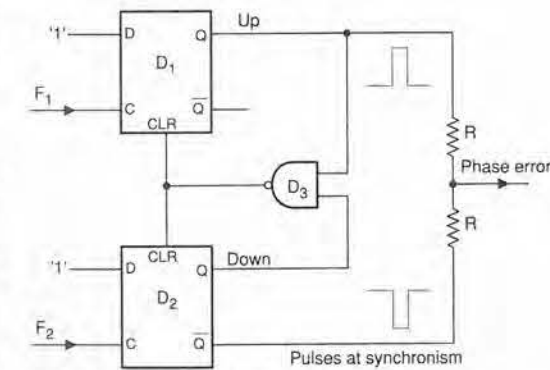


Fig. 3. Phase/frequency comparator with resistive combiner is an improvement over Fig. 2's differential output. In theory, infinite reference suppression is possible.

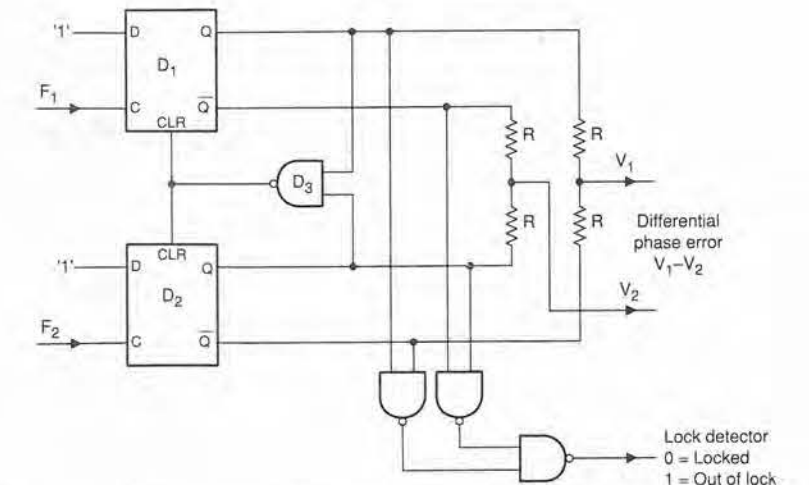
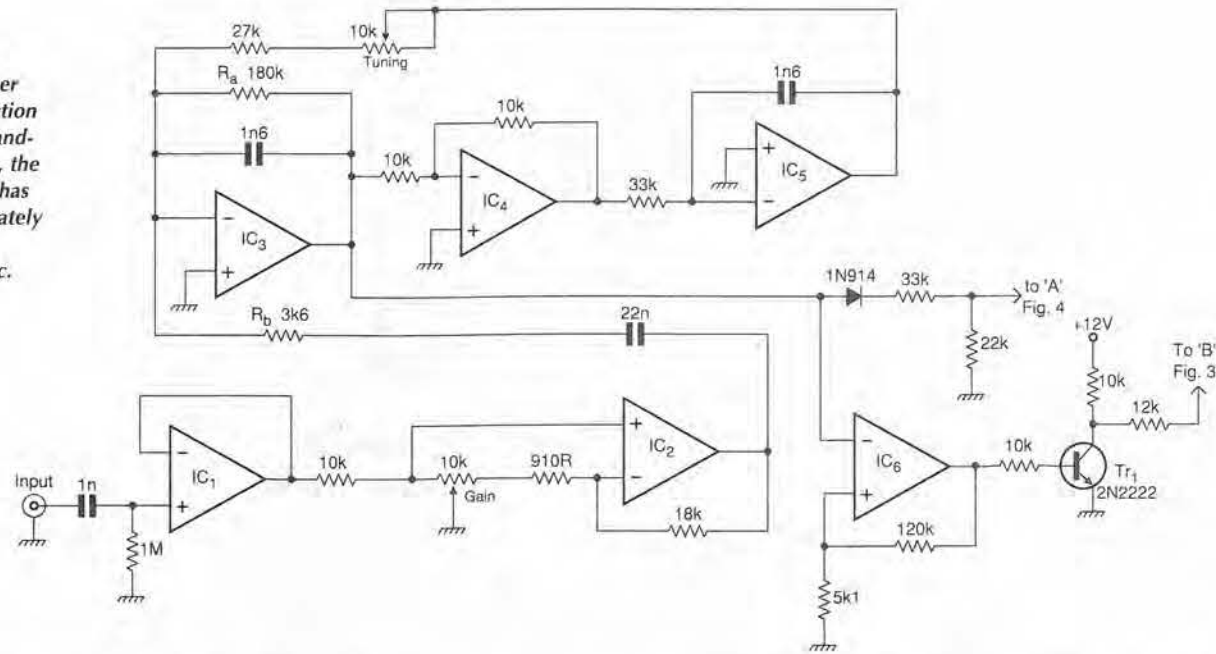


Fig. 4. Differential phase/frequency comparator allows common-mode supply noise rejection and reduces noise due to its higher gain.

Fig. 2. In the input amplifier and filter section of the wow-and-flutter meter, the gain control has an approximately exponential characteristic.



not so high that the biquad filter is overloaded. After S_1 in Figs 3, 4 & 5 is switched to the 'set' position, the biquad is tuned for maximum meter reading and the gain is adjusted for a reading anywhere in the top two-thirds of the meter scale.

The squared carrier signal is sent to the 4046 pll, IC7, Fig. 3. Demodulated output from the pll goes to two filter chains, selected by S_2 , to give either a flat or a weighted bandpass (IC8,9 or IC8,10,11).

Op-amp IC12 provides appropriate gain for 0.1%, 0.3%, or 1.0% full-scale meter deflection; range is selected by S_3 . The 180k resistor at S_{1A} causes the gain of the IC12 circuit to yield rms-reading average measurement when S_1 is set to 'average'.

The feedback resistors around IC12 are

calculated for a meter movement on which the full-scale values are 10dB apart - i.e., 1.0 on one scale and 3.16 on the other. Use of the less common movement, on which the full-scale values are 1.0 and 3.0, would require recalculation of the resistors. Output is provided for monitoring the detected, filtered signal with an oscilloscope or a spectrum analyser.

The detected signal is sent to a full-wave rectifier consisting of IC13,14, and associated parts, Fig. 4. When S_1 is set for average readings, the low impedance output of IC14 charges and discharges C_a through R_c so that the voltage across C_a is the average value of the rectified signal.

Two series diodes at IC15 limit the voltage to which C_a can be charged in off-scale conditions such as switching transients. This minimises the

settling time once normal conditions have returned. When S_1 is set for peak readings, IC15 and its feedback diode form a low impedance charging source for C_a through R_d but no discharge path; the sole discharge path is then through R_c .

Time constants

Charge and discharge time constants for average readings and the discharge time constant for peak readings are 7.3 seconds; this somewhat long time constant gives stable readings even for the 1.8s period of 33rev/min wow. The charge time constant for peak readings is 36ms, which is short enough for a significant response to only a few cycles of a 180Hz flutter signal.

Reduce R_d to decrease the peak-reading time

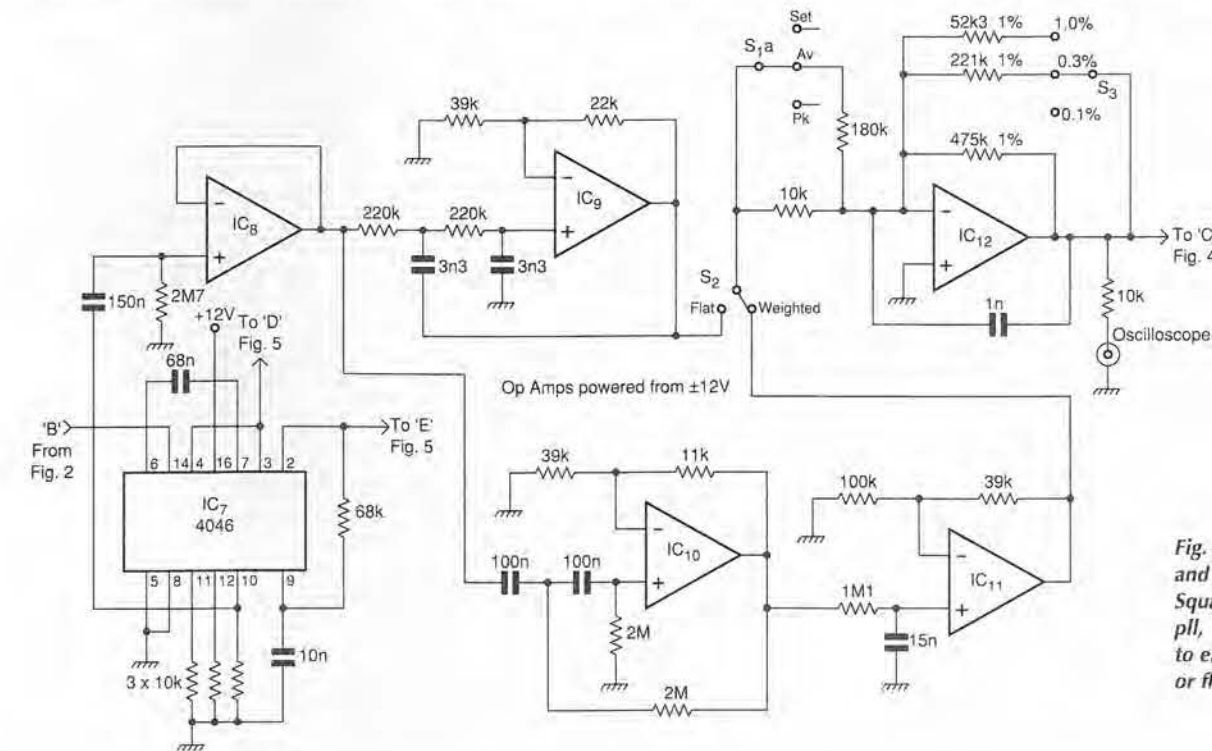


Fig. 3. Phase-locked loop and weighting filter. Squared input feeds the pll, whose output passes to either the weighted or flat filter.

constant at your own risk. Doing so will drastically increase sensitivity to impulse noise.

Charge voltage of C_a is sensed by IC16, which drives the meter movement. The movement, one that happened to be on hand, has a 200µA full-scale sensitivity and a resistance of 1k3Ω. Standards associations specify meter ballistics, but I chose not to attempt to duplicate these specifications electronically. Instead, I subjectively determined a driving impedance for the movement such that the resulting ballistics appeared close to those of a borrowed professional wow-and-flutter meter.

The movement's open-circuit ballistics are on the sluggish side, so a rather high driving impedance proved satisfactory. The resistors associated with IC16 could be changed to accommodate virtually any reasonable meter movement.

Circuit Fig. 5 serves two purposes. First, the logic circuitry detects the lock condition of the pll unambiguously. I wanted the led at Tr_2 to be completely extinguished when the pll is not perfectly locked and to be lighted without flicker when the pll is locked. I also wanted it to follow the state of lock virtually instantaneously. If this situation can be accomplished directly from the 4046 without external logic, I haven't discovered how.

Secondly, Tr_3 discharges C_a summarily when the pll falls out of lock, preventing wild swings of the meter pointer when the carrier is switched in or out or when the instrument is switched on. This function, plus the series diodes at IC15 in Fig. 4, keeps the meter tamed.

The power supply in Fig. 6 is straightforward. A separate +12V supply for the pll is necessary to prevent 3kHz spikes originating in the pll from leaking into the rest of the circuit.

Calibration options

Several options exist for calibration. When a 200µA, 1.3kΩ meter movement is used, the simplest is merely to set the calibration potentiometer, Fig. 4, to the centre of its range and forget it. Although absolute accuracy will suffer, the instrument will be perfectly satisfactory for almost all uses. Should you find this approach unacceptable, you have at least three alternatives for accurate calibration.

In the first method, the demodulator-filter sensitivity is measured, and a signal generator is used to inject a simulated flutter signal of amplitude corresponding to 1% into the amplifier preceding the meter rectifiers. Temporarily short the 150nF capacitor at IC8 of Fig. 3. Connect a stable audio generator set to 3kHz to the system input. With S_1 switched to 'set', peak the tuning control and adjust the gain control for a reading in the top two-thirds of the meter scale.

Monitor the generator frequency with a counter and the voltage at the output of IC9 with a digital voltmeter. Vary the generator frequency around 3kHz and tabulate frequency versus the IC9 output voltage. The change in voltage divided by the corresponding percentage change in frequency is the sensitivity of the demodulator and the IC9

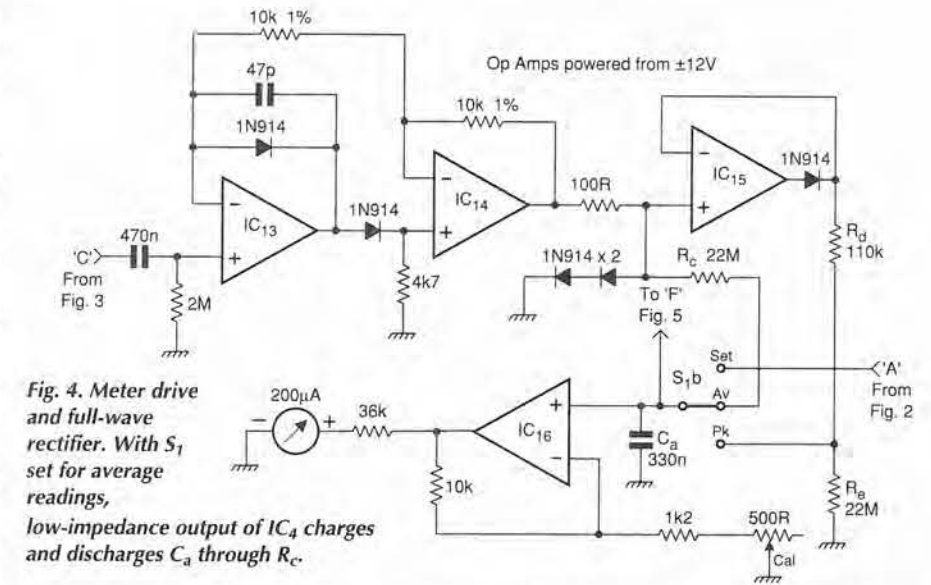


Fig. 4. Meter drive and full-wave rectifier. With S_1 set for average readings, low-impedance output of IC4 charges and discharges C_a through R_c .

lowpass filter. In my case this was 0.18V per 1.0% frequency deviation.

Remove the short from the 150nF capacitor. Temporarily ground Tr_3 's base, Fig. 5, and disconnect the pole of S_2 from the pole of S_{1A} in Fig. 3. Inject a 100Hz or so signal with an amplitude corresponding to 1.0% peak deviation, for example 0.18V peak, into the pole of S_{1A} . Set S_1 to 'peak' and S_3 to '1.0%'. Adjust the calibration pot for a 1.0 reading. If 3kHz and 3.15kHz are not well within the capture range of the pll, trim the 68nF capacitor at IC7.

If you have a good ear for music, a own a generator, but no counter, then zero-beat the generator to a piano or other fixed-pitch musical instrument. The frequency of each note is 1.059463 times that of its semitone-lower neighbour. On the International music scale, defined by $A_4=435\text{Hz}$, G_7 is 3100Hz. On the American Standard scale, where $A_4=440\text{Hz}$, G_7 is 3136Hz.

For the second method, lash up an fm generator with an 8038 function generator chip

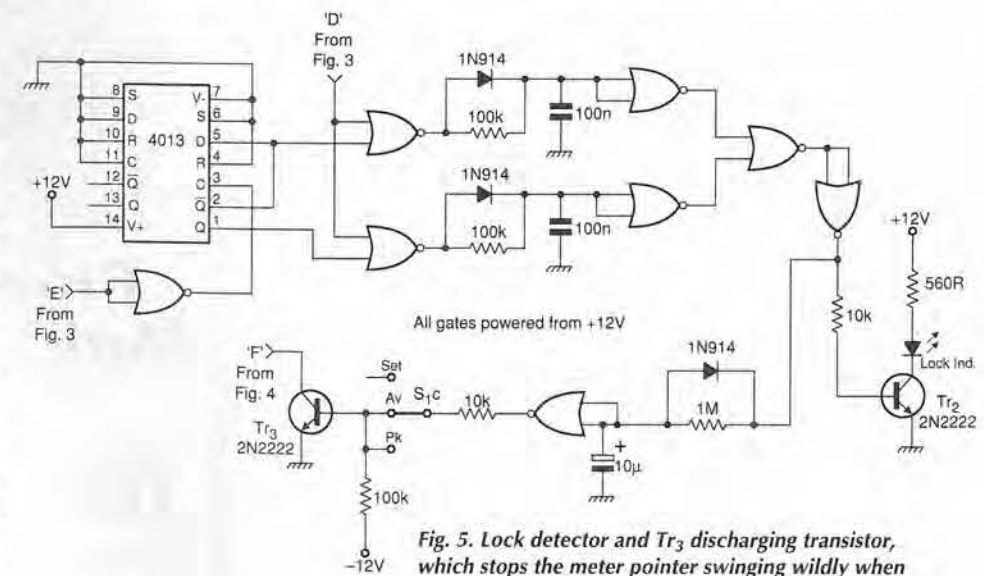


Fig. 5. Lock detector and Tr_3 discharging transistor, which stops the meter pointer swinging wildly when the detector falls out of lock.

oscillating at 3kHz and swept to 1.0% peak deviation by a low-frequency second oscillator. Initially, connect a variable dc voltage source and a dvm to the modulation terminal of the 8038, and monitor the 8038's oscillation frequency with a counter. Once the 8038's modulation sensitivity is known, its modulation terminal can be driven by a sine wave of appropriate amplitude for 1.0% peak deviation. The output of the 8038 is then used as a test signal for the wow and flutter meter.

The third method uses a desk-top computer with a sound card. A few lines of code can generate a sound file that represents a digitised 3kHz carrier modulated at any level and frequency and that can be played through the sound card. Or, to avoid the programming, use one of the several available inexpensive digital audio editors that feature flexible generators.

I have tried all three methods of calibration. The most accurate and flexible is the third, but not everyone has the necessary computer hardware and either the software or the desire

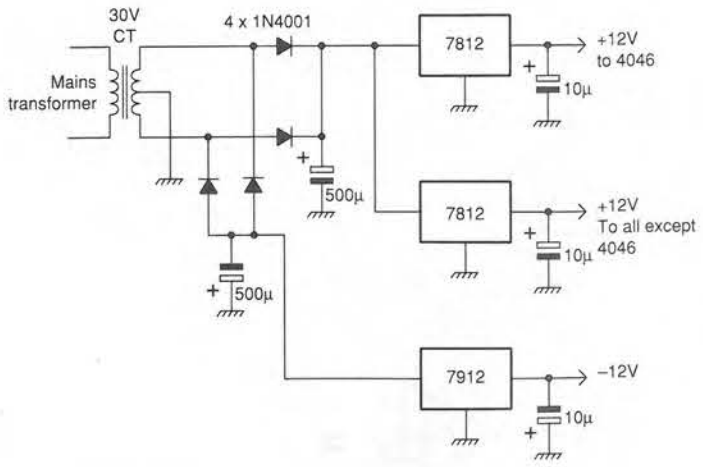


Fig. 5. Power supply. Mains-rated filtering at the transformer primary may be needed to prevent mains borne interference affecting readings.

frequency counter. This feature could be included with the addition of a dc amplifier and a second meter movement having a centre-scale zero. Provision would be required to tune the pll exactly to the carrier frequency.

One last word: if you've never measured wow and flutter, you'll be surprised at the apparent discrepancy between manufacturers' specifications and reality. The peak-reading, non-weighted settings are the most useful for trouble-shooting and for testing design changes, but new equipment is usually specified with a weighted rms-calibrated average reading. The difference between these measurements can be nearly an order of magnitude. On the other hand, a superb turntable may have less speed variation than the residual of even a good test record, especially given the difficulty of centring the record precisely on the platter. ■

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to write software. Methods two and three have the theoretical advantage of allowing for testing or calibration of the whole system at once. The first method is fast, almost as accurate as the third, and can be done in a pinch with only an audio generator and a voltmeter.

Summary

This simple meter has been in use for years with no problems. The effects of flywheel or platter weight and balance, bearing wear, and drive-belt

quality are easy to measure objectively.

Readings are within a few percentage points of those obtained on a calibrated, professional wow and flutter meter when both meters are set up similarly and driven with a sine-modulated test carrier. Congruence is sometimes rather less for non-sinusoidal impulsive functions, and this difference may be due in part to the differences in meter movement ballistics.

Drift measurement was not included in the design, because drift is easily tracked with a

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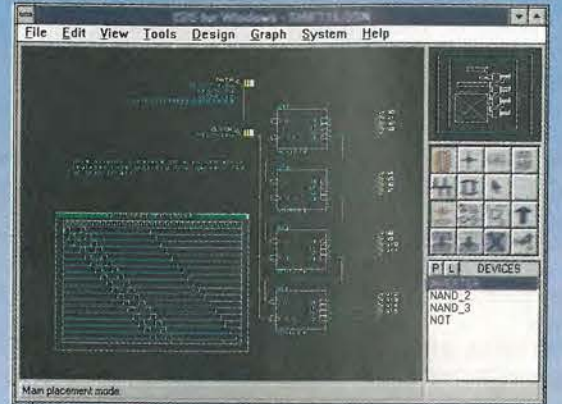
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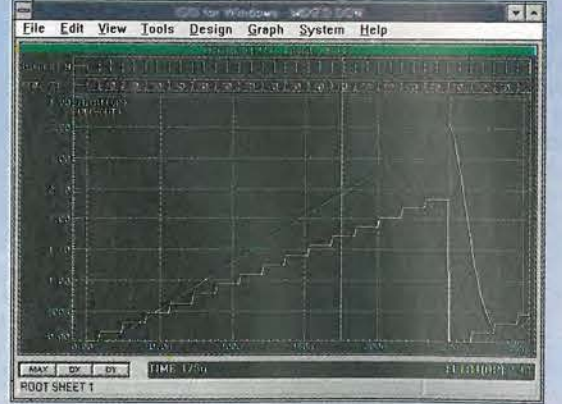
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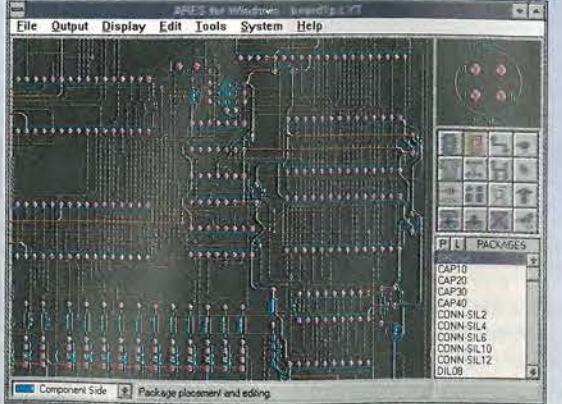
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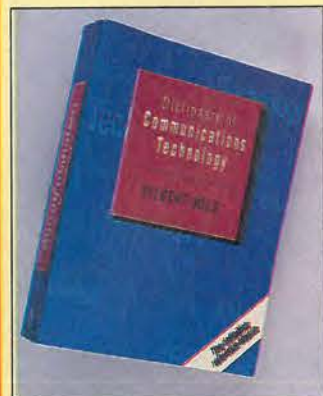
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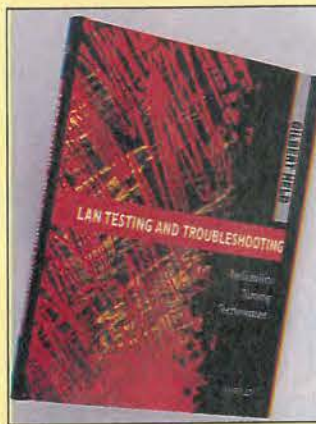
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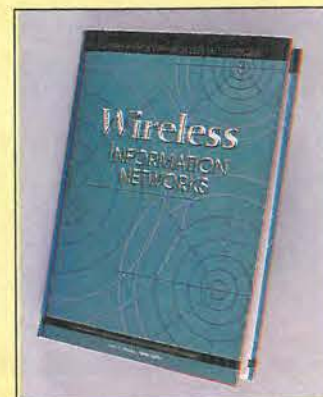
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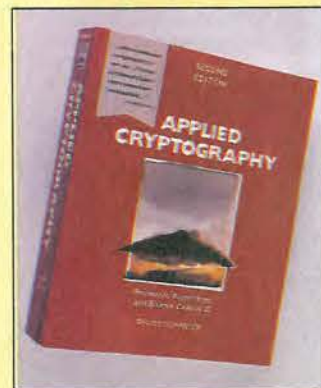
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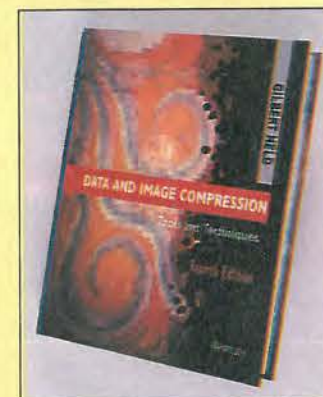
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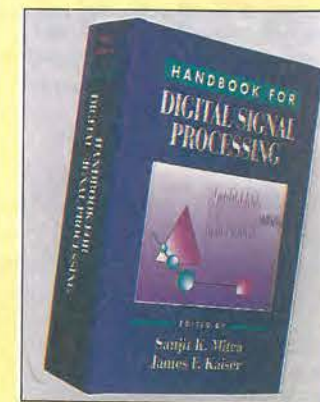


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LETTERS

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

LVD is not new

In his editorial, EW Oct '96, Rod Cooper paints a more gloomy picture than is justified - and one statement is quite wrong. The Low Voltage Directive came into force in 1976 - yes, 20 years ago. What happened on 1 January 1997 was that you had to use CE mark. But you have had to do that, for safety as well, since 1 January 1996, for anything that falls within the EMC Directive.

It is very important for the health and welfare of the electronics industry in the UK not to encourage people to close down projects or even companies through fear of the Directives. More stringent regulations have been in force in Germany and Austria for more than a decade, and their small electronics businesses have learned how to cope: so it is possible.

The chance of any large company trying to use a Directive against a very small competitor is minimal. It is certainly impossible to forecast that there will be no chief executive who suffers a fit of paranoia and goes down that route, but most would see that the game is not worth the candle - particularly if there are actually no grounds for the complaint.

It is true that the engineering profession is disunited, but it is not to be expected that they could unite in opposition to requirements introduced under the EMC Directive. For example, the requirements of EN 61000-3 (all four Sections) are bound to be looked on with favour by the power industry, which would like to preserve their 50Hz sine-waves undistorted. But the regulations are a bane to equipment designers.

The idea of eliminating immunity tests has been raised before, and seems attractive at first sight. But lack of immunity can result in quite dangerous situations, which are almost entirely unpredictable because the manufacturer often has no idea what his equipment may be used for. Of course, these are the events that captured the imaginations of the politicians in the first place, and helped to 'sell' the Directives to them.

Furthermore, complaints of lack of immunity in the field receive virtually no publicity, but the broadcasters, BT and the electricity suppliers receive thousands of complaints every year - usually directed to quite the wrong body. The cost of following these up, and solving them where possible, is quite large.

On the subject of EMC in Australia

(page 724 of the same issue), without the 'benefit' of Brussels, the SMA seems to have produced requirements which are more severe than apply in Europe at present, although they have backed off on some points.

John Woodgate
Rayleigh
Essex

Fooling stereo

In his article 'Music in mind' in the Oct. '96 issue, Ian Hickman wondered why his gyroscopically-controlled headphones refuse to work in stereo.

I wasn't totally convinced by explanations from the experts in the letters pages of subsequent issues, so here's my guess. While processing time or volume differences, the brain needs the signal from both ears to be approximately equal, as in a mono-signal. It is impossible to determine phase difference from two completely different signals.

If you mixed, say, a third of the right-hand channel's signal to the left channel and a third of the left channel's signal to the right, the brain would maybe have enough information to fool it.

Hannu Multanen
Joensuu
Finland

Preamp defence

I would like to comment on Mr Allen Wright's letter (*Electronics World*, Nov. 1996) concerning Douglas Self's pre-amp.

My comments come from two viewpoints - as a collaborator with Douglas in the design of the pcb for the pre-amp, and as a self-employed designer in the professional audio industry.

Over the past 15 years, my work has taken me into a number of UK companies specialising in the design and manufacture of top-of-the-range mixing consoles that end up in radio, television and recording studios throughout the world. All the companies use 5532 and 5534 devices; they also all use large-value electrolytics for signal coupling.

It is a fact that Douglas's preamplifier has been designed with the same criteria used to design professional mixing consoles.

Mr Wright states that budget mixing console manufacturers stopped using the 5532 years ago. I will not contest the budget manufacturers' lack of use of the 5532, as my experience is not at the budget end of the market. But may I be so bold as to suggest that the term

'budget' is significant? The 5532 is expensive. Is it financial constraints rather than sonic performance that preclude its use?

As for the existence of an "unpleasant sonic signature" from the 5532, I am surprised. I would like Mr Wright to tell us more about it.

The ultimate conclusion to his statement of the existence of a "signature" is that virtually all live and recorded sound will be afflicted with the "signature" by virtue of the recording and/or broadcast equipment used to get the sound to the listener.

No matter what devices are used in a preamplifier, if it is well designed, it will faithfully reproduce at its output what is offered to its input. The output will include Mr Wright's alleged, "unpleasant sonic signature", which will have been introduced earlier in the signal chain, and is therefore beyond the control of the pre-amp designer or the listener.

Gareth Connor
GJC Designs
London

CAD inadequacies?

The reviews on pcb cad failed to answer the one question I wanted answered. All the packages seem to work on the same basis: having designed your circuit, you use the schematic drawing package to produce a professional printable schematic and the computer does the rest - i.e. it produces for you a parts list, it captures the circuit for the layout and even simulates the circuit performance.

The problem comes in the phrase 'having designed the circuit'. A word processor package doesn't assume that you had produced hand written draft before you start typing and pcb cad shouldn't assume you have done a paper circuit first. I design the circuit on-screen. I may import bits from earlier circuits. When I draw a bias chain I don't know what the values are going to be, that is left until I'm happy that I have got the configuration right. The design process may include several changes of mind, adding components, taking the out, etc.

Q&A

Shifting phases?

I am looking for a circuit to phase shift by 90° the components of a signal with frequencies in the range 10Hz to about 350Hz. Although simple integration or differentiation can achieve this, they do so at the expense of a frequency dependent change in the signal amplitude which I cannot use.

In *Electronics World* April 1993, Terrance Finegan mentions that such 'a useful analogue function may be realised differentially with all-pass filters', but this hint has proven insufficient. Text books even mentioning all-pass filters seem to be the exception - at my level of mathematical sophistication anyhow.

Are there any readers with a solution to this problem? It would help me, and being an unusual function may inspire other interesting designs.

Alan Scrimgeour
London

(This question is repeated from p 790, Oct '96 issue - ed)

In answer to Alan Scrimgeour's query, it is only possible to produce an approximation to a wide-band 90° phase-shift. The complexity of the solution depends very much on the tolerable errors. To handle 10Hz signals, only active all-pass filters are really practicable. The design procedure is too lengthy to reproduce here but is not difficult, and a good source of the information is the 'Electronic filter design handbook' by A B Williams (McGraw Hill). The ISBN of the first edition is 0 07 070430 9, and this deals with the subject fully, but there is a later edition.

John Woodgate
Rayleigh
Essex

In answer to the question from Alan Scrimgeour: The need to phase shift audio frequencies by 90° is also required in the phasing method of sbsb generation. The article on this subject in *Electronics World* March 1994, pp202-206 covers this subject along with the merits of polyphase networks versus 'all-pass' filters.

John Crabtree
Connecticut
USA

LETTERS

When I am happy with the configuration I renumber the components, take a print, get the calculator out and work out the values and decide on whether a capacitor needs to be a 1% polystyrene, or a ceramic, or whatever...

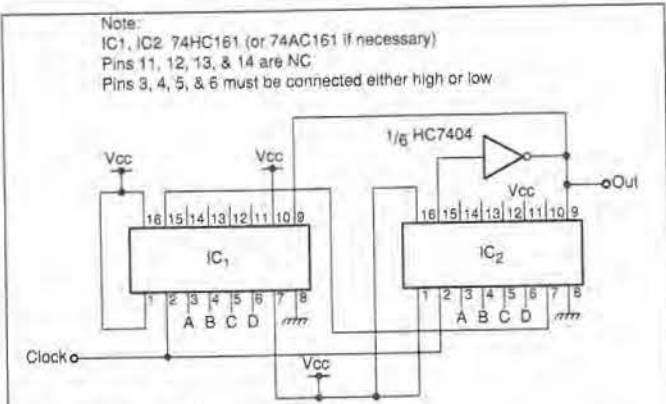
To simply re-number one component in Easy PC Professional XM requires switching from keyboard to mouse to keyboard to mouse to keyboard and finally to mouse.

What I want is a cad package which at the schematic stage does not worry about value or package size. John Kennaugh, Callington, Cornwall.

Rod replies: It think that John Kennaugh has not read the first issue, i.e. Sept. '96, in which I said, "The review is not intended to enable the reader to choose one program to suit his or her particular requirements..."

I don't think John can have read the October issue either, in which I said "I think relative to having the circuit drawn before you on real paper, all methods of representing the circuit on screen are inferior if planning a circuit from scratch rather than just drafting..."

Schematic drawing programs are only there as an entry into those things you just cannot do well with pen and paper - simulating a circuit, routing a pcb - and are regarded as a necessary evil.



Divide by integers to 256 This circuit appeared as part of Nick Wheeler's article on page 893 of the November 1996 issue. Apologies. The original circuit redraw contained an error in the connections around the output.

expensive up-market products, schematic drawing does not get any easier: you just get more features. My review was about producing pcbs, and not about using the monitor screen as an improvised circuit design tool.

Is there a need for a 'scribble-pad' type of schematic drawing program for designing rather than drafting? I'm not sure there is. If there were a need, I think that one of the current generation of software writers would have produced it by now.

Rod Cooper, Sutton Coldfield, West Midlands.

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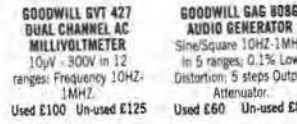


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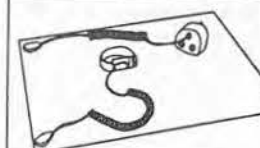


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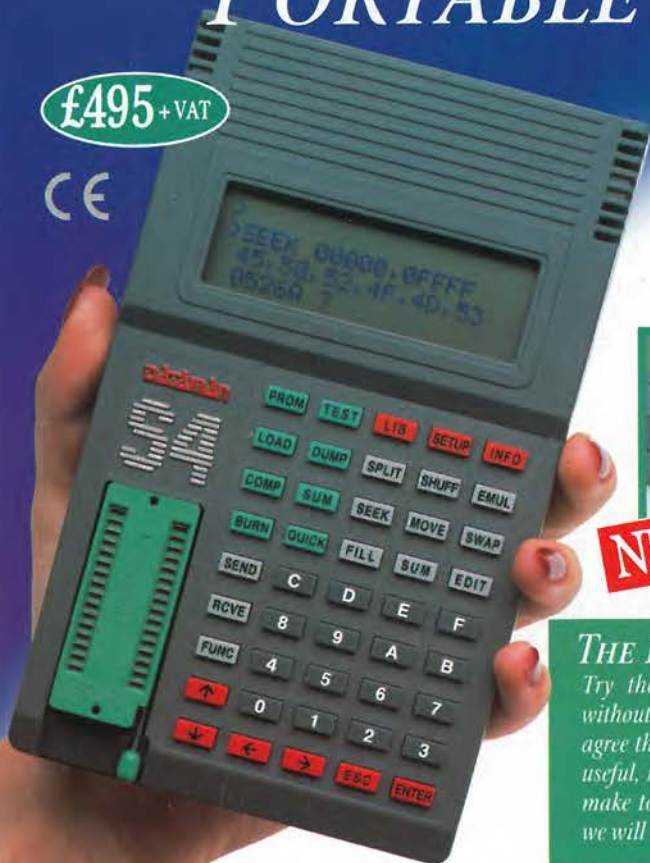
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